MASTER

**SECURITY ASSESSMENT
OF POWER SYSTEMS
INCLUDING ENERGY STORAGE**

**Progress Report
for Period July 1, 1977 - October 1, 1977**

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October 1977

Prepared For

**THE U.S. ENERGY RESEARCH AND DEVELOPMENT ADMINISTRATION
UNDER CONTRACT NO. EC-77-S-02-4206.A000**

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CONTRACT ACTIVITY

The principal investigators have devoted 25% of their workload to this contract since July, as scheduled. This is expected to continue through the remainder of the year.

TABLE OF CONTENTS

	Page
ABSTRACT	iv
SECTION 1 - SIMPLIFIED MODELS OF FORCE-COMMUTATED CONVERTERS . . .	1
1.1 Current-Fed Converter for Superconducting Magnets	1
1.2 Voltage-Fed Converter for Storage Batteries	11
SECTION 2 - POWER SYSTEM IDENTIFICATION.	13
2.1 Direct Computation of Measurement Spectra	13
2.2 Time Series Analysis of Simulated Signals	14
2.3 Work Planned.	17
2.4 References.	17

ABSTRACT

Energy Storage

Work has continued on the development of simplified models for the power converters used to interface superconducting magnets and storage batteries with an ac system. During the present phase of the project the emphasis has been on the application of force-commutated converter (FCC) circuits. Simplified analytical models, suitable for power system control studies, have been postulated for current-fed and voltage-fed converters to be used respectively in inductor-converter (IC) units and battery-converter (BC) units. Detailed computer simulations have also been developed for these converters. During the next phase of the project, the simplified FCC models will be verified using a detailed computer simulation of a power system with the FCC units incorporated.

Identification

Numerical ill-conditioning has been discovered in the direct calculation of measurement spectral characteristics. A change in system modeling has been implemented to eliminate the ill-conditioning. An analog simulation of a one machine power system with load variations has been established. Efforts are presently being directed at establishing agreement between the analog simulation and direct calculations.

SECTION 1

SIMPLIFIED MODELS OF FORCE-COMMUTATED CONVERTERS

The previous progress report C00-4206-1 was concerned with the modeling and simulation of superconducting magnets and storage battery units connected to a power system through line-commutated converter (LCC) interfaces. In that report simplified analytical models were described for inductor-converter (IC) and battery-converter (BC) units, and the accuracy of these models was verified using a detailed computer simulation of the power system.

In the present report an extension of this work is described for the application of force-commutated converter (FCC) interfaces. Simplified analytical models have been postulated for selected FCC circuits to be used in IC and BC units. Detailed computer simulations have also been developed for these converters. During the next phase of the project, the simplified FCC models will be verified using a detailed computer simulation of a power system with the FCC units incorporated.

1.1 Current-Fed Converter for Superconducting Magnets

A number of current-fed FCC circuits are suitable for operation in an IC unit. The basic capacitor-commutated converter shown in Fig. 1.1 was selected for purposes of this investigation. A steady-state analysis of this converter circuit was performed to establish a set of exact equations for the assumed conducting sequence. For convenience, the analysis is given for the upper half bridge only. The lower half bridge analysis is identical. It is assumed that the trapped charge on each capacitor is sufficient to hold off the diode in each leg until after conduction of the corresponding thyristor. As indicated in Fig. 1.2, the ac system characteristics are modeled by a

Fig. 1.1 Current-Fed Converter Bridge

Fig. 1.2 Upper Half Bridge Model

balanced voltage-behind-reactance equivalent, and the superconducting magnet is represented by a constant direct current source.

Suppose D_5 and T_5 are conducting, then

$$i_5 = I_d$$

$$V_{pn} = V_5 = e_c$$

The three capacitors are charged to some unknown voltages

$$V_{15} = V_1 - V_5 = V_o > 0$$

$$V_{31} = V_3 - V_1 = V_x$$

$$V_{53} = V_5 - V_3 = -V_o - V_x$$

At $\omega t = 0$, which is considered to be α degrees after $e_a - e_c$ crosses zero going positive, the thyristor T_1 receives a control pulse. Because of the positive voltage of the equivalent capacitor between T_1 and T_5 , the direct current is immediately transferred from T_5 to T_1 , and T_5 is shut off. The equivalent circuit shown in Fig. 1.3 is applicable for $0 < \omega t \leq \theta$.

Fig. 1.3 Capacitor Discharge Circuit

In terms of the time reference, the ac voltages may be written

$$e_a = E \sin(\omega t + 30 + \alpha)$$

$$e_b = E \sin(\omega t - 90 + \alpha)$$

$$e_c = E \sin(\omega t + 150 + \alpha)$$

From Fig. 1.3, the capacitor voltage is described by

$$\frac{d}{dt} v_{15} = - \frac{2}{3C} I_d$$

$$v_{15} = V_o - \frac{2}{3} X_c I_d \omega t$$

$$V_1 = e_c + V_o - \frac{2}{3} X_c I_d \omega t$$

This condition prevails until $\omega t = \theta$, when $V_1 = e_a$ and diode D1 starts to conduct.

$$E \sin(\theta + 30^\circ + \alpha) = E \sin(\theta + 150^\circ + \alpha) + V_o - \frac{2}{3} X_c I_d \theta$$

which gives

$$\sqrt{3} E \sin(\theta + \alpha) = V_o - \frac{2}{3} X_c I_d \theta \quad (1.1)$$

The equivalent circuit shown in Fig. 1.4 is applicable for $\theta \leq \omega t \leq \theta + u$, after commutation begins from D5 to D1.

Fig. 1.4 Diode Commutation Circuit

The diode currents are related by

$$e_a - e_c = 2L \frac{di_1}{dt} + \frac{2}{3C} \int_{\theta/\omega}^t (-i_5) dt$$

and since $i_1 + i_5 = I_d$, the equation may be written

$$\omega^2 \left[\frac{X_c}{3X_L} I_d + \frac{\sqrt{3}E}{2X_L} \cos(\omega t + \alpha) \right] = \frac{d^2 i_1}{dt^2} + \frac{X_c}{3X_L} \omega^2 i_1$$

where

$$i_1(\theta) = 0$$

$$\left. \frac{di_1}{dt} \right|_{\theta} = 0$$

The above differential equation has the solution

$$i_1 = A \cos \sqrt{\frac{X_c}{3X_L}} (\omega t - \theta) + B \sin \sqrt{\frac{X_c}{3X_L}} (\omega t - \theta) + I_d + \frac{3\sqrt{3}E}{2(X_c - 3X_L)} \cos(\omega t + \alpha)$$

where

$$A = -I_d - \frac{3\sqrt{3}E}{2(X_c - 3X_L)} \cos(\theta + \alpha) \quad (1.2)$$

$$B = 9 \sqrt{\frac{X_L}{X_c}} \frac{E}{2(X_c - 3X_L)} \sin(\theta + \alpha) \quad (1.3)$$

At $\omega t = \theta + u$, when the commutation from D5 to D1 is complete, the steady state operation of the system requires that the following equations be satisfied.

$$v_{15}(\theta + u) = v_{53}(0) = -V_o - V_x$$

$$v_{31}(\theta + u) = v_{15}(0) = V_o$$

$$v_{53}(\theta + u) = v_{31}(0) = V_x$$

In addition

$$v_{15}(\theta + u) - v_{15}(0) = \frac{-2}{3C} \int_0^{\theta+u} i_5 dt$$

$$v_{31}(\theta + u) - v_{31}(0) = \frac{1}{3C} \int_0^{\theta+u} \frac{1}{\omega} i_5 dt$$

$$v_{53}(\theta + u) - v_{53}(0) = \frac{1}{3C} \int_0^{\theta+u} \frac{1}{\omega} i_5 dt$$

The above equations give

$$V_o + \frac{1}{2} V_x = V_o - V_x = V_o + 2V_x$$

which implies

$$V_x = 0$$

Therefore

$$v_{15}(\theta + u) = [e_a - e_c - 2L \frac{di_1}{dt}]_{\theta+u} = -V_o$$

which gives

$$\begin{aligned} -V_o &= \sqrt{3}E \sin(\alpha + \theta + u) + \frac{2A}{\sqrt{3}} \sqrt{X_\ell X_c} \sin \sqrt{\frac{X_c}{3X_\ell}} u \\ &\quad - \frac{2B}{\sqrt{3}} \sqrt{X_\ell X_c} \cos \sqrt{\frac{X_c}{3X_\ell}} u - \frac{3\sqrt{3} X_\ell E}{X_c - 3X_\ell} \sin u \end{aligned} \quad (1.4)$$

Also at $\omega t = \theta + u$, it is clear that

$$i_1(\theta + u) = I_d$$

which gives

$$A \cos \sqrt{\frac{X_c}{3X_\ell}} u + B \sin \sqrt{\frac{X_c}{3X_\ell}} u + \frac{3\sqrt{3}E}{2(X_c - 3X_\ell)} \cos(\alpha + \theta + u) = 0 \quad (1.5)$$

The nonlinear equations (1.1) - (1.5) may be solved digitally for the variables V_o , θ , and u , which describe the steady-state operation of the converter bridge.

The validity of these equations has been verified by comparison with results

taken from a detailed analog simulation of the converter.

A simplified steady-state analysis similar to the line commutated converter is also possible, using various approximations. For example, if the commutation period is neglected, the average direct voltage of the FCC bridge is the same as that for the LCC bridge, that is

$$V_d = \frac{3\sqrt{3}}{\pi} E \cos \alpha \quad (1.6)$$

If the diode commutation angle u is considered only, i.e. neglecting the capacitor discharge angle θ , another expression for average direct voltage may be found by superimposing the effect on Equation (1.6). From Fig. 1.4, it is apparent that the average voltage drop due to the diode commutation is given by

$$\Delta V_{du} = \ell \frac{di_1}{dt} \Big|_{\text{average}} = \frac{3}{\pi} \int_{\theta}^{\theta+u} \ell \frac{di_1}{dt} d(\omega t)$$

and since $i_1(\theta) = 0$ and $i_1(\theta + u) = I_d$

$$\Delta V_{du} = \frac{3X_\ell}{\pi} I_d$$

Therefore

$$V_d = \frac{3\sqrt{3}}{\pi} E \cos \alpha - \frac{3X_\ell}{\pi} I_d \quad (1.7)$$

which is recognized as one form of the expression that describes the average voltage of a LCC bridge. Equation (1.7) was used in a simplified model of a force-commutated IC unit connected to a one-machine/infinite bus power system. Computer simulations of the power system with both the simplified converter model and a detailed converter representation were compared. Under transient fault conditions, where the IC unit power was modulated to provide

system damping, it was observed that for lagging converter power factors the results using the simplified model agreed with those taken from the detailed simulation. However, for leading converter power factors, the simplified model introduced significantly less damping than was observed from the detailed simulation. It was, therefore, concluded that Equation (1.7) is not an adequate representation of the FCC bridge for control studies.

The inadequacy of Equation (1.7) demonstrates the importance of representing the capacitor discharge angle θ in the average direct voltage of the bridge. This effect may be superimposed on Equation (1.7) to produce a more accurate model. From Fig. 1.3, it is apparent that the average voltage rise due to the capacitor discharge is given by

$$\Delta V_{d\theta} = v_{15} \Big|_{\text{average}} = \frac{3}{\pi} \int_0^\theta \left(V_o - \frac{2}{3} X_c I_d \omega t \right) d(\omega t)$$

which becomes

$$\Delta V_{d\theta} = \frac{\theta}{\pi} (3V_o - X_c I_d \theta) \quad (1.8)$$

Equation (1.8) may be added to Equation (1.7) to give an exact value for the average direct voltage. Unfortunately, this requires the solution of Equations (1.1) - (1.5) for V_o and θ , which precludes the possibility of a useful simplified model. However, it has been determined that in order for the FCC bridge to function properly for all leading and lagging power factors, the capacitor size is such that $X_c \gg X_\ell$. This implies that the resonant frequency of the commutating circuit is always greater than the 60 Hz system frequency.

This observation is in agreement with most applications of the current-fed FCC inverter in electric drive systems, where a practical range of operation is

$0.1 < \frac{\omega}{\omega_{\ell c}} < 0.3$. Therefore, a good approximation for purposes of developing

a simplified model is the assumption that $\omega_{lc} \gg \omega$. Under this condition the expression for the diode commutation current is greatly simplified, for example, for commutation from D5 to D1

$$i_5 \approx I_d \cos \left(\omega_{lc} t - \frac{\omega_{lc}}{\omega} \theta \right)$$

and from previous analysis

$$v_{15}(\theta + u) - v_{15}(\theta) = - \frac{2}{3C} \int_{\theta/\omega}^{\theta+u/\omega} i_5 dt$$

or

$$-2V_o + \frac{2}{3} X_c I_d \theta = - \frac{2}{3} \frac{\omega}{\omega_{lc}} X_c I_d \sin \left(\frac{\omega_{lc}}{\omega} u \right)$$

Furthermore, since $u \approx \frac{\omega}{\omega_{lc}} (\pi/2)$, the above expression may be written

$$V_o \approx \frac{1}{3} X_c I_d \theta + \frac{1}{3} \frac{\omega}{\omega_{lc}} X_c I_d \quad (1.9)$$

Equation (1.9) may be used with Equation (1.1) to eliminate the variable V_o , which gives the approximate relationship of θ vs. α

$$X_c I_d \theta + 3\sqrt{3} E \sin(\theta + \alpha) - \frac{\omega}{\omega_{lc}} X_c I_d = 0 \quad (1.10)$$

Equation (1.9) may also be substituted in Equation (1.8) to give

$$\Delta V_{d\theta} \approx \frac{1}{\pi} \frac{\omega}{\omega_{lc}} X_c I_d \theta \quad (1.11)$$

which may be used with Equation (1.10) to form an alternate expression

$$\Delta V_{d\theta} \approx \frac{3}{\pi} X_c I_d - \frac{9}{\pi} \sqrt{\frac{X_l}{X_c}} E \sin(\theta + \alpha) \quad (1.12)$$

If θ is assumed to be small, Equation (1.12) may be added to Equation (1.7) to form a new more accurate approximation for the average direct voltage,

that is,

$$V_d = \frac{3\sqrt{3}}{\pi} E \cos \alpha - \frac{9}{\pi} \sqrt{\frac{X_l}{X_c}} E \sin \alpha \quad (1.13)$$

The simplified expression given in Equation (1.13) was evaluated over the full range of bridge operation, and the results were compared with the exact average voltage function. Figures (1.5) and (1.6) show this comparison for values of $X_c = 10$ pu and $X_c = 100$ pu. The other parameters used were $X_l = 0.1$ pu, $E = 1$ pu and $I_d = 1$ pu.

Fig. 1.5 Average Direct Voltage vs. Control Angle for $X_c = 10$ pu

Fig. 1.6 Average Direct Voltage vs. Control Angle for $X_c = 100$ pu

1.2 Voltage-Fed Converter for Storage Batteries

For purposes of this project, it was assumed that impulse-commutated, voltage-fed FCC bridges would be used to interface storage batteries with a power system. Therefore, the dynamic characteristics of a voltage-fed BC unit are quite different than a current-fed IC unit. From a modeling point of view, the predominant difference is that the voltage commutation time in a voltage-fed FCC is negligible in comparison to the current commutation time for a current-fed FCC. The effect of the voltage-fed commutation on the ac

phase currents may be viewed as external to the converter circuit itself. Therefore, a basic simplified representation for a voltage-fed FCC is a sinusoidal ac voltage source, which is phase-controlled by the dc power regulator. The ac phase currents are modeled externally as a function of the network representation and the converter voltage. As a result, the voltage-fed FCC model does not require the development of a direct voltage equation that internally accounts for the effects of current commutation. A simplified model based on this approach will be verified during the next phase of the project.

SECTION 2

POWER SYSTEM IDENTIFICATION

This section of the progress report concerns identification of power system electromechanical models from system measurements, with only normal load variation present as a disturbing input.

2.1 Direct Computation of Measurement Spectra

It was previously reported that the capacity had been developed to calculate directly the spectral composition of measured signals. More has been determined about this calculation.

Direct computation of measurement spectra involves writing system equations, linearizing these about an operating point, and then determining from the linearized equations the multiple input, multiple output transfer function from the disturbing inputs to the outputs that are to be measured. Spectral density of the measured signals is easily determined from the transfer function and the assumed spectral characteristics of the disturbing input. Calculation of the matrix transfer function requires a transformation of the linearized equations to a canonical form (phase variables). It has been found that the calculation of the required transformation matrix is an extremely ill-conditioned problem if the system eigenvalues range widely in magnitude. For example, in the previous report a one machine, tenth order model of a power system was discussed. This model had eigenvalues that ranged over about four orders of magnitude at any given operating point. Speaking without rigor in regard to eigensystems, the smallest eigenvalues correspond to response of the machine field circuit and the largest eigenvalues correspond to electrical response of

the stator and network. Dynamics of electromechanical oscillation and damper action correspond to eigenvalues of intermediate magnitude. The coefficients in the characteristic polynomial thus ranged over about twenty-eight orders of magnitude. As characteristic polynomial coefficients are used in a recursive calculation of the columns of transformation matrix [1], this matrix is quite ill-conditioned. The ill-conditioning manifests itself as an inability to calculate the lower frequency zeroes of the matrix transfer function with any confidence. This was discovered when it was observed that the calculated values of the lower frequency zeroes were extremely sensitive to small variations of system parameters.

The cure for such difficulties appears to be a reformulation of the system equations, which is to "neglect electrical transients". This lowers the order of the system and has the practical effect of removing the large magnitude eigenvalues from the linearized system. It has been found that the transfer function of such a reduced order system can be calculated with confidence, as the sensitivity of zeroes to small parameter variations is reasonable. The remaining question, then, is the validity of "neglecting electrical transients" in the system equations. Only recently has an analytical basis been given to the two time scale nature of power system electromechanical motion [2]. Of importance in this matter is the damping of rotor oscillations.

2.2 Time Series Analysis of Simulated Signals

This research project involves the use of relatively detailed simulations of a power system to generate time series signals that should be representative of measurements that could be taken on an actual system. The power system simulations contemplated are to include network and stator transients, due to the above mentioned uncertainty about possible model order reductions.

So far, only a simulation of a one machine power system has been implemented. This simulation size will not be increased until methods for signal spectral analysis are implemented which are adequate. The problems with obtaining signal spectra are described below.

An analog simulation of a one machine system has been implemented which includes a load. Real and reactive power in the load vary in the manner of Brownian Motion. With analog to digital converters, power flowing in a transmission line can be sampled at regular intervals and a time series of measurements thus developed. Figure 2-1 shows an estimated autocorrelation of a signal from the simulation. This autocorrelation estimate definitely displays the dominant rotor oscillations at about 10 radians per second. However, since only 500 samples are involved in making this autocorrelation, little other information is available from it. To improve the estimate without storing a large number of samples, a recursive algorithm is required.

It is both desirable and necessary to obtain evidence of agreement between the analog simulations and direct computation of signal spectra. Such agreement is required if confidence is to be established in any reduced order system equivalents. While calculation of autocorrelation estimates provides good qualitative information about the system, this type of calculation is perhaps not the most direct way to obtain quantitative information. In order to more directly compare the sampled time series with calculated spectra, a Fast Fourier Transform algorithm has been employed but results so far are not satisfactory. If the FFT algorithm cannot be made to perform adequately in this application, then a least squares parameter estimation algorithm will be implemented. This method also is capable of providing the confirmation.

Fig. 2-1 Estimated Auto Correlation

2.3 Work Planned

The immediate task is to confirm that the analog simulation and the direct computations are in agreement regarding spectral properties of measurements. When this is confirmed, implementation of parameter estimation algorithms can proceed. Initially, it appears that the algorithms will be based on minimizing a positive definite quadratic function of model forecast errors.

2.4 References

1. Tuel, "On the Transformation to (Phase-Variable) Canonical Form," IEEE Trans., Vol. AC-11, p. 607, July 1966.
2. Krause, Nozari, Skvarenina, Olive, "The Theory of Neglecting Stator Transients," paper submitted to IEEE for publication.