A SOLID STATE VOLTAGE REGULATOR AND EXCITER FOR A LARGE POWER SYSTEM TEST MODEL

by

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We accept this thesis as conforming to the required standard

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THE UNIVERSITY OF BRITISH COLUMBIA

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ABSTRACT

This thesis is concerned with the design, construction, and testing of a solid state voltage regulator-exciter for a new power system test model. The basic elements of a power system, the prime mover and the synchronous generator, are simulated by small laboratory sized machines whose characteristics have been altered by electronic devices. The voltage regulator-exciter consists of a transfer function simulator, a selector switch for forced or linear excitation, a field amplifier, a negative resistor for the field circuit and a voltage transducer for closed loop control. Also included are a braking resistor with solid state switching and power and speed transducers.

Transient power system studies have been largely theoretical in nature because of the difficulty and danger in carrying out experimental work on actual power systems. With this model, many important and interesting tests can be easily carried out.

Chapter 2 outlines the proposed system while Chapter 3 details the realization and circuitry of the subsystems. The results of the tests described in Chapter 4 are divided into two parts. The first section deals with the testing of each subsystem while the last section, the system as a whole.
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1. INTRODUCTION

1.1 Development of Micromachines

Micromachine modelling of synchronous machines and power systems is not new. A micromachine and microréseaux system was proposed and constructed by Robert in 1950. He stated certain electromagnetic and mechanical similarities that must be observed between the real machine and the model. The per unit reactances between the actual system and the representative system must be similar, the time constants of the corresponding circuits must be equal, the saturation curves must be identical, the inertia constants must be similar, and finally, the torque-speed characteristics of the micromachine must be identical with the real machine. Robert employed a rotary machine to obtain a negative resistance in order to increase the field time constant. The cross section of the damper windings was altered to achieve the correct time constant. Variable air gaps in the rotor and stator permitted the saturation curves to be adjusted. Three different types of rotors were used to simulate a salient pole generator, motor and a turbo generator. Weights fastened to the shaft of the micromachine allowed the inertia constant to be altered over a wide range. The model prime mover was a dc machine driven by gas-filled thyatrons. Altering the firing angle of the thyatrons varied the torque-speed characteristics. Speed sensing was achieved by means of a frequency modulation system. A tone generator whose frequency was proportional to angular speed was slope-demodulated by an R-C filter. The transmission line was simulated by a three phase pi section.

Venikov in 1952 reported a thorough study of the mathematics of micromachine simulation. Specified in detail are the similarity
requirements for many electrical and mechanical quantities including time, mechanical and electrical power losses, torque, moment of inertia, self and mutual inductance, magnetic field and space harmonics, and thermal conductivity. In Venikov's micromachine, the field time constant is adjusted by means of a negative resistor realized by a commutator machine. Three different rotors were employed to achieve variations in saliency; $x_d = 0.85, 0.55, 0.40$. The damper windings were adjustable allowing various amounts of damping to be introduced. The inertia constant of the micromachine set was varied by means of a flywheel. A twelve section transmission line was used to simulate a tie line. Provision was made to simulate arc, corona, overvoltage switching and lightning surges.

A third micromachine was described by Adkins in 1960. Short-circuits, synchronizing and damping torques, swing curves, asynchronous operation, and resynchronization were investigated. A synchronous machine rather than the complete power system was simulated and tested. The machines were specially constructed so that their per unit armature resistance value was considerably lower than commercial units of the same rating. An electronic negative resistance was used to vary the field time constant. The voltage regulator was simulated by operational amplifiers. Five rotors of different design were constructed. The damper bars of the rotors could be changed or removed; an adjustable flywheel provided a wide range of inertia constants. A tachogenerator provided a speed signal. The microturbine was a separately excited dc machine with a special thyatron regulator.

1.2 Analogue Modelling of Synchronous Machines

Synchronous machines may also be modelled by analogue, either direct, operational, or hybrid. A direct analogue simulates voltage by
voltage, current by current, and impedance by impedance. The network analyzer, a direct analogue, has been used extensively in power system studies. The general analogue computer is an example of an operational analogue in which variables are represented by dc voltages and where operational amplifiers, multipliers, function generators, and resolvers carry out mathematical operations. A hybrid analogue is a simulation by both direct analogue and by a general analogue computer. An interface is needed for the interconnection.

The paper by James (4) in 1953 is an example of hybrid modelling. The investigation was carried out to optimize the transient response of a generator-regulator system by varying generator open circuit time constant, excitation shunt field time constant and feedback transformer time constants. An actual carbon pile voltage regulator was used in the system. The remaining portions of the system, generator, exciter and feedback circuits were analogued on a computer.

Van Ness (5) in 1954 built two synchronous machine operational analogues for use with a network analyzer to carry out swing curve studies. His second model allowed such factors as voltage regulator action to be taken into account whereas his first model was based on the conventional voltage-behind-reactance representation. The power angle δ was computed continuously instead of in a conventional step-by-step manner.

Riaz (6) in 1956 analogued a synchronous machine entirely on a general analogue computer. Park's equations were applied. While the armature resistance and flux variations were neglected, the effects of saturation and speed variation were included. The purpose of this paper was to construct an accurate synchronous machine model for use in the transient study of voltage regulation systems.
Aldred and Shackshaft's paper in 1958 concerned the prediction and improvement of steady-state and transient-stability of a voltage regulated synchronous generator. A general analogue computer was used to carry out the study. The mechanical power input was assumed constant and the synchronous machine was assumed ideal. The generated voltage due to the change of speed, the induced voltage due to the rate of change of armature flux linkages, and the armature and line resistances were neglected. The effects of the gains and time constants of the voltage regulating systems on stability were examined.

A hybrid power system simulator was described in a paper by Corless and Aldred in 1958. The synchronous machine was simulated on a general analogue computer while the transmission networks were simulated by a high-frequency analogue. Special coupling units were designed to transform the dc computer voltages into their ac equivalents and vice versa. Sustained faults, clearing and reclosing, and pole slipping were investigated.

A later paper by Aldred in 1962 transformed the transmission line equations into d-q coordinates thus eliminating the need for an ac analogue and its interface. A two-machine stability problem was investigated.

The two papers by Peterson et al in 1966 modelled synchronous machines together with dc links with and without parallel ac lines. All system equations were transformed into d-q coordinates and simulated on a general analogue computer. Two-machine problems were investigated.

1.3 A New Model for Large Power System Tests

In all the papers surveyed, it is apparent that there has always
been a compromise made between detail and the cost of modelling. An exceedingly complex model would be required if most details of the machine and power system were included.

The general analogue computer has been used extensively in machine and power system studies because of its versatility. Parameters may be varied rapidly and independently over a wide range. Complex mathematical models including nonlinear differential equations can be readily solved. As the entire system is centralized, instrumentation is simplified. Accidental faults are not injurious to the system as the elements are current limited.

Micromachine and micronetwork models however, are a closer representation of the real system than are the direct and operational analogues. The micromachine has many salient features of a real machine. It operates in real time and therefore is amenable to on-line instrumentation and control. It has a three-phase output and thus does not require an interface with the remainder of the system. Instrumentation and control methods developed for such a model can be readily extended to a real system.

One of the major projects of the power groups is the development of a new test model for a large regulated power system for stability and control studies. This thesis is particularly concerned with the design and construction of a solid state voltage regulator-exciter. Present power system transient stability studies have been confined to analysis and computation. Results obtained from these studies have seldom been tested on real power systems because of the possibility of damage to the system. It is hoped that such tests can be readily carried out on the proposed test model, which would permit the development of more sophisticated power system control.
2. PROPOSED SOLID STATE REGULATION AND EXCITATION SYSTEM

A block diagram of the system is shown in Figure 2.1. The system consists of eight sections: a transfer function simulator, a selector switch, a field amplifier, a negative resistor, a voltage transducer, a power transducer, a speed transducer, and a braking resistor.

The regulation and excitation system for the machine being modelled is simulated in a transfer function section by a special purpose analogue computer. Included in the computer are twelve operational amplifiers, two reference power supplies, two multipliers, three precision potentiometers, and three standard potentiometers. The computer is constructed so that the input and feedback elements of the amplifiers are readily accessible allowing considerable freedom in transfer function design.

The signal from the transfer function section is passed to a selector switch constructed of integrated circuits. The switch selects either a linear or a forced excitation mode. The forced excitation mode is further divided into two sub-modes; one outputting a positive ceiling voltage and the other a negative ceiling voltage. The signal from the selector switch is then passed to a field amplifier.

The field amplifier consists of two sections, a voltage amplifier and a current amplifier. The voltage amplifier multiplies a ±10 volt signal from the field selector switch by a factor of 10. The current amplifier increases the signal current by a factor of 1500.

A negative resistor is placed in series with the field amplifier for the purpose of adjusting the apparent resistance of the synchronous machine field circuit permitting the field circuit time constant to
FIGURE 2.1 Solid State Regulation and Excitation System
be varied. The negative resistor designed is linear, stable and continuously variable from -45 to -75 ohms at 0 to 1.2 amperes.

The three phase power output of the machine is monitored by a power transducer. The transducer generates a signal proportional to the instantaneous real power output from the synchronous machine regardless of phase balance, magnitude, or waveform of the voltages and currents. This signal, combined with a reference signal, yields a power error signal.

Following the power transducer is a voltage transducer. The armature voltage is scaled, rectified, filtered, and referenced to produce a voltage error signal.

A three-phase resistor is connected to the armature of the synchronous machine through a silicon-controlled rectifier (SCR) switch. The period of time during which the braking resistor is connected to the synchronous machine is controlled by a logic unit and is variable from .2 to 4 seconds.

A tachometer mounted on the shaft of the synchronous machine set produces a dc voltage proportional to the speed of the machine. This speed voltage, combined with a reference voltage in a summing amplifier, yields a speed error signal.

The three error signals: power, voltage and speed, are weighted and summed to produce a composite error signal. The composite signal is introduced into the transfer function section thus completing the control loop.
3. SYSTEM REALIZATION

3.1 Transfer Function Simulator

Each power system to be modelled has associated with it a characteristic transfer function describing the voltage regulator and exciter of each synchronous machine, Figure 3.1. From the literature, the general form of the transfer function of the regulator is

$$G(s) = \frac{K(1+T_A s)(1+T_B s)}{(1+T_1 s)(1+T_2 s)(1+T_3 s)} \quad (3.1)$$

and that of the exciter

$$F(s) = \frac{1}{(1+T_{ex} s)} \quad (3.2)$$

Two procedures that yield an analogue computer representation of the above are described as follows. The first, which may be called the division method, obtains the output signal $V_0(s)$ as a function of $V_i(s)$ and $V_0(s)$ by a division process. Let $H(s)$, the transfer function to be simulated, be of the form

$$H(s) = G(s) F(s) = \frac{V_0(s)}{V_i(s)} = \frac{N(s)}{D(s)} \quad (3.3)$$

where $N(s)$ and $D(s)$ are polynomials of $n$th and $i$th order respectively. For all transfer functions studied in this thesis, $n<i$. From (3.3), one has

$$D(s) V_0(s) = N(s) V_i(s)$$
FIGURE 3.1 Voltage Regulator and Exciter Transfer Function
or

\[ V_0(s) = \left( \frac{b_n}{a_1}s^n - i + \frac{b_{n-1}}{a_1}s^{n-1} - i + \ldots + \frac{b_0}{a_1}s^{-i} \right)V_1(s) - \left( \frac{a_{i-1} - 1}{a_i} \right) \]

\[ + \ldots + \frac{a_0}{a_i}s^{-i}V_0(s) \]  

(3.4)

The output \( V_0(s) \) can be constructed according to the right hand side of (3.4) resulting in Figure 3.2. Inverting amplifiers are employed to prevent positive feedback.

The second procedure, which will be called the lead-lag method, represents the transfer function by lead and lag networks as in Figures 3.3, 3.4, and 3.5.

The division procedure is employed where the time constants in the denominator of the transfer function are in excess of .1 second or where the amplifier impedance elements are not accessible. If the numerator is of greater degree than the denominator, i.e. \( n > i \), differentiator circuits result. While a reasonable approximation to a differentiator can be constructed, its frequency response is limited and access to the amplifier impedance elements is necessary. Although the division method permits rapid adjustment of parameters, large amounts of analogue computation equipment are required. For short time constants in the denominator, long time constants in the numerator, or both, large loop gains can result with attendant instability and noise problems.

The lead-lag method is employed where the transfer function contains time constants of .1 second or less in the denominator. The method can be employed only where the impedance elements of the operational amplifiers are accessible. The procedure generally results in
FIGURE 3.2 Division Method of Transfer Function Realization on Analogue Computer
FIGURE 3.3 Lead Network Realization
\[
\frac{V_o(s)}{V_i(s)} = - \frac{R_1}{R_2} \frac{1}{1 + R_1 C s}
\]
FIGURE 3.5 Lead-Lag Network Realization

\[
\frac{V_o(s)}{V_i(s)} = \frac{R_1}{R_2} \frac{1 + R_2C_2s}{1 + R_1C_1s}
\]
a saving of computational equipment but is less flexible. There are practical bounds on the size of resistor and capacitor used in the lead-lag elements. A lower bound is set on resistors by the amount of current that the operational amplifiers can deliver. The current drain through the feedback path and that into the load must not exceed the current rating of the amplifier at rated output voltage. For the 10 volt operational amplifiers, resistors larger than 5KΩ are chosen while for the 100 volt operational amplifiers, the minimum resistance is about 15KΩ. It is found that 100KΩ and 10MΩ are upper bounds for the 10 volt and 100 volt amplifiers respectively. Above these values, excessive amounts of noise, predominantly 60 Hz, is picked up by the amplifiers. The largest non-polarized capacitors suitable are limited to about 1µF. The largest time constant is thus 1 second for the 10 volt amplifier and 10 seconds for the 100 volt amplifier.

From the Portage Mountain\(^{(12)}\) and Kincaid\(^{(13)}\) machines, two examples of transfer function simulation are presented. The Portage Mountain transfer function; Figure 3.6 is simulated entirely by the lead-lag method. The Kincaid transfer function; Figure 3.7, is simulated by a combination of both methods.

3.2 Selector Switch

It is required that the field voltage be controlled either by a linear signal from the transfer function synthesizer or by a positive or negative ceiling voltage signal for forced excitation. The field voltage under forced excitation is quickly raised above its normal excitation level, undergoes several polarity reversals and then is returned to linear control. This procedure improves stability under certain fault conditions.
\[ H(s) = \frac{K}{(1+0.05s)(1+0.003s)} \]

FIGURE 3.6 Portage Mountain Transfer Function Realization
FIGURE 3.7 Kincaid Transfer Function Realization

\[
H(s) = \frac{770(1+s)}{(1+70s)(1+.001s)(1+.002s)(1+.14s)}
\]
Control of the field selector switch is achieved through integrated circuits including two binaries, an inverter and a dual NOR; Figure 3.8(a). For the purpose of this thesis, control signals are generated from two pushbutton switches. Pushbutton switch A selects either the linear or forced excitation mode while pushbutton switch B determines the polarity of the forced excitation.

With reference to Figure 3.8(a), momentarily closing switch Sw A starts Monostable A causing the voltage at pin 6 to rise. During the period, (about .8 seconds) that the monostable is in the elevated or 1 state, it is insensitive to further signals from Sw A. As the monostable returns to its ground or 0 state, Binary A changes state.

Consider first the operation of the linear gate formed by Q4. For a 1 level at pin 7 of Binary A, a 0 level results at pin 7 of the inverter. This 0 level passes to transistor Q2 turning it off. Q3 is turned off by Q2, dropping the gate G of Q4 to -12V. When G is more negative than S by 8V, Q4 is cut off and the linear signal is prevented from passing. Conversely, if the output of Binary A at pin 7 had been a 0, Q4 would be turned on and the linear signal would pass into the summing amplifier following the selector switch.

Consider now the selection of either of the forced excitation signals. The linear gate must be off. Thus, pin 7 of Binary A is at 1. The state of Binary B determines which forced excitation signal, positive or negative, will pass to the summing amplifier. Binary B is controlled by Monostable B in a like manner as described previously for Binary A. The output from pins 5 and 7 of Binary B passes to pins 2 and 5 respectively of a Dual NOR gate. The response or truth table of a single 2 input NOR
FIGURE 3.8(a) Selector Switch Circuit
The Dual NOR circuit is shown in Figure 3.8(b). Inputs at pins 1 and 3 are at 0 because of the state of Binary A. A 0 from pin 5 of Binary B to pin 2 of the Dual NOR will cause a 1 to be output to $Q_5$ turning $Q_5$ off. The top of the 5KΩ potentiometer connected to $Q_5$ will drop to -12V thus giving a negative forced excitation voltage from the tap. At the same time, pin 7 of Binary B outputs a 1. From the truth table, it is seen that a 1 at either input of the NOR will cause a 0 at the output. This 0 becomes a 1 in the Inverter and turns on $Q_1$, preventing a positive forced excitation voltage. A similar procedure may be followed in obtaining a positive forced excitation voltage for the other state of Binary B. Diodes $D_1$ and $D_2$ compensate for the saturation voltage of $Q_1$ and $Q_2$ respectively.

Four lamps indicate the states of Binaries A and B. A 1 level into the lamp driver turns $Q_6$ and $Q_7$ on, lighting the lamp, while a 0 level input turns off $Q_6$ and $Q_7$. The lamp driver inputs are connected to pins 5 and 7 of the binaries.

Finally, the three signals; linear, positive forced excitation, and negative forced excitation are passed to the Field Amplifier through a summing amplifier, Figure 3.8(c).
FIGURE 3.8(b) Dual NOR Circuit
FIGURE 3.8(c): Selector Switch Summing Amplifier
3.3 Field Amplifier

The field of the synchronous generator of the power system model is rated at 1.2A, 115V dc. The maximum output of the transfer function analogue computer is ±10 volts at 5mA. An amplifier is therefore required to match the maximum output of the computer to the maximum power absorbed by the field.

The field amplifier, Figure 3.9, consists of two parts, a voltage amplifier and a current amplifier. The ±10 volt signal from the computer is multiplied by a factor of 10 after passing through a ±100 volt operational amplifier. The resulting signal is then passed to a current amplifier, a complementary common-collector pair in the Darlington configuration. The current amplifier is composed of two symmetrical circuits, one side amplifies positive signals, the other, negative signals. With the exception of polarity, the operation of each side is identical. It has been shown\(^{(14)}\) that the current gain for a single transistor common-collector amplifier is

\[
A_i = \beta + 1
\]

where \(\beta\) is the common-emitter short circuit current gain. For a Darlington connected transistor pair, the current gain is

\[
A_i = (\beta+1)(\beta+1)
\]

The current amplifier has essentially unity voltage gain. There is, however, a small voltage drop \(V_{BE}\) across the two base-emitter junctions which is independent of the input voltage.
FIGURE 3.9 Field Amplifier Circuit
This creates a dead zone at zero of about 2.0 volts. Normal operation however will be on the positive half of the current amplifier only. For negative forced excitation, the signal passes through the dead zone within a few microseconds. Because of the leakage $I_{CB0}$ of the PNP transistors, this dead zone is considerably altered. It was found that for 0 volts into the voltage amplifier, the emitter of the current amplifier is at -0.4 volts because of the net leakage through the load. In addition, there was a noticeable discontinuity in the output signal as the input signal passed through 0 volts. A voltage divider is not employed because of stability and second breakdown. Instead, a 1.5 volt battery is inserted into the base circuit of the NPN transistors turning them partially on. The transistors are on sufficiently to absorb almost all the leakage current of the PNP transistors reducing the zero error from -0.4 volts to -0.1 volt. In addition, the zero crossing discontinuity is removed by this procedure.

Considerable difficulty was encountered in ensuring the stability of the field amplifier. The input impedance of the common-collector configuration can have a negative real part for certain ranges of capacitive loadings above a certain frequency. Stray capacity has been reduced as much as possible; the frequency response is reduced, and the real part of the input impedance is increased by placing a resistor between the voltage and current amplifiers. The voltage drop across this resistor is about 1.5 volts under a maximum load of 1.2A.

The largest apparent resistance seen by the field amplifier is 30 ohms. For the maximum load current of 1.2A, the maximum output
voltage required is 36 volts. The maximum output of the amplifier designed is 38 volts at 1.2A.

3.4 Negative Resistor

It is found that the time constant, $\frac{L_f}{R_f}$, of the field circuit of the synchronous generator is too low. The time constant of large generator exciters ranges from 3 to 7 seconds. The time constant can be increased by either increasing the inductances or by decreasing the resistance. Physical change of inductance or resistance of a winding requires redesign of the machine which is expensive and is very inflexible. Apparent change needs the addition of external devices in the field circuit that alter the field characteristics. The apparent resistance of the synchronous field winding is reduced by inserting a current dependent voltage source (negative resistor) between the field and its driving voltage, Figure 3.10. The realization is given by

\[ v + v_i = (R_f + sL_f)i, \quad v_i = ki \]

or

\[ v = (R_f - k)i + sL_f i \]

The apparent driving point resistance becomes $R_f - k$. By adjusting $k$, the field time constant can be varied over the required range. A simplified scheme to realize the negative resistor is shown in Figure 3.11. The current dependent voltage is given by

\[ v_i = k_1 k_2 k_3 R_s i = ki \]

The field current $i$ develops a voltage across resistor $R_s$. This current dependent voltage is amplified in operational amplifier $A_1$. The signal is
FIGURE 3.10  Synchronous Machine Field Winding Circuit

\[ v_f = ki \]
FIGURE 3.11 Simplified Scheme to Realize Negative Resistor
then passed to a power amplifier made up of voltage amplifier $A_2$ and
current amplifier $A_3$. The power amplifier is similar to the field
amplifier described in Section 3.3.

The actual negative resistor circuit is shown in Figure 3.12. The power amplifier, $A_2$ and $A_3$, while similar to that of the field amplifier, has one important difference: The maximum $V_{ce}$ that may appear across any current amplifier transistor is 180V. PNP transistors having a collector-emitter breakdown voltage base open circuited, $BV_{CEO}$, greater than 160V and capable of passing the required load current, are difficult to obtain. Accordingly, an NPN transistor, $Q_5$, with a $BV_{CEO}$ of 300V is placed in the return of the PNP power supply. Whenever the input to the current amplifier swings more than 10V positive, $Q_5$ turns off disconnecting $Q_3$ and $Q_4$ from the power supply.

A number of difficulties were encountered with the negative resistor current amplifier. The problems of oscillation and dead zone were dealt with in a similar manner as that described in 3.3 for the field amplifier. In addition, the problem of secondary breakdown was encountered. Secondary breakdown is a destructive phenomenon not yet well understood. Beyond a certain critical dissipation, the collector and emitter will short circuit. This critical dissipation can be as small as 10% of the rated dissipation of the transistor depending on $V_{ce}$. The problem is overcome by reducing the supply voltage from ±125 to ±90V.

Drift in $k$ due to thermal variation of parameters is not appreciable. Large heat sinks and a fan prevent the temperature of the current amplifier transistors from increasing greatly. When $k = R_f$, the amplifier is sufficiently stable after about 10 minutes of operation.
FIGURE 3.12 Negative Resistor Circuit
The negative resistor is designed to operate between -45 and -75 ohms at a maximum current of 1.2A.

3.5 Voltage Transducer

For a closed loop voltage control, it is necessary to generate an error signal proportional to the difference between the terminal voltage of the synchronous machine and a reference voltage. The terminal voltage is first converted to dc, compared with a reference, and then passed through an active filter to remove the ac component. The voltage transducer block diagram is shown in Figure 3.13, and its rectifier circuit in Figure 3.14.

3.6 Power Transducer

For control studies, it is required that an analogue voltage be generated proportional to the instantaneous real power flow from the synchronous machine. It is desirable that the device has a fast response and that the signal produced be compatible with the ±10 volt level of the operational amplifiers. The two wattmeter method is employed using two quarter-square multipliers as the main elements, Figure 3.15.

The power of a three-phase three-wire system is given by

\[ P = i_a v_{ac} + i_b v_{bc} \]

The four signals are obtained as in Figure 3.16. The average real power flow from the synchronous machine is obtained by passing the instantaneous power signal through a low pass filter. An active filter \(^{(15)}\) is used because of its flexibility and sharp cutoff. The details of construction are shown in Figure 3.17.

3.7 Speed Transducer

Various excitation feedback signals have been considered \(^{(15)}\)
FIGURE 3.13: Voltage Transducer Block Diagram
FIGURE 3.14 Voltage Transducer Rectifier
FIGURE 3.15 Power Transducer Block Diagram

Multiplier A

Multiplier B

Summing Amplifier

Filter

$P_{\text{inst.}}$

$P_{\text{av}}$

$i_a$

$v_{ac}$

$i_b$

$v_{bc}$
FIGURE 3.16 Generation of Power Transducer Voltage and Current Signals

\[ V_A = \frac{K_a}{a} \]

\[ V_B = \frac{K_b}{b} \]

\[ V_C = K_c \cdot V_{ac} \]

\[ V_d = K_d \cdot V_{bc} \]
FIGURE 3.17: Power Transducer Active Filter

\[ R = 80\,k \]
\[ C = 0.1\,\text{ufd} \]
\[ f = 20\,\text{Hz} \]
to improve stability of the power system. They are speed error, acceleration, rate of change of terminal voltage, armature current, field current, and direct and quadrature subtransient field currents. Digital and analogue computer studies revealed that the speed error signal was the most effective for improving system stability.

A dc tachometer is attached to the synchronous machine shaft through a 1:5 gear train. The output of the tachometer at a shaft speed of 1800 rpm is 155 volts. A negative reference voltage of 155 volts is added to the tachometer signal in a summing amplifier and the resulting error signal passes through an active filter. The filter is similar to that described in the power signal section. Protection of the input circuit of the summing amplifier is provided by diodes D₁ and D₂ as in Figure 3.18.

3.8Braking Resistor

Studies(12) have indicated that a resistive load connected directly to the terminals of a synchronous machine after a transmission line is disconnected immediately after a fault improves the transient stability of the system.

A block diagram of the Braking Resistor is shown in Figure 3.19 and details appear in Figure 3.20. In Figure 3.20(a), the first monostable is used to eliminate false triggering in the pushbutton by introducing a time delay of about .8 seconds. The contact period of the second monostable is variable and can be set from .2 to 4 seconds so that the unijunction oscillator, and in turn, the switch are turned on for the period set.

Four silicon controlled rectifiers, SCR's, arranged as in Figure 3.20(b) are selected as switching devices. Each will carry up to
FIGURE 3.18 Speed Transducer Circuit
FIGURE 3.19  Braking Resistor Block Diagram
FIGURE 3.20 Braking Resistor Circuit
(a) Control Circuit
(b) SCR Switch
three times the rated current of the synchronous machine. After started, the SCR's are continually fired by a 10KHz pulse train from the unijunction oscillator. The starting delay is about $10^{-4}$ sec. It is felt that the complexity and expense involved in having all SCR's commutate or turn off at the same time is not warranted. The maximum time that only one SCR is conducting is half a cycle.

3.9 Power Supplies

Ten power supplies are required for the operation of the system. Power Supply 1, which supplies ±15V to most of the low voltage operational amplifiers, is required to have low ripple and drift. A commercial supply is selected for this purpose. In order to reduce costs, the remaining supplies are designed and constructed. Power Supply 2 provides ±15V to the low voltage operational amplifier of the negative resistor. Two separate amplifier supplies, 2 and 9, are necessary as there is a difference in ground potential between the main computer and the negative resistor. Power Supply 3, rated at ±12V, operates the selector switch and the braking resistor. Power Supply 4 provides a ±12V reference. Supplies 5 and 6 provide plus and minus 40V respectively for the output stage of the field amplifier while supplies 7 and 8 respectively supply plus and minus 90V for the negative resistor output stage. Power Supply 9 supplies ±120V to the high voltage operational amplifier of the negative resistor. Power Supply 10 supplies ±120V for the high voltage operational amplifier of the field amplifier. Figure 3.21 is the circuit of supplies 2, 3, 4, 9 and 10, Figure 3.22 that of supplies 5 and 6, and Figure 3.23 that of 7 and 8.

Consider first, Figure 3.21. The transformer supplying the bridge rectifier is centre tapped so as to provide two voltages opposite
FIGURE 3.21  Circuit for Power Supplies 2, 3, 4, 9 and 10
FIGURE 3.22  Circuit for Power Supplies 5 and 6
FIGURE 3.23 Circuit for Power Supplies 7 and 8
in polarity. As both the positive and negative circuits of the supply are similar, with the exception of polarity, only the positive portion will be examined in detail. Resistor $R_c$ and $R_b$ and capacitor $C$ form the first filtering stage. Filter capacitor $C$ is made as large as possible in order to achieve low ripple without an expensive regulator. Capacitors of this size require a series surge resistor $R_c$ of a few ohms to prevent the diode bridge and the capacitor from being damaged when the supply is first energized. No such surge resistor is required for the large filter capacitor on the output of the supply since the current overload protection circuit, $R_1$ and $D$, acts to limit the charging current. A bleeder resistor $R_b$ is placed across the filter capacitors in order to minimize the danger of shock after the supply has been turned off. As large electrolytic capacitors can have an appreciable impedance at radio frequencies, capacitor $C_f$ is added to bypass to ground any high frequency signals from the load. Capacitor $C_L$ was placed across the primary of the transformer to prevent high frequency signals from entering the supply from the line.

The regulator of the power supply in Figure 3.21 consists of a transistor in series with the rectifier and the load. The voltage across the transistor is automatically adjusted so as to keep the voltage across the load constant. The output voltage is determined by an internal voltage reference circuit formed by $R_z$ and zener diode $D_z$. The voltage across the zener diode is essentially independent of the supply voltage. This reference voltage is applied to the base of the regulating transistor. The transistor acts to keep the emitter at almost the same voltage as that of the base. Thus, since the base is connected to a constant reference voltage, the emitter is maintained at almost the same voltage.

Resistor $R_1$ and diode $D$ of the same figure form a current
overload protection circuit to prevent the regulator transistor from being destroyed under fault conditions. As the current through $R_1$ approaches the limit of the supply, the diode $D$ gradually becomes forward biased connecting the base of the regulating transistor to the load end of $R_1$. An increasing amount of available base current is diverted through $D$ causing the transistor to cut off. In the extreme case where the output voltage is zero, the transistor is almost completely cut off. All supplies of this type will sustain continuous short circuits without damage. Knowing the forward transfer characteristics of diode $D$, the base-emitter voltage of the regulating transistor and the maximum current, the resistor $R_1$ can be estimated.

Supplies 5 and 6, Figure 3.22, are similar in operation to those of Figure 3.21. Because of increased power requirements, it is necessary to supply the symmetrical positive and negative field amplifier voltages from two supplies. The first filtering section is made up of a large capacitor $C$ and a bleeder resistor $R_b$. A surge resistor is not required as the overload current is limited by the resistance in the transformer. The regulator is identical in operation with that of Figure 3.21 except that two transistors in the Darlington configuration are used for increased current gain. The current limiting circuit and the filter of the first stage are identical in operation to those in Figure 3.21.

The output power required of supplies 7 and 8, Figure 3.23, is large. Readily available transformers will just meet the required voltage output. The introduction of a voltage regulator to those power supplies would drop the available output voltage excessively. Thus, supplies 7 and 8 are unregulated. The slight increase in ripple is not critical as the supply feeds a common-collector amplifier. Capacitor
surge resistors are not needed as current-limiting takes place in the transformer.

3.10 System Arrangement

The completed system built is shown in Figure 3.24 (front view) and Figure 3.25 (rear view). Rack A contains computation, control and most power supplies. Rack B contains the voltage and power transducers, braking resistor, and field and negative resistance current amplifiers. Room is left for future expansion.
FIGURE 3.24 Completed System, Front View
4. TEST RESULTS

4.1 Section Tests

Transfer Function Simulator

Two transfer functions are set on the simulator and their step responses compared with digital computation results. The step responses of the Portage Mountain and Kincaid transfer functions are shown in Figure 4.1(a) and (b) respectively. Analytic digital computer solutions for the same transfer functions are shown in Figures 4.2 and 4.3 respectively.

Selector Switch

Figure 4.4 displays the operation of the selector switch. The sine wave represents a linear signal from the transfer function simulation section and the step voltages, the forced excitation. Switching between modes and between levels takes place in less than $10^{-6}$ seconds which is negligible compared with the operating time of the system to be studied.

Field Amplifier

The step response of the field amplifier is shown in Figure 4.5. The rise time of approximately $10^{-5}$ seconds justifies the assumption that the time delay in the amplifier is negligible compared with the time constants being simulated.

Negative Resistor

The variation in field time constant with increasing negative resistance is shown in Figure 4.6. A step voltage is applied through the field amplifier. The final current is maintained constant. The 63% level is marked by a second trace. The rise time of the power
FIGURE 4.1(a) Simulator Step Response of Portage Mountain Transfer Function
FIGURE 4.1(b) Simulator Step Response of Kincaid Transfer Function
FIGURE 4.2 Digital Computer Solution of Step Response of Portage Mountain Transfer Function
FIGURE 4.3 Digital Computer Solution of Step Response of Kincaid Transfer Function
FIGURE 4.4 Selector Switch Operation

Selector Voltage Output (5V/div)

Time (ms/div)
FIGURE 4.5 Field Amplifier Rise Time
FIGURE 4.6 Variation in Field Time Constant with Negative Resistance
FIGURE 4.6 Variation in Field Time Constant with Negative Resistance

(c) Time (1s/div) $\tau = 3s$

(d) Time (1s/div) $\tau = 7s$
amplifier section of the negative resistor is shown in Figure 4.7. Figure 4.8 represents the field current variations under forced excitation. The top trace is proportional to the applied voltage while the lower trace is proportional to the field current. The very slight discontinuity in the upper portion of the lower trace is caused by the negative power supply switching out of the circuit.

**Voltage Transducer**

The voltage transducer generates a dc signal proportional to the terminal voltage of the synchronous generator. Figure 4.9 is a calibration curve obtained with an AVO meter relating the terminal voltage to the output of the voltage transducer. The quality of transformer results in a slight nonlinearity at voltage extremes.

**Power Transducer**

The power transducer generates a dc signal proportional to the power output of the synchronous generator. A calibration curve relating the power output of the generator to the output of the power transducer is shown in Figure 4.10. Wattmeters are used to calibrate the transducer. The rise time of the transducer is shown in Figure 4.11.

**Speed Transducer**

Figure 4.12 is the generator speed error signal and represents a 4% decrease resulting from a load current increase from 1 to 3 amperes.

**Braking Resistor**

Figure 4.13(a) displays the closing of the SCR switch. The three traces are proportional to the phase currents in the load resistor. Figure 4.13(b) indicates the maximum and minimum con-
FIGURE 4.7 Negative Resistor Power Amplifier Rise Time
FIGURE 4.8 Negative Resistor Under Forced Excitation
FIGURE 4.9 Voltage-Transducer Calibration Curve

AVO METER VOLTAGE READING (VOLTS AC)

VOLTAGE TRANSUCER OUTPUT (VOLTS DC)
FIGURE 4.11 Power Transducer Rise Time
FIGURE 4.12 Generator Speed Error
FIGURE 4.13(a) SCR Switch Operation
FIGURE 4.13(b) Braking Resistor Operation Time
duction period of the SCR switch. The trace is the output of the power transducer.

4.2 System Tests

In order to test the solid state voltage regulator-exciter built, not just as separate components, but also as an integrated part of the complete power system test model, the voltage regulator-exciter is connected to a real machine. However, because the remaining parts of the system, the prime mover, the speed governor, the generator, and the transducer line have not yet been completed, the experiments are constrained to a constant speed drive without using the selector switch, power transducer, or braking resistor.

Two tests are carried out. The first is an open loop test with a step load application without a voltage error feedback. The voltage error signal versus time is observed. The second is a closed loop test with a voltage error feedback control. The voltage regulator-exciter transfer function for the Kincaid machines is simulated. For both tests the field voltage is adjusted so that the terminal voltage is 150V line-to-line with an initial load of 1A. The synchronous generator is driven by a synchronous motor. The lower trace of Figure 4.14 displays the transient speed change after the load application. The maximum speed change is about .66%. The upper trace of the same figure is the voltage error signal.

Figure 4.15(a), (b), and (c) are voltage errors with load application without voltage control; time constants are .014, 1, and 3 seconds respectively. A constant voltage is fed directly to the field amplifier and adjusted so that the terminal voltage is 150V line-to-line for an initial load of 1A. The spike in the traces is caused by the
FIGURE 4.14 Speed Variation and Voltage Error for a Load Application
FIGURE 4.15 Voltage-Error with Load Application without Voltage Control
Synchronous Machine Terminal Voltage Error (.05V/div)

Time (1s/div) \( t = 3.05 \)
rapid change of current due to switching.

Figure 4.16(a) through (f) are voltage errors with load application with voltage control. Figure 4.16(a) and (b) have a field time constant of 1 second, (c) and (d) a time constant of 3 seconds, and (e) and (f) a time constant of 5 seconds. Further, Figure 4.16(a), (c), and (e) have a gain of 20% of that shown in Figure 3.7 while Figure 4.16(b), (d), and (f) have a gain of 40%. The spike due to switching is again observed. Note the decrease in error voltage with increased gain for a constant field time constant. The error voltage is also reduced as the negative resistance is increased since a smaller field voltage is required in order to pass the same amount of field current.

The load application is the same for both tests, from 1A to 3A or from 18% to 55% of the full load current of the generator. A load application to 100% was not possible because synchronous speed could not be maintained without speed control.
FIGURE 4.16 Voltage-Error with Load Application with Voltage Control
FIGURE 4.16 Voltage-Error with Load Application with Voltage Control
FIGURE 4.16 Voltage-Error with Load Application with Voltage Control
5. CONCLUSION

A solid state voltage regulator-exciter for a new test model of a power system has been designed, constructed, and tested. This device is capable of simulating the response characteristics of almost any known excitation system. Provision has been made for forced excitation in both positive and negative directions. A variable-negative resistor has also been designed and constructed to allow the micro-generator field time constant to be varied. A power transducer was developed which outputs a signal voltage proportional to the instantaneous and average real power flow from the synchronous generator. A high-speed solid state switch for a braking resistor has been designed and built for load-rejection resistance-braking studies. Component tests indicate that all the subsystems meet design requirements. Tests of the voltage regulator-exciter system indicate that the various subsystems work in concert.

Future development may include a more accurate, noise-free speed measuring system. Literature suggests that the system should be capable of resolving a speed change of .01%. As most analogue systems are capable of resolving changes of .1% at best, a digital system is therefore required. The very slight nonlinearities and phase shifts in the power transducer transformers could be corrected by better quality transformers and suitable phase shifting networks. Finally, the push-button control for the selector switch and the braking resistor will be replaced with a computer directed control circuit.
REFERENCES


