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**EXPERIMENTAL STUDIES AND COMPUTER SIMULATION OF THE CONTROL OF ENERGY
TRANSFER USING INDUCTOR-CONVERTER BRIDGES**

by

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Fusion Power Program

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EXPERIMENTAL STUDIES AND COMPUTER SIMULATION OF THE CONTROL OF ENERGY TRANSFER USING INDUCTOR-CONVERTER BRIDGES

by

Masaru Hirano and Robert L. Easton

ABSTRACT

An inductor-converter bridge (ICB) is a solid state DC-AC-DC power converter system for bidirectional, controllable, energy transfer between two coils. The ICB is suitable for supplying large pulsed power to such magnets as the superconducting equilibrium field coil of the proposed tokamak power reactors from another superconducting energy storage coil.

By using double data-word instructions, the resolution of the microcomputer is increased up to 12 bits. The magnitude of the current oscillation for different values of resolution of the microcomputer are examined. The magnitude of the current oscillation is shown to be determined by the size of the quantization band where all magnitudes falling within an interval are equated to a single bit of the microprocessor.

The control system of simultaneous open-loop with the guided phase difference and closed-loop with a derivative controller is introduced. The performance of the bridge is analyzed using a small signal analysis. The steady-state signals, or fundamental signals, are determined from the open-loop control using phase angles determined from the measured value of the storage current, called the guided phase difference. As a result of small signal analysis, the derivative control action is deduced and the derivative controller is implemented in the closed-loop. The coefficient of the derivative controller is a function of the storage current. According to the experimental results, noise provides this system with frequent voltage and current oscillations.

The control system of simultaneous open-loop with guided phase difference and closed-loop with a bang-bang controller is introduced. The experimental and simulated results shown the quick damping of the output signal. This system has a feature that, if the open-loop control using the guided phase difference keeps the digitized error signal to be zero, the bang-bang control action will not take place.

1. INTRODUCTION

Pulsed power loads of increasing magnitudes up to several hundred megawatts or more must be supplied in the near future. High energy physics research laboratory experiments⁽¹⁾ and the longer range foreseeable needs for successful nuclear fusion reactors⁽²⁻³⁾ are representative examples. If pulsed power is supplied directly from the power grid, power flow between the power grid and the pulsed power load may cause unacceptable disturbance in the power grid. In order to eliminate disturbance, an energy storage unit can be placed as a buffer between the power grid and the pulsed power load and then energy transfer takes action between the storage unit and the pulsed power load. An inductor-converter bridge (ICB) has been proposed for such a purpose.⁽⁴⁻⁸⁾

The ICB is a solid state DC-AC-DC converter system for reversible energy transfer between two coils. In an ICB system, the load constitutes one of the ICB coils, the other being the storage coil. The converter thyristors⁽⁹⁾ are naturally commutated by a set of wye-connected capacitors on the ac lines of the circuit. The circuit is designed so that, in every converter cycle, a very small fraction of the magnet energy is stored in these capacitors.

The main objective is analysis and experiments for control of the three-phase ICB system. Our last experiment which was conducted during the summer in 1982 has been reported in an Argonne National Laboratory report.⁽¹⁰⁾ Since then, the experimental equipment was moved from Argonne National Laboratory to the University of Wisconsin-Madison for future cooperation with the Wisconsin Superconductive Energy Storage Project Group.⁽¹¹⁾ By use of the computer and experimental facilities at the University of Wisconsin, further analysis and experiments are done and their results are reported in this paper.

In Section 1.2, we review the state equation of the ICB for the energy transfer and the open-loop control system with the guided phase difference. The phase difference is defined as the phase by which the load-side switching sequence leads the storage-side. The state equation which consists of the two first-order differential equations describes power flow in the ICB system. The state equation includes terms associated with losses due to the wiring resistance and the thyristor forward voltage. By adjusting parameters associated with losses, a mathematical model of the ICB system is determined and used for simulation purpose.

Based on the state equation, the guide equation was derived and, as a result, the open-loop control system with the guided phase difference were developed. As previously indicated in the guided phase difference system, the phase difference varies according to the magnitude of the storage current.

In Section 2, importance of resolution of the microcomputer is discussed. The control system is structured by the microcomputer (Intel 8080A), its interfaces, and the ICB. A single and double data-words have word lengths of 8 bits and 16 bits, respectively. By using double data-word instructions for microcomputer computation, the available length of the data word is increased from 8 bits to 16 bits. The effective length of the data word is limited to 12 bits by the resolution of the analog-to-digital (A/D) converter. After error calculation is made, 8 bits are extracted from the effective length of the data-word. The resolution of the microcomputer is determined by the location of 8 bits in the effective length of the data word. By using bang-bang feedback control which was introduced by Ehsani⁽¹²⁾, the magnitude of the current oscillation for different values of resolution is measured and is also calculated from the measured data. Then, using the measured and calculated magnitude of the current oscillation, the effect of the resolution of the microcomputer on control performance of the ICB system is examined.

In Section 3, the control system of simultaneous open-loop with guided phase difference and closed-loop with a derivative controller is introduced. Again, the guided phase difference is a technique in which the phase difference is adjusted according to the amplitude of the measured storage current. The fundamental component of the phase difference is determined from open-loop control using the guided phase difference, while the deviation component of the phase difference is determined from closed-loop control using the derivative controller. The two components are added together to structure the phase difference. The derivative controller is deduced by using small signal analysis. By use of the small signal analysis, the deviation component of the phase difference is described as a function of the derivative of the deviation component of the load current. The coefficient of the derivative is a function of the storage current. Then the coefficient of the derivative controller has the magnitude equal to the one derived by small signal analysis but the opposite sign. In order to utilize the derivative controller for microcomputer control, a table of values of the coefficient of the derivative controller

versus the magnitudes of the storage current is stored in the microcomputer memory. The dynamic behavior of this control system is discussed.

In Section 4, the control system of simultaneous open-loop with guided phase difference and closed-loop with a bang-bang controller is introduced. The phase difference is determined by the open-loop control using the guided phase difference. On the other hand, an error signal is used for bang-bang control action. The output of the controller determines the converter frequency selected from three frequencies. The control characteristics of this control system is discussed.

1.1 System Configuration

Figure 1-1 shows the ICB experimental system. The wye-connected capacitor bank links the two six-pulse thyristor bridges, one of which is connected to the storage coil and the other connected to the load coil. The two coils are superconductive with the equal inductance. A current power supply, PS1, and its series-connected thyristor, SS7, are provided for the purpose of initially charging the storage coil. Table 1-1 shows the characteristics of the superconducting coils. The load and storage coils have identical specifications. Table 1-2 shows the characteristics of the thyristor. Table 1-3 shows the system parameters in the experimental system. In order to start the energy transfer, precharging is required. Before the energy transfer, the current power supply PS1 charges the storage coil through the thyristor SS7. In the mean time, the voltage power supplies PS2 and PS3 charge their parallel-connected capacitors C_1 and C_3 with the opposite polarities, as indicated by the signs on each capacitor in Figure 1-1. The magnitude of the initial capacitor voltage, V_o , is

$$V_o = \frac{I_o T}{6C}$$

where I_o is the initial value of the storage current and T is the converter period. When the thyristor firing starts and the energy transfer starts to take action, the current thyristor SS7 is switched off due to the back bias and the power supply PS1 is electrically disconnected from the circuit. During the energy transfer, the resistors R_2 and R_3 limit the current flow at the power supplies PS2 and PS3. At the beginning of the energy transfer, one of

Figure 1-1. Experimental ICB system.

TABLE 1-1. CHARACTERISTICS OF THE SUPERCONDUCTING COIL

Dimensions:	
Outer Radius	13.3 cm
Inner Radius	11.0 cm
Length	40.6 cm
Conductor:	
Inductance	4 H
Critical Current (I_C)	265 A
Peak Magnetic Field at I_C	4.3 T
Composition	NbTi
Stabilizer	Copper
Cooling Type	Pool Cooling

TABLE 1-2. CHARACTERISTICS OF THE THYRISTOR

Type	IR-151RF30
Maximum Repetitive Off-State Voltage (VDRM)	300 V
Maximum Static On-State Current (IT)	235 A
Maximum Hold Current	200 mA
Maximum On-State Voltage	1.9 V
Typical Off-Switching Time	20 μ sec
Rise Rate of Recovery Voltage	200 V/ μ sec

TABLE 1-3. SYSTEM PARAMETERS

<u>Power Supplies</u>		
PS1:	Maximum Voltage	10 V
	Maximum Current	100 A
PS2 and 3:	Maximum Voltage	600 V
<u>Resistors</u>		
	R_S and R_L	250 Ω
	R_2 and R_3	100 k Ω
<u>Capacitors</u>		
	C_1 , C_2 , and C_3	234.5 μF
<u>DC Current Transducer</u>		
	Rated Current	± 250 A dc
	Rated Output (RO)	± 10 V dc
	Response Time to 99%	< 150 μsec
	Linearity	$< \pm 0.5\%$ RO

the coils has no current, and thyristors in the low-current side do not receive enough holding current. In order to keep enough holding current, the power transistor in the low-current side is switched on and the resistor R_{S1} or R_{L1} conducts enough current to hold the thyristors on.

The rate and direction of the energy transfer are controlled by the relative phase between the two converters. The phase difference is defined as the phase by which the load side switching sequence leads the storage-side. The thyristors are fired in the order shown by thyristor numbers in Figure 1-1. At first, the thyristors of numbers 1 and 6 at each converter are switched on at the same time. At this moment, the phase difference is zero degree. Then it is moved from zero degree to the desired via transition of the phase shifting. After transition of the phase shifting or frequency shifting, each capacitor voltage must be kept balanced; each average capacitor voltage must be zero. In order to avoid permanent unbalance of each capacitor voltage, the

four-step switching sequence for the phase shifting and the three-step switching sequence for the frequency shifting are used.⁽¹²⁾ During the transition of the phase shifting, the duration of the switching interval is modulated in the sequence of 60° -half of the desired phase shift, 60° , 60° , and 60° minus half of the desired phase shift. The switching interval is an interval between the two successive switching events. During the transition of the frequency shifting, the duration of the switching interval is modulated in the sequence of the new frequency, the old frequency, and the new frequency.

Figure 1-2 shows a block diagram of the microcomputer and its interfaces. Table 1-4 shows the features of the microcomputer and the A/D converter. The storage current, i_s , and load current, i_L , are detected by the dc current transducers (DCCTs). The reference signal, i_R , is supplied from a function generator. The converter frequency, f , and the phase difference, ϕ , can be adjusted by potentiometers on an external control board. Each analog signal is connected to its assigned input port of the multiplexer and is converted into a digital number by the A/D converter. Values of the converter frequency and the phase difference are shown on external LED displays. The microcomputer accepts an interrupt signal from the pulse sequencer only during the transition of the frequency shifting or phase shifting. Otherwise, an interrupt mask prevents the interrupt signal from interfering with the control command sequence.

Source programs are compiled by an Intel 8080 cross assembler⁽¹⁵⁾ located on a HP-1000 minicomputer. After cross assembly, object programs are punched out in a hexadecimal machine code on paper tapes and later loaded to the microcomputer memory through a tape reader of a teletypewriter (TTY). The monitor program which resides in the EPROM receives instructions from the teletype keyboard for data entry, memory manipulation, or program loading.

Digital simulations have been done by aid of a VAX-11 minicomputer.⁽¹⁶⁾ The state equation of the ICB is solved by using a subroutine from the package library, International Mathematical and Statistical Library (IMSL).⁽¹⁷⁾ The simulated results are plotted by a Varsatec printer with aid of the plotting software, Varsaplot-07.⁽¹⁸⁾

Figure 1-2. Block diagram of the microcomputer and its interface.

TABLE 1-4. FEATURES OF THE MICROCOMPUTER AND THE A/D CONVERTER

Microcomputer

CPU	8080 A
Word Length	8 bit
Clock Rate	2.15M Hz
Memory Size:	
PROM	2K byte
RAM	34K byte

Additional:

- Eight-level prioritized interrupts
- 48 lines of parallel I/O programmable for uni-directional input/output or bidirectional ports

A/D Converter

Resolution	12 bits
Number of Channels	16 ports
Rated Input	± 10 V Bipolar
Type of Output	Two's Complement
Total Conversion Time	33 μ sec
System Accuracy	± 0.025 % FSR

1.2 Mathematical Description of the Inductor-Converter Bridge and the Open-Loop Control System with the Guided Phase Difference

Figure 1-3 shows a block diagram of energy transfer process for the positive phase difference. When the phase difference is positive, energy is transferred from the lossless storage coil to the lossless capacitor bank, and from the lossless capacitor bank to the lossless load coil. The lossless coils and the lossless capacitor bank mean the purely inductive coils and the purely capacitive capacitor bank, respectively. The average power of the lossless capacitor bank is zero; i.e., the average power delivered from the storage converter is equal to the one delivered to the load converter. During the energy transfer, energy is dissipated due to the losses associated with the wiring resistance, the thyristor forward voltage, the thyristor switching circuit, and the capacitor resistance. Since the ICB is symmetrically structured, the losses which exist before and after the lossless capacitor bank result from the same causes mentioned above.

The average power delivered from the load converter is derived in Appendix. By neglecting higher harmonics, the average power, $P(t)$, is

Figure 1-3. Block diagram of energy transfer process.

$$P(t) = \frac{54 t_{SW} i_S(t) i_L(t)}{\pi^3 C} \sin \phi, \quad [1-1]$$

where ϕ is the phase difference and t_{SW} is the switching interval. The wiring resistive loss and the thyristor forward voltage drop are expressed in terms of the coil currents. The thyristor switching loss and the capacitor resistive loss are not mathematically expressed. The power flow for slow changes can be mathematically described as follows:

$$-\frac{d}{dt} \frac{L_S i_S^2(t)}{4} = P(t) + R_S i_S^2(t) + 2V_f i_S(t), \quad [1-2]$$

$$\frac{d}{dt} \frac{L_L i_L^2(t)}{2} = P(t) - R_S i_S^2(t) - 2V_f i_S(t), \quad [1-3]$$

where V_f is the thyristor forward voltage; and R_S and R_L are wiring resistances in the storage side and in the load side, respectively. The negative time derivative of energy in the storage coil is equal to a sum of the average power delivered from the storage converter to the capacitor bank, power due to the wiring resistive loss, and power due to the thyristor forward voltage drop. On the other hand, the time derivative of energy in the load coil is equal to the average power $P(t)$ minus a sum of power due to the wiring resistive loss and power due to the thyristor forward voltage drop. Substituting Eq. [1-1] into $P(t)$ and performing the differentiation in the left-hand side of Eqs. [1-2] and [1-3], we have the state equation that consists of the two first-order differential equations:

$$\frac{di_S(t)}{dt} = -\frac{54 t_{SW} \sin \phi}{\pi^3 L_S C} i_L(t) - \frac{R_S}{L_S} i_S(t) - \frac{2V_f}{L_S}, \quad [1-4]$$

$$\frac{di_L(t)}{dt} = \frac{54 t_{SW} \sin \phi}{\pi^3 L_L C} i_S(t) - \frac{R_L}{L_L} i_L(t) - \frac{2V_f}{L_L}. \quad [1-5]$$

The measured wiring resistance in the entire electrically connected loop and the measured thyristor forward voltage are 0.014 Ω and 1.0 V, respectively. The state equation of the ICB only contains terms associated with the wiring resistive loss and the thyristor forward voltage drop, but not terms associated with the thyristor switching loss and the capacitor resistive loss. By

adjusting loss terms in the state equation and comparing the simulated and experimental dynamic behavior of the open-loop control system with the constant phase difference and the converter frequency, a mathematical model of the ICB is determined. In the mathematical model, the wiring resistance and the thyristor forward voltage are 0.1 Ω and 1.5 V, respectively.

Let us describe the open-loop control system with the guided phase difference. When the desired rate of rise of the load current for the period of ramping up and the desired magnitude of the load current for the period of holding, the phase difference is expressed as a function of the storage current for both periods. The equation which expresses the phase difference is derived from the state equation of the ICB. During the energy transfer, the value of the phase difference is determined from the magnitude of the storage current. In other words, the phase difference is guided by the storage current. In this paper, the system where the phase difference is guided by the storage current is called the open-loop system with the guided phase difference and the equation used for control of this system is called the guide equation.

At first, the guide equation for the period of ramping up is derived. In a lossless energy transfer case, the last two terms in Eq. [1-5] are eliminated. Then Eq. [1-5] is reduced to

$$\frac{di_L(t)}{dt} = \frac{54 \tau_{SW} \sin \phi}{\pi^3 L_L C} i_S(t) \quad [1-6]$$

From the desired rate of the current rise, the constant which is designated by D_L is determined. The time derivative of the load current in Eq. [1-6] is substituted by the constant D_L . After some manipulation, the guide equation for the period of ramping up is derived. The guide equation that expresses the phase difference as a function of the storage current is given as follows:

$$\phi = \sin^{-1} \left[\frac{\pi^3 L_L C}{54 \tau_{SW}} D_L \frac{1}{i_S(t)} \right]. \quad [1-7]$$

The guide equation for the period of holding is also derived by using Eq. [1-5]. When the desired magnitude of the load current for the period of holding is given, the constant which is designated by I_L^0 is determined. By

substituting the constant I_L^0 into the load current and zero into the time derivative of the load current, Eq. [1-5] becomes

$$0 = \frac{54 t_{SW} \sin \phi}{\pi^3 L_L C} i_S(t) - \frac{R_L}{L_L} I_L^0 - \frac{2V_f}{L_L}, \quad [1-8]$$

After some manipulation, the guide equation for the period of holding is given as follows:

$$\phi = \sin^{-1} \left[\frac{\pi^3 L_L C}{54 t_{SW}} \left(\frac{R_L}{L_L} I_L^0 - \frac{2V_f}{L_L} \right) \frac{1}{i_S(t)} \right] \quad [1-9]$$

In order to utilize Eqs. [1-7] and [1-9] for microcomputer control, a table of values of the phase difference versus the magnitude of the storage current for the given constant is precalculated and loaded into the successive microcomputer memories. Calculations are made for each least significant bit change in current. During the energy transfer, the storage current is detected by the DGCT and converted into a digital number by the A/D converter. Then the microcomputer searches the memory location corresponding to the digital number of the storage current and determines the value of the phase difference by using the contents of the memory.

Figure 1-4 shows the dynamic behavior of the experimental open-loop control system with the guided phase difference. The operation modes of the load current are characterized by a period of ramping up, a period of holding, and a period of ramping down. The guide equation for the period of ramping down is the same as the one for the period of ramping up. The constant D_L for the period of ramping up and the one for the period of ramping down are 20 A/sec and -20 A/sec, respectively. They represent the desired rates of the current change, too. The constant I_L^0 is 60 A. The load current of with a rate of 20 A/sec, it is held at about 60 A, and it falls at a rate of -23 A/sec. The duration of the period of holding is regulated by a combination of program instructions and a signal from a function generator. The duration is about 3 sec. The desired waveform of the load current is approximately obtained by control using the guided phase difference. During the period of ramping down, the difference between the desired rate and the actual rate is relatively large. The loss terms which are neglected in the guide equation of Eq. [1-7] may affect the rate of the current change when the storage current is decreased.

Figure 1-4. Dynamic behavior of the experimental open-loop system with the guided phase difference.

2. IMPORTANCE OF RESOLUTION OF THE MICROCOMPUTER

By using double data-word instructions, the length of the data word is increased to 16 bits. Some of double data-word instructions are available in a set of original instructions and the others are made in the form of a subroutine. But the effective length, limited by the resolution of the analog-to-digital (A/D) converter, is 12 bits. In order to avoid increasing the number of program instructions, computation using double-word instructions is done until error calculation is made. After that point, the length of the data word is reduced to 8 bits and computation using single data-word instructions is carried on. The 8 bits are extracted from the effective length of the data word. The location of the 8-bit portion in the effective length of the data word depends on the magnitude of an error and determines the resolution of the microcomputer.

Figure 2-1 shows a flow chart of the program algorithm for error calculation using double data-word instructions. The reference signal and the feedback signal are converted into 12-bit two's complementary numbers by the A/D converter. Then each 12-bit number is stored in two successive memories, where the upper 4 bits of the 12-bit data are located at the lower 4 bits in the first memory and the remaining bits in the second memory. The upper 4 bits in the first memory are set to the same value as the most significant bit of the data. Then each 16-bit number is shifted to left by four. Each 16-bit number is now structured by the 12-bit data at the upper 12 bits and 4 zeros at the lower 4 bits. The subtraction of the 16-bit number for the feedback signal from the 16-bit number for the reference signal yields the 16-bit number for the error signal. The 16-bit number of the error signal can be further shifted to left by a number of between 0 and 4. Then 8 bits in the first of the two memories are extracted. The resolution of the microcomputer is determined by the shift number. The relation between the microcomputer resolution, q , and the shift number, p , is given as follows: $q=8+p$. Further computation is carried on using single data-word instructions and the 8-bit error signal. Even after switching to single data-word arithmetic, a value of current corresponding to each bit is not changed, so is the resolution of the microcomputer.

Figure 2-1. Flow chart of program algorithm for error calculation using double data-word instructions.

The error calculation using double data-word instructions is programmed and tested in the bang-bang feedback control system of the ICB. In this control strategy,⁽¹²⁾ the operating phase difference is switched from one extreme, 90° (or -90°), to the other, -90° (or 90°), according to the sign of the error quantity at any given time. The bang-bang control strategy renders no decision when the value of the digitized error is zero.

Figures 2-2, 2-3, and 2-4 show the dynamic behaviors of the bang-bang feedback control system, where the resolution of the microcomputer is 8, 10, and 12 bits, respectively. The reference signal rises at a rate of 20 A/sec until it reaches current of 60 A. Then it is held at 60 A. The triangular signal, supplied from a function generator, is used as the reference signal for the period of ramping up. After the load current reaches 60 A, the initially given constant is used as the reference signal for the period of holding. When the phase difference is 90° , energy is transferred from the storage coil to the load coil, the storage voltage is at the positive extreme, and the load voltage is at the negative extreme. When the phase difference is -90° , the direction of the energy transfer is reversed, so is the sign of the both coil voltages. By the bang-bang control, the load current moves around the reference signal within the range which depends on the resolution of the microcomputer.

At first, we describe the general waveform of the both coil currents, ignoring the current oscillation. The storage coil is initially charged with 100 A. During the period of ramping up, the load current rises at a rate of 20 A/sec, while the storage current is slowly decreased. During the period of holding, the load current varies around a current of 60 A within the range depending on the resolution of the microcomputer, while the storage current droops in order to keep the average load current constant and to supply energy dissipated due to losses.

We measure the magnitude of the current oscillation during the period of holding. Figures 2-5, 2-6, and 2-7 show magnified views of waveform of the load current during the period of holding. The resolutions of the microcomputer are 8, 10, and 12 bits in Figs. 2-5, 2-6, 2-7, respectively. By use of a digital scope manufactured by Nicolet Inc., the waveform of the load current is magnified around the reference signal which is not shown here. The time and voltage scales are shown at the lower left and right of each figure, respectively. The voltage of 1V corresponds to a value of 25 A. Noise structures a band around the real signal. The peak-to-peak magnitude of the band is about 10 mV or current of 0.25 A. In order to find causes of noise, we placed the DCCT outside the conducting wire during the energy transfer. We found noise of the same waveform and magnitude as ones on the real signal. In addition, we found that noise coincided in time with thyristor switchings.

Figure 2-2. Dynamic behavior of the bang-bang feedback control system, where the resolution of the microcomputer is 8 bits.

Figure 2-3. Dynamic behavior of the bang-bang feedback control system, where the resolution of the microcomputer is 10 bits.

(a) During the period of ramping up.

Figure 2-4. Dynamic behavior of the bang-bang feedback control system, where the resolution of the microcomputer is 12 bits.

(b) During the period of holding.

Figure 2-4. Dynamic behavior of the bang-bang control system, where the resolution of the microcomputer is 12 bits.

Figure 2-5. Magnified view of waveform of the load current. The resolution is 8 bits.

Figure 2-6. Magnified view of waveform of the load current. The resolution is 10 bits.

Figure 2-7. Magnified view of waveform of the load current. The resolution is 12 bits.

From these observations, we conclude that noise results from thyristor switchings, but that they do not exist on the real signal. We have placed a 1 kHz low-pass filter between the DCCT and the A/D converter for reducing noise. The magnitude of the current oscillation is measured by taking the difference between a middle point of the band at the upper peak and a middle point of the band at the lower peak. The size of the quantization band and the magnitude of the current oscillation for the resolutions of 8, 10, and 12 bits are tabulated in Table 2-1. The difference between the measured magnitude of the current oscillation and the size of the quantization is within 0.1 A.

TABLE 2-1. SIZE OF THE QUANTIZATION AND THE MAGNITUDE OF THE CURRENT OSCILLATION FOR THE RESOLUTIONS OF 8, 10, AND 12 BITS

Resolution	Size of the Quantization Band	Magnitude of the Current Oscillation
8 bits	1.97 (A)	2.07 (A)
10	0.49	0.488
12	0.123	0.175

When the resolution of the microcomputer is increased, the effect of noise on the current oscillation cannot be ignored. As a result of noise, the period of the voltage oscillation which is synchronized in time with the current oscillation is not kept constant (shown in Fig. 2-4b). Noise causes the band around the real signal and, especially for the higher resolution, we have difficulty in measuring the magnitude of the current oscillation. In addition to measuring the magnitude of the current oscillation, we calculate it by using an average value of the measured period of the voltage oscillation. Calculation using the average value gives us the average magnitude of the current oscillation. The equation for this calculation is derived from the state equation for the lossless energy transfer.

The state equation without terms associated with losses are

$$\frac{di_S(t)}{dt} = -a i_L(t), \quad [2-1]$$

$$\frac{di_L(t)}{dt} = a i_S(t), \quad [2-2]$$

where

$$a = \frac{54 t_{SW}}{\pi^3 LC}.$$

The Laplace transformation of Eqs. [2-1] and [2-2] gives us

$$sI_S - i_S(0) = -aI_L,$$

$$sI_L - i_L(0) = aI_S,$$

or

$$aI_L + sI_S = i_S(0), \quad [2-3]$$

$$sI_L - aI_S = i_L(0). \quad [2-4]$$

Adding Eq. [2-3] multiplied by a to Eq. [2-4] multiplied by s yields

$$(s^2 + a^2) I_L = a i_S(0) + s i_L(0),$$

or

$$I_L = \frac{a}{s^2 + a^2} i_S(0) + \frac{s}{s^2 + a^2} i_L(0). \quad [2-5]$$

The inverse Laplace transformation of Eq. [2-5] gives us

$$i_L(t) = i_S(0) \sin at + i_L(0) \cos at. \quad [2-6]$$

The magnitude of the current oscillation, Δi_L , is

$$\Delta i_L = i_L(t_1) - i_L(0) \quad [2-7]$$

where t_1 is the half period of the voltage oscillation.

The magnitude of the coil currents and the half period of the voltage oscillation are measured 4, 6, and 8 sec after the energy transfer starts. The half period of the voltage oscillation is an average of the half period over more than ten cycles. For the higher resolution, the number of cycles is increased. By substituting the measured magnitude of the coil currents and the measured half period of the voltage oscillation into Eq. [2-6], the magnitude of the load current at the end of the half period of the voltage oscillation is calculated. By using Eq. [2-7], the magnitude of the current oscillation is calculated. The size of the quantization band, the measured half period of the voltage oscillation, the measured magnitude of the storage current, the measured magnitude of the load current, and the calculated magnitude of the current oscillation for the resolutions of 8, 10, and 12 bits are tabulated in Table 2-2.

TABLE 2-2. CALCULATED MAGNITUDES OF THE CURRENT OSCILLATION
FOR THE RESOLUTIONS OF 8, 10, AND 12 BITS

Time After the Energy Transfer Starts	Measured Half Period	Measured i_S	Measured i_L	Calculated Δi_L
(a) 8-bit resolution (where the size of the quantization band is 1.98 A)				
4 (sec)	80 (msec)	70 (A)	57.5 (A)	1.97 (A)
6	90	63.75	57.5	2.016
8	95	57.5	57.5	1.914
(b) 10-bit resolution (where the size of the quantization band is 0.49 A)				
4 (sec)	22 (msec)	70 (A)	57.5 (A)	0.547 (A)
6	27	62.75	57.5	0.601
8	30	57.5	57.5	0.612
(c) 12-bit resolution (where the size of the quantization band is 0.123 A)				
4 (sec)	7.5 (msec)	67.5 (A)	57.5 (A)	0.180 (A)
6	8.47	61.875	57.5	0.187
8	10.42	55	57.5	0.204

The calculated magnitudes of the current oscillation for the 10-bit and 12-bit resolutions are larger than their corresponding size of the quantization band. The measured half period includes a period for control computation and the following phase shifting. During a period of the phase shifting, the load current is assumed to be kept constant. We calculated the period for control computation and the following phase shifting, subtract this period from the measured half period, and recalculated the magnitude of the current oscillation.

The transient period consists of the period for microcomputer computation and the two converter periods for the phase shifting. The computation time is calculated by summing the number of instructions times each instruction execution time which equals a sum of time periods multiplied by the microcomputer clock time. The clock time in our system is approximately 465 n sec. The number of instructions varies, depending on the number of instructions used in branches following computation for comparison and whether a reference signal ramps up or it is kept constant. The computation time with and without the phase shifting required and during the period of ramping up and during the period of holding is tabulated in Table 2-3.

TABLE 2-3. COMPUTATION TIME WITH AND WITHOUT PHASE SHIFT REQUIRED AND DURING THE PERIOD OF RAMPING UP AND DURING THE PERIOD OF HOLDING

Phase Shifting	Period of Ramping Up	Period of Holding
Required	1.126 (msec)	1.203 (msec)
Not Required	0.60	0.75

Following microcomputer computation, the phase shifting occurs if it is required. The phase shifting requires two converter cycles. Therefore, when the phase shifting is required, the duration of the period for control computation and the following phase shifting are 3.43 msec for the period of ramping up and 3.51 msec for the period of holding. Here the converter frequency is 868 Hz. By using the measured magnitude of the coil currents from Table 2-2 and the measured half period excluding the period for control computation

and the following phase shifting, the magnitude of the current oscillation is recalculated and tabulated in Table 2-4.

TABLE 2-4. CALCULATED MAGNITUDES OF THE CURRENT OSCILLATION FOR THE RESOLUTIONS OF 8, 10, AND 12 BITS. THE MEASURED MAGNITUDE OF THE COIL CURRENTS ARE GIVEN FROM TABLE 2-1.

Time	Half Period Excluding a Period for Control Computation and Following Phase Shifting Transient	Calculated Δi_L
(a) 8-bit resolution (where the size of the quantization band is 1.98 A)		
4 (sec)	76.5 (msec)	1.887 (A)
6	86.5	1.938
8	91.5	1.845
(b) 10-bit resolution (where the size of the quantization band is 0.49 A)		
4 (sec)	18.5 (msec)	0.460 (A)
6	23.5	0.524
8	26.5	0.541
(c) 12-bit resolution (where the size of the quantization band is 0.123 A)		
4 (sec)	3.99 (msec)	0.096 (A)
6	4.96	0.109
8	6.91	0.135

Since the magnitude of the current change during the transient period is ignored, the magnitude of the current oscillation becomes relatively lower than the size of the quantization band. The difference between the magnitude of the current oscillation and the size of the quantization band is within a range of 0.1 A. A value of 0.1 A is within an acceptable measurement error. From measurement and calculation of the magnitude of the current oscillation, we can conclude that the magnitude of the current oscillation is determined by the size of the quantization band.

By using the computation time without phase shifting required during the period of holding, we calculate the magnitude of the current change during one

computation period where signals cannot be controlled. The computation time is 0.75 msec. When the load current is 57.5 A, the maximum magnitude of the current change during one computation period of 0.75 msec is about 0.025 A. The maximum magnitude is calculated by using the storage current of 100 A. On the other hand, the size of the quantization band is 0.123 A for the resolution of 12 bits. Then, even for the resolution of 12 bits, it takes the load current 4 to 5 computation times to pass through one of the quantization band. Therefore, even for the resolution of 12 bits, the computation time does not limit control performance in our system.

3. THE CONTROL SYSTEM OF SIMULTANEOUS OPEN-LOOP WITH THE GUIDED PHASE DIFFERENCE AND CLOSED-LOOP WITH THE DERIVATIVE CONTROLLER

Since open-loop poles of the control system are located near an imaginary axis on the s -plane, the proportional-plus-integral (PI) control action does not provide the control system with the quick damping. On the other hand, the derivative (D) control action may improve the transient response of the control system to the change of the reference signal. In practice, the derivative controller may amplify the noise signal and cause saturation effect in the actuator. But, since the two coils are quite inductive, the coil currents hold low-magnitude noises generated mainly due to the thyristor switchings. The derivative control action is deduced from by small signal analysis.

Simultaneous open-loop control with the guided phase difference and closed-loop control with the derivative controller is based on setting the fundamental component of phase difference determined by the magnitude of storage current and adding or subtracting an increment in phase to the fundamental determined by the derivative control. The error signal that is the difference between the reference signal and the feedback signal is used for the derivative control action. The time constant of the derivative controller is expressed as a function of the storage current.

According to the small signal analysis, the variables are represented by the fundamental and deviation components, the representatives are substituted into their corresponding variables, the state equation is linearized, and then the relation among the deviation components or small signals is approximately obtained. The load current, $i_L(t)$, and the phase difference, ϕ , are given as follows: $i_L(t) = i_R(t) + \Delta i_L(t)$ and $\phi = \phi_0 + \Delta\phi$. The fundamental component of the phase difference is expressed as follows:

$$\phi_0 = \sin^{-1} \left[\frac{\pi^3 CL}{54 t_{SW}} D_L \frac{1}{i_S(t)} \right] \quad [3-1]$$

for a period of ramping up and

$$\phi_0 = \sin^{-1} \left[\frac{\pi^3 CL}{54 t_{SW}} \left(\frac{R_L}{L} I_L^0 + \frac{2V_f}{L} \right) \frac{1}{i_S(t)} \right] \quad [3-2]$$

for a period of holding. Eqs. [3-1] and [3-2] are derived from the guide equation.

One of the two first-order differential equations is

$$\frac{di_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \sin \phi - a_2 i_L(t) - a_3, \quad [3-3]$$

where

$$a_2 = \frac{R_L}{L} \text{ and } a_3 = \frac{2V_f}{L}.$$

By substituting $i_R(t) + \Delta i_L(t)$ into $i_L(t)$ and $\phi_0 + \Delta \phi$ into ϕ , Eq. [3-3] becomes

$$\frac{d}{dt} (i_R(t) + \Delta i_L(t)) = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \sin(\phi_0 + \Delta \phi) - a_2 i_L(t) - a_3,$$

or

$$\frac{di_R(t)}{dt} + \frac{d\Delta i_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \sin(\phi_0 + \Delta \phi) - a_2 i_L(t) - a_3 \quad [3-4]$$

The trigonometric expansion is

$$\sin(\phi_0 + \Delta \phi) = \sin \phi_0 \cos \Delta \phi + \cos \phi_0 \sin \Delta \phi.$$

By aid of the trigonometric expansion, Eq. [3-4] becomes

$$\frac{di_R(t)}{dt} + \frac{d\Delta i_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) (\sin \phi_0 \cos \Delta \phi + \cos \phi_0 \sin \Delta \phi) - a_2 i_L(t) - a_3. \quad [3-5]$$

The trigonometric functions are approximated as follows: $\cos \Delta\phi \approx 1$ and $\sin \Delta\phi \approx \Delta\phi$. By use of this approximation, Eq. [3-5] becomes

$$\frac{di_R(t)}{dt} + \frac{d\Delta i_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) (\sin \phi_0 + \cos \phi_0 \Delta\phi) - a_2 i_L(t) - a_3. \quad [3-6]$$

The fundamental component ϕ_0 depends on the operational period. For the period of ramping up, we have ϕ_0 in the form of Eq. [3-1]. Substituting Eq. [3-1] into ϕ_0 , Eq. [3-6] becomes

$$\frac{di_R(t)}{dt} + \frac{d\Delta i_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \left(\frac{\pi^3 CL}{54 t_{SW}} \frac{1}{i_S(t)} D_L + \cos \phi_0 \Delta\phi \right) - a_2 i_L(t) - a_3,$$

or

$$\frac{di_R(t)}{dt} + \frac{d\Delta i_L(t)}{dt} = D_L + \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \cos \phi_0 \Delta\phi - a_2 i_L(t) - a_3. \quad [3-7]$$

During the period of ramping up, the constant D_L can be selected to be equal to the rate of rise of the reference signal. Then Eq. [3-7] becomes

$$\frac{d\Delta i_L(t)}{dt} = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \cos \phi_0 \Delta\phi - a_2 i_L(t) - a_3. \quad [3-8]$$

By adding $a_2 i_L(t) + a_3$ to the both sides of Eq. [3-8] and dividing it by the coefficient of the $\Delta\phi$ term, $\Delta\phi$ is expressed as follows:

$$\Delta\phi = \frac{1}{a_1} \left(\frac{d\Delta i_L(t)}{dt} + a_2 i_L(t) + a_3 \right), \quad [3-9]$$

where

$$a_1 = \frac{54 t_{SW}}{\pi^3 CL} i_S(t) \cos \phi_0.$$

The coefficient a_1 is a function of the storage current.

For the period of holding, we have the component ϕ_0 in the form of Eq. [3-2]. We substitute Eq. [3-2] into ϕ_0 of Eq. [3-6]. After manipulation in the way similar to derivation of Eq. [3-9], the deviation component of the phase difference is expressed as follows:

$$\Delta\phi = \frac{1}{a_1} \left(\frac{d\Delta i_L(t)}{dt} - a_2 i_L(t) \right). \quad [3-10]$$

The term $a_2 i_L(t)$ term in Eqs. [3-9] and [3-10] can be eliminated because the value of the wiring resistance in the term $a_2 i_L(t)$ is very small. The term a_3 can also be eliminated by adjusting the constant D_L in the guide equation. Then, by eliminating these terms, the deviation component of the phase difference is expressed in the similar form for the both periods:

$$\Delta\phi = \frac{1}{a_1} \frac{d\Delta i_L(t)}{dt}, \quad [3-11]$$

Here the coefficient a_1 is different for the period of ramping up and for the period of holding.

In order to cancel the deviation of the phase difference, a controller needs to generate a signal opposite to the one in Eq. [3-11]. By using the error signal $e(t) = i_R(t) - i_L(t)$, we have

$$-\Delta\phi = \frac{1}{a_1} \frac{de(t)}{dt}. \quad [3-12]$$

By using Eq. [3-12], the derivative controller is introduced. From now on, $\Delta\phi$ is used for an output of the controller instead of for a small signal or a deviation component of the phase difference. Therefore the $\Delta\phi$ output of the controller is

$$\Delta\phi = T_D \frac{de(t)}{dt}. \quad [3-13]$$

where $T_D = \frac{1}{a_1}$ and T_D is called the derivative time constant.

Figure 3-1 shows a block diagram of the control system of simultaneous open-loop with the guided phase difference and closed-loop with the derivative

WHERE $\alpha = \frac{\pi^3 CL}{54 i_{S0}^2}$ AND $T_D = \frac{\alpha}{i_S(t) \cos \phi_0}$

Figure 3-1. Block diagram of the control system of open-loop with the guided phase difference and closed-loop with the derivative controller for the period of ramping up.

controller. The fundamental component of the phase difference is determined by open-loop control using the guided phase difference. The error signal is used for the derivative control action and the controller determines its output, according to Eq. [3-13]. The phase difference is determined by a sum of the fundamental component of the phase difference and the output of the controller. For the period of holding, the guide equation is Eq. [3-2] and the coefficient a_1 is changed with the guide equation.

We transform Eq. [3-13] by the Laplace and z transformations. The Laplace transform of Eq. [3-13] is

$$\Delta\phi(s) = T_D sE(s),$$

or

$$E(s) = \frac{1}{T_D} \frac{1}{s} \Delta\phi(s). \quad [3-14]$$

The z transform of Eq. [3-14] is

$$E(z) = \frac{1}{T_D} \frac{1}{1 - z^{-1}} \Delta\phi(z),$$

or

$$\Delta\phi(z) = \frac{1}{T_D} (1 - z^{-1}) E(z). \quad [3-15]$$

By use of the inverse z -transformation, we have

$$\Delta\phi(k) = T_D (e(k) - e(k - 1)), \quad [3-16]$$

where k is an integer. The derivative controller is programmed by using Eq. [3-16]. According to Eq. [3-16], the output of the controller at the instant, $k\Delta t$, is determined from the difference between the two error signals at instants of $k\Delta t$ and $(k-1)\Delta t$, where Δt is a sampling time.

In order to utilize Eq. [3-16] for computer control, a table of the values of the derivative time constant versus the magnitudes of the storage current is precalculated and loaded into the successive memories. The derivative time

constant varies with the storage current. In precalculation of the table, the value of the derivative time constant becomes infinite when the magnitude of the storage current is near zero. Infinite numbers are avoided by limiting the range of the derivative time constant. In our application, the range is from 30° to -30° . In other applications, these limits can be changed. The output of the controller is determined from a product of the derivative time constant and the difference of the two error signals. The multiplication may not be necessary in our application because of slow movement of the load current. As has been discussed in Section 2, it takes the load current 4 to 5 computation times to cross one quantization band even though the phase difference is held at its extreme, 90° or -90° , and the resolution of the microcomputer is 12 bits. This result is calculated from the experimental data measured 5 sec after the energy transfer starts. But, in any case, at least a few computation times are required. Therefore, if noise does not change the magnitude of the error signal, the value of the difference of the two error signals cannot exceed one in one computation time. By limiting it to one, effect of noise on control performance is decreased and calculation for multiplying the derivative time constant by the difference of the two error signals is avoided. The output of the controller has a value of the derivative time constant when the value of the difference of the two error signals is not zero.

Figures 3-2 and 3-3 show the dynamic behaviors of the control system of simultaneous open-loop with guided phase difference and closed-loop with the derivative controller. The resolutions of the microcomputer are 10 and 12 bits in Figs. 3-2 and 3-3, respectively. The reference signal rises at a rate of 20 A/sec until it reaches 60 A. Then, the reference signal is held at 60 A. In the waveform of the load voltage, the load voltage trace is partially disappeared because of the ink problem of the pen. The voltage oscillation causes a band, the center of which is determined by the fundamental component of the phase difference. While the magnitude of the storage current is large during the period of ramping up, the derivative time constant is small so is the output of the derivative controller. Then the magnitude of the voltage oscillation corresponding to the width of the band is small. During the period of holding, the magnitude of the voltage oscillation is increased from 5 V at the beginning of this period to 20 V at the end of the energy transfer.

Figure 3-2. Dynamic behavior of the control system of open-loop with the guided phase difference and closed-loop with the derivative controller. The resolution of the microcomputer is 10 bits.

Figure 3-3. Dynamic behavior of the control system of open-loop with the guided phase difference and closed-loop with the derivative controller. The resolution of the microcomputer is 12 bits.

Compared with the coil voltages in Fig. 3-2, the coil voltages in Fig. 3-3 show the continuous band during the period of holding. The derivative control action is induced due to noise on the coil current measurements. Especially when the resolution is increased, the size of the quantization is decreased and, as a result, the cycle of the voltage oscillation is increased.

Figure 3-4 shows the waveform of the instantaneous load voltage during the period of holding. The resolution of the microcomputer is 12 bits. A band generated due to the thyristor switching has the peak-to-peak magnitude of about 80 V. The coil voltages in Figs. 3-2 and 3-3 only show their average voltages. The peak-to-peak magnitude of the band is changed due to the derivative control action.

4. THE CONTROL SYSTEM OF SIMULTANEOUS OPEN-LOOP WITH THE GUIDED PHASE DIFFERENCE AND THE CLOSED-LOOP WITH THE BANG-BANG CONTROLLER

The control system of simultaneous open-loop with the guided phase difference and the closed-loop with the bang-bang controller is based on the open-loop control system with the guided phase difference. The feedback control is used for reducing the deviation of the output signal from the reference signal. If only the phase difference is regulated for control action, the phase difference is given by a sum of the fundamental component determined by the open-loop control using the guided phase difference and the deviation component by the feedback control. One of the above control system has been discussed in Section 3. But, as a result of summing the two components, the operational range of the deviation component is limited. The fundamental component, depending on the storage current, varies over a range between 90° and -90° . On the other hand, the range of the phase difference is limited by the two extremes, 90° or -90° . Then, when the value of the fundamental component is near 90° or -90° , the available range of the deviation component is quite limited and the control action is not effective.

On the other hand, the converter frequency as well as the phase difference can be controllable in the ICB. When the converter frequency is changed, the time interval corresponding to the phase difference must be adjusted. When the desired phase difference is maintained after the frequency shifting, the converter frequency and the phase difference are independent parameters. The range of the converter frequency and the range of the phase difference do not depend

Figure 3-4. Instantaneous load voltage during the period of holding. The resolution of the microcomputer is 12 bits.

on each other. By using the two independent parameters, the control system of simultaneous open-loop with the guided phase difference and closed-loop with the bang-bang controller is implemented. The error is used for the bang-bang control action. The output of the bang-bang controller changes the switching interval.

At first, we describe the pulse sequencer and the method for the frequency modulation. Figure 4-1 shows a block diagram of the pulse sequencer. The pulse sequencer generates a series of trigger pulses for the thyristor triggering at the both converters. The pulse sequencer consists of a 5 MHz crystal clock, a 4-bit down counter, 8-bit down counters, and two distributors. The mark, CK, at the blocks of the counters shows the location of the clock input. The 5 MHz clock is used for a clock of the 4-bit down counter. The 4-bit down counter receive the number from the microcomputer, counts down from this number to zero, and then sends a pulse to clock terminals of the both 8-bit down counters. The both 8-bit down counters receives numbers from the microcomputer. In a steady state, both numbers are 120's. The count period at each 8-bit down counter determines the duration of the switching interval at its corresponding converter. Each 8-bit down counter counts down from the number given by the microcomputer to zero, sends a pulse to the distributor, and loads the new number from the microcomputer. The distributor determines which thyristor is to be fired and delivers a pulse to a gate driver of the selected thyristor. The same time, it sends a pulse to the microcomputer for the purpose of interrupting the program if the interruption is required. The thyristor gate driver amplifies pulses and sent them to thyristor gates through a pulse transformer, which isolates the computer system from the experimental system.

The frequency modulation that is the change of the converter frequency but not of the phase difference can be done by the change of the numbers loaded in the two 8-bit down counters or the change of the number loaded in the 4-bit down counter. The method of changing numbers loaded in the both 8-bit down counters has been implemented and tested.⁽¹⁰⁾ By using this method, the change of the converter frequency is practically continuous. However, in order to maintain the desired phase difference, the phase shifting must be done along with the frequency shifting. On the other hand, the method of the phase shifting is designed in the case that the number loaded in each 8-bit down counter is 120 in a steady state. When the number loaded in each 8-bit down counter in a steady

Figure 4-1. Block diagram of the pulse sequencer.

state is changed, the program algorithm for the phase shifting is changed and the available number for the phase difference is reduced. In addition, the additional phase shift occurs due to our hardware design. During the transition of the frequency shifting, the number loaded in each 8-bit down counter is changed according to the three-step switching sequence for the frequency shifting. The frequency shifting occurs at one of the two converters and then at the other. The delay of control period causes the additional phase shift which depends on the phase difference before the transition, the converter frequency before the transition, and the converter frequency after the transition. The additional phase shift cannot easily be corrected by modifying the hardware design or software algorithm.

When the number loaded in the 4-bit down counter is changed according to the three-step switching sequence for the frequency shifting, the frequency shifting occurs at both converters at the same time and the desired phase difference is maintained without phase adjustment after the frequency shifting. When the number loaded in each 8-bit down counter is 120, the value of the converter frequency depends on the number stored in the 4-bit down counter. The discrete number of available converter frequencies is fifteen in the range from 460 to 7 kHz. In our control system, the frequency modulation is done by changing the number loaded in the 4-bit down counter. In order to use frequency modulation for bang-bang control action, three converter frequencies are required, selected from the available range. When the value of digitized error signal is zero, the converter frequency is set at the middle value of the three selected values. This value is used in the guide equation. When the value of the digitized error signal is positive and negative, the converter frequency is set at the lower and the higher values, respectively.

Figure 4-2 shows a block diagram of the control system of simultaneous open-loop with the guided phase difference and the closed-loop with the bang-bang controller for the period of ramping up. The phase difference is determined by open-loop control using the guided phase difference. On the other hand, the error signal $e(t)$ is used for the bang-bang control action. The controller determines the duration of the incremental interval from the three values, depending on the value of the digitized error signal. By adding the incremental interval to the initial switching interval t_{SW}^0 , the switching interval is changed, so is the converter frequency.

WHERE $\alpha = \frac{\pi^3 CL}{54 t_{SW}^0}$

Figure 4-2. Block diagram of the control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller for the period of ramping up.

Figure 4-3 shows the dynamic behavior of the simulated control system of simultaneous open-loop with the guided phase difference and closed-loop with the bang-bang controller. Figures 4-3(a), (b), (c), and (d) show the coil currents, the phase difference, the switching interval, and the error signal, respectively. The time step is 0.05 sec. The reference signal rises at a rate of 20 A/sec until it reaches 60 A. Then it is held at 60 A. In Fig. 4-3(a), the load current follows the reference signal. The magnitude of the storage current is decreased from 100 A to 56 A during the period of ramping up and then from 56 A to 48 A during the period of holding. In Fig. 4-3(b), the magnitude of the phase difference is increased from 30° to 65° during the period of ramping up and from 3.6° to 4.1° . The guide equation is changed with the period, so that, at the transition of the period, the magnitude of the phase difference is changed from 65° to 3.6° . Figure 4-3(c) shows the waveform of the switching interval. The duration of the switching interval is determined from the value of the digitized error signal and selected from 222, 264, and 333 μ sec. The least significant bit of the digitized error signal corresponds to a value of 0.2 A. It is noted that, in plotting the waveform of the switching interval, each point is connected continuously. But the switching interval is changed only among the three selected switching intervals. In Fig. 4-4(d), the error signal varies over a range between -0.04 A and 0.1 A during the period of ramping up and a range between -0.2 A and -0.19 A during the period of holding. During the period of ramping up, switching interval is changed even though the digitized error signal is less than the size of the quantization band. The change of the switching interval is due to digitization of the signals. When the reference signal and the feedback signal are increased with time, the value of the digitized error signal may not be zero at times when either the value of the digitized reference signal or the value of the digitized feedback signal is incremented.

Figures 4-4, 4-5, and 4-6 show the dynamic behaviors of the control system of simultaneous open-loop with the guided phase difference and closed-loop with the bang-bang controller. The resolutions of the microcomputer are 10, 11, and 12 bits in Figs. 4-4, 4-5, and 4-6, respectively. The reference signal rises at a rate of 20 A/sec until it reaches 60 A. Then it is held at 60 A. The fundamental waveform of the load current is determined by the open-loop control using the guided phase difference. The frequency modulation, regulated by the bang-bang controller, changes the magnitude of the average load voltage and then

INDUCTOR-CONVERTER BRIDGE BANG-BANG IN FREQUENCY

○ → STORAGE CURRENT
△ → LOAD CURRENT

(a) Coil Currents

Figure 4-3. Dynamic behavior of the simulated control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller.

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BANG-BANG IN FREQUENCY

⊙ → PHASE DIFFERENCE

(b) Phase difference

Figure 4-3. Dynamic behavior of the simulated control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller.

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BANG-BANG IN FREQUENCY

⊙ → SWITCHING TIME

(c) Switching interval

Figure 4-3. Dynamic behavior of the simulated control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller.

INDUCTOR-CONVERTER BRIDGE
GANG-BANG IN FREQUENCY

⊙ → ERROR

(d) Error

Figure 4-3. Dynamic behavior of the simulated control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller.

Figure 4-4. Dynamic behavior of the experimental control system of the open-loop with the guided phase difference and the closed-loop with the bang-bang controller. The resolution of the microcomputer is 10 bits.

Figure 4-5. Dynamic behavior of the experimental control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller. The resolution of the microcomputer is 11 bits.

Figure 4-6. Dynamic behavior of the experimental control system of open-loop with the guided phase difference and closed-loop with the bang-bang controller. The resolution of the microcomputer is 12 bits.

the rate of rise of the load current. By measuring the magnitude of the average load voltage, the rate of rise of the load current can be estimated. During the period of ramping up, the average load voltage varies over a range of -80 ± 8 V or the rate of rise of the load current varies over a range of 20 ± 2 A/sec. During the period of holding, the rate of rise of the load current varies within the range of ± 0.6 A/sec. The cycle of the voltage oscillation is increases with the resolution, but the range of the voltage oscillation is not changed with the resolution. As compared with the experimental results of the bang-bang control system (shown in Figs. 2-2 to 2-6), the range of the voltage oscillation of this control system becomes less than one-tenth.

This control system has a feature that, if the open-loop control using the guided phase difference keeps the digitized error signal to be zero, the bang-bang control action will not take place. Then, the cycle of the voltage oscillation during the period of holding is reduced, compared with the one during the period of holding.

Figure 4-7 shows the waveforms of the instantaneous coil voltages during the period of ramping up. The resolution of the microcomputer is 12 bits. The thyristor switching cause high-magnitude harmonic voltages and, as a result, a band in the oscillograph. The peak-to-peak magnitude of the band is changed with the converter frequency. The lower converter frequency causes the high peak-to-peak magnitude of the band while the higher converter frequency does the low peak-to-peak magnitude of the band.

5. CONCLUSIONS

The following results are summarized.

1. By using double data-word instructions, the length of the data word is increased from 8 bits to 16 bits, but the effective length is limited by the 12 bit resolution of the A/D converter.

Figure 4-7. Waveforms of the instantaneous coil voltages during the period of ramping up. The resolution of the microcomputer is 12 bits.

In order to minimize the number of program instructions, computation using double data-word instructions is done until error calculation is made. After that, the word length is reduced to 8 bits and computation using single data-word instructions is done. The resolution of the microcomputer depends on the location of 8 bits in the effective length of data word of the error.

Computation using double data-word instructions is used for control of the bang-bang feedback system. For the different resolution of the microcomputer, the magnitude of the current oscillation is measured and it is also calculated from the measured half period of the voltage oscillation and the measured magnitude of the coil currents. The magnitude of the current oscillation is recalculated by using the measured half period excluding the period for microcomputer computation and following phase shifting. The period for computer computation and following phase shifting is 3.51 msec during the period of holding.

According to the experimental results, the magnitude of the current oscillation depends on the resolution of the microcomputer and it is almost equal to the size of the quantization band. The difference is within an acceptable error. Noise is observed on the signals corresponding to the coil currents. Noise which appears to be associated with the thyristor switching has the peak-to-peak magnitude of about 0.5 A. Noise causes difficulty in measuring the magnitude of the current oscillation especially for the higher resolution of the microcomputer. For the higher resolution, the calculated and the recalculated values are useful because the effect of noise on calculation is reduced by using an average of the period of the voltage oscillation over more than ten cycles.

The magnitude of the current rise during one computation period (0.75 msec for the period of holding) is about 0.025 A. This magnitude is the upper limit when the load current is 57.5 A. On the other hand, the value of current corresponding to each bit for the 12-bit resolution of the microcomputer is 0.123 A. Then, even though the resolution is 12 bits, 4 to 5 computation times are required for crossing one of the quantization band. Therefore, the computation time does not limit control performance in our system.

2. The control system of simultaneous open-loop with guided phase difference and closed-loop with a derivative controller is introduced. The fundamental component of the phase difference is determined by the open-loop control using the guided phase difference, while the derivative component of the phase difference is determined by the closed-loop control using the bang-bang controller. The two components are added together to structure the phase difference. The derivative control action is derived by small signal analysis. By use of small signal analysis, the deviation component of the phase difference is expressed as a function of the time derivative of the deviation component of the load current. The coefficient of the time derivative is a function of the storage current. In order to reduce the error, the derivative controller is implemented. The coefficient of the derivative controller has the magnitude equal to the one derived from small signal analysis but the opposite sign.

In order to utilize the derivative controller for microcomputer control, a table of the values of the coefficient of the derivative controller versus the magnitudes of the storage current is stored in the successive memories. If noise does not amplify the signals, the difference of the two digitized error does not increment more than one in one computation period because of slow movement of the error signal. Then computation for multiplying the difference of the two errors by the coefficient of the derivative controller is avoided. Whenever the value of the digitized error signal is not zero, the output of the derivative controller is determined from the contents of the memory in which the location corresponds to the value of the digitized storage current. This output is added to the value given from the guide equation to generate the value of phase difference.

According to the experimental results, the derivative control action results in oscillation of the coil voltages due to noise. The range of the voltage oscillation is increased as the storage current is decreased. The cycle of the voltage oscillation is increased with the resolution of the microcomputer.

3. The control system of simultaneous open-loop with the guided phase difference and closed-loop with a bang-bang controller is introduced. The phase difference is determined by open-loop control using the guided phase

difference. On the other hand, the error signal is used for bang-bang control action. The output of the bang-bang controller determines the converter frequency which is selected from three frequencies. This control system has a feature that, if the open-loop control using the guided phase difference maintains the value of the digitized error signal to zero, the bang-bang control action will not take place.

According to the simulation results, the error signal moves over a range between -0.04 A and 0.1 A during the period of ramping up and a range between -0.2 A and -0.19 A during the period of holding. Each least significant bit of the error signal corresponds to a value of 0.2 A. Even though the error signal moves within the values, ± 0.2 A during the period of ramping up, the value of the digitized error signal is not always zero. Nonzero values occur at times when either the value of the digitized reference signal or the value of the digitized feedback signal is incremented, but not both. When the value of the digitized error is positive, the switching interval is its maximum. When the digitized value is negative, the switching interval is its minimum. During the period of holding, the switching interval is kept constant.

According to the experimental results, the frequency modulation causes the change of the average load voltage and the rate of rise of the load current. During the period of ramping up, the average load voltage varies over a range of -80 ± 8 V or the rate of rise of the load current varies over a range of 20 ± 2 A/sec. During the period of holding, the rate of rise of the load current varies over a range of ± 0.6 A/sec. The cycle of oscillation of the average load voltage is increased with the resolution of the microcomputer, but the range of the oscillation is not changed with the resolution. The cycle of the oscillation is decreased during the period of holding compared with the one during the period of ramping up.

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APPENDIX

A. AVERAGE POWER⁽¹²⁾

The waveforms of the load-side converter current, $i_{CL}(t)$, and the storage-side converter current, $i_{CS}(t)$, in one of the three phases in a steady state is shown in Fig. A-1. The rectangular waveform can be represented by the Fourier series as follows:

$$i_{CS}(t) = \sum_{n=1}^{\infty} a_n \cos n\omega t, \quad [A-1]$$

$$i_{CL}(t) = \sum_{n=1}^{\infty} b_n \cos n(\omega t + \phi) \quad [A-2]$$

where

$$a_n = \frac{2I_S}{\pi n} [1 - (-1)^n] \sin \frac{n\pi}{3}, \quad [A-3]$$

$$b_n = \frac{2I_L}{\pi n} [1 - (-1)^n] \sin \frac{n\pi}{3}, \quad [A-4]$$

and

ϕ = angle by which load switching sequence leads
storage switching sequence.

All even terms of a_n and b_n are zero due to $[1 - (-1)^n]$ term and all terms that are multiples of the phase number, 3, are zero. By using Eqs. [A-1] and [A-2], the capacitor voltage is obtained as follows:

Figure A-1. Waveforms of the converter currents in one of the three phases in a steady state.

$$v_C(t) = \frac{1}{C} \int_0^t [i_{CS}(t) + i_{CL}(t)] dt + v_C(0)$$

$$= \frac{1}{C} \sum_{n=1}^{\infty} a_n \left[\frac{\sin n\omega t}{n\omega} \right]_0^t + \frac{1}{C} \sum_{n=1}^{\infty} b_n \left[\frac{\sin n(\omega t + \phi)}{n\omega} \right]_0^t + v_C(0)$$

$$= \frac{1}{C} \sum_{n=1}^{\infty} \frac{1}{n\omega} [a_n \sin n\omega t + b_n \sin n(\omega t + \phi) + b_n \sin n\phi] + v_C(0)$$

[A-5]

In normal operation, the balance of each capacitor voltage is maintained; i.e., each average capacitor voltage is zero. On the other hand, the sinusoidal functions have an integral value equal to zero over a cycle, but the $\int \sin n\phi$ term does not. Therefore, the constant $v_C(0)$ is equal and opposite in value $\int \sin n\phi$, so the third and fourth terms of Eq. [A-5] are eliminated. By substituting Eqs. [A-3] and [A-4] into a_n and b_n , we have:

$$v_C(t) = \sum_{n=1}^{\infty} \frac{1}{C} \left(\frac{2}{\pi n^2 \omega} \right) [1 - (-1)^n] \sin \frac{n\pi}{3} [i_S \sin n\omega t + i_L \sin n(\omega t + \phi)] \quad [A-6]$$

The average power into the load coil is the time average value of the negative product of $v_C(t)$ and $i_L(t)$ over a converter period. The average power P_0 is

$$P_0 = -\frac{1}{T} \int_0^T v_C(t) i_{CL}(t) dt$$

$$= -\frac{1}{T} \int_0^T \left\{ \sum_{n=1}^{\infty} \frac{1}{C} \left(\frac{2}{\pi n^2 \omega} \right) [1 - (-1)^n] \sin \frac{n\pi}{3} [i_S \sin n\omega t + i_L \sin n(\omega t + \phi)] \right\} i_L \sin n(\omega t + \phi) dt.$$

$$\left\{ \sum_{n=1}^{\infty} \left(\frac{2i_L}{\pi n} \right) [1 - (-1)^n] \sin \frac{n\pi}{3} \cos n(\omega t + \phi) \right\} dt, \quad [A-7]$$

where T is a converter period.

The integral value of the product of the second term of $v_C(t)$ and of $i_{CL}(t)$ is zero, whether m is equal to n , or not. Computation of the integral value of the first term of $v_C(t)$ and $i_{CL}(t)$ is aided by the trigonometric expansion of $i_{CL}(t)$;

$$\cos m(\omega t + \phi) = \cos m\omega t \cos m\phi - \sin m\omega t \sin m\phi \quad [A-8]$$

The first term of the expansion in Eq. [A-8] leads to $\cos m\omega t \sin n\omega t$ term in the product of $v_C(t)$ and $i_{CL}(t)$. This term has zero over the integral for either m equal to n or m not equal to n . The second term of expansion in Eq. [A-8] leads to a $\sin m\omega t \sin n\omega t$ term in the product of $v_C(t)$ and $i_{CL}(t)$ that has a nonzero integral value only when m is equal to n . Therefore, Eq. [A-8] reduces to

$$P_o = -\frac{1}{T} \int_0^T \sum_{n=1}^{\infty} \frac{1}{C} \frac{4i_L i_S}{\pi^2 n^3 \omega} [1 - (-1)^n]^2 \left(\sin \frac{n\pi}{3}\right)^2 (-\sin^2 n\omega t) \sin n\phi \, dt$$

On the other hand, we have

$$\int_0^T \sin^2 n\omega t \, dt = \int_0^T \frac{1 - \cos 2n\omega t}{2} \, dt = \frac{1}{2} \left(t - \frac{\sin 2n\omega t}{2n\omega} \right) \Big|_0^T = \frac{T}{2}$$

Hence, the average power is

$$P_o = \frac{\infty}{n=1} \frac{1}{C} \frac{2i_L i_S}{\pi^2 n^3 \omega} [1 - (-1)^n]^2 \left(\sin \frac{n\pi}{3}\right)^2 \sin n\phi \quad [A-9]$$

The total power into the load coil is three times the value indicated in Eq. [A-9]. The contribution to power flow for increasingly higher harmonics is inversely proportional to n^3 . Since the next nonzero term after the first is

the fifth, the power to a very high degree of accuracy can be calculated by neglecting higher harmonics. By taking only the fundamental term and substituting $2 T_L/6 t_{SW}$ into ω , the total power P in the circuit is

$$P = 3 \frac{1}{C} \frac{2I_L I_S}{2\pi} \frac{1}{6t_{SW}} 4 \frac{3}{4} \sin\phi$$

$$= \frac{54I_L I_S t_{SW}}{\pi^3 C} \sin\phi$$

[A-10]

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