A DIGITAL POWER SYSTEM STABILIZER

by

Tom Mayor

A Thesis
presented to the University of Manitoba
in partial fulfillment of the
requirements for the degree of
Master of Science
in
Department of Electrical Engineering

Winnipeg, Manitoba, 1981

A DIGITAL POWER SYSTEM STABILIZER

BY

TOM MAYOR

A thesis submitted to the Faculty of Graduate Studies of the University of Manitoba in partial fulfillment of the requirements of the degree of

MASTER OF SCIENCE

I hereby declare that I am the sole author of this thesis.

I authorize the University of Manitoba to lend this thesis to other institutions or individuals for the purpose of scholarly research.

Tom Mayor

I further authorize the University of Manitoba to reproduce this thesis by photocopying or by other means, in whole or in part, at the request of other institutions or individuals for the purpose of scholarly research.

Tom Mayor

The University of Manitoba requires the signatures of all persons using or photocopying this thesis. Please sign below, and give address and date.

ABSTRACT

The use of microprocessors in implementing digital filters for power system stability control loops was investigated. A real system model which exhibits two natural frequencies of oscillation, one associated with dynamic stability, the other with transient stability, was developed and a digital filtering system, acting as a dynamic stability compensator, was designed and tested. The system can serve as a basis for continued work. Results indicate that digital filters are well suited for acting as controllers for power systems. As a single compensator cannot provide optimal stabilization for all operating conditions, a form of adaptive filtering, easily performed with a microprocessor based system, should be implemented.

ACKNOWLEDGEMENT

I would like to express my gratitude to Dr. G.W. Swift for suggesting this topic and for his encouragement and guidance throughout its development. The financial assistance provided by Manitoba Hydro is acknowledged with thanks. I would also like to give a special thank you to my wife, Christine, for her moral support throughout my education.

TABLE OF CONTENTS

LIST OF FIGURES

Chapter I

INTRODUCTION

Conflicting requirements often limit the effectiveness of a system design. Transient and dynamic stability criteria of synchronous machines in a power system often pose such conflicting requirements. Modern machines employing high speed static exciters have an inherent reduction in their dynamic stability margin requiring supplementary stabilizing signals. These signals generally provide an optimized damping of power system transients to maintain dynamic stability. If additional signals are introduced, optimized for a transient stability condition, they may produce unacceptable modes of operation to the system.

To solve these conflicting requirements,Bayne,Kundar,and Watson (3) inserted logic into their stabilizing scheme. A continuously acting stabilizer was used to enhance the dynamic stability of the system, while an additional stabilizing signal was added only when transient stability was threatened. The above mentioned paper served as a stimulus for this thesis.

This investigation concerns a system which exhibits two natural frequencies of oscillation, one associated with dynamic stability, the other with transient stability. As done by Bayne et al., logic could be added to introduce the required stabilizing signals. However, a power system is a dynamic network so any particular solution to a stability problem will probably have only a limited range of application unless some means of updating the solution to changing operating variables is found. As additional machines are added to the system and interconnections between other systems increase in number, the transfer functions of the required stabilizers change. In a hard - wired system, this would require the replacement of part if not all of the stabilizing system and also require the tuning of the new system. Here, we implement all stabilizers and logic functions with a microprocessor system. As all stabilizer functions and the logic to choose them are performed in software, changes are relatively simple.

To give a better understanding of the stability problems involved, this chapter gives a brief overview of transient and dynamic stability. The Heffron - Phillips machine model, a linearized small perturbation model of a single synchronous machine connected to an infinite bus through an external impedance, is introduced. This model assists in showing how excitation control can affect synchronous machine stability. Due to many excellent publications on the subject, (6 ,10), only the most important aspects are covered.

Chapter II describes the conditions for the occurrence of two natural frequencies of oscillation. Ontario Hydro's solution to the problem is explained as is our proposed solution. The two-frequency power system model that was developed is described. Chapter III gives an introduction to digital filtering and discusses the advantages, disadvantages, and problems associated with digital filters. Chapter IV discusses the realization of the continuous-acting digital filter and the discontinuous system as required by the two frequency power system model. The test results are shown in Chapter IV.

1.1 SYNCHRONOUS STABILITY

Power system generators connected through a transmission network must run in synchronism. The ability of a system to return to normal or synchronous operation after having been subjected to some form of disturbance is called stability. Synchronous stability can be divided into two areas, dynamic stability and transient stability, dynamic stability being concerned with small disturbances to the system such as continual load changes and transient stability being concerned with major disturbances such as caused by faults.

The understanding of synchronous stability is simplified by considering a single synchronous machine connected to an infinite bus through an external reactance as shown in Figure 1.1 .

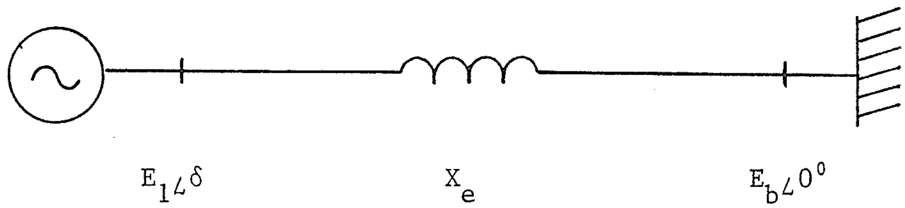


Figure 1.1: Synchronous Machine - Infinite Bus One Line Model

This circuit provides insight when studying the stability performance of one machine in a system. The external reactance and infinite bus represent the system as seen from the terminals of the machine in question. The stability model of this generator is governed by three equations which are presented here without development as this may be found in most texts on power systems.

$$\frac{d\bar{\omega}}{dt} = \frac{P_m - P_e}{2H} \qquad \ldots (1.1)$$

$$P_e = \frac{E_1 \, E_b \sin \delta}{X_e} \qquad \ldots (1.2)$$

$$\frac{d\delta}{dt} = \omega_o(\bar{\omega} - 1) \qquad \ldots (1.3)$$

where H = inertia constant

$\bar{\omega}$ = per unit speed = speed / rated speed

$\omega_0 = N\,\omega_r$

N = number of pole pairs

ω_r = rated speed

P_m = mechanical power

P_e = electrical power

E_1 = machine terminal voltage

E_b = infinite bus voltage

δ = angle between the quadrature axis and the infinite bus

X_e = equivalent system reactance

Equation (1.2) is plotted in Figure 1.2 . The system is in equilibrium when $P_e = P_m$, thus a given P_m will result in two equilibrium points δ_1 and δ_2 . However, for $P_m > P_e$, the system will respond by accelerating thus increasing δ. For $P_e > P_m$, the system will decelerate thus decreasing δ. From these characteristics it is obvious that for small disturbances only δ_1 is a stable equilibrium point. The power level P_{max} , called the steady state stability limit, represents the stable limit of power transfer between the generator and infinite bus.

Transient stability analysis is mainly concerned with the effects of transmission line faults on generator synchronism. When a fault occurs the power output of the synchronous generator is greatly reduced. However, the

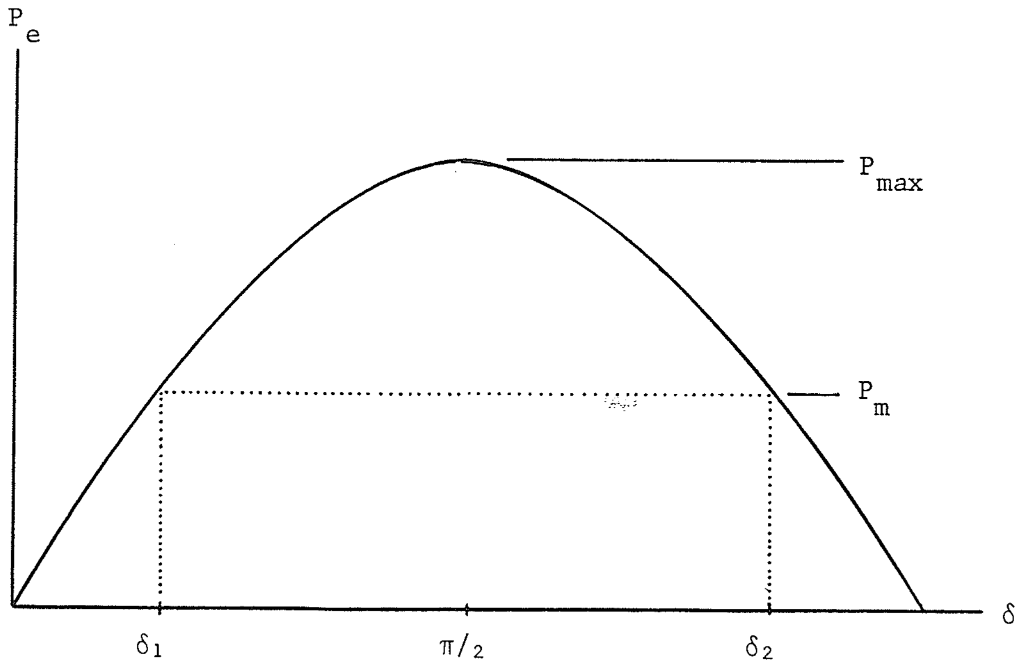


Figure 1.2: Power - Angle Diagram

input power P_m is relatively constant. The rotor gains speed in order to store the excess energy and hence the rotor angle δ increases. If the rotor is unable to return to the system the energy gained during this acceleration period, the generator will lose synchronism.

Mathematically,this can be looked at through (1.1)-(1.3). By differentiating (1.3) and using the result in (1.1) and (1.2) we get the system swing equation:

$$\frac{d^2\delta}{dt^2} = \frac{\omega_o}{2H}(P_m - P_e) \qquad \ldots \quad (1.4)$$

Multiplying both sides by $2(d\delta / dt)$ gives

$$2 \frac{d\delta}{dt} \left(\frac{d^2\delta}{dt^2} \right) = 2 \frac{d\delta}{dt} \frac{\omega_o}{2H} (P_m - P_e) \qquad \ldots (1.5)$$

or

$$\frac{d}{dt} \left(\frac{d\delta}{dt} \right)^2 = \frac{\omega_o}{H} (P_m - P_e) \frac{d\delta}{dt} \qquad \ldots (1.6)$$

Integrating both sides gives

$$\left(\frac{d\delta}{dt} \right)^2 = \frac{\omega_o}{H} \int_{\delta_o}^{\delta_p} (P_m - P_e) d\delta \qquad \ldots (1.7)$$

If a generator is to remain synchronized, the rotor will swing until its angular acceleration reaches zero.

$$\therefore \int_{\delta_o}^{\delta_p} (P_m - P_e)\, d\delta = 0 \qquad \ldots (1.8)$$

Thus the criterion for the machine to remain stable is that the area between the P-δ curve and the line representing the input power must equal zero. This is called the equal area criterion. The power angle diagram for the system shown in Figure 1.3 clarifies these equations. The power angle curves are shown in Figure 1.4. The system is a generator connected to an infinite bus through two transmission lines in parallel. A fault occurs on one line and is cleared by opening breakers at both ends of the line. The integral in (1.8) may be separated into two parts.

$$\int_{\delta_o}^{\delta_p} (P_m - P_e)d\delta = \int_{\delta_o}^{\delta_c} (P_m - P_{1_{max}} \sin \delta)d\delta + \int_{\delta_c}^{\delta_p} (P_m - P_{2_{max}} \sin \delta)d\delta = 0 \quad \ldots (1.9)$$

$$\therefore \int_{\delta_o}^{\delta_c} (P_m - P_{1_{max}} \sin \delta)d\delta = \int_{\delta_c}^{\delta_p} (P_{2_{max}} \sin \delta - P_m)d\delta \quad \ldots (1.10)$$

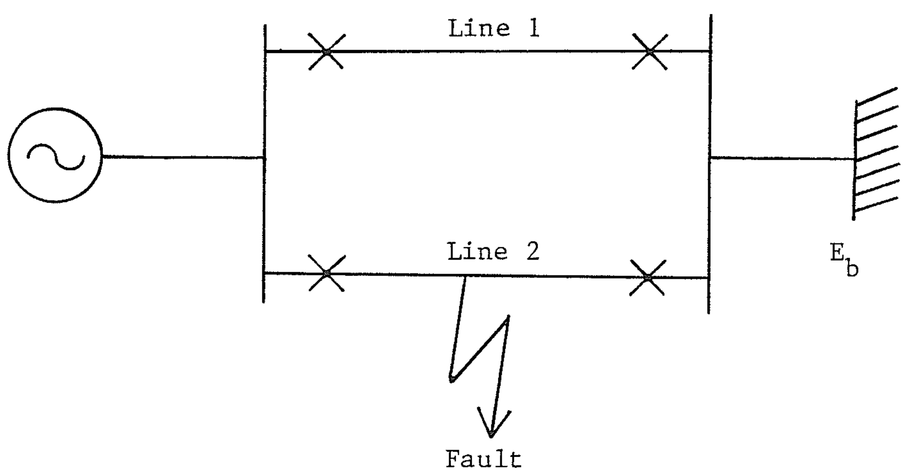


Figure 1.3: Fault on One Line of Two Lines in Parallel

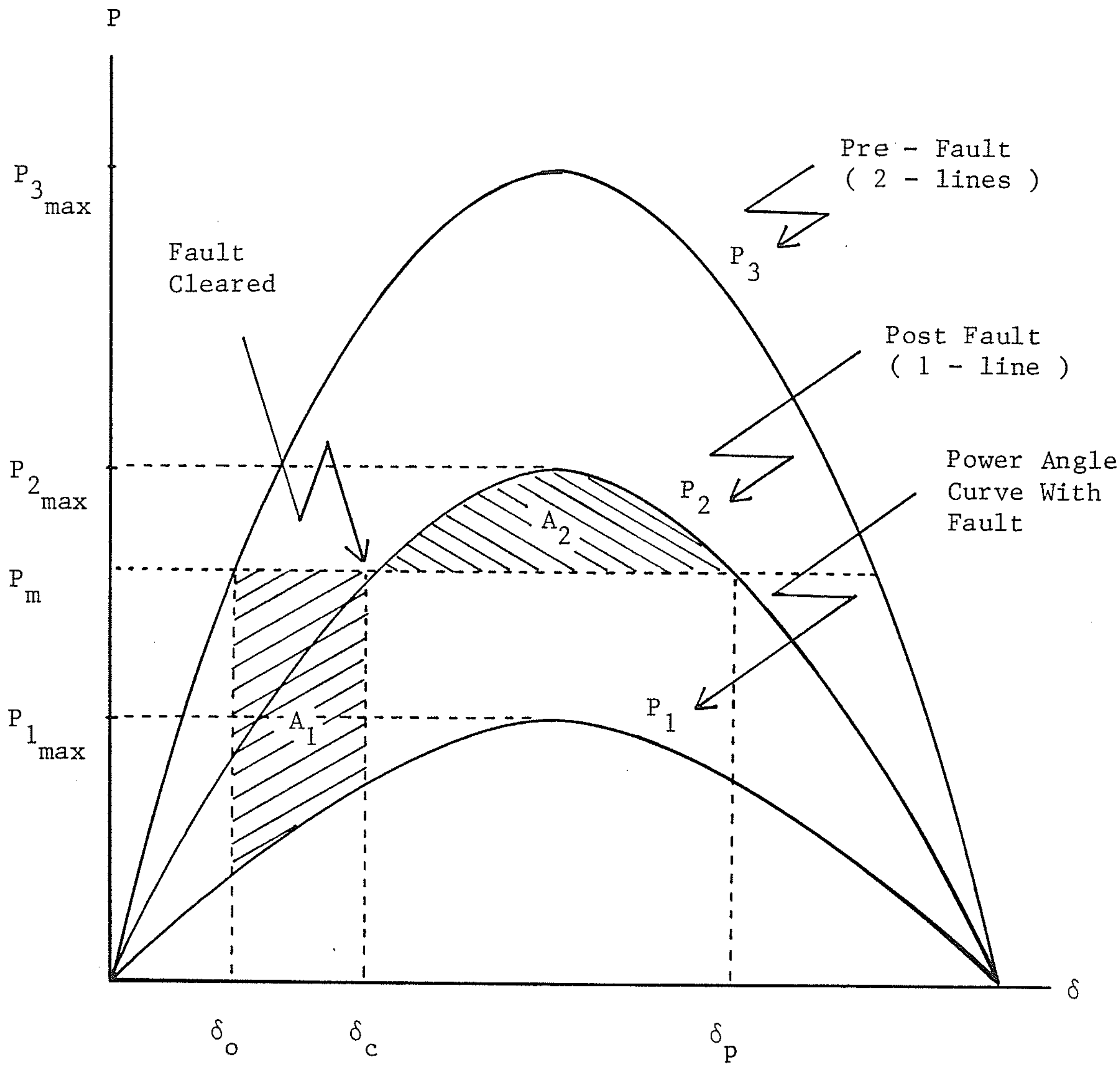


Figure 1.4: Equal Area Criterion

The left hand side of (1.10) is area A_1 in Figure 1.4 and the right hand side is area A_2. If area A_1 is larger than area A_2 then the rotor is unable to return its excess energy to the system and hence the machine goes unstable.

High initial response excitation with high ceiling voltage as provided by static exciters is an effective method of improving transient stability. However, the use of high gain static exciters results in decreased system damping and may even cause small signal dynamic instability. This is shown in the next section through the analysis of the Heffron - Phillips machine model.

1.2 HEFFRON - PHILLIPS MODEL

The linearized small perturbation model of a single machine connected to an infinite bus through an external reactance is shown in Figure 1.5 . This model applies to a 2-axis machine representation with the field circuit in the direct axis but without armortisseur effects. As this model is well known, it will not be developed here. The interested reader may refer to either (6) or (10). The parameters K_1 to K_6, called the Heffron - Phillips constants, are functions of machine and system impedances and operating points. The equations for these parameters are shown in Appendix A as given in (6). Due to their dependancy on the operating point, K_1 - K_6 can vary considerably. Thus the dynamic behavior of a machine can vary greatly over its range of operating conditions. It should also be noted that as this model is based upon small perturbations around a set operating point, sections of the model can be isolated by assuming a variable to be constant.

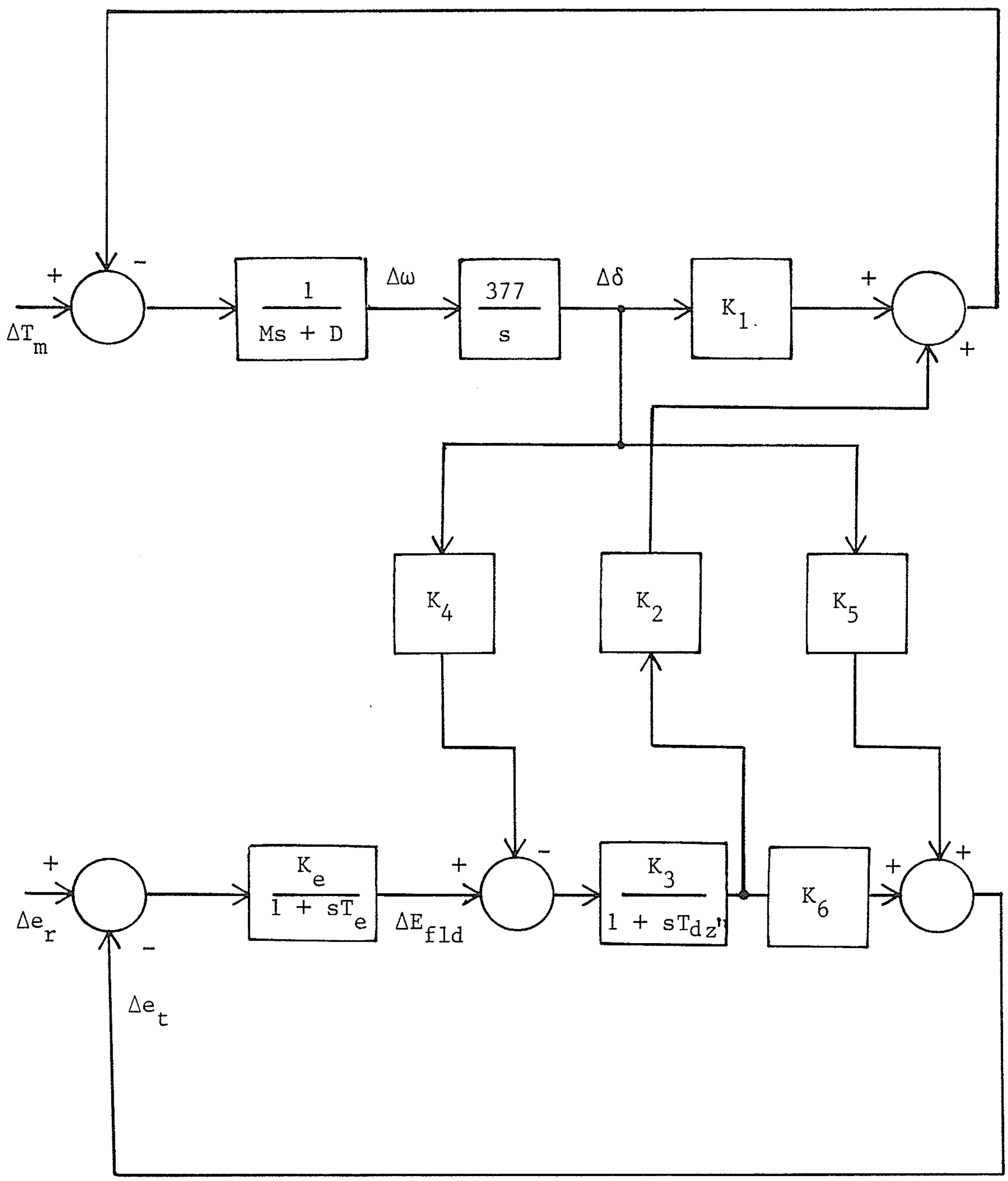


Figure 1.5: Heffron - Phillips Model

Our primary interest when studying the stability of a system is in the torque - angle relationship. If the forces acting on a system restore the rotor angle of a machine following a small displacement of this angle, the system is considered dynamically stable. Figure 1.6 shows the torque - angle relationship for the condition of constant flux linkages in the d-axis. The characteristic equation of this system is

$$S^2 + \frac{D}{M} S + \frac{377K_1}{M} = 0 \qquad \ldots (1.11)$$

The response of the system to a change in mechanical torque would be a damped oscillation with a frequency of $\omega_n \sqrt{1 - \zeta^2}$ where ω_n is the natural frequency of oscillation of the system and ζ is the damping ratio.

$$\omega_n = \sqrt{\frac{377K_1}{M}} \qquad \ldots (1.12)$$

$$\zeta = \frac{D}{2\sqrt{377K_1 M}} \qquad \ldots (1,13)$$

Damping in large systems is often almost negligible and so the frequency of oscillation approaches ω_n . Note from

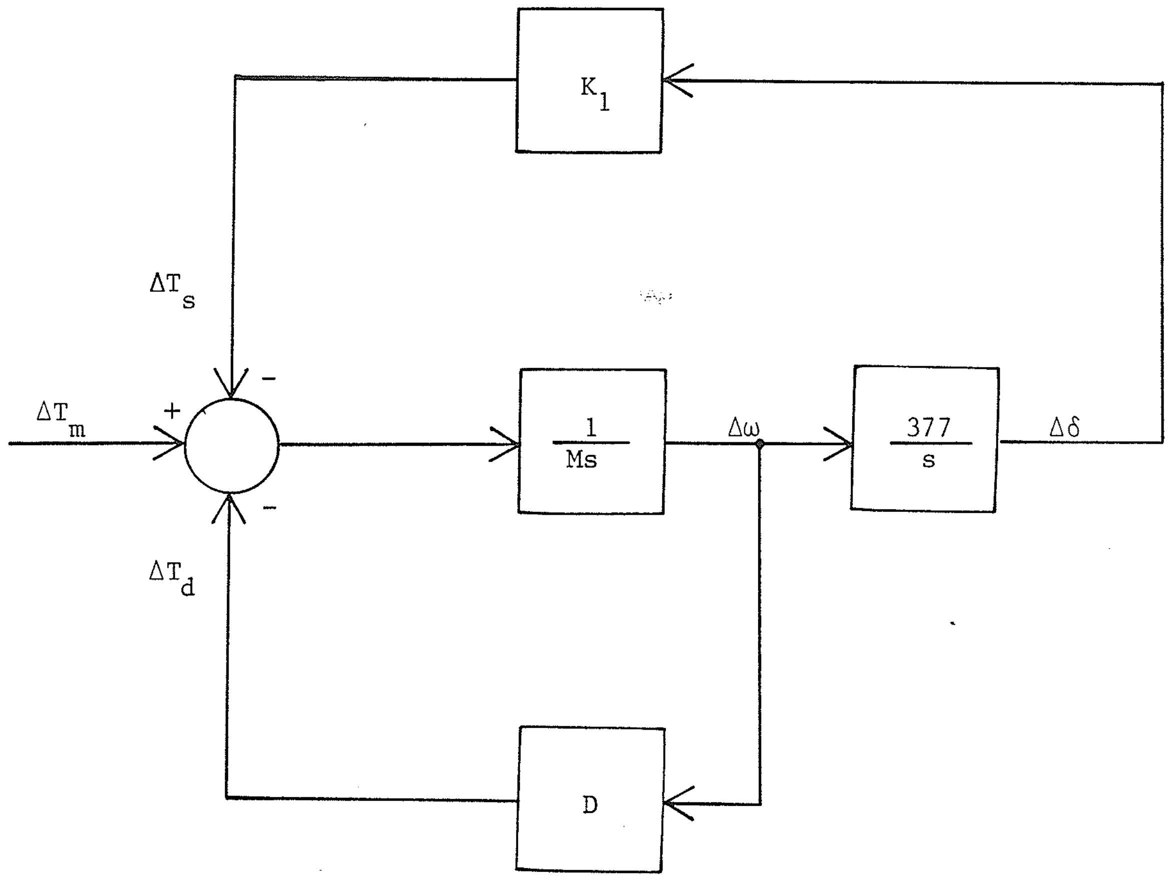


Figure 1.6: Linearized Torque - Angle Relationship

(1.12) that ω_n^2 is proportional to K_1 and inversely proportional to M. As a result of this, small,light inertia machines will have a higher frequency of oscillation than larger machines. For normal ranges of inertia, impedances, and loading values the oscillation frequency can vary from 0.1 Hz. to 4 Hz.

From Figure 1.6 it is seen that braking torques in phase with both machine rotor speed and machine rotor angle are developed. These torques are called damping torques ($\Delta T_d = D\Delta\omega$) and synchronizing torques ($\Delta T_s = K_1\Delta\delta$) respectively. Lack of synchronizing or damping torque or the condition of a negative damping torque can cause a machine to go unstable.

The field circuit of a generator contributes to the braking torques. Figure 1.7 shows the torque - speed - angle loop including the effect of field losses for the case of constant field voltage. The effect on the braking torque is described by

$$\frac{\Delta T}{\Delta\delta} = - \frac{K_2\,K_3\,K_4}{1 + sT_{dz'}} \qquad \ldots(1,14)$$

In the steady state, $\Delta T = -K_2K_3K_4\Delta\delta$ and is a pure synchronizing torque, opposed in sign to the synchronizing torque $\Delta T_s = K_1\Delta\delta$ that was previously discussed. Stability can only be maintained for $K_1 - K_2K_3K_4 > 0$. For high oscillation frequencies ($\omega >> 1 / T_{dz'}$) the torque component is phase shifted by +90 degrees and is almost totally damping torque. At these frequencies, though, the magnitude of the torque is small and the effect on damping is negligible.

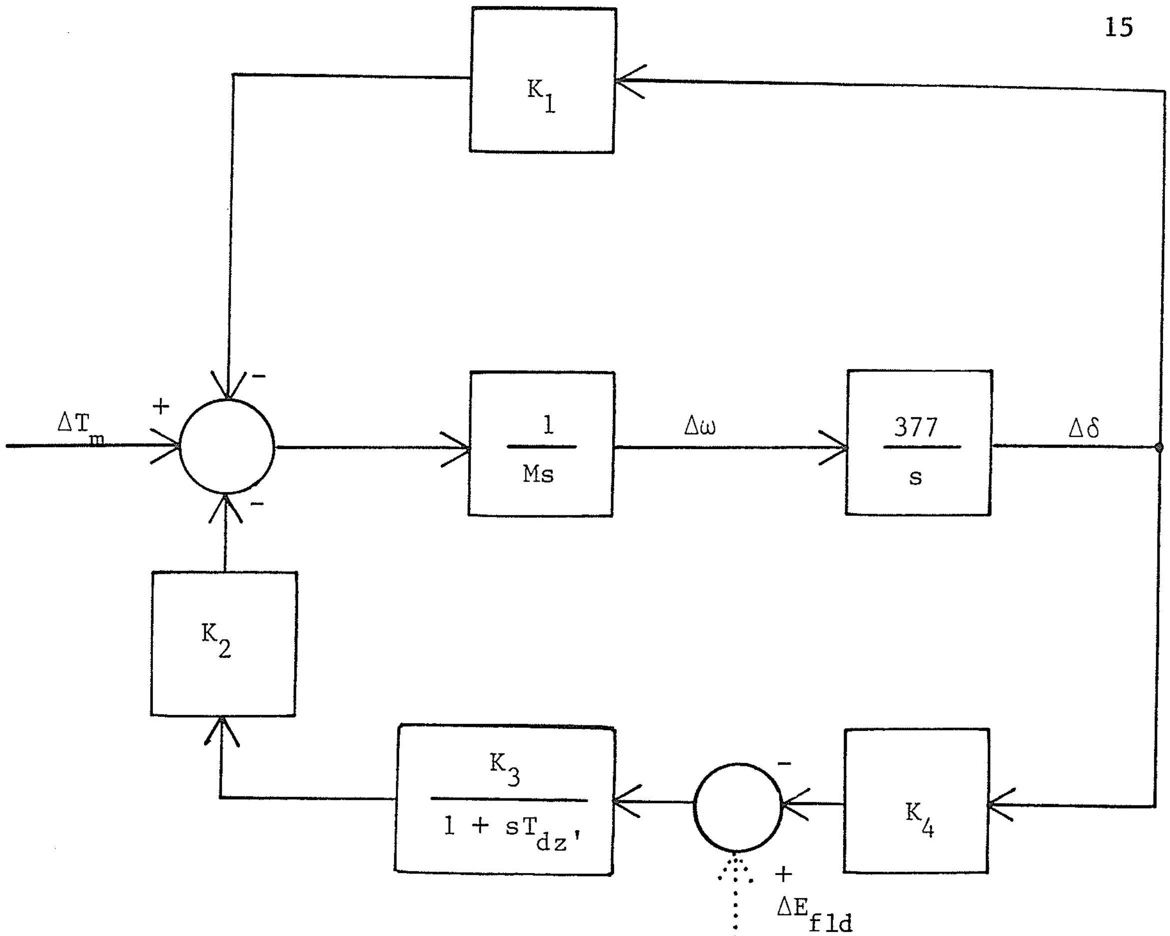


Figure 1.7: Torque - Angle Loop with Direct Axis Field Effects

Finally, the effect of the voltage regulator on the damping and synchronizing torques must be examined. Only high speed static type exciters are considered as these tend to improve transient stability as previously mentioned. An exciter of this type has a typical transfer function of

$$G_e(s) = \frac{K_e}{1 + sT_e} \qquad \ldots (1.15)$$

where K_e is the exciter gain and T_e the time constant (typically 0.03 to 0.05 sec.). Two torque - angle paths are apparent. These are shown in Figure 1.8 .

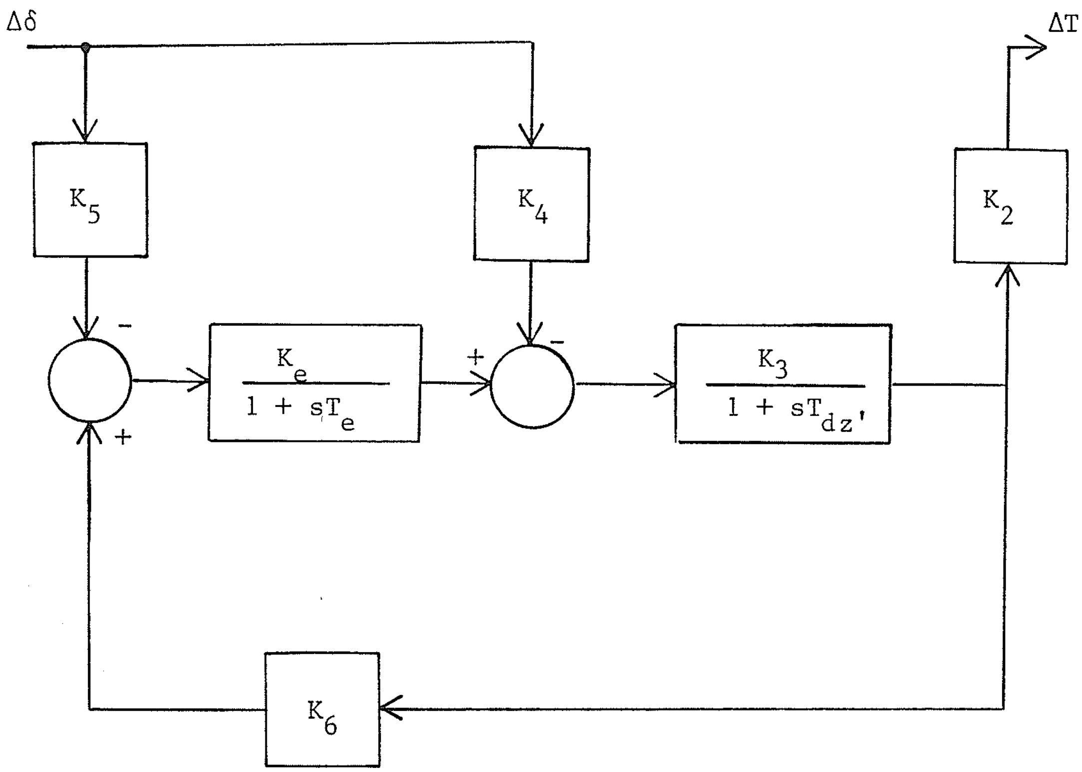


Figure 1.8: Torque-Angle Effects as a Result of the Excitation System

The torque - angle relationship through the K_4 branch is

$$\frac{\Delta T}{\Delta \delta} = - \frac{K_2 K_3 K_4 (1 + sT_e)}{(1 + sT_{dz'})(1 + sT_e) + K_6 K_3 K_e} \qquad \ldots (1.16)$$

Since $T_e << 1$, this can be shown as

$$\frac{\Delta T}{\Delta \delta} \approx - \frac{K_2 K_3 K_4}{(1 + s\, T_{dz'}) + K_6 K_3 K_e} = \frac{K_2 K_4}{K_6 K_e \left(1 + \frac{s T_{do'}}{K_6 K_e}\right)} \qquad \ldots (1.17)$$

(recalling that $T_{dz'} = K_3 T_{do'}$). With the exciter loop included, the net steady state synchronizing torque due to the K_4 branch is now $K_1 - K_2 K_4 / K_6 K_e$ as opposed to $K_1 - K_2 K_3 K_4$ for the case of no voltage regulator. This improvement in stability is due to K_e being high. The effective field time constant is reduced to $T_{do'} / K_e K_6$. The oscillation frequency must be very high to produce a phase lag of 90°. Therefore, the damping torque through the K_4 branch due to the exciter is negligible.

Referring to Figure 1.8, the torque-angle relationship through the K_5 branch gives the following expression:

$$\frac{\Delta T}{\Delta \delta} = - \frac{K_2\, K_5\, K_e}{\frac{1}{K_3} - K_6 K_e + s\left(\frac{T_e}{K_3} + T_{do'}\right) + s^2 (T_e T_{do'})} \qquad \ldots (1.18)$$

Substituting $j\omega$ for s in (1.18) and applying approximations for the usual range of constants gives expressions for the synchronizing and damping torque due to the K_5 branch.

$$\frac{\Delta T_s}{\Delta \delta} \approx - \frac{K_2 \ K_5/K_6}{1 + \omega^2 (\frac{T_{do'}}{K_6 K_e})^2} \qquad \cdots \ (1.19)$$

$$\frac{\Delta T_d}{\Delta \delta} \approx \frac{(\frac{T_{do'}}{K_6 K_e})(\frac{K_2 K_5}{K_6})}{1 + \omega^2 \ (\frac{T_{do'}}{K_6 \ K_e})^2} \qquad \cdots \ (1.20)$$

The K_5 parameter is important to these two equations. If K_5 is positive, which is the case for low to medium loading and external impedance, the synchronizing torque is decreased but the damping torque is increased. For negative K_5, which is the case for moderate and high loading and external impedance, instability can occur due to the negative damping destroying the small natural damping of the system.

Thus a trade-off exists. High speed static exciters are effective in increasing transient stability. However, in doing so, they cause decreased damping of the generators,possibly causing dynamic instability . An effective method of counteracting this effect is to introduce a stabilizing signal, optimized to give maximum system damping. This stabilizing signal is generally derived from rotor speed, power, or terminal frequency.

Chapter II

POWER SYSTEM OSCILLATIONS

2.1 CONDITIONS FOR TWO NATURAL FREQUENCIES OF OSCILLATION

As discussed previously, the natural frequency of a synchronous machine connected to an infinite bus is

$$f_n = \frac{1}{2\pi}\sqrt{\frac{377K_1}{M}} \qquad \ldots(2.1)$$

Typical values for K_1 of 0.6 p.u. and for M of 5 - 10 p.u. give an oscillation frequency $f_n \simeq 1$ Hz. This oscillation can be considered a product of local generation against the rest of the system.

For two areas of generation connected as shown in Figure 2.1 , tie - line oscillations may occur. Here, the base power, the tie line power,equals 1 p.u. Area A could have power generation of 20 p.u. K_1 remains typically about 0.6 p.u., but since inertia is directly proportional to power, M in (2.1) is now typically 100 - 200 p.u. giving an oscillation frequency $f \simeq 1/5$ Hz. This low frequency oscillation is that of Area A against its neighbouring area through the tie-line. An impact to the system could cause substantial oscillation on generators close to the tie - line. A typical example is shown in Figure 2.2.

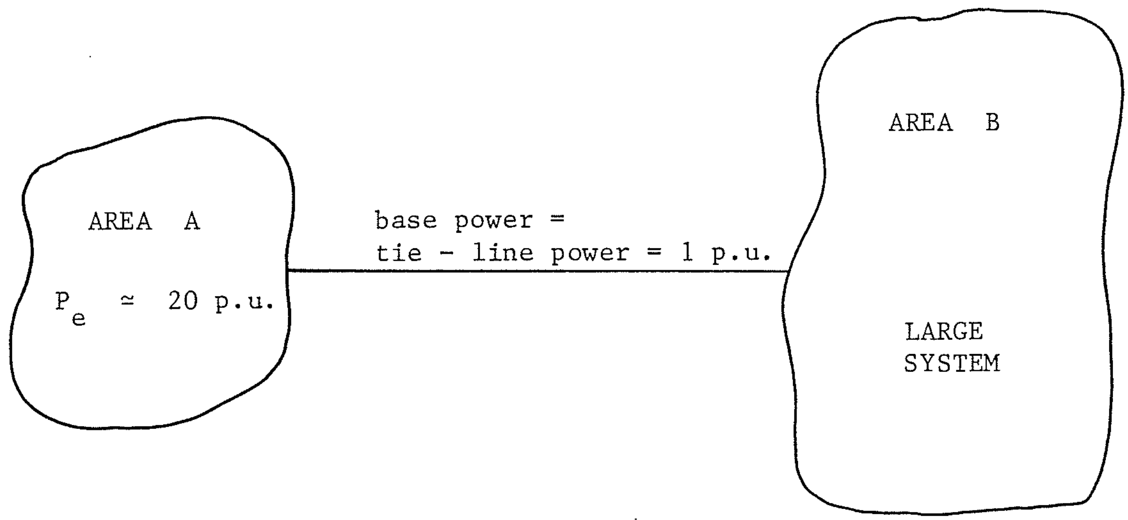


Figure 2.1: Two areas of generation connected by a tie - line

2.2 ONTARIO HYDRO'S STABILIZING SCHEME

Synchronizing torque is needed to counteract the low frequency oscillation shown in Figure 2.2. A stabilizer using a signal in phase with δ, easily obtained by integrating speed or frequency, could be added to the excitation system to provide the synchronizing torque. Due to the low frequency, the reset time of the stabilizer would have to be long for the signal to be effective. This contradicts the requirement on excitation systems that sustained frequency changes should not alter the terminal voltage.

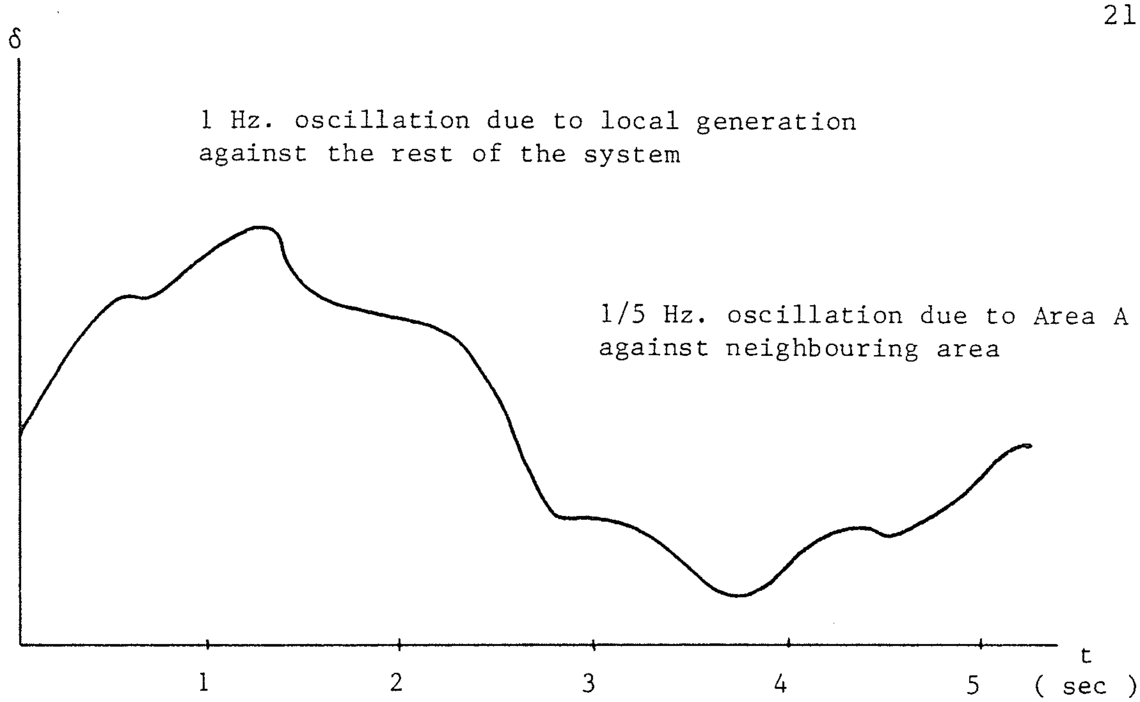


Figure 2.2: Typical Rotor Angle - Time plot following impact to system

Ontario Hydro's solution to this problem was to use a discrete stabilizer that adds its signal to the excitation system only when transient stability is threatened. The stabilizer for the dynamic stability control remains in continuous operation. By removing the need to compromise between the conflicting demands for transient and dynamic stability , each stabilizer can be optimized separately.

The logic to connect the discontinuous stabilizer to the system is controlled by three factors.

1. Terminal voltage drops by a specified amount.

2. The exciter is in positive saturation .

3. Generator speed increases by a specified amount.

The reasoning behind these criteria is as follows. A fault causes a drop in the terminal voltage , the magnitude of which indicates the severity of the fault and hence whether a potentially dangerous condition to transient stability exists. The exciter goes into saturation and the generator speeds up as a consequence of the electrical load decreasing. By making speed change above a preset level one of the criteria , the stabilizer will not be connected for a temporary voltage drop such as occurs when a large load is suddenly applied. Once the stabilizer has been connected , it is removed by either the exciter coming out of saturation or the generator speed dropping below the preset level. A rise in the terminal voltage will not cause the stabilizer to become disconnected. This is because the stabilizer acts to keep the field voltage and hence the terminal voltage high , minimizing the effect of the fault. When the logic acts to disconnect the stabilizer from the exciter the stabilizer output continues in an exponential decay preventing sudden jolts to the system. The stabilizing system is shown in Figure 2.3.

Ontario Hydro's modified stabilizers have proven to be effective in improving transient stablity in the type of systems where a low frequency inter - area oscillation is dominant over the machine versus system local oscillation.

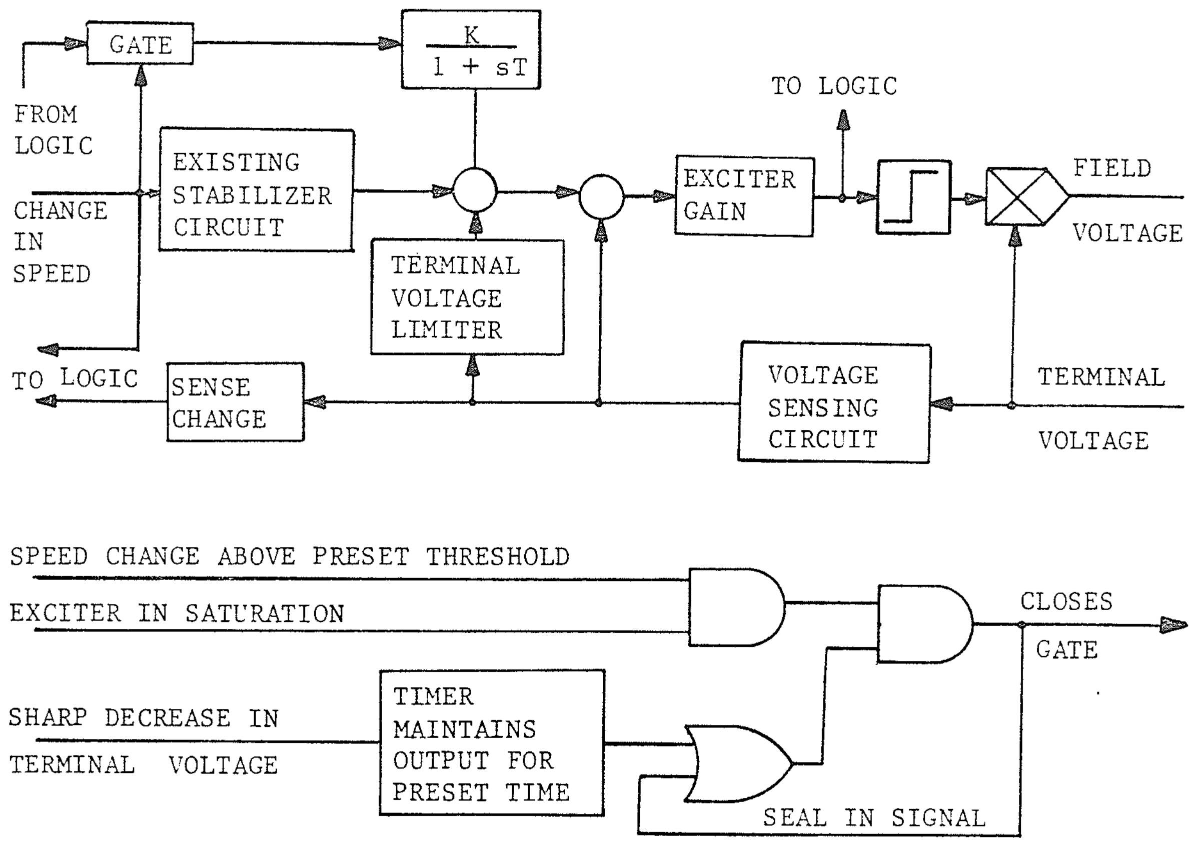


Figure 2.3: Ontario Hydro's Stabilizing System

In one study they were able to increase the critical clearing time from 3.75 to 7.05 cycles. For a fault cleared after 3.6 cycles, the peak rotor angle was decreased from 115 ° to 61 °.

2.3 MICROPROCESSOR BASED STABILIZER

This investigation was intended to improve on Ontario Hydro's solution by implementing both the continuously acting stabilizer needed for dynamic stability and the discrete stabilizer needed for transient stability using a microprocessor. Any changes in stabilizer functions, necessitated by a system change, could be achieved through a software change rather than a hardware change. The project was divided into four major sections, the first three which were successfully achieved , the last still requiring extensive work. The divisions were:

1. Implementation of a system with two natural frequencies of oscillation .
2. Implementation of an interface between the system and a microprocessor.
3. Implementation of a digital filter for the case of one natural frequency of oscillation.
4. Implementation of digital filters for the case of two natural frequencies of oscillation.

2.4 SYSTEM EXHIBITING TWO NATURAL FREQUENCIES OF OSCILLATION

A real system model was needed to investigate the use of microprocessors as stabilizers. The scheme chosen is shown in Figure 2.4. Machine 1 is BICEPS (Basic Instrumented Controllable Electric Power System) and is the machine that is to be stabilized. BICEPS is a 120 watt base power system which exhibits characteristics similar to those of a real power system , including a natural frequency of oscillation of ≃ 1Hz. Full documentation on BICEPS is available in

(17). Machine 2 , representing the large system shown in Figure 2.1, is a large laboratory synchronous machine driven by a D.C. motor.

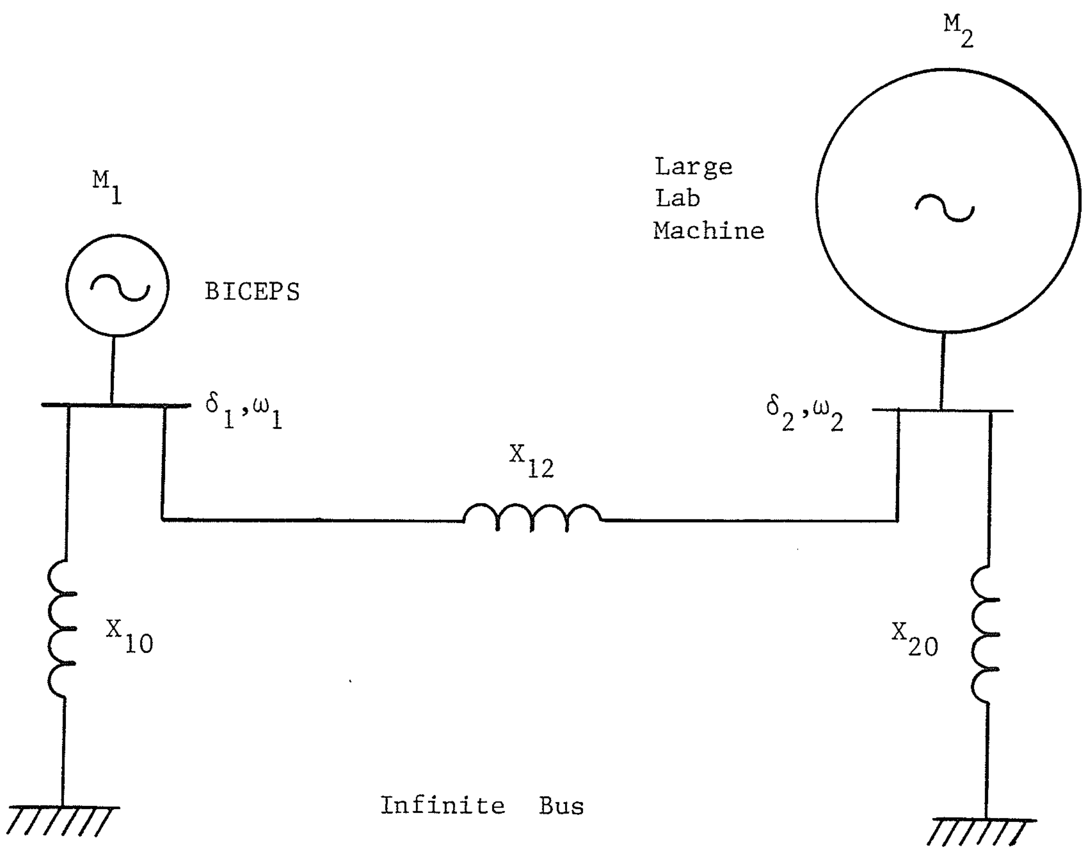


Figure 2.4: Basic System exhibiting two natural frequencies

For a realistic simulation , Machine 2 had to be made to oscillate at ≃ 1/5 Hz. Attempts to achieve this by adjusting its operating point (hence changing K_1) and by adding large flywheels to increase the machine inertia proved unsuccessful. The solution came through the control of the D.C. motor driving the synchronous machine. The motor is operated as a shunt wound machine. A 1/5 Hz decaying exponential signal is fed into the series field causing the machine set to oscillate at the required frequency. The 1/5 Hz signal is initiated simultaneously with a change in the transmission line length between BICEPS and the infinite bus causing both the 1Hz and 1/5 Hz. oscillations to occur. The circuit is shown in Figure 2.5.

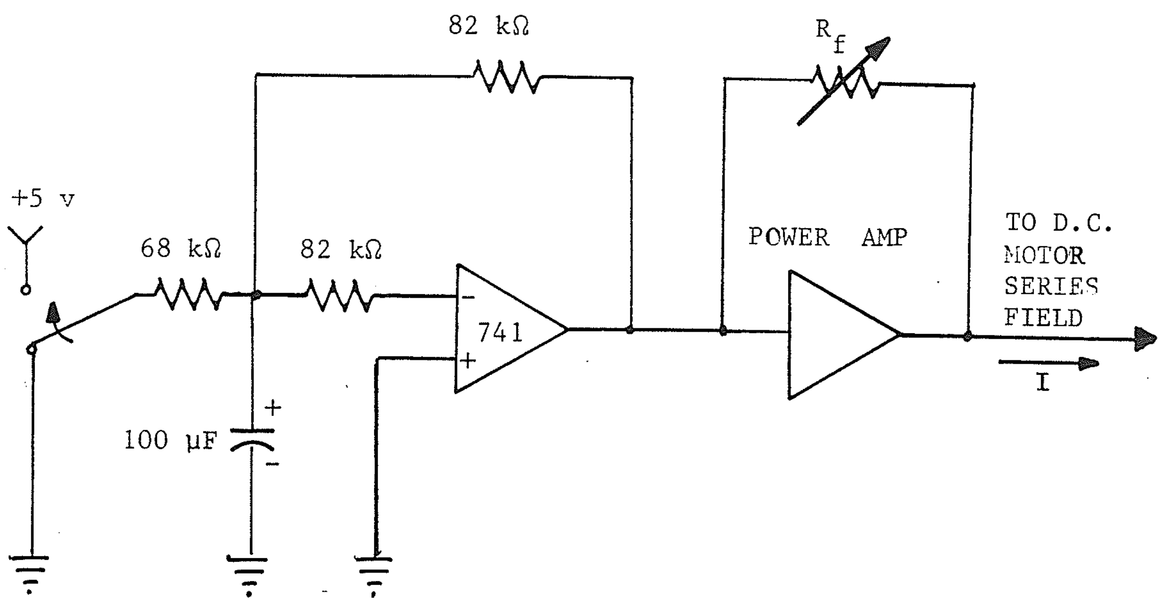


Figure 2.5: 1/5 Hz. oscillation circuit

2.5 MICROPROCESSOR - SYSTEM INTERFACE

The microprocessor system had to be interfaced to BICEPS since this was the machine to be controlled. The interfacing was facilitated by BICEP's transducers and the operational amplifier input of the regulator being at voltage levels compatible to A/D and D/A converters. A block diagram of the interface is shown in Figure 2.6. More detailed representations are given in Appendix B. Frequency , terminal voltage , and an on/off signal indicating whether or not the exciter is in saturation are sufficient for implementing the dynamic and transient stability filters.

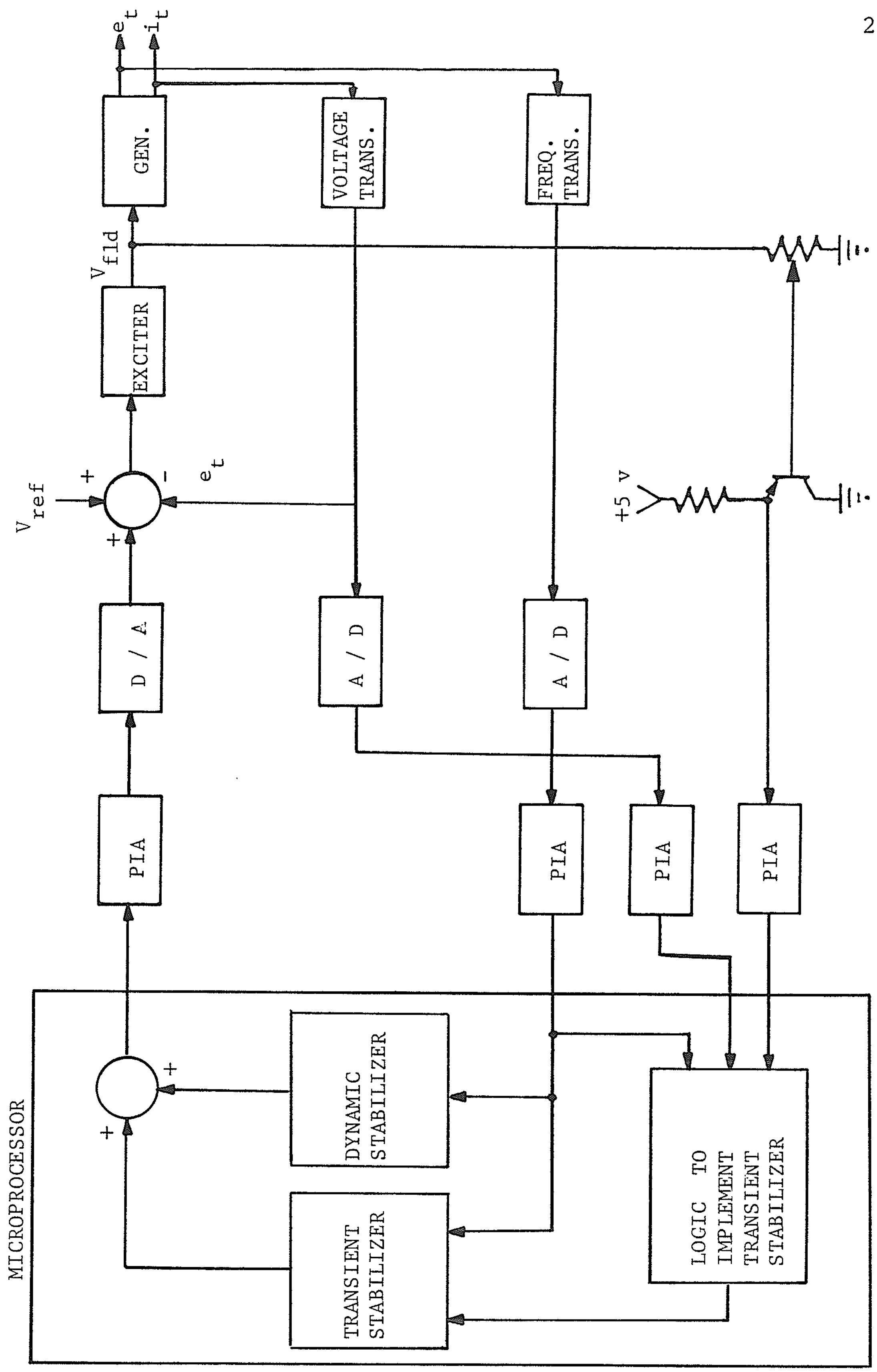


Figure 2.6: Interface Block Diagram

Chapter III

DIGITAL FILTERS

Digital filters have certain inherent properties that make them more suitable than analog filters for certain applications. These properties are:

1. Accuracy can be precisely controlled through a filter's word length.

2. Components do not drift.

3. Component matching is unnecessary as the performance of identical devices is non - varying.

4. Adaptive filtering is relatively simple.

5. Reliability is high.

These properties make digital filters ideal for low frequency applications where analog components usually become large and for high precision applications where analog filters must be individually tuned to match a performance standard.

A digital filter is a digital signal processor that converts a sequence of numbers (input) to another sequence of numbers (output). As the signals to be filtered are generally continuous, they must be sampled at discrete times and quantized to obtain the finite length sequence of numbers that serve as input to the digital filter. The rate and method of sampling the input signal affect the amount of error that is incurred when the input signal is analog reconstructed from the digital data samples. A sampled data processing system is shown in Figure 3.1 . The various elements of this system are described in the following sections.

3.1 SAMPLED DATA SYSTEMS

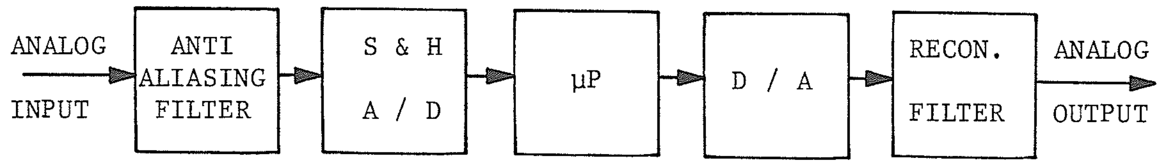
A sampled signal, denoted as $x^*(t)$, is generated by sampling a continuous signal x(t) with a train of impulse functions as shown in Figure 3.2. The impulse train is defined as:

$$\delta_T(t) = \sum_{n=-\infty}^{\infty} \delta(t - n\,T) \qquad \ldots(3.1)$$

where T is the interval between samples.

The sampled signal is thus

$$x^*(t) = x(t)\,\delta_T(t) \qquad \ldots(3.2)$$



ANTI ALIASING FILTER :	Bandlimits input signal to reduce distortion due to sampling.
S & H :	Holds data sample for sufficient time to process.
A / D :	Analog to digital conversion generates a digital word from the sampled analog signal.
μP :	Performs the digital filtering.
D / A :	Converts the digital word back to an analog signal.
RECONSTRUCTION FILTER :	Smooths the quantized signal from the D / A.

Figure 3.1: Elements of a Sampled Data Processing System

or for $t \geq 0$

$$x^*(t) = \sum_{n=0}^{\infty} f(n\,T)\,\delta(t - n\,T) \qquad \ldots(3.3)$$

as $(t-nT) = 0$ outside the sample moments.

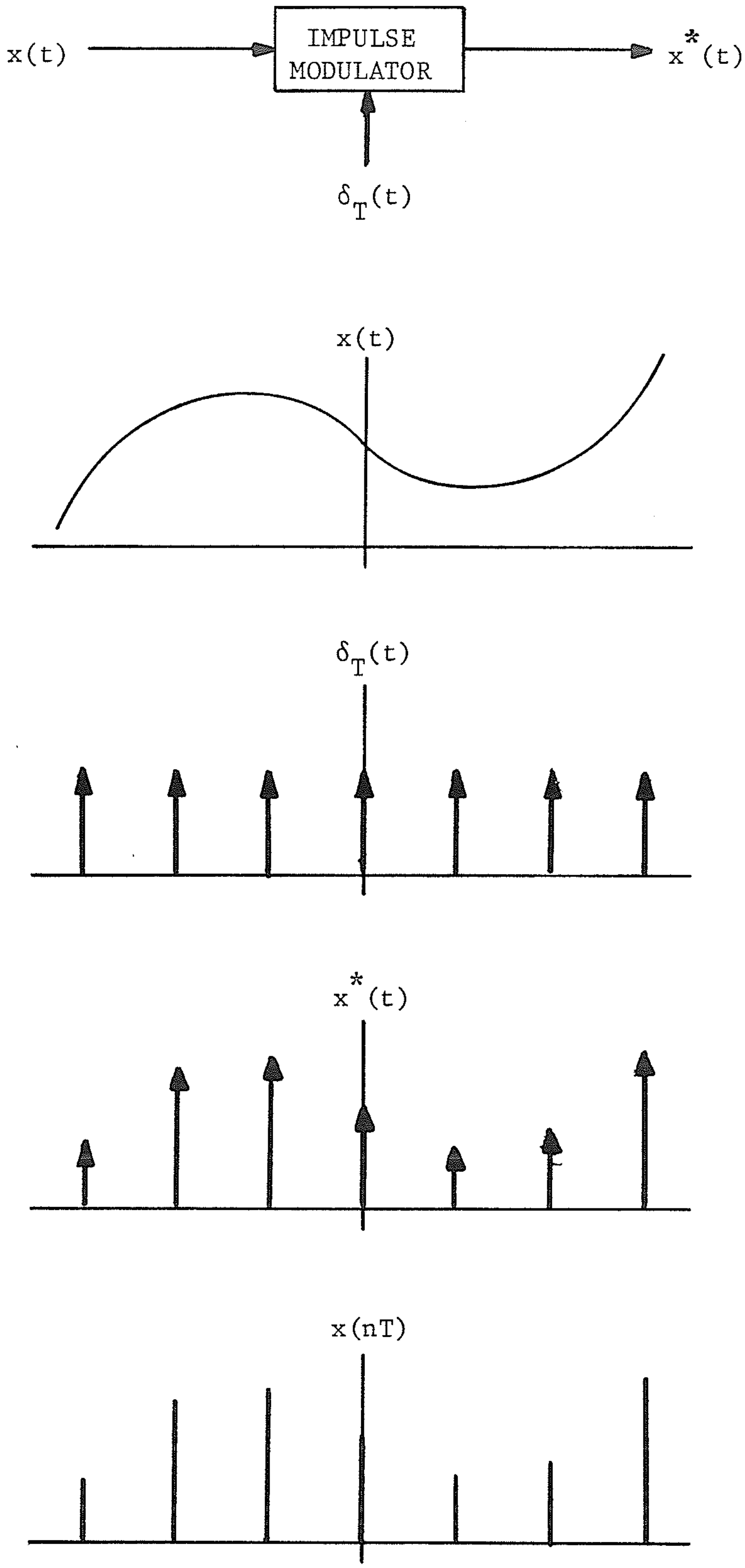


Figure 3.2: Generation of a Sampled Signal

If (3.1) is expanded as a Fourier series,

$$\delta(t) = \sum_{n=-\infty}^{\infty} C_n e^{jn\omega_s T} \qquad \ldots(3.4)$$

where

$$C_n = \frac{1}{T} \int_0^T \delta_T(t) e^{-jn\omega_s t} dt \qquad \ldots(3.5)$$

Since the area of an impulse function is unity

$$C_n = \frac{1}{T} \qquad \ldots(3.6)$$

and

$$\delta_T(t) = \frac{1}{T} \sum_{n=-\infty}^{\infty} e^{jn\omega_s t} \qquad \ldots(3.7)$$

Substituting into the sampled signal equation, (3.2), we get

$$x^*(t) = \frac{1}{T} \sum_{n=-\infty}^{\infty} x(t)\, e^{jn\omega_s t} \qquad \ldots(3.8)$$

Taking the Laplace transform and applying the shifting theorem gives

$$X^*(s) = \frac{1}{T} \sum_{n=-\infty}^{\infty} X(s - jn\omega_s) \qquad \ldots(3.9)$$

or

$$X^*(j\omega) = \frac{1}{T} \sum_{n=-\infty}^{\infty} X[j(\omega - n\omega_s)] \qquad \ldots(3.10)$$

From (3.9) and (3.10) it is evident that as a result of sampling, the frequency spectrum of x(t) is repeated at intervals of $n\omega_s$. This is shown in Figure 3.3.

An important outcome of this is the need to sample at a frequency greater than twice the highest frequency component of x(t). If this is not done the frequency spectra will overlap one another and the original signal cannot be completely recovered. The spectral overlap is referred to as aliasing noise and is shown in Figure 3.4 .

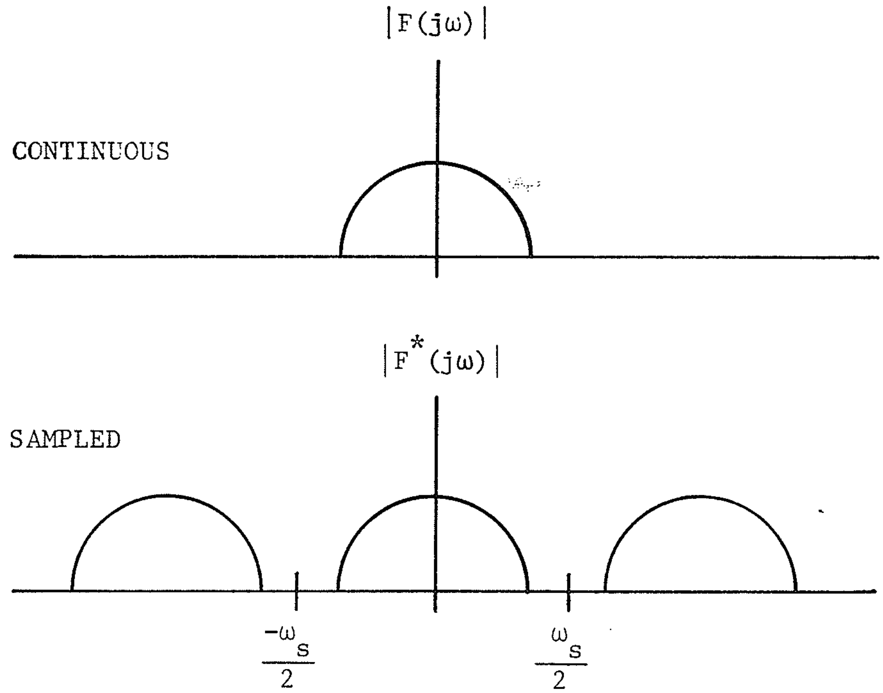


Figure 3.3: Frequency spectra of continuous and sampled signals

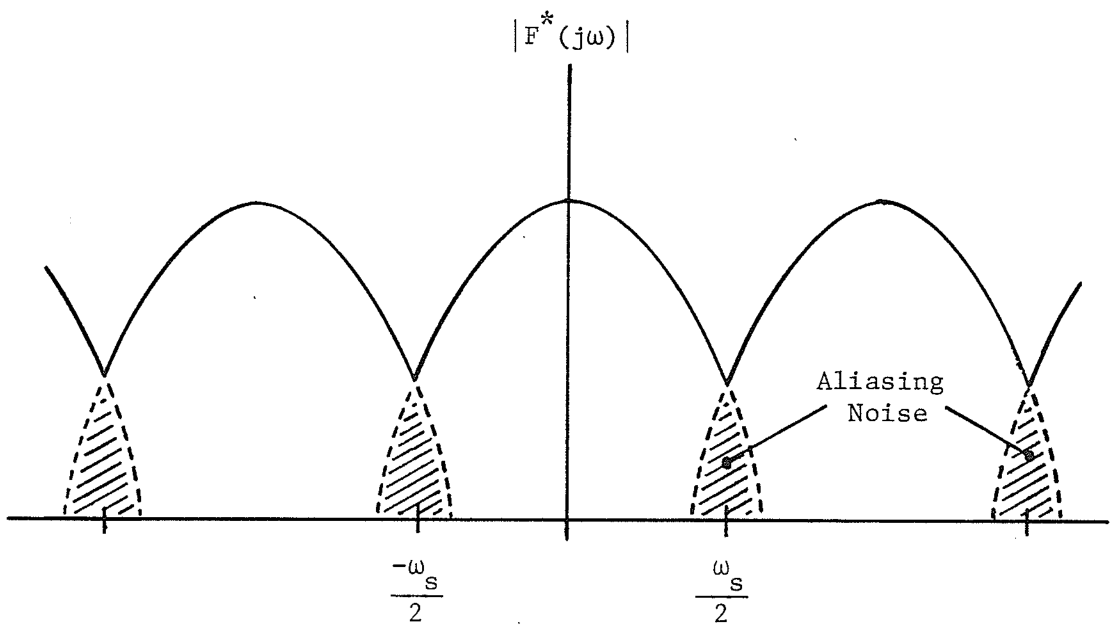


Figure 3.4: Aliasing of frequency spectra

Generally, two measures are taken to prevent aliasing: an anti-aliasing (lowpass) filter is included in the system before the Sample and Hold / Analog -to- Digital converter (S&H - A/D) combination as shown in Figure 3.1; the input signal is sampled at a rate of 5 - 10 times the highest frequency component.

Because of the infinite number of frequency spectra of a sampled data signal, there is an infinite number of poles and zeros in the s - plane representation, making analysis of a system in the s - plane quite difficult. This is shown for a simple transfer function in Figure 3.5. The Z-transform is used to simplify the analysis and design of discrete systems much in the same way as Laplace and Fourier transforms are used to simplify continuous systems. The transformation is defined as

$$X(z) \triangleq Z[x(n\ T)] \triangleq \sum_{n=-\infty}^{\infty} x(n\ T)z^{-n} \qquad \ldots(3.11)$$

where Z is the complex variable

$$z = e^{sT} = e^{-(-\sigma - j\omega)T} \qquad \ldots(3.12)$$

or

$$|z| = e^{\sigma T} \ ; \quad \angle z = \omega T \qquad \ldots(3.13)$$

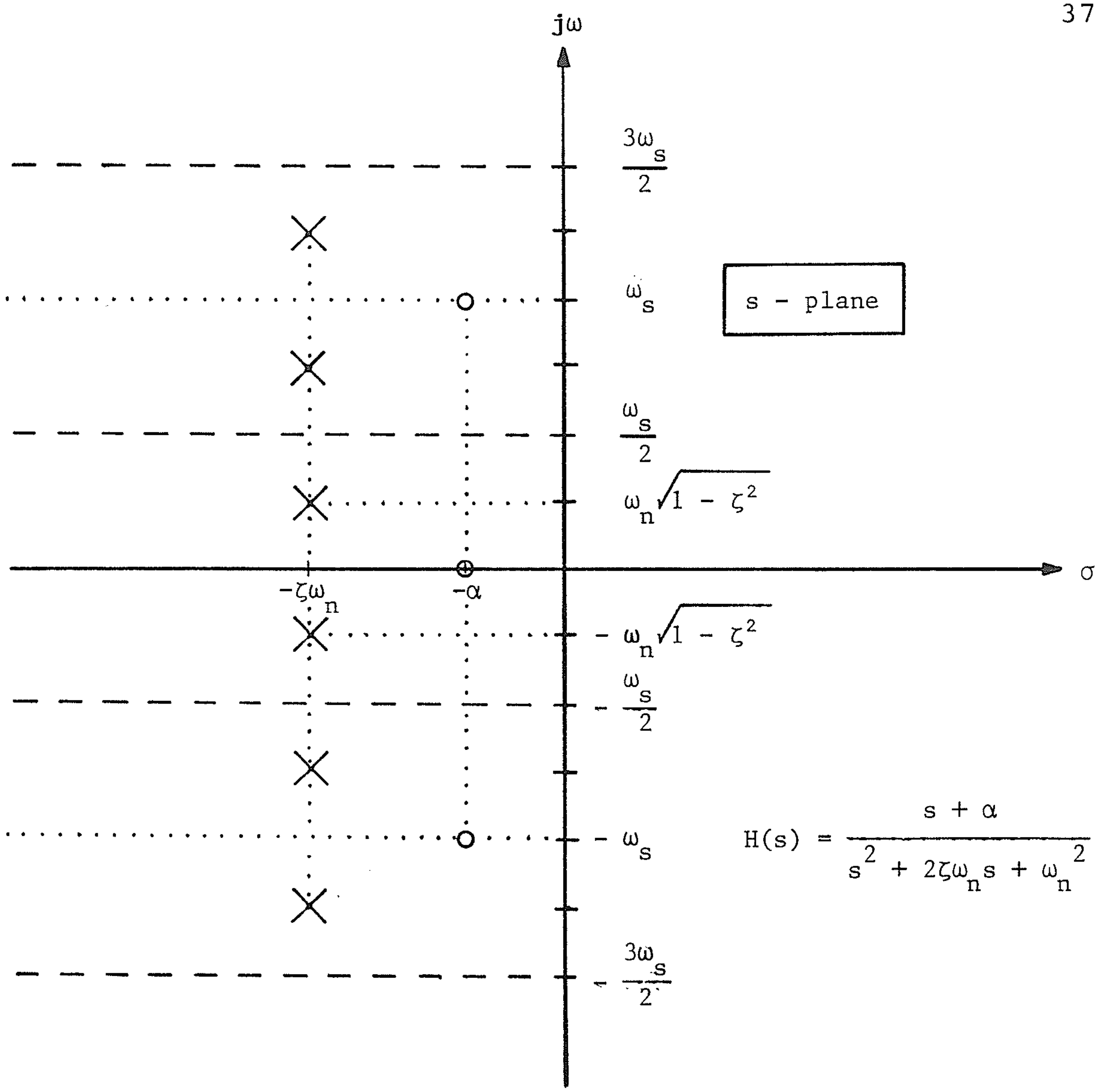


Figure 3.5: Multiple Pole - Zero Pattern for a Sampled System

From (3.13) it is seen that the imaginary axis in the s - plane maps into the unit circle in the z - plane and that the left hand plane in the s- plane maps into the interior of the unit circle in the z - plane. Thus all stability criteria of continuous systems with respect to left hand plane poles apply to digital systems with respect to poles inside the unit circle.

The most important effect of the Z - transformation, however, is that the infinite number of poles and zeros of a sampled data system (Figure 3.5) are now superimposed in the z - plane to a single set of poles and zeros.

3.2 DIGITAL FILTER CONFIGURATIONS

Digital filters can be characterized by either a difference equation or by a transfer function. The transfer function is the Z-transform, H(z), of the impulse response to the linear time - invariant system. In general the transfer function has the form

$$H(z) = \frac{a_0 + a_1 z^{-1} + \ldots + a_n z^{-n}}{1 + b_1 z^{-1} + \ldots + b_n z^{-n}} = \frac{Y(z)}{X(z)} \qquad \ldots(3.14)$$

From (3.14) it is evident that three types of devices are needed to construct digital filters: adders, multipliers, and delay units or shift registers (the term z^{-1} corresponds to a time delay of one sampling period since $z^{-1} = e^{-sT}$). These devices may be implemented through either an algorithm (software) or digital hardware. In this investigation only the software implementation was considered. The computational algorithm for (3.14) is more clearly seen by writing the difference equation that realizes H(z) directly

$$y(m) = \sum_{i=0}^{n} \alpha_i \, x(m-i) - \sum_{j=1}^{n} \beta_j \, y(m-j) \qquad \ldots(3.15)$$

where x(m) is the input sequence

y(m) is the output sequence

α and β are the filter coefficients

This direct form is shown in Figure 3.6.

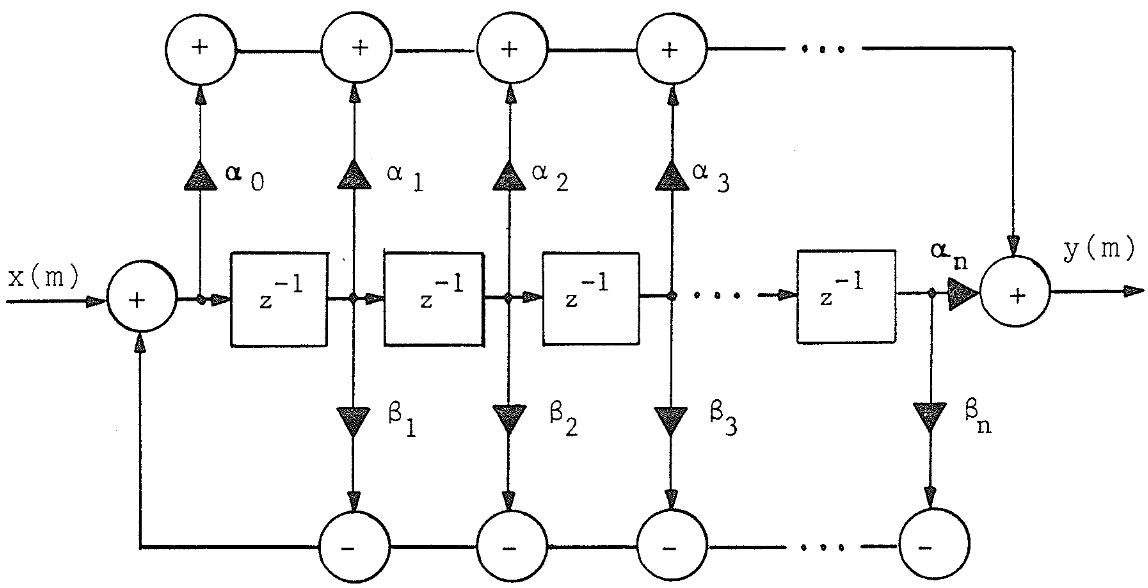


Figure 3.6: Direct Form

Digital filters may be divided into two classes, recursive and nonrecursive filters. Nonrecursive filters produce an output that is a function only of present and previous inputs, hence the β_j terms of (3.15) and Figure 3.6

would all be zero. As these types of filters have a finite impulse response they are called finite impulse response (FIR) filters. Recursive filters produce an output that is a function of present and previous inputs and previous outputs. The feedback in these filters causes an infinite impulse response, hence the name infinite impulse response (IIR) filters. IIR filters are more efficient than FIR filters in terms of implementing high order filters and thus are the only type considered here.

Direct form realizations become numerically inaccurate for high order difference equations. Designs are generally implemented using first and second order subfilters either in a cascaded or parallel form. The second order section of Figure 3.7 can be used to realize any desired filter and is the type used here since it has been shown (12) to be one of the forms least susceptible to instability due to filter coefficient (α, β) roundoff.

3.3 FILTER DESIGN

Once the filter specifications have been set, the problem is to determine the filter coefficients ,α's and β's, that will satisfy them. The most common approach to this problem is to determine a suitable analog filter transfer function,

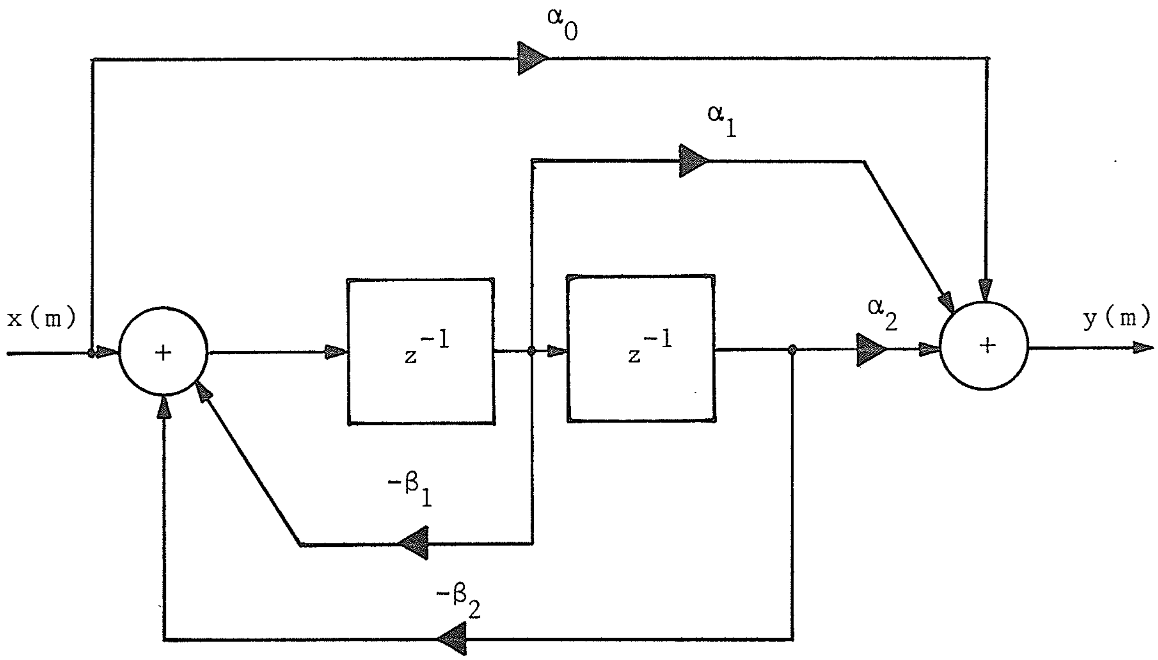


Figure 3.7: Recursive Second - Order Section

H(s), and to then digitize the analog filter. The standard Z-transform can be used for certain types of filters (low - pass and most band - pass) but is inadequate for high pass and bandstop filters. For these non - bandlimited filters it is impossible to prevent aliasing noise using the standard Z-transform, hence an alternative method such as the bilinear Z-transform must be used.

The mapping between the s - plane and z - plane by the bilinear z-transform is described by

$$s = \frac{2}{T}\left[\frac{z-1}{z+1}\right] \qquad \ldots(3.16)$$

and is bandlimiting in nature since it maps the frequency range 0 to ∞ in the continuous case to 0 to π/T in the sampled case. The frequency mapping of the bilinear Z - transform is not linear and so the original cutoff frequency of the analog design must be prewarped before applying the transformation.

$$\omega_c = \frac{2}{T}\tan\left(\frac{\omega_1 T}{2}\right) \qquad \ldots(3.17)$$

where ω_1 = desired digital cutoff frequency

ω_c = prewarped analog cutoff frequency

The standard Z - transform and the bilinear Z - transform are not the only transformations used, but are among the most common. Using the appropriate transformation, it is possible to completely design a filter as an analog filter and then transform it to a digital filter. Alternatively, it is possible to design a lowpass prototype, transform it to a digital prototype and then use appropriate transformations to the type of digital filter desired. Transformation tables for both methods are listed in Chen (4).

3.4 SOURCES OF ERROR AND INSTABILITY

There are three basic sources of error when finite length digital hardware is used to implement a digital filter:

1. Signal quantization in A/D converters.

2. Quantization of filter coefficients (α's, β's).

3. Arithmetic errors due to roundoff or overflow.

Quantization in the A/D converter is not a serious problem but the errors due to (2) and (3) may cause instability or limit cycle oscillations.

3.4.1 Quantization of Filter Coefficients

The standard Z-transform and the bilinear Z-transform both theoretically yield stable digital filters if the transformation is performed on a stable analog model. However, this is only necessarily true if the filter coefficients can be represented to a high degree of precision. For finite, fixed-point, processor word lengths, small changes in the β_j coefficients could cause a pole to move outside the unit circle causing the filter to become unstable.

The exact pole locations are given by the characteristic equation

$$1 + \sum_{j=1}^{N} \beta_j z^{-j} = 0 \qquad \ldots(3.18)$$

If one or more of the filter coefficients changes to a new value $\beta_j + \Delta\beta_i$ due to quantization, the characteristic equation becomes

$$1 + \sum_{j=1}^{N} \beta_j z^{-j} + \Delta\beta_i z^{-i} = 0 \qquad \text{...(3.19)}$$

Referring to (3.13) it is seen that for high sampling frequencies , all the poles are located near the unit circle boundary at z = 1. Therefore, a value of $\Delta\beta_i$ which could cause a pole to move outside the unit circle is given by

$$\Delta\beta_i = 1 + \sum_{j=1}^{N} \beta_j \qquad \text{...(3.20)}$$

A comparison of the $\Delta\beta_i$ term with the largest β_j coefficient ($\Delta\beta_i/\beta_j$) will give the resolution needed (hence the required word length) to avoid instability.

$$\frac{\beta_{j_{max}}}{\Delta\beta_i} \leq 2^{(w-1)} \qquad \text{...(3.21)}$$

where w is the word length necessary.

3.4.2 Limit Cycles

Fixed data word lengths cause a problem with multiplication. An M-bit word multiplied by an N-bit word produces an (M+N)-bit product. This product must be shortened by either rounding or truncation. A consequence of rounding the data is that small scale limit cycles may occur when there is a zero or constant input.

Arithmetic operations with fixed data word lengths may also produce an output value outside the allowed range of values, ie. overflow (or underflow) occurs. Because of the two's complement addition used by most processors, the overflow results in a "wraparound" characteristic, ie. a sudden jump from the largest to the smallest value possible. When overflow occurs it is possible for large scale limit cycles to occur regardless of input. Ebert et al. ([8]) have shown that if the two's complement adder is modified so that it saturates when overflow occurs, no limit cycles will occur.

Chapter IV

DYNAMIC STABILITY FILTER: REALIZATION AND TESTING

Stage three of this project was the implementation of a digital stabilizer for the condition of one natural frequency of oscillation, ie. for the machine versus the system case. The stabilizer had to meet specific criteria of both the power system and of digital filters. The power system criteria were:

1. The stabilizer signal should be in phase with $\Delta\omega$ to give optimum damping.

2. The stabilizer should be a lead-lag function optimized to provide damping at approximately 1 Hz.

3. A washout term should be included so that sustained frequency changes don't alter the terminal voltage.

The above criteria can be met using a bandpass filter. However, it must be realized that to implement an optimum compensator, the system transfer function must be known exactly. As this was not the case, a trial and error approach had to be used for part of the design. An analog computer model of BICEPS as given in (17) was set up and various second order bandpass filters of the form

$$G_c(s) = G \frac{Bs}{s^2 + Bs + \omega_0^2} \qquad \ldots(4.1)$$

were tried to give an indication of the best bandwidth and centre frequency. The best damping was achieved with:

$$B \simeq 10 \text{ Hz.}$$
$$\omega_0 \simeq 4\pi \text{ rad/sec}$$
$$G \simeq 80$$

From these values and the power system criteria for the digital compensator, the following design specifications were derived:

1. Three db. down points of 0.5 Hz. and 10 Hz.

2. Attenuation for frequencies smaller than 0.05 Hz. and greater than 50 Hz. should be at least 20 db.

3. Sampling frequency was chosen to be approximately 10 times the highest frequency component which in this case was the 60 Hz. ripple of the frequency transducer. A clock signal of 580 Hz. was available and so used as the sampling frequency.

4.1 DESIGN PROCEDURE

The bilinear transformation was chosen as it is more suitable for bandpass filters than the standard Z-transformation. The design sequence was as follows:

H(s) normalized analog lowpass Butterworth filter

↓ frequency warping

H'(s) denormalized prewarped analog lowpass filter

↓ bilinear transformation

H(z) digital lowpass prototype

↓ Chen p.256

H(z) digital bandpass filter

As indicated in Section 3.3, a different order of procedure could have been chosen with the resulting filter having the same characteristics though not necessarily the same coefficients.

Figure 8.5 in Chen (4) shows the stop-band attenuation of Butterworth filters of varying degrees. From this chart it was determined that a second order Butterworth lowpass prototype was needed to achieve the required digital filter characteristics. After frequency warping the analog lowpass prototype and performing the bilinear transformation, the resulting digital lowpass prototype was

$$H(z) = \frac{2.4664298 \times 10^{-3} (z + 1)(z + 1)}{z^2 - 1.922410939 z + 0.864553317} \qquad \ldots(4,2)$$

This lowpass prototype met the required gain characteristics but one of its poles was outside the unit circle and hence unstable. This unstable pole would be carried through the lowpass to bandpass transformation causing an unstable realization. This was alleviated by replacing the problem pole with its reciprocal pole. The validity in doing this is shown by considering a digital filter

$$H(z) = \frac{N(z)}{D(z)\,(z - re^{j\theta})} \qquad \ldots(4.3)$$

where $re^{j\theta}$ is the pole outside the unit circle (hence $r > 1$). Replacing $re^{j\theta}$ by its reciprocal pole gives

$$H'(z) = \frac{N(z)}{D(z)\,(z - r^{-1}e^{j\theta})} = \frac{N(z)}{D(z)\,(rz - e^{j\theta})} \qquad \ldots(4.4)$$

where $r^{-1}e^{j\theta}$ is the reciprocal pole of $re^{j\theta}$. Thus all poles of H'(z) are within the unit circle. If $|H'(z)| = |H(z)|$ for all ω then H'(z) has the same gain characteristics as H(z) and is also stable. Expanding the denominator of H(z) and H'(z) and eliminating the common D(z) gives

$$
\begin{aligned}
|z - re^{j\theta}| &= |e^{j\omega T} - re^{j\theta}| \\
&= |\cos\omega T - j\sin\omega T - r\cos\theta - jr\sin\theta| \\
&= \{(\cos\omega T - r\cos\theta)^2 + (\sin\omega T - r\sin\theta)^2\}^{1/2} \qquad \ldots(4.5)
\end{aligned}
$$

$$
\begin{aligned}
|rz - e^{j\theta}| &= |re^{j\omega T} - e^{j\theta}| \\
&= |e^{j\omega T}e^{j\theta}(re^{-j\theta} - e^{-j\omega T})| = |e^{-j\omega T} - re^{-j\theta}| \\
&= \{(\cos\omega T - r\cos\theta)^2 + (\sin\omega T - r\sin\theta)^2\}^{1/2} \qquad \ldots(4.6)
\end{aligned}
$$

Thus using reciprocal poles does not change a filter's gain characteristics. Using the reciprocal pole, the lowpass digital prototype becomes

$$
H(z) = \frac{2.466298 \times 10^{-3}\,(z+1)(z+1)}{z^2 - 1.547539\,z + 0.5955600} \qquad \ldots(4.7)
$$

Transforming the lowpass prototype to a bandpass filter yields

$$H(z) = \frac{0.0101716\ (z^2 - 1)\ (z^2 - 1)}{(z^2 - 1.829441\ z + 0.8299787)\ (z^2 - 1.717056\ z + 0.7175607)} \quad \ldots (4.8)$$

The poles of this filter are

$$0.9147207 \pm 0.0820684$$

$$0.8585282 \pm 0.1396780$$

and are inside the unit circle and thus stable.

4.2 MICROPROCESSOR REALIZATION

Rearranging (4.8) gives the second order section filter coefficients

α_G	$= 0.0101716$		
α_{10}	$= 1.0$	α_{20}	$= 1.0$
α_{11}	$= 0.0$	α_{21}	$= 0.0$
α_{12}	$= -1.0$	α_{22}	$= -1.0$
β_{11}	$= -1.829441$	β_{21}	$= -1.717056$
β_{12}	$= 0.8299787$	β_{22}	$= 0.7175607$

These coefficients, with the exception of α_G , are the multipliers shown in Figure 3.7. α_G is a gain constant that allows the α_0's of all second - order sections to be equal to 1.0. To be realistically implemented in a microprocessor, the processor would need a word length greater than the eight bits of the M6800 (the microprocessor used in this design) or else multiple precision arithmetic would have to be performed. Multiple precision arithmetic, however, would severely decrease the realizable filter bandwidth due to the increased computing time, and was consequently eliminated as a solution.

The multiplications can be realized to the resolution of one bit through an alternating sequence of shift and add operations on partial sums. A standardized routine could have been developed for all multiplications but this would not have been as efficient as a separate routine for each multiplication in terms of execution time. Since speed was more important than overall memory size, it was decided to write a separate routine for each multiplication. To simplify the multiplications and hence further decrease the execution time, computer analyses were performed to find the frequency response when the β coefficients were rounded to values requiring fewer shifts and adds. Suitable results were obtained for the following values:

$$\beta_{11} = -1.875 \qquad \beta_{21} = -1.75$$

$$\beta_{12} = 0.75 \qquad \beta_{22} = 0.75$$

The α values were left unchanged as they were already integer values.

The digital filter program that was developed (Appendix C) was made as general as possible with each second - order section placed in a subroutine. A previously set parameter tells the program how many second - order sections exist. A sampling clock issues an interrupt to the system and causes an analog value to be read and digitized. This signal serves as the input to the first second - order section. The output of each section serves as the input to the next. The final second - order section output goes through a D/A converter and a smoothing filter to become the analog output of the stabilizer.

The suppression of overflow oscillations is desirable since they are a large scale disturbance. As discussed previously, this can be achieved through the use of saturation arithmetic. This was implemented in the digital filter design by making the summer at the end of each second - order section a saturation device. Though this did reduce the accuracy of the filter, it was necessary. Tests made on filters without saturating summers showed that they tended to have large scale limit cycles except for the case of very low level input signals.

4.3 <u>TESTING</u>

The digital filter was first tested apart from the power system to ensure its proper operation. The testing was performed using a wave generator as an input signal. The filter performed as desired and exhibited the multiple passbands around integer values of the sampling frequency as shown previously in Figure 3.3. One interesting fact to note is that even though the signal being inputted to the filter at the various passbands is from $n\omega_s - \omega_p$ to $n\omega_s + \omega_p$ (where n is an integer, ω_p is the passband bandwidth, and ω_s is the sampling frequency), the actual output signal will be a lower frequency signal varying from ω_p at the $n\omega_s - \omega_p$ and $n\omega_s + \omega_p$ frequencies to a D.C. value at ω_s. This is more clearly shown in Figure 4.1. However, this is merely academic as a properly constructed digital signal processor will have an anti - aliasing filter prior to the A/D converter preventing the appearance of these higher frequencies at the filter's input.

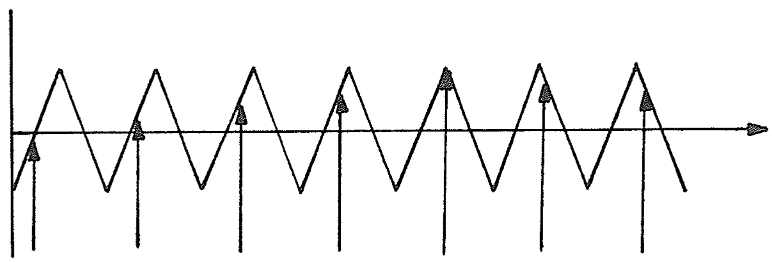
The in-system testing was comprised of two basic tests:

1. An IN - OUT - IN test of the stabilizer to determine its effectiveness in increasing the stability limits of the generator.

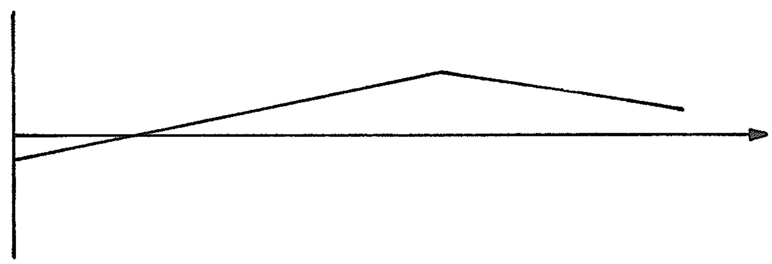
2. A transmission line switching operation.

The tests were repeated using a second order analog filter to use as a comparison.

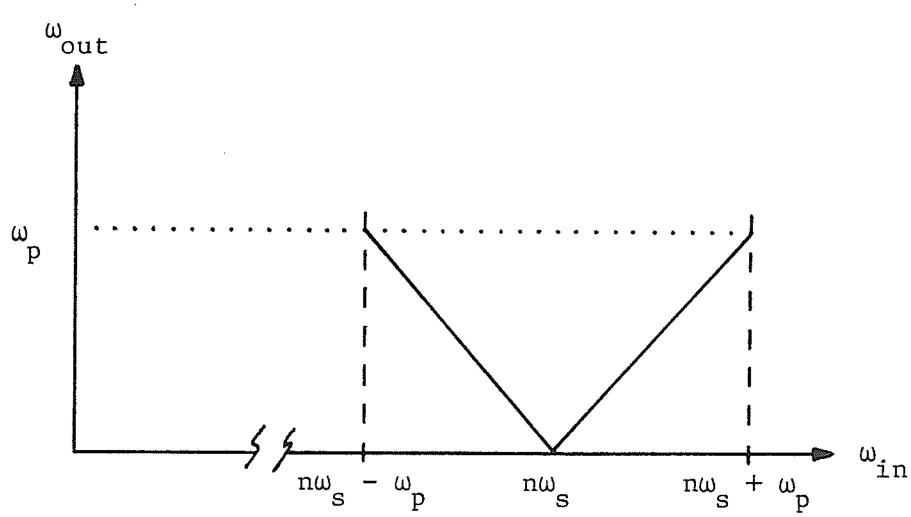
The following oscillograms show the signal from the power transducer of BICEPS. Figure 4.2 shows the ability of a stabilizer to increase the dynamic stability limit of a gen-



a) signal frequency is close to sampling frequency



b) apparent signal frequency



c) frequency output in one passband

Figure 4.1: Frequency Output of a Digital Filter

erator. With the stabilizer in, the system is stable; with it out, the power oscillations increase and the generator would lose synchronism if the stabilizer was not put back in. There is a difference in the amplitude of the power oscillations between the system when using the analog filter and the system when using the digital filter. This could be due to a phase difference between the two filters. Figure 4.3 shows a transmission line switching operation, with the equivalent reactance being switched from 1.5 p.u. to 0.5 p.u. and then back to 1.5 p.u. This is equivalent to suddenly reducing the length of a line and is really a transient stability problem, whereas the stabilizer is designed to act as a dynamic stabilizer. However, it is useful for showing that the stabilizer is not effective for all operating conditions. It would probably be worthwhile to investigate modifying the filter (through software) to implement gain changes for different operating conditions.

In comparing the digital and analog results, it is seen that there is little difference in their effectiveness. However, the digital filter is more easily modified, if desired, to include gain compensation for different operating conditions.

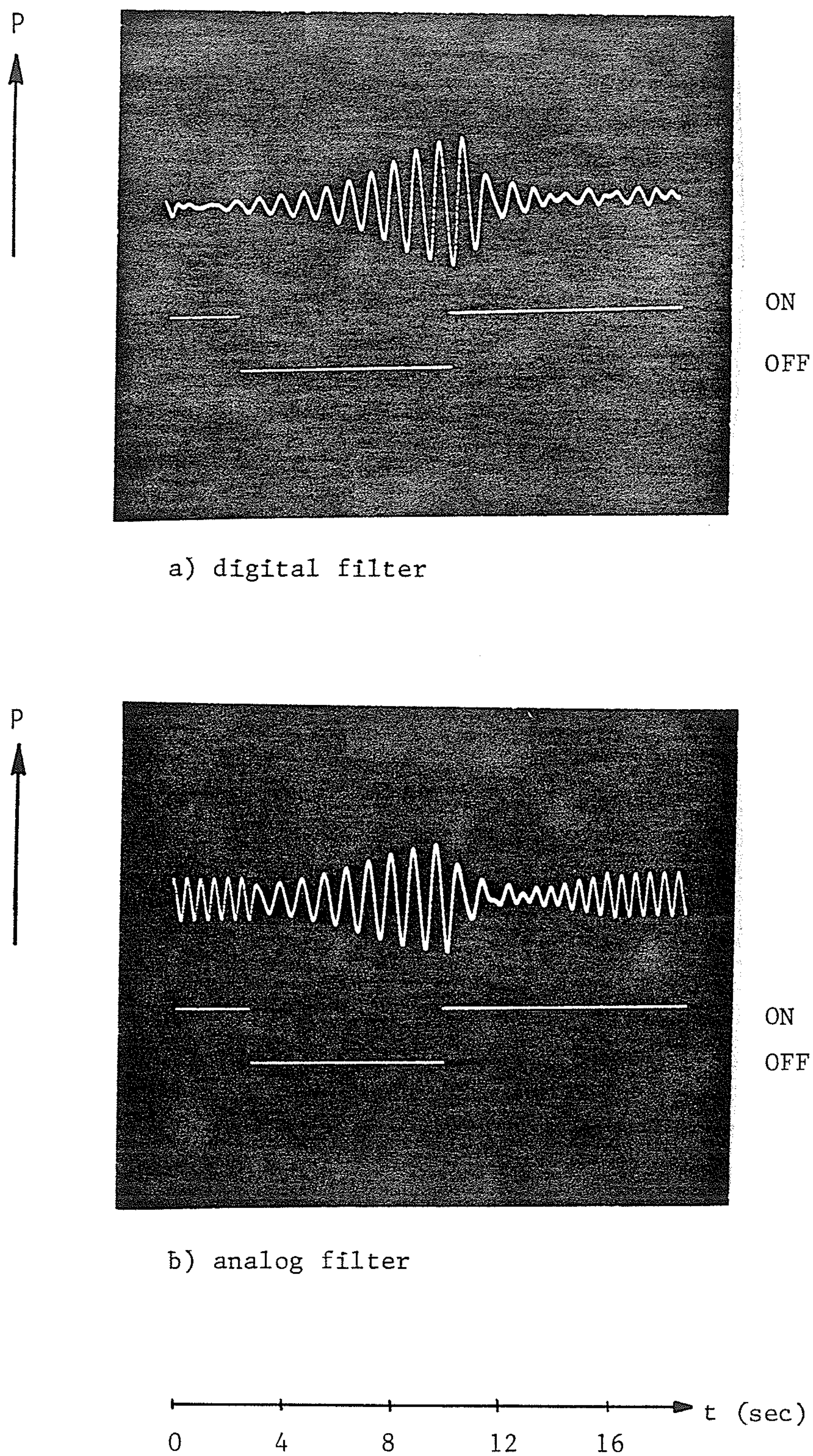
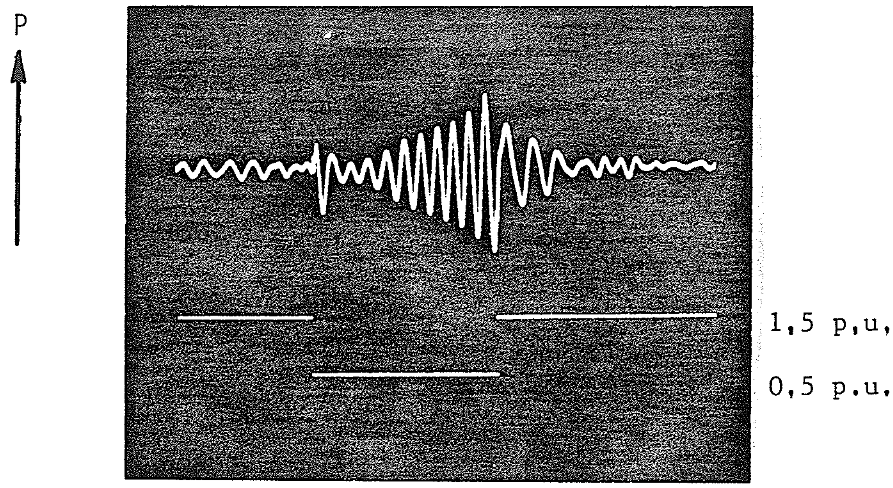
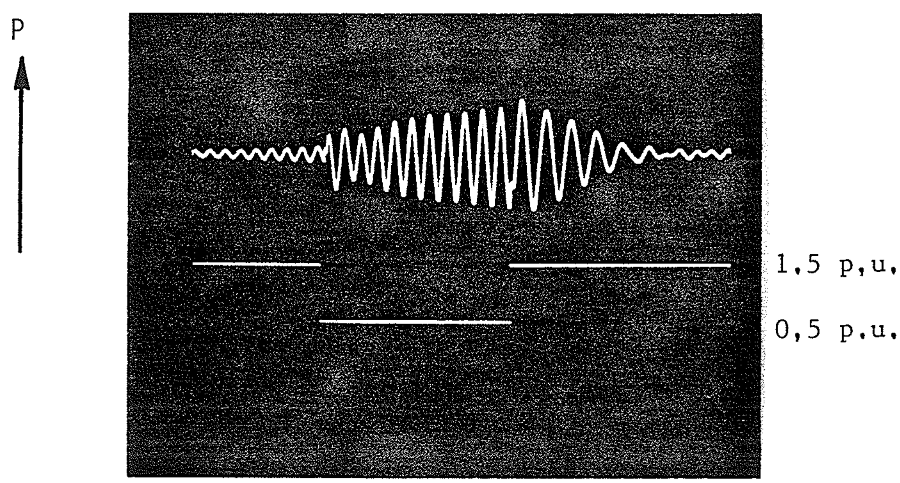


Figure 4.2: Dynamic Stabilizer Operation



a) digital filter



b) analog filter

Figure 4.3: Transmission Line Switching

Chapter V

CONCLUSIONS AND RECOMMENDATIONS

The purpose of this project was to investigate the use of digital filters, implemented through the use of microprocessors, in power system stability control loops. It was desired to use a microprocessor to control and implement multiple stabilizer functions for a system in which two natural frequencies of oscillation occur. Though the project was not completed to the stage originally planned, a solid basis has been developed for continued work. A real system model was developed that exhibits characteristics similar to those of a real power system. An interface between the system model and a microprocessor was implemented. Digital filter structures were studied and methods suitable for use with microprocessors were noted. A digital filter for the case of one natural frequency of oscillation was designed and successfully implemented into the system. The digital stabilizer was compared to an analog stabilizer.

The following conclusions and recommendations became evident during the course of this study. The recommendations are included in hope that they will assist in future work.

1. A software digital filter worked satisfactorily for the continuously acting part of the power system stabilizer. The implementation of an Ontario Hydro type two mode controller is feasible using an M6800 microprocessor and is worth pursuing further.

   As a single compensator cannot provide optimal dynamic stabilization for all operating conditions encountered on a system, better stabilization might be achieved by varying the gain for different operating conditions. Variable gain is easily implemented with a microprocessor and future projects should consider this possibility. The same situation exists for the transient stabilizer(s) that will be utilized.

2. A large part of this project became a study of digital filtering, a relatively new area. In the future there will be more literature available to assist in the design of digital filters, especially with respect to microprocessor implementation. Better software packages should be developed to assist in design, perhaps with some optimization routines for coefficient roundoff. This would be most beneficial for microprocessor implementation.

3. Analog filters are necessary for systems dealing with high frequencies, but as power systems have low frequency oscillations, digital filters are quite suitable for use as controllers. Eight bit microprocessors such as the M6800 are sufficient for these low frequency ranges. The precision and stability of a digital filter can be increased by using a wider word size (say 16 bits instead of 8 bits), but this can also be achieved using an 8 bit microprocessor. The developed filter used a sampling frequency of 580 Hz. With an anti-aliasing filter added to the system, the sampling frequency could be lowered to 60-100 Hz. This would allow more computing time and multiple precision mathematics would then be possible, eliminating the need for a 16 bit microprocessor.

4. It is difficult to simulate large power system characteristics on laboratory sized machines. The 1/5 Hz. oscillation of the large laboratory machine in the developed system is artificial in that it is totally controlled. A real system can give many unexpected results. For this reason a better laboratory model, such as a motor - generator set like BICEPS, should be developed to replace the presently used large laboratory machine.

BIBLIOGRAPHY

1. Analog Devices. Analog - Digital Conversion Notes, Analog Devices Inc. 1977

2. Antoniou, A. Digital Filters: Analysis and Design, McGraw-Hill Inc. 1979

3. Bayne, J.P., Kundar, P., Watson, W. "Static Exciter Control to Improve Transient Stability". IEEE Trans. PAS, Vol. 94, No. 4, pp. 1141-1146, July 1975

4. Chen, C.T. One-Dimensional Digital Signal Processing, Marcel Dekker Inc. 1979

5. Dandeno, P.L., Karas, A.N., McClymont, K.R., Watson, W. "Effect of High - Speed Rectifier Excitation Systems on Generator Stability Limits". IEEE Trans. PAS, Vol. 87, No.1, January 1968

6. DeMello, F.P., Concordia, C. "Concepts of Synchronous Machine Stability as Affected by Excitation Control". IEEE Trans. PAS, Vol. 88, No. 4, pp. 316-329, April 1969

7. Dorf, R.C. Modern Control Systems, 2nd Edition, Addison-Wesley, 1974

8. Ebert, P.M., Mazo, J.E., Taylor, M.G. "Overflow Oscillations in Digital Filters". Bell Systems Technical Journal, November 1969

9. El-Sherbiny, M.K., Mehta, D.M. "Dynamic System Stability Part 1 - Investigation of the Effect of Different Loading and Excitation Systems". IEEE Trans. PAS, Vol. 92, No. 5, September 1973

10. Heffron, W.J., Phillips, R.H. "Effect of Modern Amplidyne Voltage Regulators on Underexcited Operation of Large Turbine Generators". AIEE Trans. (Power Apparatus and Systems), Vol. 71, pp. 692-697, August 1952

BIBLIOGRAPHY

11. Jackson, L.R., Kaiser, J.F., McDonald, H.S. "An Approach to the Implementation of Digital Filters". IEEE Trans. A.U. ,Vol. 16, No. 3, September 1968

12. Kaiser, J.F. "Digital Filters". Systems Analysis by Digital Computer , Kuo, F.F., Kaiser, J.F. (editors). Wiley 1966

13. Kaiser, J.F. "Some Pratical Considerations in the Realization of Linear Digital Filters". Proceedings of the 3rd Annual Alterton Conference on Circuit and System Theory , 1965

14. Motorola Semiconductor Products Inc. M6800 Microprocessor Applications Manual , Motorola Inc. 1975

15. Oppenheim, A.V., Schafer, R.W. Digital Signal Processing , Prentice-Hall, 1975

16. Parker, S.R., Hess, S.F. "Limit Cycle Oscillations in Digital Filters", IEEE Trans. C.T. , Vol. 18, pp. 687-697, 1971

17. Ryczowski, E.M. Some Compensation Techniques Applied to a Laboratory Power System , M.Sc. Thesis, University of Manitoba, 1977

18. Watson, W., Manchur, G. "Experience With Supplementary Damping Signals for Generator Static Excitation Systems". IEEE Trans PAS , Vol. 92, No. 1, pp. 199-204, January 1973

Appendix A

HEFFRON - PHILLIPS CONSTANTS

K_1 = change in electrical torque for a change in rotor angle with constant flux linkages in the d - axis

K_2 = change in electrical torque for a a change in d - axis flux linkages with constant rotor angle

K_3 = impedance factor

K_4 = demagnitizing effect of a change in rotor angle

K_5 = change in terminal voltage with change in rotor angle for constant E_{fld}

K_6 = change in terminal voltage with change in E_{fld} for constant rotor angle

- -

$T_{dz'}$ = effective field time constant under load

$T_{do'}$ field open circuit time constant

Appendix B

PROCESSOR - POWER SYSTEM INTERFACE

User Memory Addresses (HEX)

Heathkit

0000 - 00FF (Refer to Heathkit manual)

Additional memory

0800 - 0BFF
or
0C00 - 0FFF

PIA Addresses (HEX)

PIA A 8#*0 - 8#*3

PIA B 8#*8 - 8#*B

0 → 7
* don't care state

Notes:

1. A / D is located at PIA A DRA (8000)
2. D / A is located at PIA A DRB (8002)
3. System is designed to run off of a Heathkit ET-3400 Trainer. No additional power supplies are needed.
4. Analog input should be limited to + 4.7 volts.
5. Analog output is limited to + 4.7 volts.

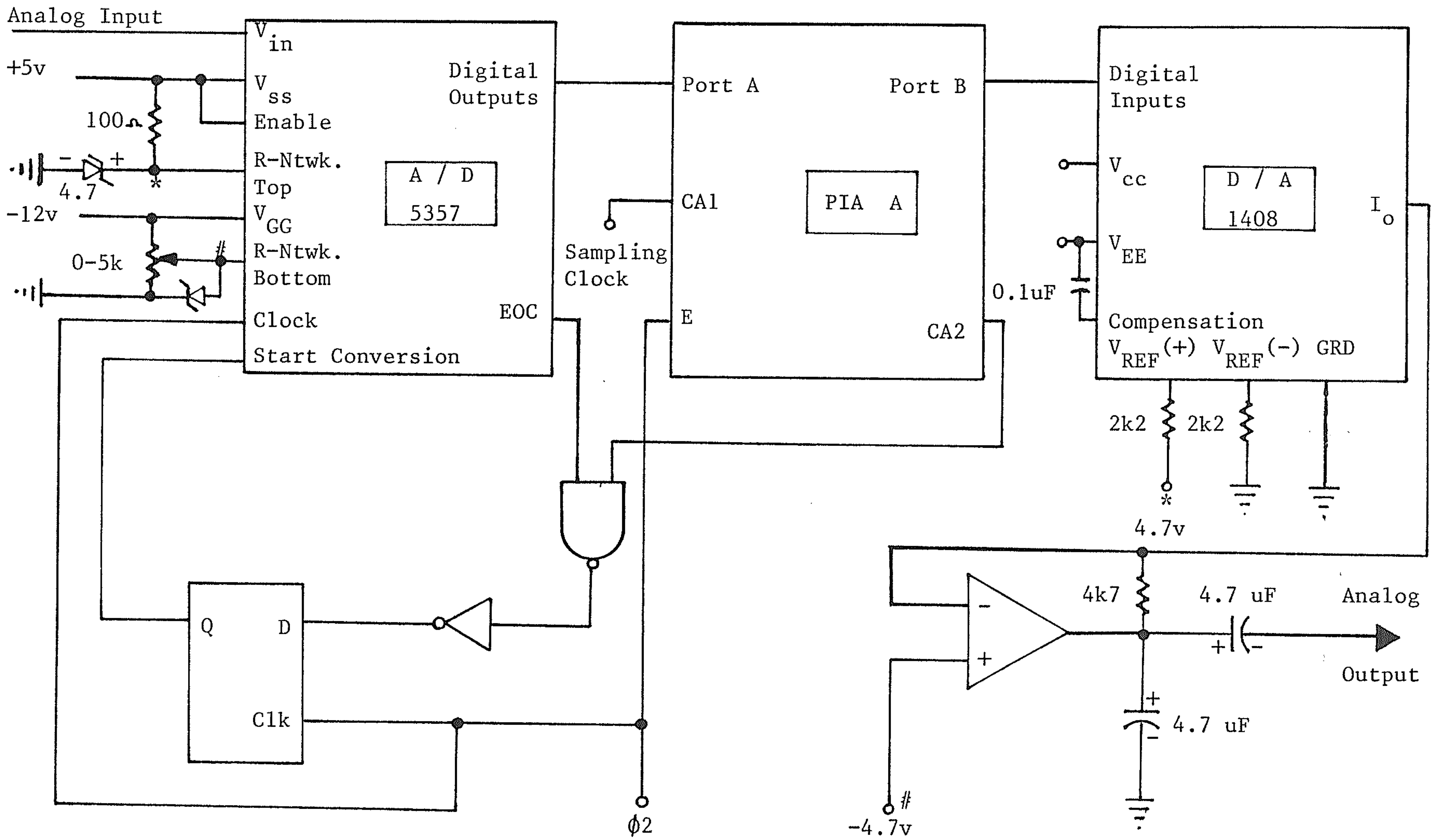


Figure B1: A / D - PIA - D / A Interface

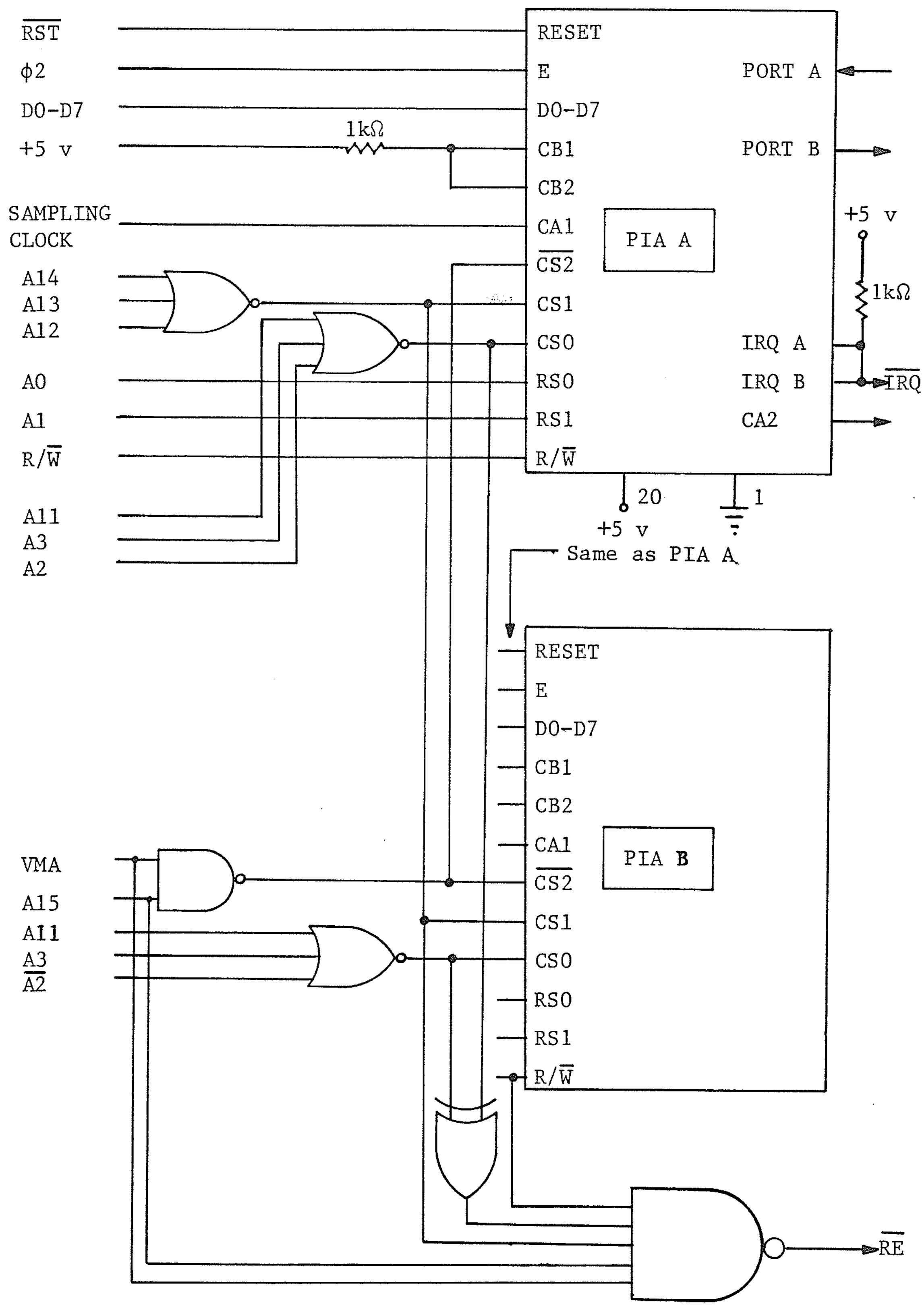


Figure B2: Microprocessor - PIA Interface

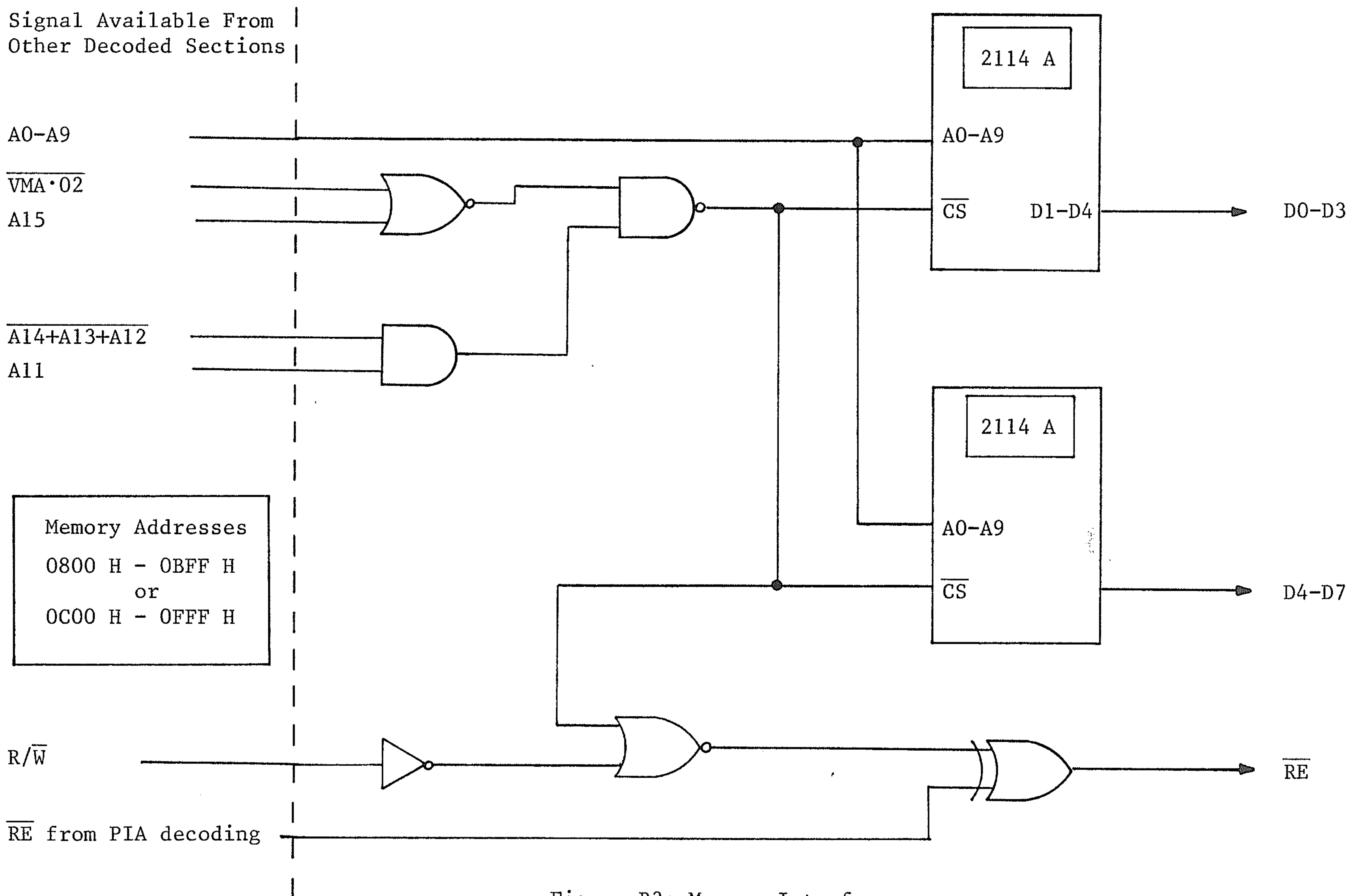
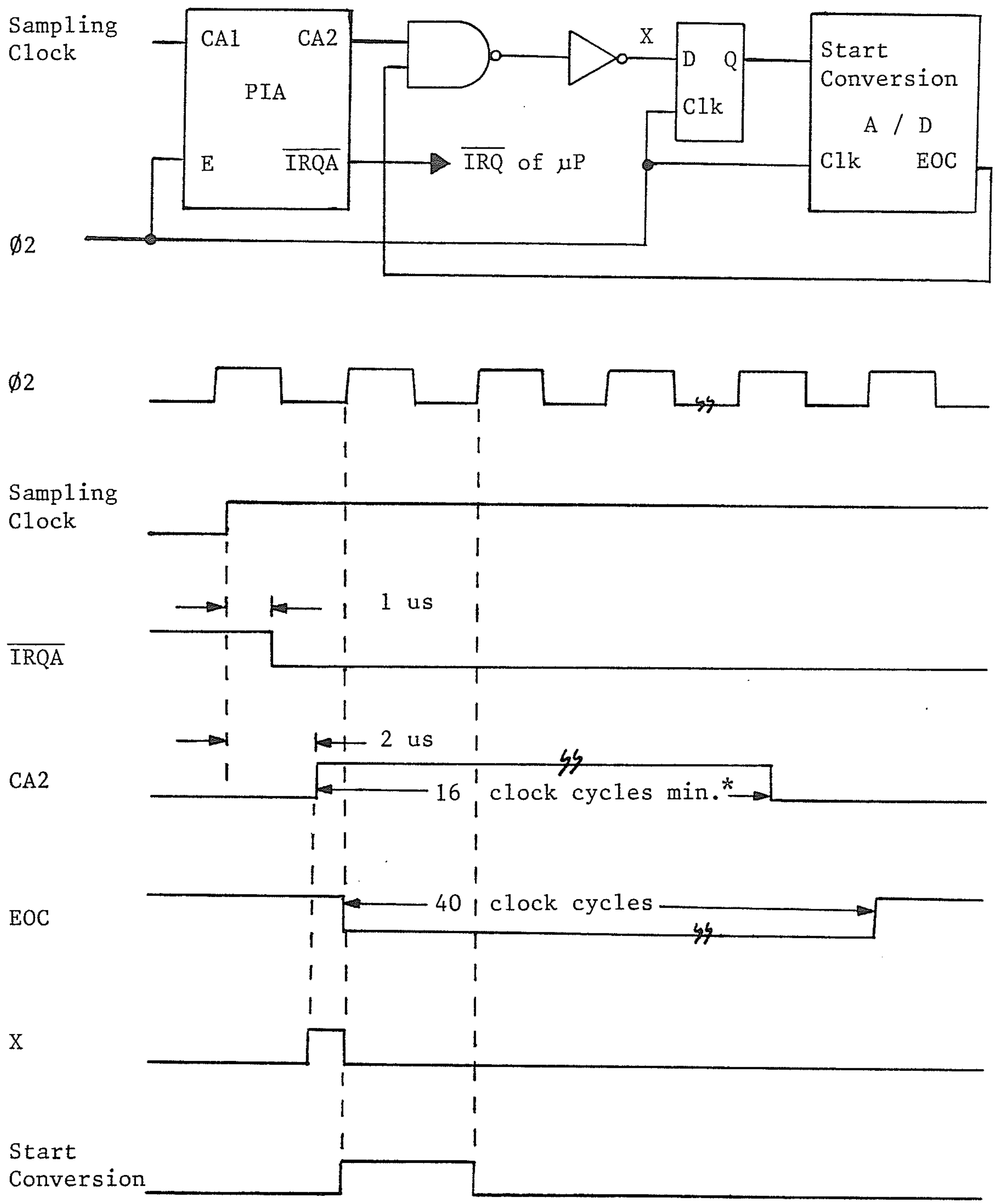


Figure B3: Memory Interface



* Minimum time for CA2 remaining high is 16 machine cycles (clock cycles for 6800). This comes from the 12 cycles for $\overline{\text{IRQ}}$ and 4 cycles for LDAA (DRA), ie. the PIA data register A must be read in order to change CA2 from high to low state. This should be done immediately once the interrupt routine has been accessed.

Figure B4: PIA - A / D Interface Timing

Appendix C

FILTER SOFTWARE

MOTOROLA M68SAM CROSS-ASSEMBLER

MOTOROLA M6800 CROSS ASSEMBLER, RELEASE 1.3

```
                         NAM   FILTER
0000                     ORG   0000H
                  *
                  *      THIS  ROUTINE REPEATS FOR EACH OF  THE N
                  *      SECOND ORDER SECTIONS. IT ROTATES THE
                  *      PSEUDO STACK FOR EACH SECTION, DROPPING
                  *      OFF THE OLD VALUE AND SAVES THE NEW
                  *      VALUE (A). A,B,C ARE USED TO CALCULATE
                  *      THE OUTPUT OF EACH SECTION WHICH THEN
                  *      SERVES AS THE INPUT TO THE NEXT SECTION.
                  *      IN THE CASE OF  THE FINAL SECOND ORDER
                  *      SECTION, THE  OUTPUT IS THE OUTPUT
                  *      OF THE FILTER. THE ALPHA AND BETA
                  *      COEFFICIENTS OF EACH SECTION HAVE BEEN
                  *      PREVIOUSLY CALCULATED. EACH SECTION HAS
                  *      ITS OWN STACK IMMEDIATELY FOLLOWED BY
                  *      ROUTINE TO CALCULATE THE SECTION OUTPUT.
                  *      THE FIRST STACK AND ROUTINE STARTS AT
                  *      0800H. INPUT IS THROUGH A 5357 A/D AND
                  *      THE OUTPUT IS THROUGH A 1408 D/A. THE
                  *      MICROPROCESSOR USED IS AN M6800.


0000 0002         LOC1   RMB   02
0002 0001         LOC2   RMB   01
0003 0001         LOC2A  RMB   01
0004 0001         LOCX   RMB   01
0005 0001         LOCB   RMB   01
```

```
0800                     ORG    0800H
0800 00                  FCB    00H,00H,00H
0801 00
0802 00
0A00                     ORG    0A00H
0A00 00                  FCB    00H,00H,00H
0A01 00
0A02 00



0010                     ORG    0010H
0010 01                  NOP
0011 01                  NOP
0012 01                  NOP
0013 0F                  SEI
0014 86 7E               LDA A #$7E
0016 97 F7               STA A $F7        INTERRUPT VECTOR
                                          LOCATION



0018 CE 0040             LDX    #$0040
001B DF F8               STX    $0F8
                *
                *     PIA INITIALIZATION
                *
001D CE 0027             LDX    #$0027
0020 FF 8004             STX    $8004
0023 CE FF04             LDX    #$0FF04
0026 FF 8006             STX    $8006
0029 0E                  CLI

                *     OUTPUT Y(T) OF FILTER

002A 96 04      CONT     LDA A LOCX
002C 88 7F               EOR A #$7F       2'S COMP.- OFFSET
                                          BINARY
002E B7 8006             STA A $8006
0031 20 F7               BRA    CONT
0040                     ORG    0040H
                *
                *     INTERRUPT ROUTINE
                *
0040 B6 8006             LDA A $8006      CA2 GOES LOW
0043 9F 00               STS    LOC1
0045 C6 01               LDA B #$01       NUMBER OF 2ND ORDER
                                          SECTIONS
```

```
0047 D7  05                STA B LOCB
0049 8E  0800              LDS   #$0800
004C 30          LOOP      TSX
004D 09                    DEX               SET X SAME AS S.P.
004E 32                    PUL A             ROTATES B TO C
004F 34                    DES               DROPPING OFF OLD C
0050 36                    PSH A             AND ROTATES A TO B
0051 31                    INS               S.P. IS AT  ORIGINAL
0052 31                    INS               POSITION AT TOP OF
0053 32                    PUL A             SECOND ORDER
0054 34                    DES               STACK WHEN
0055 36                    PSH A             FINISHED
0056 6E  03                JMP   03,X        CALC A AND Y
0058 7A  0005 RTN          DEC   LOCB
005B 27  0C                BEQ   RTI
005D 9F  02                STS   LOC2
005F 96  02                LDA A LOC2
0061 8B  02                ADD A #$02
0063 97  02                STA A LOC2
0065 9E  02                LDS   LOC2
0067 20  E3                BRA   LOOP
0069 9E  00      RTI       LDS   LOC1
006B 3B                    RTI



0A03                       ORG   0A03H
                 *         SECOND  2ND ORDER FILTER SECTION
                 *         PROGRAM INCLUDES OVERFLOW MINIMIZATION


0A03 96  04                LDA A LOCX        INPUT FROM FIRST
                                             SECTION
0A05 01                    NOP
0A06 47                    ASR A



0A07 2C  03                BGE   LOCOK
0A09 7E  0AA4              JMP   LOCF
0A0C E6  00      LOCOK     LDA B 00,X        REGISTER C
0A0E 2C  1E                BGE   LOCG
0A10 E6  01                LDA B 01,X        REGISTER B
0A12 2D  3C                BLT   LOCH
0A14 1B                    ABA
0A15 28  03                BVC   LOCPK
0A17 7E  0A9C              JMP   LOCXX
0A1A 57          LOCPK     ASR B
```

```
0A1B 1B              ABA
0A1C 29 7E           BVS   LOCXX
0A1E 57              ASR B
0A1F 1B              ABA
0A20 29 7A           BVS   LOCXX
0A22 E6 00           LDA B 00,X     REGISTER C
0A24 57              ASR B
0A25 10              SBA
0A26 29 74           BVS   LOCXX
0A28 57              ASR B
0A29 10              SBA
0A2A 29 70           BVS   LOCXX
0A2C 20 47           BRA   LOCA

0A2E 57        LOCG  ASR B
0A2F 10              SBA
0A30 57              ASR B
0A31 10              SBA
0A32 E6 01           LDA B 01,X     REG B
0A34 2E 0D           BGT   LOCJ
0A36 1B              ABA
0A37 29 67           BVS   LOCY
0A39 57              ASR B
0A3A 1B              ABA
0A3B 29 63           BVS   LOCY
0A3D 57              ASR B
0A3E 1B              ABA
0A3F 29 5F           BVS   LOCY
0A41 20 32           BRA   LOCA

0A43 1B        LOCJ  ABA
0A44 29 56           BVS   LOCXX
0A46 57              ASR B
0A47 1B              ABA
0A48 29 52           BVS   LOCXX
0A4A 57              ASR B
0A4B 1B              ABA
0A4C 29 4E           BVS   LOCXX
0A4E 20 25           BRA   LOCA

0A50 1B        LOCH  ABA
0A51 2D 10           BLT   LOCK
0A53 57              ASR B
0A54 1B              ABA
0A55 57              ASR B
0A56 1B              ABA
0A57 E6 00           LDA B 00,X     REG C
```

```
0A59 57              ASR B
0A5A 10              SBA
0A5B 29 3F           BVS   LOCXX
0A5D 57              ASR B
0A5E 10              SBA
0A5F 29 3B           BVS   LOCXX
0A61 20 12           BRA   LOCA

0A63 E6 00    LOCK   LDA B 00,X       REG C
0A65 57              ASR B
0A66 10              SBA
0A67 57              ASR B
0A68 10              SBA
0A69 E6 01           LDA B 01,X       REG B
0A6B 57              ASR B
0A6C 1B              ABA
0A6D 29 31           BVS   LOCY
0A6F 57              ASR B
0A70 1B              ABA
0A71 29 2D           BVS   LOCY
0A73 20 00           BRA   LOCA

0A75 A7 02    LOCA   STA A 02,X       NEW A REGISTER VALUE
0A77 2C 10           BGE   LOCP
0A79 E6 00           LDA B 00,X
0A7B 2F 09           BLE   LOCQ
0A7D 10              SBA
0A7E 29 14           BVS   LOCS
0A80 47       END    ASR A
0A81 97 04           STA A LOCX       OUTPUT REGISTER
0A83 7E 0058         JMP   RTN

0A86 10       LOCQ   SBA
0A87 20 F7           BRA   END
0A89 E6 00    LOCP   LDA B 00,X
0A8B 2F 02           BLE   LOCR
0A8D 20 F7           BRA   LOCQ
0A8F 10       LOCR   SBA
0A90 29 06           BVS   LOCT
0A92 20 EC           BRA   END
0A94 86 80    LOCS   LDA A #$80
0A96 20 E8           BRA   END
0A98 86 7F    LOCT   LDA A #$7F
0A9A 20 E4           BRA   END
0A9C 86 7F    LOCXX  LDA A #$7F
0A9E 20 D5           BRA   LOCA
0AA0 86 80    LOCY   LDA A #$80
0AA2 20 D1           BRA   LOCA
0AA4 E6 00    LOCF   LDA B 00,X
0AA6 2E 22           BGT   LOCL
0AA8 57              ASR B
```

```
0AA9 10                 SBA
0AAA 57                 ASR B
0AAB 10                 SBA
0AAC E6 01              LDA B 01,X
0AAE 2E 0D              BGT   LOCM
0AB0 1B                 ABA


0AB1 29 ED              BVS   LOCY
0AB3 57                 ASR B
0AB4 1B                 ABA
0AB5 29 E9              BVS   LOCY
0AB7 57                 ASR B
0AB8 1B                 ABA
0AB9 29 E5              BVS   LOCY
0ABB 20 B8              BRA   LOCA

0ABD 1B        LOCM     ABA
0ABE 29 DC              BVS   LOCXX
0AC0 57                 ASR B
0AC1 1B                 ABA
0AC2 29 D8              BVS   LOCXX
0AC4 57                 ASR B
0AC5 1B                 ABA
0AC6 29 D4              BVS   LOCXX
0AC8 20 AB              BRA   LOCA

0ACA E6 01     LOCL     LDA B 01,X
0ACC 2F 25              BLE   LOCN
0ACE 1B                 ABA
0ACF 2F 12              BLE   LOCO
0AD1 E6 00              LDA B 00,X
0AD3 57                 ASR B
0AD4 10                 SBA
0AD5 57                 ASR B
0AD6 10                 SBA
0AD7 E6 01              LDA B 01,X
0AD9 57                 ASR B
0ADA 1B                 ABA
0ADB 29 BF              BVS   LOCXX
0ADD 57                 ASR B
0ADE 1B                 ABA
0ADF 29 BB              BVS   LOCXX
0AE1 20 92              BRA   LOCA

0AE3 57        LOCO     ASR B
```

```
0AE4 1B                ABA
0AE5 57                ASR B
0AE6 1B                ABA
0AE7 E6 00             LDA B 00,X
0AE9 57                ASR B
0AEA 10                SBA
0AEB 29 B3             BVS   LOCY
0AED 57                ASR B
0AEE 10                SBA
0AEF 29 AF             BVS   LOCY
0AF1 20 82             BRA   LOCA

0AF3 1B        LOCN    ABA
0AF4 29 AA             BVS   LOCY
0AF6 57                ASR B
0AF7 1B                ABA
0AF8 29 A6             BVS   LOCY
0AFA 57                ASR B

0AFB 1B                ABA
0AFC 29 A2             BVS   LOCY
0AFE E6 00             LDA B 00,X
0B00 57                ASR B
0B01 10                SBA
0B02 29 9C             BVS   LOCY
0B04 57                ASR B
0B05 29 99             BVS   LOCY
0B07 7E 0A75           JMP   LOCA

0803                   ORG   0803H
               *
               *      FIRST 2ND ORDER SECTION
               *
0803 B6 8004           LDA A $8004     INPUT FROM A/D
0806 88 7F             EOR A #$7F      COMP. OFFSET BINARY
                                       TO  2'S COMP.
0808 47                ASR A
0809 47                ASR A
080A 47                ASR A
080B 47                ASR A
080C E6 01             LDA B 01,X
080E 1B                ABA
080F 1B                ABA
```

```
0810 57          ASR B
0811 57          ASR B
0812 10          SBA
0813 57          ASR B
0814 10          SBA
0815 E6 00       LDA B 00,X
0817 10          SBA
0818 57          ASR B
0819 57          ASR B
081A 1B          ABA
081B A7 02       STA A 02,X
081D E6 00       LDA B 00,X
081F 10          SBA
0820 97 04       STA A LOCX
0822 7E 0058     JMP   RTN
                 END
```

SYMBOL TABLE

```
0000  LOC2   0002  LOC2A  0003  LOCX   0004  LOCB   0005
002A  LOOP   004C  RTN    0058  RTI    0069  LOCOK  0A0C
0A1A  LOCG   0A2E  LOCJ   0A43  LOCH   0A50  LOCK   0A63
0A75  END    0A80  LOCQ   0A86  LOCP   0A89  LOCR   0A8F
0A94  LOCT   0A98  LOCXX  0A9C  LOCY   0AA0  LOCF   0AA4
0ABD  LOCL   0ACA  LOCO   0AE3  LOCN   0AF3
```