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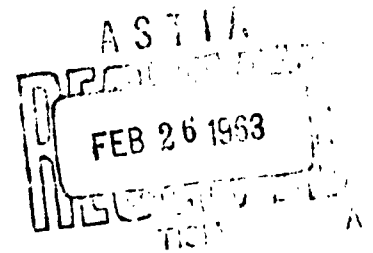
Development of Sound Analysis Equipment

for Sonar Research

Part III

FINAL REPORT ON CONTRACT Nonr-2129

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WOODS HOLE, MASSACHUSETTS

WOODS HOLE OCEANOGRAPHIC INSTITUTION
Woods Hole, Massachusetts

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Part III

FINAL REPORT ON CONTRACT Nonr-2129

by

Lincoln Baxter

and Vernon Chi

Submitted to Undersea Warfare Branch, Office of Naval
Research, under Contract Nonr-2129(00) (NR261-104).

October 1961

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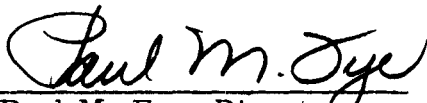

Paul M. Fye, Director

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ABSTRACT

An eight-channel analog computer has been developed for hydro-acoustic and geophysical oceanographic studies. This computer drives Sanborn pens to produce a deflection proportional to the total energy received at an input after any desired instant. Alternatively, a deflection proportional to either amplitude, rectified amplitude, or power, averaged over a selected time interval or proportional to certain functions useful for studies of the correlation of signals at two inputs may be produced. These latter functions are the product of the amplitudes at the inputs and the sum of the squares of these amplitudes.

High gain D. C. amplifiers with mixed capacitive and resistive feedback or with pure capacitive feedback are used to integrate with respect to time various quantities with or without a time exponential weighting factor. With the weighting factor, the integral approximates a running average over a time interval selected in 2 to 1 steps between 4 seconds and 10^{-3} seconds. Without the weighting factor, all contributions are weighted equally. A deflection proportional to the logarithm of any of the possible integrals may also be produced. Computation to an accuracy of 5% or better is obtained for input signals having all significant components between 20 cycles and 10 kilocycles per second. Under certain conditions 5% accuracy is obtained for frequencies between 5 cycles and 50 kilocycles per second.

I. INTRODUCTION

1.1 Hydroacoustic transmission to ranges of more than several thousand yards frequently involves many paths. Because of the multiplicity of paths a single short emitted pulse is received as a complex sound consisting of a series of pulses of varying length and number. Correspondingly a radiated sine wave is received with an amplitude that depends on the phase difference between different paths.

1.2 Fourier analysis of emitted and received signals is an extremely useful technique. The radiated signal $P(t)$, source pressure as a function of time, may be expressed as a source pressure frequency density, $\phi(f)$, by:

$$\phi(f) = \int_{-\infty}^{+\infty} e^{-2\pi i f t} P(t) dt \quad (1)$$

$\phi(f)$, a function of frequency, is the Fourier transform of $P(t)$, a function of time. Similarly the received signal, $p(t)$, may be transformed to $\varphi(f)$ by:

$$\varphi(f) = \int_{-\infty}^{+\infty} e^{-2\pi i f t} p(t) dt \quad (2)$$

The ratio of $\varphi(f)/\phi(f)$ is defined as $T(f)$, the frequency transfer function between the source and receiver locations, i. e.

$$\varphi(f) = \phi(f) T(f) \quad (3)$$

In general $\varphi(f)$, $\phi(f)$ and $T(f)$ have both real and imaginary

parts to express both amplitude and phase. If the sound intensities are not too great and if the transmission paths do not change appreciably, the $p(t)$ that corresponds to any $P(t)$, regardless of the shape of $P(t)$, can be computed from equations (1) and (3) and the inverse Fourier transform,

$$p(t) = \int_{-\infty}^{+\infty} e^{2\pi i f t} \phi(f) T(f) df \quad (4)$$

when $T(f)$ has been previously determined by means of another signal or by theoretical analysis.

1.3 In theoretical analysis it is customary to express $T(f)$ as a sum of components. These components may be chosen in different ways. Two ways are common. In one the components of $T(f)$ correspond to different paths of sound rays; in the other, the components correspond to different normal modes (Worzel, Ewing, and Pekeris, 1948; Officer, 1958). The normal mode description is useful for regions that are shallow relative to the wavelength and where surface, bottom, and sub-bottom layers are nearly parallel, so that ray paths at long ranges grade so continuously into each other that they cannot be conveniently resolved. When transmission occurs over a relatively limited number of ray paths, the common deep water situation, the interference effects that occur can be resolved without resorting to the normal mode description. If travel time differences between fairly discrete ray paths are resolved by the measuring technique, the ray description is usually more convenient.

1.4 If a single short radiated pulse $P(t)$ is received as a series of n pulses, i. e.

$$p(t) = \sum_{x=1}^{x=n} p_x(t) \quad (5)$$

where $p_x(t)$ is the pressure wave of the x th pulse and $p(t)$ is zero except when a pulse is received, and if $p(t) \equiv p_x(t)$ for $t_i < t < t_n$,

then from equations (2) and (5)

$$\varphi(f) = \sum_{\alpha=1}^{\alpha=n} \int_{-\infty}^{+\infty} e^{-2\pi i f t} p_{\alpha}(t) dt = \sum_{\alpha=1}^{\alpha=n} \varphi_{\alpha}(f) \quad (6)$$

where the n integrals in the summation are each the transform of a component pressure pulse $p_{\alpha}(t)$.

We choose n components of $T(f)$ such that the α^{th} component

$$T_{\alpha}(f) = \frac{\varphi_{\alpha}(f)}{\phi(f)} \quad (7)$$

From equations (6) and (7)

$$\varphi(f) = \sum_{\alpha=1}^{\alpha=n} \phi(f) T_{\alpha}(f) = \phi(f) \sum_{\alpha=1}^{\alpha=n} T_{\alpha}(f) \quad (8)$$

and from equations (3) and (8)

$$T(f) = \sum_{\alpha=1}^{\alpha=n} T_{\alpha}(f) \quad (9)$$

The components, $T_{\alpha}(f)$, are the transfer functions that correspond to certain ray paths or combinations of ray paths.

1.5 Detailed measurement of $\varphi(f)$ is not normally made because of time limitations. For many applications, a quadratic form derived from Plancherel's theorem, i. e., the Fourier-integral energy theorem (Goldman, 1949, p. 431), gives sufficient information. In the symbols above Plancherel's theorem is as follows:

$$\int_{-\infty}^{+\infty} p^2(t) dt = \int_{-\infty}^{+\infty} \varphi(f) \varphi^*(f) df = 2 \int_0^{\infty} |\varphi(f)| df \quad (10)$$

where $*$ denotes the complex conjugate. One may derive a quadratic transfer function as follows:

$$|\varphi(f)|^2 = \phi(f) \phi^*(f) T(f) T^*(f)$$

or

$$|\varphi(f)|^2 = |\phi(f)|^2 |T(f)|^2 \quad (11)$$

therefore

$$\int_{-\infty}^{\infty} p^2(t) dt = 2 \int_0^{\infty} |\phi(f)|^2 |T(f)|^2 df \quad (12)$$

also

$$\int_{-\infty}^{+\infty} p^2(t) dt = 2 \int_0^{\infty} |\phi(f)|^2 df \quad (13)$$

Consider component pressure pulses $p_{ab}(t)$ and $p_{ab}(t)$ obtained by restricting $\phi(f)$ and $\varphi(f)$ to a frequency band $f_a < f < f_b$ by means of a filter. If $p_{ab}(t)$ is zero before t , and

$P_{ab}(t)$ is zero after t_2 ,

$$\int_{t_1}^{t_2} P_{ab}^2(t) dt = 2 \int_{f_a}^{f_b} |\varphi(f)|^2 df \quad (14)$$

by equation (10) and

$$\int_{t_1}^{t_2} P_{ab}^2(t) dt = 2 \int_{f_a}^{f_b} |\phi(f)|^2 df \quad (15)$$

by equation (13). Since the average results for the frequency band approach the results for the center frequency,

$$f_m = \frac{f_a + f_b}{2}$$

when the band limits approach f_m , we shall define average energy spectral densities as follows:

$$E'_{ab}(f_m) = \frac{1}{\rho c (f_b - f_a)} \int_{t_1}^{t_2} P_{ab}^2(t) dt = \frac{2}{\rho c (f_b - f_a)} \int_{f_a}^{f_b} |\varphi(f)|^2 df \quad (16)$$

$$E_{ab}(f_m) = \frac{1}{\rho c (f_b - f_a)} \int_{t_1}^{t_2} P_{ab}^2(t) dt = \frac{2}{\rho c (f_a - f_b)} \int_{f_a}^{f_b} |\phi(f)|^2 df \quad (17)$$

where ρ_c is the specific acoustic impedance of the sea water. $E'_{ab}(f_m)$ and $E_{ab}(f_m)$ are the equivalent plane wave energies per unit area per cycle per second at the receiver and source respectively. The equivalent plane wave energy per unit area is defined as the energy per unit area that would have been carried across a surface parallel to the wavefront by a plane wave with the same acoustic pressure as that actually measured (Horton, 1957, Art. 2A9). In the remainder of this report, except where otherwise noted, any mention of energy or power will refer to energy or power of an equivalent plane wave. A mean energy transfer coefficient,

$$T_{ab}^2(f_m)$$

for the band f_m is defined as follows:

$$T_{ab}^2(f_m) = \frac{E'_{ab}(f_m)}{E_{ab}(f_m)} \quad (18)$$

If $\phi(f)$ is constant between f_a and f_b equation (12) becomes

$$\int_{t_1}^{t_2} p_{ab}^2(t) dt = 2 |\phi(f_m)|^2 \int_{f_a}^{f_b} |T(f)|^2 df \quad (19)$$

and equation (13) becomes

$$\int_{t_1}^{t_2} P_{ab}^2(t) dt = 2 |\phi(f_m)|^2 (f_b - f_a) \quad (20)$$

Then from equations 16, 17, 18, 19, and 20

$$T_{ab}^2(f_m) = \frac{1}{f_b - f_a} \int_{f_a}^{f_b} |T(f)|^2 df \quad (21)$$

Use of equivalent plane wave energy for the source may be justified by the usual practice of measuring the source levels at a range where the radiated wave has become nearly plane and the acoustic pressure has decreased to a level for which transmission of the wave is linear as required for Fourier analysis.

1.6 We shall now see that the power transfer for a long steady pulse of the given bandwidth is given by the same quadratic transfer function. Let $W_{ab}(f_m)$ be the power radiated in the frequency band during a long steady pulse $P(t)$. Let $W'_{ab}(f_m)$ be the received power. As $P(t)$ becomes very long,

$$(f_b - f_a) \frac{E_{ab}(f_m)}{t_2 - t_1} \longrightarrow W_{ab}(f_m) \quad , \quad (22)$$

and

$$(f_b - f_a) \frac{E'_{ab}(f_m)}{t_2 - t_1} \longrightarrow W'_{ab}(f_m) \quad , \quad (23)$$

therefore, by equation (18),

$$\frac{W'_{ab}(f_m)}{W_{ab}(f_m)} = \frac{E'_{ab}(f_m)}{E_{ab}(f_m)} = T_{ab}^2(f_m) \quad (24)$$

Since equation (24) is true regardless of the pulse length or the phase relationship of the components, the quadratic transfer function obtained with energy measurements for short pulses is the same as the quadratic transfer function obtained with power measurements in the steady state condition.

1.7 If separate paths or combinations of paths are resolved by separate pulses as discussed in paragraph 1.4, a quadratic transfer coefficient, $T_x^2(f_m)$, may be measured for each component, $T_x(f_m)$, of the frequency transfer function $T(f)$, and

$$\sum_{x=1}^{x=n} T_x^2(f_m) = T^2(f_m) \quad (25)$$

The $T^2(f_m)$ determined from equation (25) is, as shown above, the same as that which would be obtained from power measurements of long steady pulses of the same bandwidth.

1.8 Quantities frequently measured in ray studies of acoustic propagation include travel time difference between ray paths, transmittance or transmission loss, and reflectivity or reflection loss at boundary surfaces. These quantities are functions of frequency, incident angle, and geographic location. All but the first are obtained by comparison of the measured for known paths.

1.9 Unless transmission measurements are made circumspectly it is difficult or impossible to determine the separate contributions of interference and transmission loss. In ray studies it is desirable to deal with these factors separately because interference effects for single frequencies or narrow frequency bands may change quite rapidly with frequency and range without great changes in the number and relative intensity of the sound from contributing paths. The characteristics of individual paths usually change much more slowly.

1.10 One may conceive of eliminating interference effects from transmission measurements by averaging sine wave measurements of differing frequency over a frequency band broad enough to include equal numbers of all phase differences. However, the time required to assemble the required number of sine wave measurements and to compute the averages would be impractically long. Computation of the travel time differences of the individual paths from sine wave measurements would also be extremely difficult.

1.11 Corliss (1958) discusses the time bandwidth conditions for resolving discrete transient signals with power sensitive detectors. Her results are applicable to analysis by filters or by correlation techniques. Both types of analysis select components of a signal. Signals compete with each other and with noise in a space spanned by the coordinates power (W), frequency (f), and time (t). Presence of a transient signal of bandwidth Δf within a time interval Δt is detected by a power fluctuation ΔW . Analysis systems have equal potential ability to distinguish a given number of separate signals if they are sensitive to equal volumes $\Delta W \Delta f \Delta t$ in this signal space. The natural performance of real analyzers limits the minimum size of this volume, but the shape can be altered to match the analysis system to the characteristics of the signals which one desires to distinguish from noise and from other signals.

1.12 The theory outlined here and in Corliss's paper implies that the minimum time bandwidth product required for the average discussed in paragraph 1.10 is that of a pulse short enough to resolve the interfering paths. The minimum bandwidth for such resolution requires the optimum combination of phases of the frequency components of the pulse. Pulses with the same bandwidth with other phase relationships will be longer, but as shown in paragraphs 1.6 and 1.7, if the time bandwidth product is adequate, the total quadratic transfer coefficient for the paths in question gives the total energy transferred whether or not the pulses due to the separate paths are resolved. Goldman (1948, Chapter 4) gives a more detailed discussion of the effect of bandwidth on pulse resolution. As discussed above (paragraph 1.5) the transfer coefficient is computed from

$$T_{ab}^2(f_m) = \frac{\int_{t_1}^{t_2} p_{ab}^2(t) dt}{\int_{t_1}^{t_2} P_{ab}^2(t) dt} \quad (26)$$

If the $p_x(t)$ are resolved

$$T_x^2(f_m) = \frac{\int_{t_1}^{t_k} p_x^2(t) dt}{\int_{t_1}^{t_2} P^2(t) dt} \quad (27)$$

1.13 Short pulses are more convenient than other wide band signals, because it is easy to measure transmission via ray paths whose travel time differences are greater than the pulse length. A short pulse tests all dimensions of the frequency transfer function in a short time. Thus, the probability of significant changes in the experimental conditions during the measurement is smaller than for a longer pulse. All narrow band analyses as well as wide band analyses can be made at the receiving location. Such measurements are equivalent to multiple sine wave measurements in a medium having a frequency transfer function equal to that of the real ocean at the time of the pulse. Signal-to-noise advantages of correlation may be secured by repeating the pulse at precise intervals if the intervals are short relative to the rate of change of the transfer function of the medium. If the repetition intervals are long relative to the travel time details, the advantages of pulse transmission for travel time measurements are retained. When it is difficult or impossible to satisfy both of the above conditions simultaneously, pseudo-random noise signals may be more advantageous than pulses. In all other cases, computation of travel time differences is easier with pulses.

1.14 In normal mode studies one deals with interference effects instead of attempting to average them out, but the time-bandwidth problem is still fundamental. In problems where normal mode descriptions are most convenient, a pulse source gives rise to received pulses that are associated with interference maxima rather than rays. In the ocean, the interference produces standing waves in the vertical direction and traveling waves in the horizontal. The propagating sound is classified in modes by the number of pressure maxima in the standing wave. One is interested in measuring transmittance and group velocity (Worzel, Ewing, and Pekeris, 1948) for each mode as a function of frequency and range. This is normally practical only for shallow water or very low frequencies where there are only a few modes and the dispersion is great enough to cause them to arrive with sufficient time separation. To make these measurements conveniently one must use frequency bands wide enough to permit short interference pulses for accurate travel time measurements yet ones narrow enough to prevent appreciable energy from other modes (those that have the same group velocity at a different frequency) from being included in the measurement.

1.15 The mode to which the separate interference pulses belong can best be identified by examination of a plot in which the Cartesian coordinates are frequency and time with acoustic intensity represented by shading or contours. Such plots have been widely used in speech analysis (Potter, Kopp, and Green, 1947). These plots may be generated intermittently by a Kay Electric Co. Vibralyzer or continuously by an instrument that combines Raytheon Rayspan filters with an Alden event recorder (Real Time Spectrum Analyzer). An example of the patterns obtained with the Vibralyzer in shallow water transmission measurements is given by Baxter (1957).

1.16 The advantages of short pulses as broad band sources with optimum time resolution have lead to the widespread employment of explosive charges, electric sparks, and other impulse generators such as the Edgerton Thumper or Boomer as sonic sources for hydroacoustic measurements (Weston, 1960; Worzel, Ewing, and Pekeris 1948; Hersey, Edgerton, Raymond, and Hayward 1960; Hoskins and Knott, 1961; Knott and Hersey, 1956). The extremely short rise time and exponential decay of at least some parts of the signal emitted by most of these sources lends itself to accurate analysis of very small travel time differences by the use of wide band filtering at medium to high frequencies. Individual peak measurements are very unreliable indications of transmission loss because the peak level of the received pulse is very sensitive to the presence and phase relationship of the high frequency components which are altered by the transmission path.

1.17 In consideration of the above facts Officer and Dietz (1953) pioneered measurement of the energy received as a function of time. They used simple analog computing components to compute the time integral of the square of filter output potential. Where pressure-sensitive hydrophones are used this is proportional to

$$\int_{t_0}^{t_1} p^2(t) dt$$

where t_0 and t_1 are the initial and final time limits of the integration. For sound waves that are plane or nearly so and for pulse bandwidths that are large enough, the above integral is proportional to the component (in the filter band) of the received energy at the hydrophone. The instrument reported here was originally conceived to carry out the computation of this integral. During its development other capabilities have been added as described below.

II. PAST AND POSSIBLE FUTURE DEVELOPMENT OF THE 45.001 OCEANOGRAPHIC COMPUTER

2.1 Late in 1955 the Deep Water Propagation Committee, a group of hydroacousticians organized at the request of the Office of Naval Research, recognizing the importance of this type of analysis and aware that new instruments can be developed more economically for quantity production, recommended that ten sets of sound analysis equipment representing a further development of the principles used by Officer and Dietz (1953) be distributed to laboratories working in the field of hydroacoustics.

2.2 Multichannel computers were recommended because analysis of data from one hydrophone for one frequency band at a time is much too slow in view of the large amount of data to be processed. Eight channels per computer were chosen because this was the largest number of channels available in the desired type of direct writing recorder.

2.3 Further investigation of the abilities of analog computers and the requirements, present and projected future, of hydroacoustic analyses led to the development of an instrument called the 45.001 Oceanographic Computer, (Electronics Associates Inc.) that was much more versatile than originally planned. Many suggestions were received from prospective users, especially Dr. J. B. Hersey of the Woods Hole Oceanographic Institution and Dr. V. C. Anderson of Marine Physical Laboratories, and

from the engineers who created the electronic and mechanical design of the system, Mr. Harry Turner and Mr. Gerhart Klaiber of Electronic Associates. The suggestions were evaluated and coordinated and the final design was specified by the senior author of the present paper.

2.4 The final version of the computer drives Sanborn pen motors to produce a deflection proportional to the total energy received at an input after any desired instant. Alternatively a deflection proportional to amplitude, rectified amplitude, or power averaged over a selected time interval, or proportional to certain functions useful for studies of the correlation of signals at two inputs may be produced. The functions of two inputs are derived from the product of the amplitude at each input with that at the other or the sum of the squares of the amplitudes at the same inputs. All functions may be integrated with respect to time, with or without a time exponential weighting factor. In the first case the integral approximates a running average over a time interval selected in 2 to 1 steps between 4 seconds and 10^{-3} seconds. In the second case all contributions are weighted equally. The exact equations of these integrals are given later. A deflection proportional to the logarithm of any of the possible integrals may also be produced. Computation to an accuracy of 5% or better is obtained for input signals having all significant components between 20 cycles and 10 kilocycles per second. Under certain conditions 5% accuracy is obtained for frequencies between 5 cycles and 50 kilocycles per second. Nearly all of these functions are useful in analysis and presentation of either transient or steady state signals of different frequency bandwidths to determine Fourier transfer functions in hydroacoustic and geophysical research. The equations of operation of the computer are given in Part III in which the various programs and their applications are defined in detail and discussed.

2.5 In late November and early December 1960, five computers, Production Model I, were distributed under the original plan. The number was reduced from ten to five because costs were much higher than anticipated. Woods Hole Oceanographic Institution, Marine Physical Laboratory, Scripps Institution of Oceanography, Hudson Laboratories, and Lamont Geological Observatory each received a computer.

2.6 Three development versions of this computer were built before production Model I. Modification to meet present standards was possible only for the third prototype. This eight-channel computer was not considered standard under the original plan because extensive improvements in the pre-amplifiers, the scale factors, and the logarithmic circuits were made in the computers that followed. Rebuilt later, under another

contract, with further improvements over Model I, it is designated as Model II. Except where otherwise noted the computer described in this paper is Model II. Two other Model I computers procured by the Woods Hole Oceanographic Institution under a third contract are being modified to meet Model II specifications.

2.7 It was found that computation was greatly speeded and errors reduced by automatically recording control settings of the computers, associated amplifiers, and filters. This has been accomplished by attaching switch encoders to the controls. The encoders are designed to read the control settings in machine language. The outputs of the encoders are recorded simultaneously on punched cards and typewritten sheets with IBM automatic process recording machinery. An alternate process, for applications where space and money are limited, generates digit codes in a switch connected to a multichannel event recorder and records up to 140 digits selected by switch encoders identical to those used in the IBM system. The digits are recorded as coded markings on the event recorder paper. A photoelectric reader would be required to machine-process such records, but the recorder used in the field would be cheap and compact and the field records could be read by an observer when necessary during the experiment. Reading of such records would be only slightly more difficult than reading of the typewritten records of the IBM system. A third process using combination punched paper tape and typewritten records is possible and might have advantages for certain experiments. We recommend that switch encoders be procured for all oceanographic computers and that the most suitable automatic recording process be used in any future hydroacoustic experiments employing this instrument.

2.8 Study of sound-scattering layers in the ocean could be greatly facilitated by automatic computation of the scattering coefficient (Machlup and Hersey, 1955). Preliminary study indicates that special input equipment to adapt the oceanographic computer to this program and other reverberation studies may be feasible. Further study is needed to determine whether development and use of such equipment would be more practical than digital computation of the same data.

2.9 A very small general purpose analog computer was purchased early in the development of the oceanographic computer for study and evaluation of its components as possible elements for the special purpose oceanographic computer. While none of the elements of this general purpose computer

entered into the final design we feel that the experience gained was very valuable and believe that it would be wise to continue to study the application of analog computing techniques to oceanographic problems. A small modern general purpose transistorized analog computer with at least 20 amplifiers and several universal function generators, although no substitute for a digital computing facility may nevertheless solve many problems more quickly and easily. The above mentioned very small computer that we purchased for study is not sufficiently accurate and does not have enough amplifiers for further progress in this direction.

2.10 The state of the computer art has improved greatly since the Oceanographic Computer was designed. If sufficient interest from the various hydroacoustic laboratories were expressed, a fresh start and a new design could produce an Oceanographic Computer with significantly greater accuracy, dynamic range and reliability.

III. DESCRIPTION OF THE ELECTRONIC ASSOCIATES OCEANOGRAPHIC COMPUTER AND ITS USES

3.1 The Electronic Associates Oceanographic Computer, Model II (manufacturer's No. 4.038) is shown in Figure 1. Although the computer is designed as an eight-channel group to match a standard Sanborn Model 158-100-B recorder, modular construction is employed for flexibility, reliability, and ease of repair. The basic program unit, containing all the computing circuits is the dual-channel module (Manufacturer's Part #45.001), the front panel of which is illustrated in Figure 2. Four of these dual-channel modules are included in a computer group. Each dual-channel module is independently supplied with all but one of the necessary operating potentials by a regulated power supply (Manufacturer's Part #10.067) whose front panel is illustrated in Figure 3. The reset potential for the integrators is supplied by the limit and storage indicator as described below. Since one power supply is required for each module there are four power supplies in the computer group. The remaining major component of a computer group is the limit and storage indicator (Manufacturer's Part #20.458) illustrated in Figure 4 which supplies a potential to relays in all the dual-channel modules to reset the integrators to zero at the beginning of each computation, and indicates the peak levels in the computing circuits to permit the operator to recognize and reject inaccurate computations and to set the

controls for the most accurate computation. One limit and storage indicator is employed in each computer group. A description of its use is given in paragraphs 3.14 and 3.15.

3.2 All connections into and out of the computer group are found on the rear of the computer rack on the bottom panel which is shown in Figure 5. Eight resistive 600-ohm inputs are provided, one per channel. The inputs are wired for single-ended operation with one side grounded, but may easily be rewired for balanced or differential operation. There are eight single-ended outputs capable of delivering from 100 volts positive to 100 volts negative at 25 milliamperes. This is sufficient to drive Sanborn recording pen motors. Eight monitor outputs for connection to a high impedance indicator such as a cathode ray oscilloscope are connected to the highest level linear point of the computer circuits, that is to the output of the preamplifiers immediately before the non-linear computing elements.

3.3 Each dual-channel module has two inputs and two outputs. The inputs, A and B are designed to be driven by any 600-ohm source at levels between 20 millivolts and 14 volts. The outputs, C and D, are each designed to drive a Sanborn pen motor. For most programs C is a function of A, and D is a function of B. However, several programs are provided in which both C and D are functions of both A and B.

3.4 The block diagram of a dual-channel module (Figure 6) illustrates the functional elements and shows how they are connected and switched to set up the various programs whose formulas are given in Table I. The effects of the major program controls are explained in paragraph 3.5. Paragraph 3.6 deals with overload limit and centering controls. The symbols used in the formulas of Table I will be explained in paragraphs 3.7 through 3.12. Uses of the various functions in oceanographic and geophysical research are explained in paragraphs 3.13 through 3.19. The equations of Table I all assume, for simplicity, that all significant signal frequencies are within the pass bands given in Figures 7-12 for the various computer components and programs. The flow through the block diagram, and the frequency responses of the computer in the various programs are discussed in paragraph 3.20. The dynamic range of the computer is discussed in paragraphs 3.21 and 3.22.

Table I Equations of standard programs of Model 45. 001
Oceanographic Computer, Production Model II

FUNCTION C (G) OR D (G)	a	b	c	d	e	f	g	h	FUNCTION NUMBER
$\frac{f\beta}{\tau} \int_{t_0}^{t_1} (N+G) e^{\frac{t-t_1}{\tau}} dt$	$\frac{A(t)}{\gamma_1}$ ■	$\frac{ A(t) }{\gamma_1}$	$\frac{A^2(t)}{\gamma_1^2}$	$\frac{B(t)}{\gamma_2}$ ■	$\frac{ B(t) }{\gamma_2}$	$\frac{B^2(t)}{\gamma_2^2}$	$\frac{A(t)B(t)}{\gamma_1 \gamma_2}$	$\frac{A^2(t)}{\gamma_1^2} + \frac{B^2(t)}{\gamma_2^2}$	1
$K \log it \left[\frac{\beta}{L\tau} \int_{t_0}^{t_1} (N+G) e^{\frac{t-t_1}{\tau}} dt \right]$	$\frac{A(t)}{\gamma_1}$ ★ ■	C	C	$\frac{B(t)}{\gamma_2}$ ★ ■	D	D	C +	D	2
$\frac{f}{a\tau_0} \int_{t_0}^{t_1} (N+G) dt$	$\frac{A(t)}{\gamma_1}$ ★ ■	C	C	$\frac{B(t)}{\gamma_2}$ ★ ■	D	D	C	D	3
$K \log it \left[\frac{1}{aL\tau_0} \int_{t_0}^{t_1} (N+G) dt \right]$	$\frac{A(t)}{\gamma_1}$ ★ ■	C	C	$\frac{B(t)}{\gamma_2}$ ★ ■	D	D	C +	D	4

★ NOT RECOMMENDED
 + RECOMMENDED ONLY WHEN INTEGRAL IS ALWAYS POSITIVE
 ■ NOT AVAILABLE ON PRODUCTION MODEL I

3.5 Three switches in each channel set up the program. They are labeled "mode", "decay constant" and "lin-log". The upper mode switch selects functions a, b, c, and g in that order. A final position, "off", disconnects the integrator from the preamplifiers and function generators to permit independent checks of the circuitry that follows these elements or to prevent large signals from blocking channels that are, for the moment, out of use. In like manner the lower mode switch selects d, e, f, h, and off. The action of the decay constant switch will be discussed in paragraph 3.9. The "lin-log" switch selects function 1 or 3 of Table I in the "lin" position or functions 2 or 4 in the log position.

3.6 The limit clamp (labeled "clamp" on the front panel) and the "output centering pot" must be set to a position dependent on the program to prevent amplifier overload and to position the pen on scale. In the "clamp on" position a diode network restricts the integrator output to values between -100 and 0 volts. In the "clamp off" position the limits are -100 and +100 volts. The later position is used only in the A, B, or AB modes and for tests. The signs of these limits are reversed in the output because of phase reversal in the amplifier that follows the integrator. The output centering control is used with the output scale control (a screwdriver adjustment on the back of the chassis) to adjust the final computer stage so that the pen deflections at the limits of the integrator coincide with the limits of the recording chart. This adjustment is described in paragraph 4.10.

3.7 Besides the output scale factor just mentioned there are three controllable scale constants α , β , and γ in each channel. The output scale factor setting is not treated separately in the equations because it is used only to compensate for pen motor variations in alignment of the computer. The subscripts 1 and 2 of α , β , and γ in the equations of Table I respectively indicate a scale constant in the A channel or the B channel of the computer. α , β , and γ are selected by controls as discussed in the following section 3.8.

3.8 The quantity represented by " α " in the equations is controlled by the knob labeled "Integration Constant". In functions 1 and 2 this controls the averaging time and in functions 3 and 4 determines a scale factor. The dial is indexed in a decibel-like form, i. e., $I = 10 \log \alpha$ where α is one (1) for 0 db. This is most useful for presentation of data in the form

$$\int_{t_0}^{t_1} V^2 dt$$

where V is proportional to the sound pressure at the hydrophone. The integral is then proportional to the plane wave energy received in time t_0 to t_1 . For two different signals V_1 and V_2 which give the same deflection using different dial settings α_1 and α_2 ,

$$\frac{1}{\alpha_1} \int V_1^2 dt = \frac{1}{\alpha_2} \int V_2^2 dt \quad \text{then } 10 \log \frac{\alpha_2}{\alpha_1} = 10 \log \frac{\int V_2^2}{\int V_1^2}$$

or the energy ratio level is the difference in the dial settings expressed in I. This holds for functions 3c and 3f. For the other functions in row 3 it is usually best to convert I to α . Similar comments hold for functions in row 4.

The averaging time τ for functions in row 1 and 2 is determined from α by the equation $\tau = \alpha \beta \tau_0$, where β is numerically 1 or 4, and τ_0 is the basic averaging time of the computer, equal to 0.001 sec.

3.9 The control labeled Decay Constant "1", "off", "4" determines the value of the quantity represented in the equations by β . In the positions "1" and "4" functions 1 or 2 are computed and β is a scale factor the value of which is given in Table II. In these positions the integral is a time average weighted by the factor $e^{\frac{t-t_0}{\tau}}$ where t_0 is the instant at which the integral is read, t is any earlier time, e is the Naperian constant, and τ is the time constant as defined in 3.8. The factor, $e^{\frac{t-t_0}{\tau}}$, approaches zero if t is much earlier than t_0 . Two values of β are required to permit maximum dynamic range of the computer for signals with different peak to root mean square ratios. For sine waves or signals filtered to a narrow frequency band, the Decay Constant "1" position gives a full scale output when the peak level in the preamplifiers and function generators is nearly full scale. For signals like white noise, or wide band exponential pulses, where the peak values are relatively higher, the Decay Constant "4" position gives a similar output level without overload in the preamplifiers and function generators. In the "off" position of the switch, functions 3 and 4 are computed. In this case, all times after t_0 , the initial time when the computer reset was last released, are weighted equally. Any input at a time before t_0 has zero weight in the integral regardless of the position of the decay constant switch.

TABLE II

Constants f , L , and β employed in standard programs of model 45.001 Oceanographic Computer are listed. The given values of β apply only to Function Numbers 1. and 2. which contain β .

FUNCTION

G	1. or 3	2. or 4
a.	$f = 4 \times 10^3$ mm/volt $\beta = 1$ or 4	not recommended
b.	$f = 4 \times 10^3$ mm/volt $\beta = 1$ or 4	$L = 2.9 \times 10^{-4}$ volts $\beta = 1$ or 4
c.	$f = 1.9 \times 10^5$ mm/volt ² $\beta = 1$ or 4	$L = 1.2 \times 10^{-3}$ volts ² $\beta = 1$ or 4
d.	$f = 4 \times 10^3$ mm/volt $\beta = 1$ or 4	not recommended
e.	$f = 4 \times 10^3$ mm/volt $\beta = 1$ or 4	$L = 2.9 \times 10^{-4}$ volts $\beta = 1$ or 4
f.	$f = 1.9 \times 10^5$ mm/volt ² $\beta = 1$ or 4	$L = 1.2 \times 10^{-3}$ volts ² $\beta = 1$ or 4
g.	$f = 9.5 \times 10^4$ mm/volt ² $\beta = 1$ or 4	$L = 1.2 \times 10^{-5}$ volts ² $\beta = 1$ or 4
h.	$f = 1.9 \times 10^5$ mm/volt ² $\beta = 0.5$ or 2	$L = 2.4 \times 10^{-3}$ volts ² $\beta = 0.5$ or 2

3.10 The control labeled "input gain" determines the value of the quantity represented in the equations by γ . This is always a scale factor; $20 \log \gamma = P$, where P is the setting of the attenuator in decibels.

3.11 The controls labeled "noise cancel" are used to set the quantity represented in the equations by N . This is an adjustable steady quantity added to the input of the integrator to cancel system noise, or any small steady state signal. If this control is set properly it is easier to keep the desired computations on scale.

3.12 The quantities represented in the equations by f , L , and K are basic constants of the computer which are fixed for any given program. Values of the basic scale constant f and the basic reference level L in the various programs are given in Table II. The scale constant K of the logarithmic circuits is approximately 2 mm per logit for quadratic type functions and 1 mm per logit for other functions.

3.13 The A and B programs are primarily useful in low frequency seismic measurements. The equation of operation given in Table I is:

$$C = \frac{f\beta}{\tau} \int_{t_0}^{t_1} \left(N + \frac{A(t)}{\gamma_i} \right) e^{\frac{t-t_1}{\tau}} dt$$

where C is the deflection of the pen in millimeters. In this type of operation the pen is set at the center of the chart with the "clamp" off. Within a frequency range bounded at the low end by the low frequency cut-off of the AC preamplifier and at the high end by $1/\tau$, the deflection of the pen is proportional to the instantaneous input potential. That is, within the given frequency band the output as a function of time is a replica of the input sound pressure. If the pens are used as output for this program, there is an additional low pass filtering action by the dynamic characteristics of the pens which prevents reproduction of frequencies above 200 cycles.

3.14 The "ABS VAL" or $|A|$ and $|B|$ programs are primarily useful in functions 2b and 2e which are used to measure in decibels and approximation to input power. These absolute value functions are less accurate representations of input power than the quadratic functions 2c and 2f, but the dynamic range of the presentation obtained without changing control settings is 50 decibels for the absolute value functions while that for the

quadratic functions is 25 decibels. Because the time average in the former functions is computed with a rectified rather than a squared signal there is an error dependent on waveform if the result is interpreted as power. If the signal is confined to a narrow band or is a sine wave and τ is small compared to the period over which a significant change in the signal level occurs, the wave form error will not be serious, and, if a large dynamic range is wanted, it is desirable to use an absolute value program. The equation of function 2b is

$$C = K \log_{10} \left[\frac{\beta}{L\tau} \int_{t_0}^{t_1} \left(N + \frac{|A(t)|}{\gamma_1} \right) e^{\frac{t-t_1}{\tau}} dt \right]$$

where logit stands for $10 \log_{10}$

3.15 The "A²" and "B²" programs are used to compute either power or total energy of all forms of acoustic waves in order to measure approximate Fourier transfer functions as discussed in Section I, and paragraphs 3.8 and 3.9. Function 3c is the one used by Officer and Dietz (1953) to measure approximate transfer functions by means of the total energy received in each of several different spectral bands. The equation is

$$C = \frac{f}{\alpha \tau_0} \int_{t_0}^{t_1} \left(N + \frac{A^2(t)}{\gamma_1^2} \right) dt$$

Function 3f is the same function in the B-channel of the computer. The shape of this function for a succession of exponential pulses is shown in Figure 17 in which time is plotted from left to right and deflection, which, on the computer Sanborn trace, is proportional to energy, is plotted from bottom to top. This figure is used in Section IV to illustrate a calibration method. Despite the fact that the computer must frequently be reset to t_0 to prevent overload from the continuous accumulation of received energy, functions 3c and 3f are very frequently used for hydroacoustic measurements with pulse signals because the integral over any desired time interval is easily read from this presentation. t_0 is out of the field of view to the left in Figure 17 but the integral from t_0 to any time t is the height of the trace at t . The integral between any times t_1 , and t_2 is the difference between the heights of the trace at these times.

3.16 Function 1c measures power averaged over a time interval τ as explained in paragraphs 3.8 and 3.9. The equation is:

$$C = \frac{f\beta}{\tau} \int_{t_0}^{t_1} \left(N + \frac{A^2(t)}{\gamma_1^2} \right) e^{\frac{t-t_1}{\tau}} dt$$

This function is useful for measuring transfer functions for steady state or slowly varying signals, or for pulses in cases where it is more inconvenient to reset t_0 in function 3c than it is to interpret power measurements. The area under the curve of function 1c may be used in the same way that the deflection of function 3c is used to compute transfer functions for pulses. It is easier to compute transfer functions of separate arrivals if τ is short enough to allow the response from one arrival to decay before the onset of the next.

3.17 Function 2c presents power in decibels over a 25-decibel dynamic range. Power is averaged over a time τ in the same manner as it is for function 1c. The equation is:

$$C = K \log_{10} \left[\frac{\beta}{L\tau} \int_{t_0}^{t_1} \left(N + \frac{A^2(t)}{\gamma_1^2} \right) e^{\frac{t-t_1}{\tau}} dt \right]$$

Computation of total energy from the trace is more difficult than for function 1c but the logarithmic scale is advantageous for measurements with a single setting of signals covering a large dynamic range.

3.18 Function 4c presents energy in decibels over a 25-decibel range. This function is useful to present, with a wide dynamic range, the total energy accumulated between t_0 and t provided that the noise accumulated during this period is negligible. Because the scale is logarithmic, it is difficult to interpret partial integrals from times later than t_0 in terms of energy of individual arrivals or arrival groups. Because any noise accumulated before the signal forces one to make an interpretation of a partial integral and drastically decreases the dynamic range, it is necessary to reset the integrator just before the signal begins. This is very difficult unless some form of automatic reset is employed.

3.19 The AB and $(A^2 + B^2)$ programs are designed to permit computations of the correlation between two signals. An example of the results that might be obtained are as follows: Given two signals A and B, $\int AB dt / \int (A^2 + B^2) dt$ is a measure of their correlation. If a time delay of either of the signals is provided the correlation is a function of this time delay. Auto-correlation may be studied if the two signals differ only by the variable time delay. In the common situation where one wishes to study multipath transmission using a sound source having an output that is a somewhat complicated function of time (such as an explosion and its bubble pulses) one might perhaps use a replica of the source signal in one channel and the received pulses in the other. The time delays required to produce maxima in the correlation function are the travel times of the separate paths.

3.20 The signal flow in the block diagram Figure 6 is as follows: After passing through an attenuator, the signal in each channel is amplified by an AC preamplifier whose frequency response is given in Figure 7. Two monitor lines not shown in the diagram connect to the output of each preamplifier. One monitor line for each preamplifier is connected to the limit and storage indicator as described in paragraph 3.21; the other is connected to the high impedance monitor outputs described in paragraph 3.2. In addition, the preamplifier outputs are connected to various diode function generators and to a resistor. The appropriate diode function generator or resistor is selected by the mode switch and applied to a D. C. amplifier used as an integrator. Figure 8 shows the frequency response of mode A^2 or B^2 . Figure 9 shows the frequency response of mode $|A|$ or $|B|$. Figure 10 shows the frequency response of mode AB . Figure 11 shows the frequency response of mode $A^2 + B^2$. The output of the integrator is connected to a logarithmic diode function generator and to a resistor. Of the two, the appropriate element, to serve as input to the final DC amplifier output driver, is selected by the lin-log switch. The output current is adjusted to the requirements of the readout pen by a variable resistor.

3.21 The limit and storage indicator shown in Figure 4 is used to monitor the output of the preamplifiers which is also the input to all diode function generators except the logarithm. If the computation is to be accurate it is necessary to set the input attenuators so that all significant signal potentials lie within a certain range of levels. This range is about 30 decibels for other modes A^2 , B^2 , AB , and $A^2 + B^2$ and about 50 decibels for other modes. The narrower range is the most important since these functions are used frequently. Since the input signal is a function of time and the low level portions may be important if their duration is long enough, it is necessary to

set the highest peak of the signal near the top of the 30-decibel range of the diode function generators. The limit and storage indicator is designed to measure and store an indication of the highest peak level. This indication may be consulted by the operator at any time after the peak has passed. Since it is recommended that the highest peak be between full scale of the diode function generators and 10 db below full scale, the limit and storage indicator has a 10-decibel range. If the peak potential was in the proper range the minimum lights are lighted and the maximum lights are not. If the peak potential was higher than it should have been both lights are lighted. If the peak potential was lower than it should have been neither light is lighted. The indicator is reset before each measurement by means of a front panel push button. The frequency response of the limit and storage indicator is shown in Figure 12.

3.22 For many waveforms deterioration in accuracy of the computed result is not serious until the peak is more than 6 db higher than the nominal full scale of the diode function generators or 10 db lower. The narrow range of the indicator is intended to serve as a timely warning of a change in optimum conditions. It does not represent the actual dynamic range of the computer.

IV. INSTALLATION AND TESTING OF THE ELECTRONIC ASSOCIATES EIGHT-CHANNEL OCEANOGRAPHIC COMPUTER GROUP 4.038

4.1 This computer requires 11 amperes of regulated 117 volt, 60 cycle power. Frequency variations from 58 to 62 cycles may be tolerated but power line potential changes can cause serious integrator drifts and potentials less than 110 volts do not give a high enough heater potential to some of the tubes. A type EM4106, 6.0KVA Stabaline voltage regulator, manufactured by Superior Electric Company of Bristol, Connecticut has been successfully used for shipboard operation from ships AC power.

4.2 Power and signal cables should be made up for the computer, and an appropriate indicating device should be provided for reading or recording the output. At WHOI a Sanborn eight-channel recorder is used for permanent records of analyses, while a panel with eight galvanometers (shown in Fig. 13) is used for immediate readout. Both may be connected at the same time without affecting the operation of the computer.

4.3 To connect the output to a Sanborn eight-channel recorder, the cables and connectors shown in Figure 14 are used. The octal plug and socket are wired as an adapter which is plugged into the pen motor power supply socket.

The octal plug of the pen motor can then be inserted in the adapter. The input cables used at WHOI are shown in Figure 15. For convenience, General Radio plugs are used.

4.4 After connecting all the inputs, all the outputs and the power, the computer is ready to be tested. The master power switch is located behind the bottom front panel of the rack. In addition, each power supply has a switch for filaments and plates. All switches should be turned on. A warm-up period of at least one-half hour is advisable before testing or use. At WHOI, the computers are left with the filament power on over nights and weekends. For longer idle periods, it may be advisable to turn everything off.

4.5 During the initial warm-up time, the voltages of all the power supplies should be checked and adjusted. This is accomplished by setting the meter switch on the front of the power supply to each of the five ranges and adjusting the corresponding pot (all of which are located on the front of the power supply) so that the meter reads exactly on the red line. Much greater accuracy can be obtained if an external precision meter is used for these adjustments, but the panel meter can be used if this degree of accuracy is not required. The +100 and -100 volt adjustments are the most critical since they are used as references for the diode function generators.

4.6 If the reading is much lower than that necessary for proper operation, and cannot be brought up from this value by means of the adjustment, it probably indicates a high resistance short to ground either within the power supply or in the corresponding computer unit. Extended operation of this nature damages the power supply and perhaps components of the computer due to heavy current flow. Therefore, if the power supply meter indicates very low, the power supply should be turned off until the trouble is discovered and corrected.

4.7 Often some ranges will not quite make the red line, but operation is usually satisfactory if these ranges fall within the green region. The procedure should be performed on each of the four power supplies.

4.8 After a suitable warm-up period, preliminary tests and adjustments can be made as follows:

For channel 1-A, the controls should be set:

Mode: off
Integration constant: 15 db.
Decay constant: on
Clamp: off
Lin-Log: linear
Noise cancel pots: coarse and fine set at 500.

Depress the amplifier 2 balance button and balance, with the potentiometer directly below it, so that the corresponding overload light goes out. This operation is not critical, and it is normal to make an approximate setting, as it is almost impossible to make the overload indicator stay out while the balance button is depressed. After the button is released, the light should go out fairly quickly if the amplifier is properly balanced. Perform the same operation with the amplifier 1 balance.

4.9 With the computer reset button (on the limit and storage indicator) depressed, adjust the output centering pot so that the recorder pen (or readout meter) is deflected to the extreme low side of the scale (about -50 volts). This driver output (the zero point) will represent zero volts at the integrator output. Release the reset button. If the pen moves to a new deflection, adjust the coarse noise cancel pot until there is no motion as the reset is pressed and released. Turn the decay constant off, and adjust the coarse noise cancel pot until the recorder pen ceases drifting. If the pen drifts off scale before this is accomplished, the pen can be returned to the zero point by pushing the computer reset button. After having made this adjustment set the integration constant to 6 db and adjust the fine noise cancel pot so that the pen drifts across the entire scale in the positive direction in about a second. After adjusting the fine noise cancel pot turn the "clamp" switch on. The "clamp" circuit is a negative limit which prevents the integrator output from becoming positive. This prevents the pen from drifting beyond the zero point in a negative direction.

4.10 In the rear of the computer, there is access to two scale factor pots for each computer 45.001. These are labelled A and B. For channel 1-A, adjust the A pot of the top pair. This controls the width of the full scale as recorded on the output indicator. Using this control, and the output centering pot on the front panel, adjust the computer so that deflection is at maximum on full scale of the readout indicator, and zero when the computer is being reset (button on the limit and storage indicator). Figure 16 illustrates how a Sanborn record would appear during such an adjustment. This procedure should

be applied to all eight channels of the computer. Once the scale factor pots in the rear are properly adjusted, they should require no further attention.

4.11 For accurate operation, the limit and storage indicator lights should be adjusted. For this, and for most other adjustments and tests, an accurate oscilloscope and signal generator capable of producing a good sine wave are necessary. In the analysis laboratory at WHOI, a Hewlett-Packard model 130 BR oscilloscope and a Hewlett-Packard model 202D signal generator are used.

4.12 There are eight monitor points located near the input and output jacks on the rear of the computer group. These monitor points are standard phone jacks and are connected to the output of the AC preamplifiers. Each jack corresponds to a channel. Using any channel, connect the signal generator to the input, and the scope to the monitor jack. Apply about a volt at 1000 cycles to the input of the channel with the attenuator (input gain knob on the front panel) for that channel set to 0 db. Then, decrease the attenuation until three to six volts peak to peak appears at the monitor point (as measured with the scope). Then adjust the signal generator to obtain exactly 6.0 volts peak to peak at the monitor point.

4.13 The diode function generators are designed for a 10-volt peak signal. However, signals up to nearly 30 volts peak are computed with little error. The main source of error in high level signals is overload in the preamplifiers between 25 and 30 volts peak. The monitor point is on the secondary of a transformer with a turns ratio of 4.7:1. Therefore the voltage applied to the function generators is 4.7 times the monitor voltage. 6.0 volts peak to peak is 3.0 volts peak. $3.0 \times 4.7 = 14.1$ volts which is well within the range of accurate computation.

4.14 Adjust the signal indicator level pot on the limit and storage indicator so that the level is barely sufficient to operate the maximum light for the channel you are using. Note that once a light has lit it will remain lit after the signal is removed (or diminished) until the indicator reset button is pushed. The lights are designed to remain lit so that levels can be read after the operation is complete. If the lights did not remain lit, it would be impossible to note which lights fired during rapid operation. All channels should indicate the same on the limit and storage indicator for the same monitor point voltage. Normally, the minimum lights fire at inputs about 10 db. below the firing voltage of the maximum lights. Once the signal indicator level has been set, the lock down nut should be tightened; the control will not need readjustment.

4.15 This completes the preliminary testing and adjustment of the computer. In order to obtain accurate results in analysis, however, the computers may be thoroughly tested and the adjustments of all modules can be perfected. Some special equipment is necessary for most of these adjustments and measurements. Among the most necessary items for these fine internal adjustments and measurements are, in addition to the signal generator and scope, a multimeter (10,000 ohm/v.) or VTVM, and a service shelf, supplied for operation of computer units outside of the rack. The service shelf is supplied by the manufacturer. For specific tests and adjustments, special adapters are necessary for the operation of certain modules external to the computer unit. The following portion of this report will show, on a per-channel basis, the method of adjusting the computer. The procedure should be followed for each of the eight channels. The reader is referred to the book of schematics, assembly and wiring diagrams supplied by the manufacturer.

4.16 The AC preamplifier should be adjusted both for gain and for DC balance of the output stages. Set the input attenuator to 0. With a multimeter, measure the voltage between terminals two and four of the preamplifier monitor transformer (T1 for channel A, T2 for channel B). Then, adjust trimpot R23 (on the preamplifier card (6.185)) so that the voltage between terminals two and four is zero as indicated on the most sensitive range of the instrument. This completes the DC balance of the output stages of the preamplifier.

4.17 Connect the signal generator to the input of the channel being tested, and adjust it for exactly 8.0 volts peak to peak at 1000 cycles. Set the input attenuator to 54 db. and measure the voltage across R82 (R84 for channel E). Adjust trimpot R12 for voltage of exactly 20 volts peak across R82 (R84). Likewise, measure the voltage across R83 (R85 for channel B) and adjust trimpot R13 for exactly 20 volts peak to peak across the resistor. Repeat the last two adjustments for highest accuracy. These trimpots are found on the front panel in Model II computers on the preamplifier card in Model I.

4.18 The frequency response of the AC preamplifier can be checked by observing the voltage at the monitor point for an input voltage of constant peak to peak value as the frequency of the input voltage is varied. The response curve should be similar to the one shown in Figure 7.

4.19 To test the operation of the DC operational amplifiers set up the controls as follows:

Input attenuator: 54 db.
Mode: off
Noise cancel pots: both coarse and fine at 500
Integration constant: off
Decay constant: $\beta = 1$
Lin-Log: linear

Balance the two amplifiers of the channel being evaluated. The other channel of the same computer module should be set up similarly to the channel being measured and care taken that the DC amplifiers of neither channel are in overload. There is a slight interaction between channels when one channel overloads due to the inability of the regulated power supply to hold the voltage perfectly constant.

4. 20 Holding the reset button in, zero the output with the output centering pot. Release the reset and adjust the coarse noise cancel pot until the output is zeroed. Frequently operation with the integration constant off will cause the first DC amplifier to oscillate slightly. This instability can be overcome by setting the integration constant to 0 db. , or even 3 db. in stubborn cases. This will effect the time constant of the integrator. . . . See discussion on constants, para. 4. 33. This slight instability is not serious and is not considered abnormal. Set the mode switch to the absolute value position and rebalance the noise cancel pot. Then, using a 100-cycle signal sufficient to light the minimum indicator light, observe the wave-form measured between the output pin and ground, with the scope. The output for channel A is on pin three of J9, that of channel B is on pin three of J11. The signal should appear as a full wave rectified sine wave. Other than the rectification, there should be no distortion apparent. On the readout device, such as a Sanborn recorder, the signal should appear similar to that observed with the scope, at the point just measured. This checks the linearity of both DC amplifiers.

4. 21 In the evaluation of the α^2 Fixed Diode Function Generator (FDFG), three methods are available. All three will be described, and theoretically, each should be accurate over the total range of operation. Practically, the first method described (with the decay constant on) is most convenient, but least accurate on the low signal end. The slope method (measuring the time it takes to reach full scale deflection), and the pulse generator method are accurate over the entire range of operation. But values of I will have to be changed during the run so as to keep the deflections within measureable limits.

4. 22 The first, most convenient method, is performed as follows:

Set the input attenuator to 30 db. Set up the other switches as follows:

Mode: square
Integration constant = 9 db
Decay constant = ∞ = 1
Noise cancel = adjusted for zero drift
Clamp = on
Lin-Log = linear

Apply a sine wave signal to the input (1000 cycles is convenient) and adjust the signal generator to yield 6.0 volts peak to peak at the monitor point of the channel being used. The readout indicator should be deflected to full scale. Now, increase the input attenuation by 3-db steps, and observe the readout deflection, which should halve each time the attenuator is advanced one step. By the time an attenuation of 27 db is reached, it becomes difficult to measure how close to half the output deflection is from the preceding step.

4. 23 The slope method is as follows: turn the decay constant off and pick a value of I suitable for an easily observed deflection rate. Each time the input attenuator is advanced a step it should take twice as long for the output deflection to run from zero (when reset button is released) to maximum. That is, there will be a ratio of one-half between the slope of the trace on one setting and that on the next larger setting. Each time the input attenuator is reduced a step there should be a ratio of two. A change in the integration constant causes a similar change but does not check the X^2 function or cause any change in the signal level through the X^2 FDFG. Therefore, the integration constant may be set as desired to facilitate measurement of this ratio. The X^2 Fixed Diode Function Generator should operate properly down to 24 db below 6.0 volts peak to peak on the monitor point, and up to 6 db above this reference level.

4. 24 The pulse method is as follows: A repetitive pulse having identical positive and negative excursions which follow each other at a shorter interval than the pulse repetition rate is connected to the input of the computer and adjusted to a level such that with the input attenuator at any convenient position (for example 30 db) the peak to peak potential of the pulse is 6.0 volts as measured at the monitor point. The decay constant should be off. The integration constant may be set at any convenient point. The output deflection

should, after the computer is reset and then allowed to operate, jump up in double steps (see Fig. 17). The input pulse used in Figure 17 is one that we have found especially useful since it is similar in wave form to the basic component of explosive source outputs. It will be described in more detail below. However, any short pulse having equal positive and negative steps, spaced about as shown may be used. Over the range of operation of the X^2 diode function generator any 3-db decrease in the input attenuation should cause the height of the steps to double and any 3-db increase in the input attenuation should cause the height of the steps to halve. As in the previous method, a change in the integration constant causes a similar change but does not change the signal level in the X^2 FDFG. Accordingly, the integration constant may be changed in the same manner as above. One should note whether the steps caused by the positive and negative pulse are equal. A difference between them usually indicates overload in the preamplifier caused by too high a level, a defective tube, or by a poor adjustment of the gain and balance pots.

4.25 The pulse we prefer for the above test is generated in the following manner: A Tectronix 161 pulse generator and a Tectronix 162 waveform generator are connected to produce a rectangular pulse with a repetition rate of 1 pulse per second and a pulse width of 10^{-1} sec. The output is differentiated by a resistive capacitive network with a time constant of 10^{-3} sec. followed by a cathode follower. The resulting signal is a double pulse having a positive exponential part followed by an identical negative part as in Figure 17.

4.26 The quarter square FDFG is evaluated by similar techniques to those described for the X^2 FDFG. The difference is that both channel inputs are involved. The computer unit should be set up as follows:

Channel A controls

Mode = AB
Integration constant = 9 db
Decay constant β = 1
Clamp = off
Lin-Log = linear
Centering pot adjusted for zero at center of scale

Channel B controls

Mode = $A^2 + B^2$

Integration constant = 9 db

Decay constant = β = 1

Clamp = on

Lin-Log = linear

Centering potentiometer adjusted for zero at lower edge of scale

All noise cancel pots should be adjusted for zero drift before decay constants are switched on.

4. 27 First, set both input attenuators to 30 db and adjust the signal generator so that 6.0 volts peak to peak appears at the monitor point of both channels. The channel B readout should now indicate full scale deflection. Holding channel A at 30 db., vary the attenuation on channel B. Then, holding channel B at 30 db, vary the attenuation on channel A. The results from the channel A readout should be the same as those obtained from the X^2 test with the single exception that here, a change of 6 db (2 steps of the attenuator) corresponds to 3 db of the X^2 test.

4. 28 It will be noticed that if a strong signal is fed into one input while none exists in the other, the readout will indicate slightly negative. This is normal, and is acceptable if the deflection so obtained is less than five percent of full scale. It should be understood that on both the $A^2 + B^2$ operation and the AB operation, the scale factor of the readout is automatically halved, that is, the readout scale corresponds to twice the A^2 or B^2 ranges. An example is given: if an input to channels A and B is equal, the readout deflection of the AB operation will be half that of the A^2 or B^2 operation, while the deflection of the $A^2 + B^2$ operation will be the same as that for the A^2 or the B^2 operation.

4. 29 In the $A^2 + B^2$ mode, the halving of the scale factor occurs only when the decay constant is on, i. e., β = 1 or β = 4 position. For details of the changes in scale factor with different control settings see Table II.

4. 30 To test the accuracy of the multiplier circuit with respect to phase, a ~~test~~ device capable of supplying 2 sine waves with a variable, controlled phase difference is necessary. The WHOI analysis laboratory uses a device, specially designed and built, which will operate at spot frequencies over the entire frequency range of the computer. This device, which will give any phase difference from 0 to 2π , is described in Appendix A.

4. 31 In tests of the multiplier circuit, the zero reference of the readout should be center scale (set by adjusting the output centering pot). 6.0 volts should appear on both monitor points while the inputs are fed by the two sine waves. Then, the output of the computer is observed as the phase relationship of the two inputs is varied. The output should vary by the equation: $D = D_0 \cos \phi$, where D is the deflection of the readout, D_0 is the deflection of the readout with $\phi = 0$, and ϕ is the phase difference between the two sine waves. Care should be taken in this test to insure that the peak values of the sine waves are constant for this set of measurements.

4. 32 Testing the Log X FDFG is less complex than either of the previous two tests. Set up the channel to be tested in the following manner:

Input attenuator: 15 db.
Mode: abs. val.
Noise cancel pots: adjusted for zero drift
Integration constant: 9 db
Decay constant: on
Clamp: on
Lin-Log: log

4. 33 Adjust the signal generator to provide 6.0 volts at the monitor point. Then, set the input attenuator control to ∞ . Then, decrease the attenuator setting, 3 db at a time, and observe the readout deflection. Over the operating range of the computer, the increase in deflection due to the reduction of attenuation should occur in equal steps. It is often found that the steps deviate near the ends of the scale. This is permissible provided that five percent accuracy can be maintained.

4. 34 If any one of the FDFG circuits does not operate within the desired tolerance, adjustments can be made, as recommended in the latest issue of the instructions presented by the manufacturer.

4. 35 At this point, all of the operations of the computer have been tested and adjusted. In order to significantly evaluate data processed by this machine, one need only know the constants of operation. The remainder of this report is devoted to the evaluation of these constants. It should be borne in mind that the computer should be in perfect operating condition before evaluation of these constants is made: viz, the previous procedures should be observed prior to the final systems check.

4.36 There is not much difference between the same constants of different 45.001 computers. Only if exact values are desired is it necessary to perform the following evaluations. A list of average values of constants in present production computers has been given in Part III of this report. In the following measurements unless otherwise specified the lin-log switch is to be in the linear position.

4.37 α is determined by the equation $10 \log_{10} \alpha = I$ or $\alpha = \text{antilog}(I/10)$, where I is the integration constant setting in db.

4.38 β , the decay constant, is determined by the feedback and input resistances of the integration amplifier. The defining equation for τ is used to measure β :

$$\tau_{\alpha} = \alpha \beta \tau_0 \quad \text{or} \quad \beta = \tau_{\alpha} / \alpha \tau_0$$

The quantities τ_{α} and $\alpha \tau_0$ are measured directly as follows: For τ_{α} , with $\beta = 1$, a steady signal of a magnitude such that the computer output approaches an asymptotic value of nine-tenths of full scale is applied. Nine-tenths of full scale is 45 mm on a Sanborn recorder. The time required for the computer to integrate from zero deflection to $(1 - e^{-1})$ times the asymptotic value is τ_{α} . $(1 - e^{-1})$ 45 mm is 28.4 mm. For τ_0 , $\alpha \tau_0$ is measured as follows and the result is divided by α : $\alpha \tau_0$ is measured with the decay constant off and the mode switch in the A^2 or B^2 position. With a 1000-cycle input signal, a setting of P such that 6.0 volts peak to peak is shown at the monitor point, and the integration constant I set to 30 db, $\alpha \tau_0$ is the time it takes the computer to integrate from zero to full scale. Noise cancel potentiometers must, of course, have been set to zero drift before the measurement begins.

4.39 γ is determined by the equation $20 \log_{10} \gamma = P$ or $\gamma = \text{antilog}(P/20)$, where P is the input attenuator setting.

4.40 f is evaluated with the decay constant on, i. e. $\beta = 1$ or $\beta = 4$. For the absolute value mode $f = D \left[\frac{\text{antilog}(P/20)}{V} \right]^{\beta}$

mm/volt where D is the deflection of the readout in mm and V is the rms potential of a sine wave input. For the quadratic modes

$$f = D \left[\frac{\text{antilog}(P/20)}{V} \right]^2$$

mm/volt² where D and V are as above.

4.41 K is evaluated by applying a constant input to the computer with the lin-log switch in the log position and $\beta = 1$. The input attenuator is switched as a logit reference while the output is observed. The output will increase in approximately equal steps as the attenuation is decreased a step at a time. There will be approximately eight of these steps in the interval from zero scale to full scale in the A^2 or B^2 modes and approximately 16 such steps over the same scale in the absolute value modes. The value of K is the quotient of the average size of a step in mm. by the decibel change of the input attenuator per step.

4.42 The basic reference level L is measured with the lin-log switch in the log position. L is the value of rms input potential for which the deflection of the pen is at the lower edge of the chart. The value obtained for the quadratic mode of operation is different from that obtained for the absolute value. For the absolute value measurement it is necessary to use a sine-wave signal.

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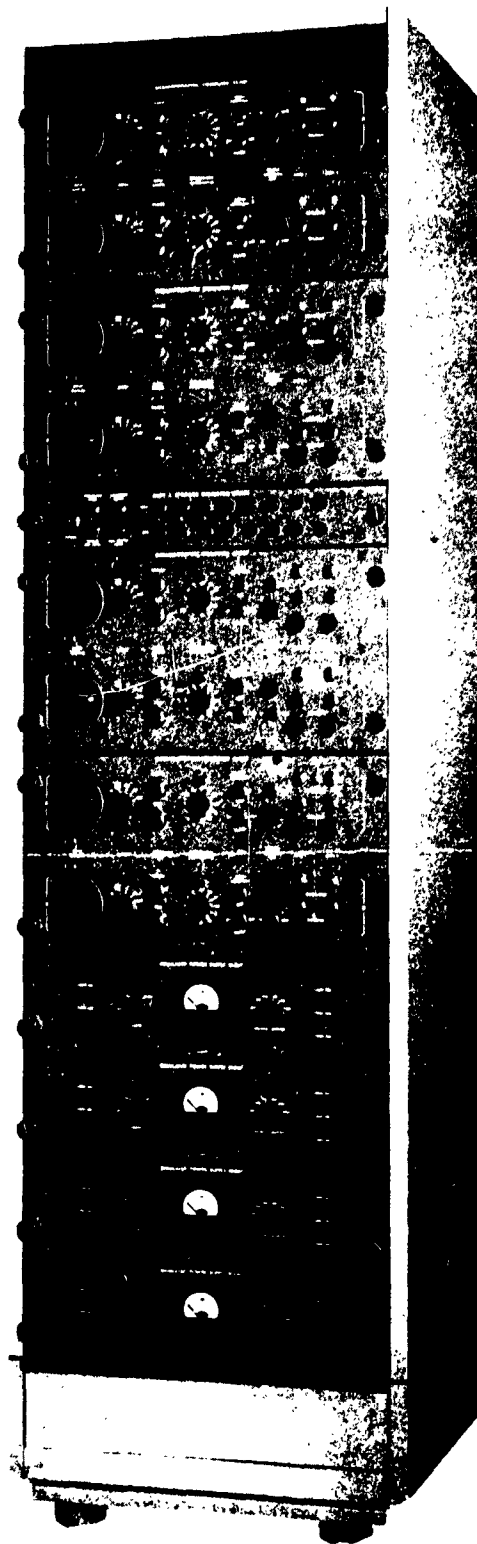


Figure 1 Oceanographic Computer Group mounted in rack.

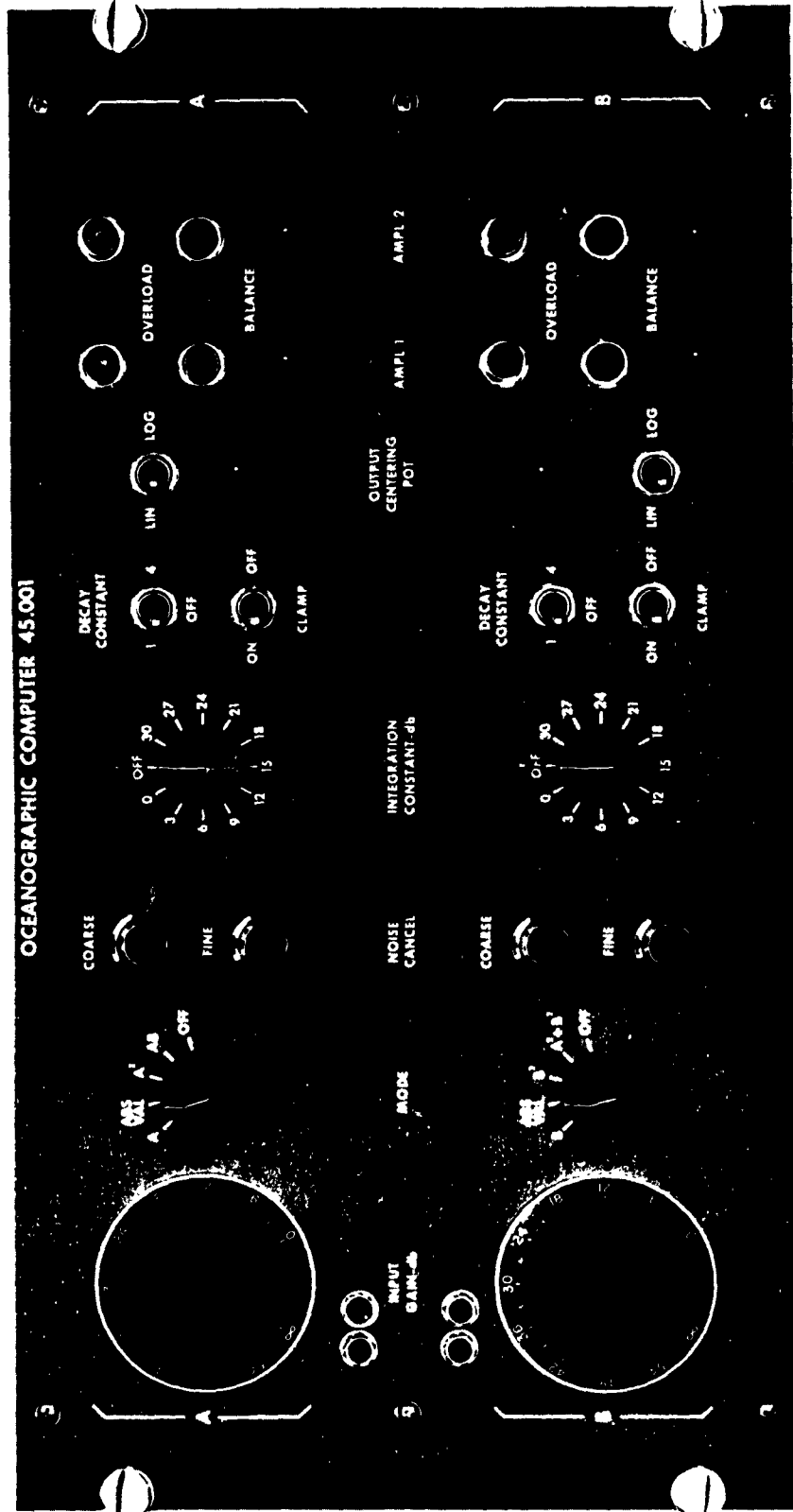


Figure 2 Dual Channel Oceanographic Computer Module #45.001.

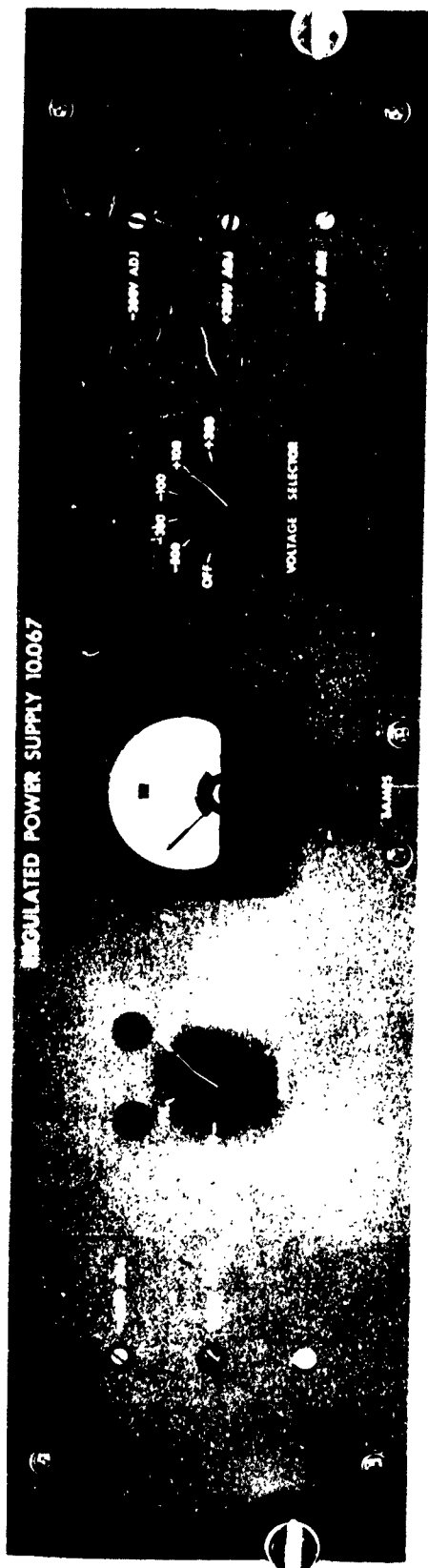


Figure 3 Regulated Power Supply #10.067.

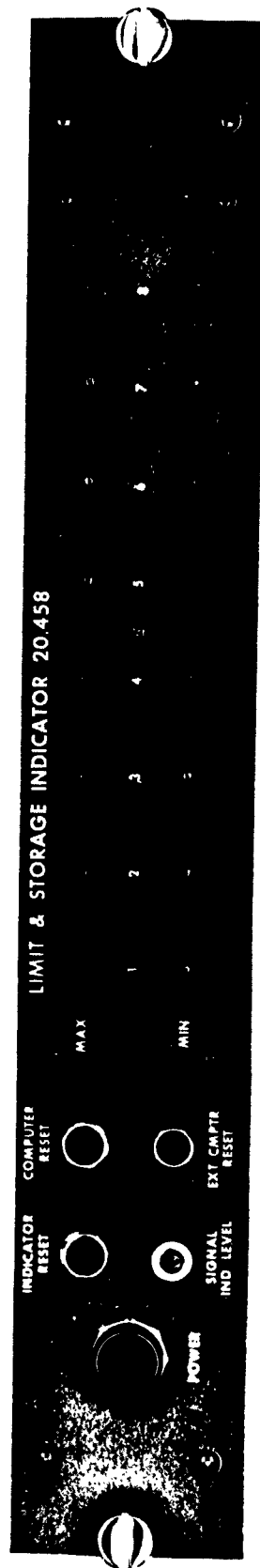


Figure 4 Limit and Storage Indicator #20.458.

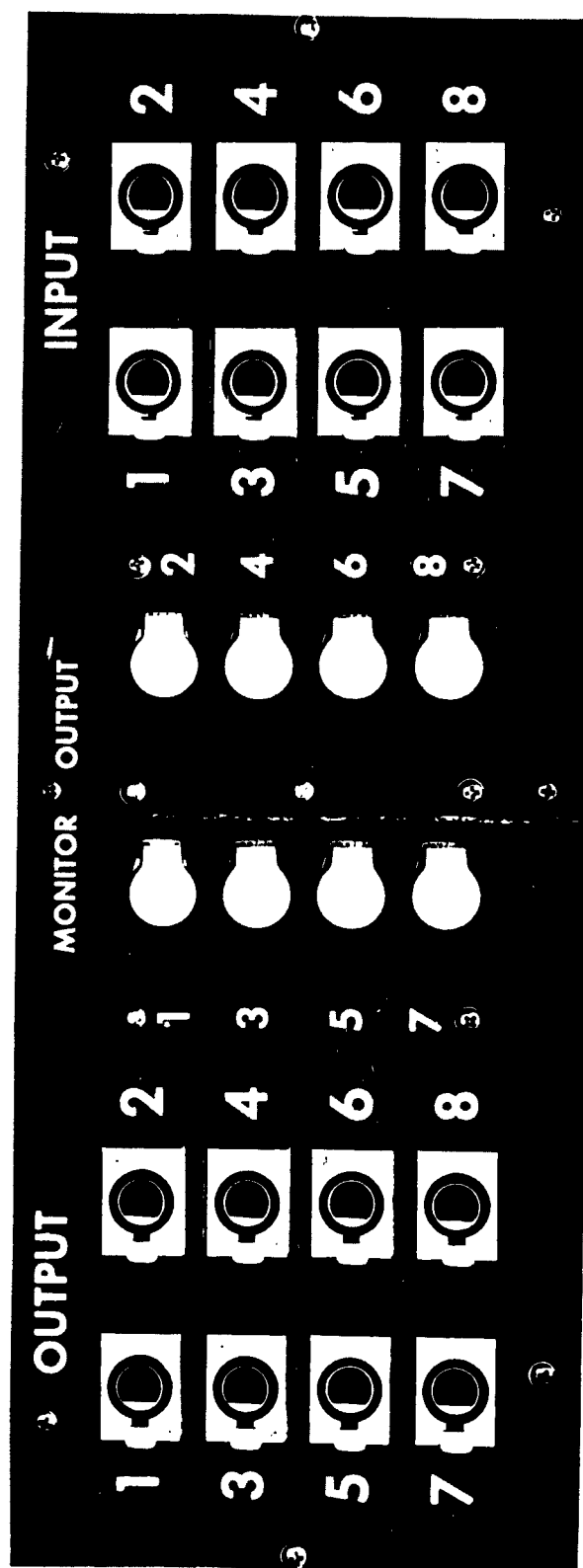


Figure 5 Connection panel for Oceanographic Computer Group.

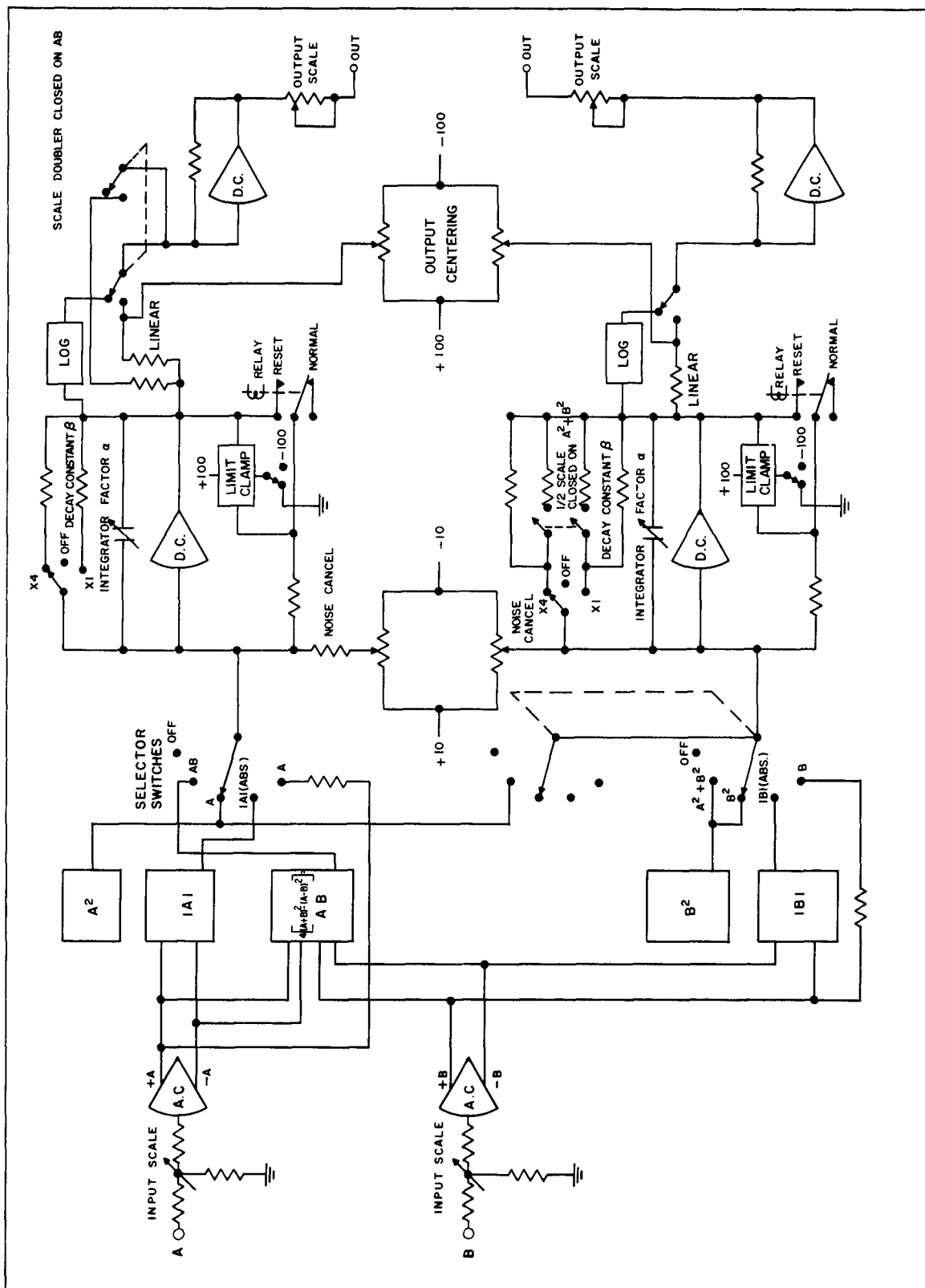


Figure 6 Block diagram of Oceanographic Computer Module.

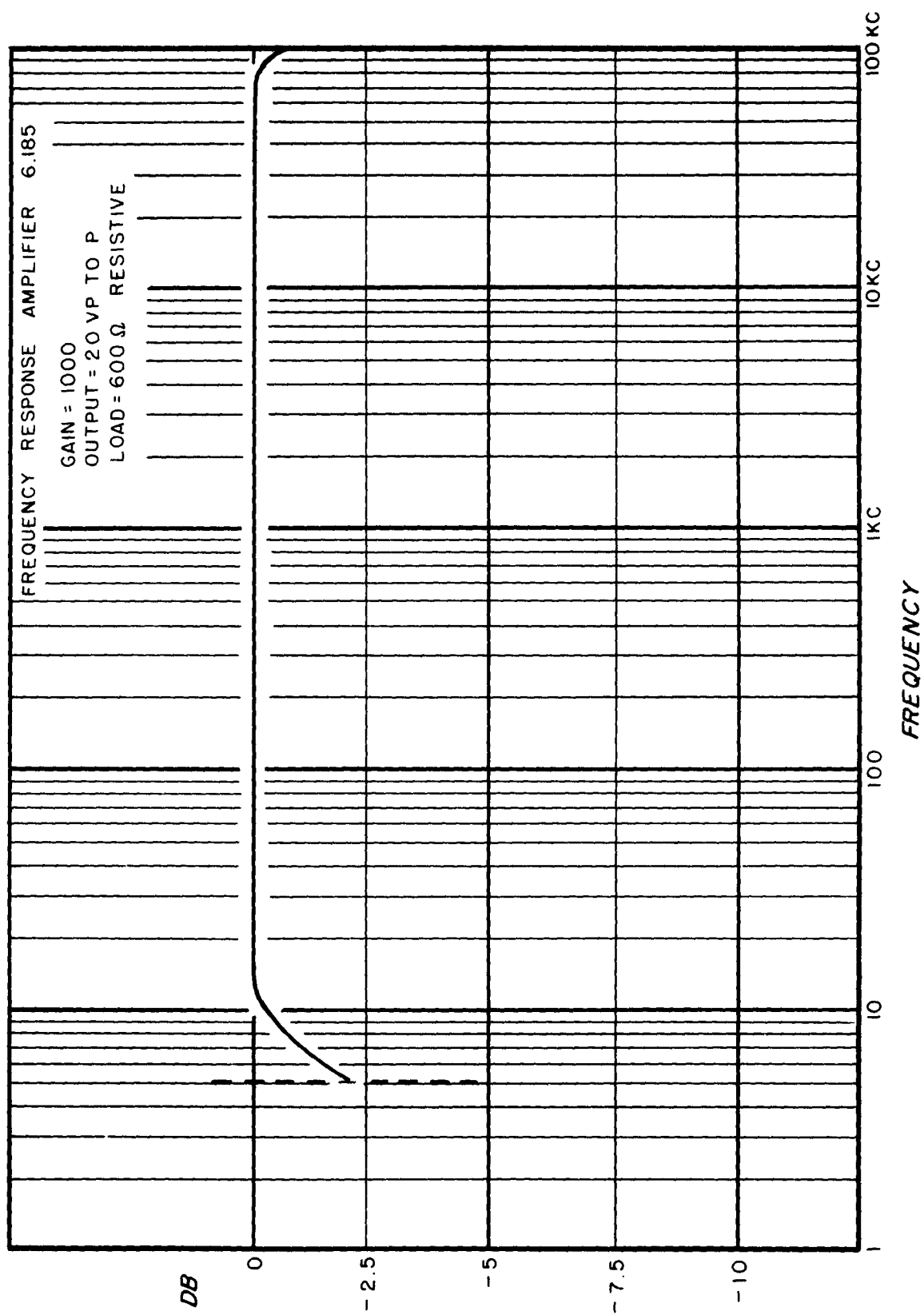


Figure 7 Frequency response of AC Preamplifier.

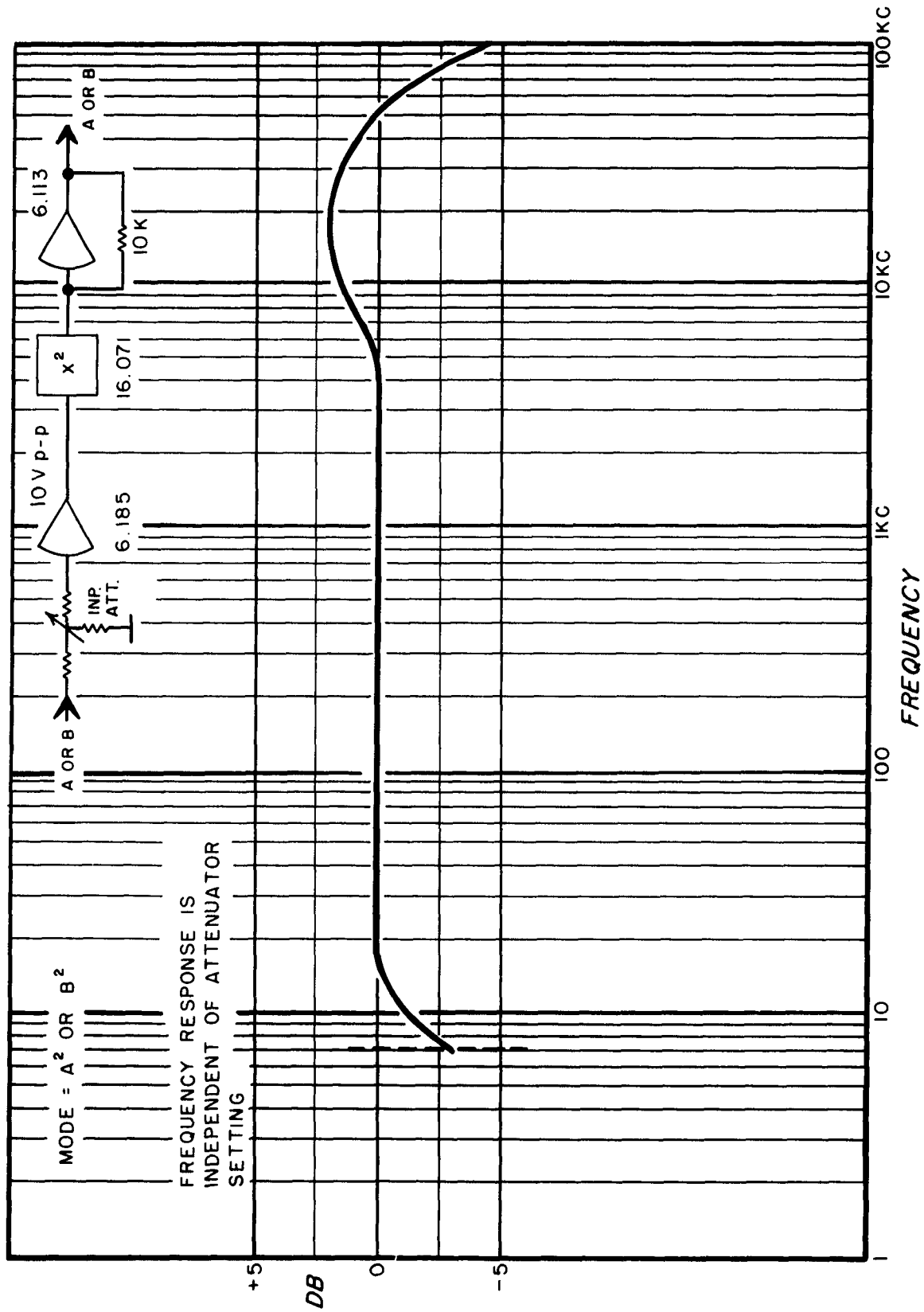


Figure 8 Frequency response of Mode A^2 or B^2 .

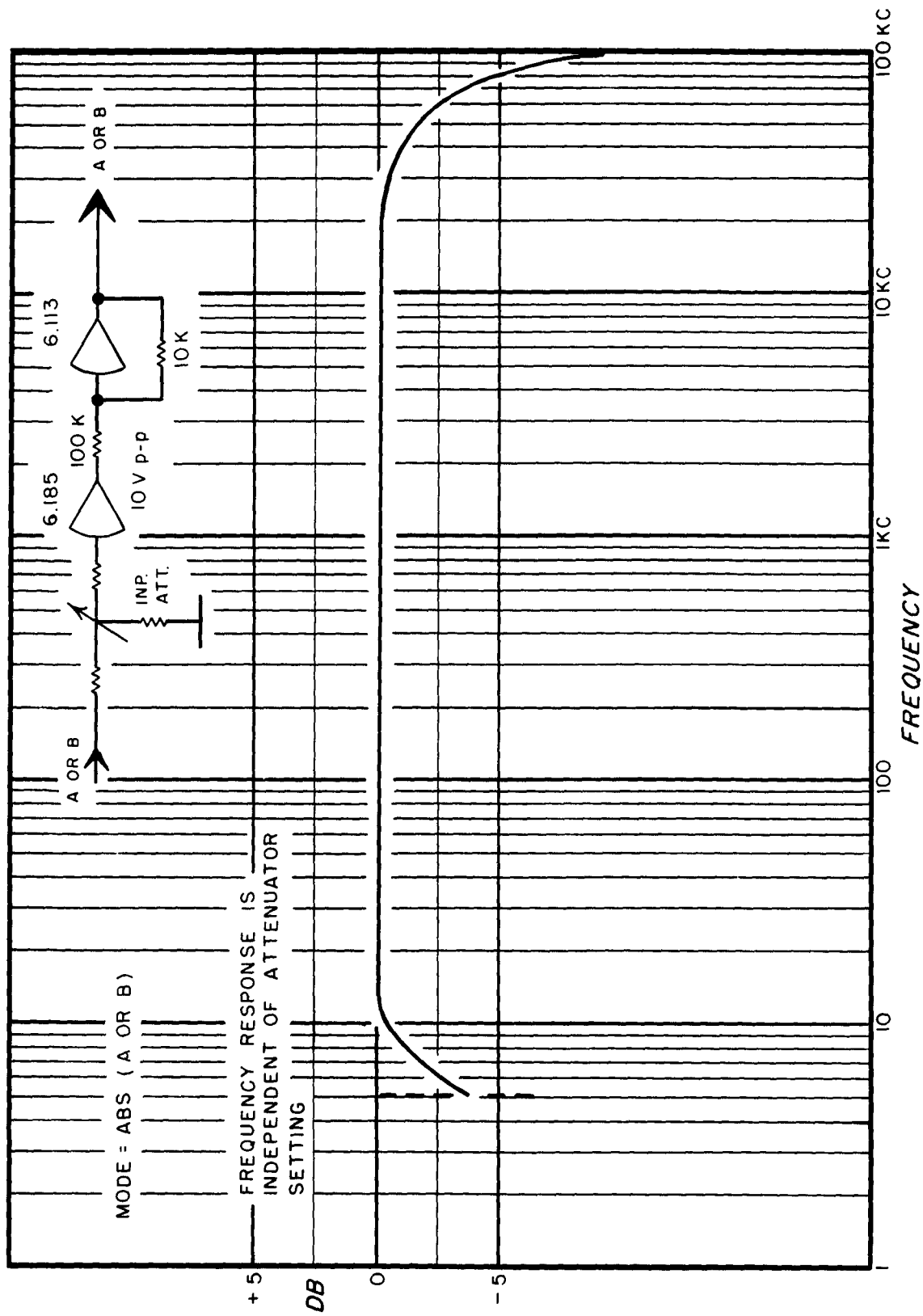


Figure 9 Frequency response of Mode |A| or |B|.

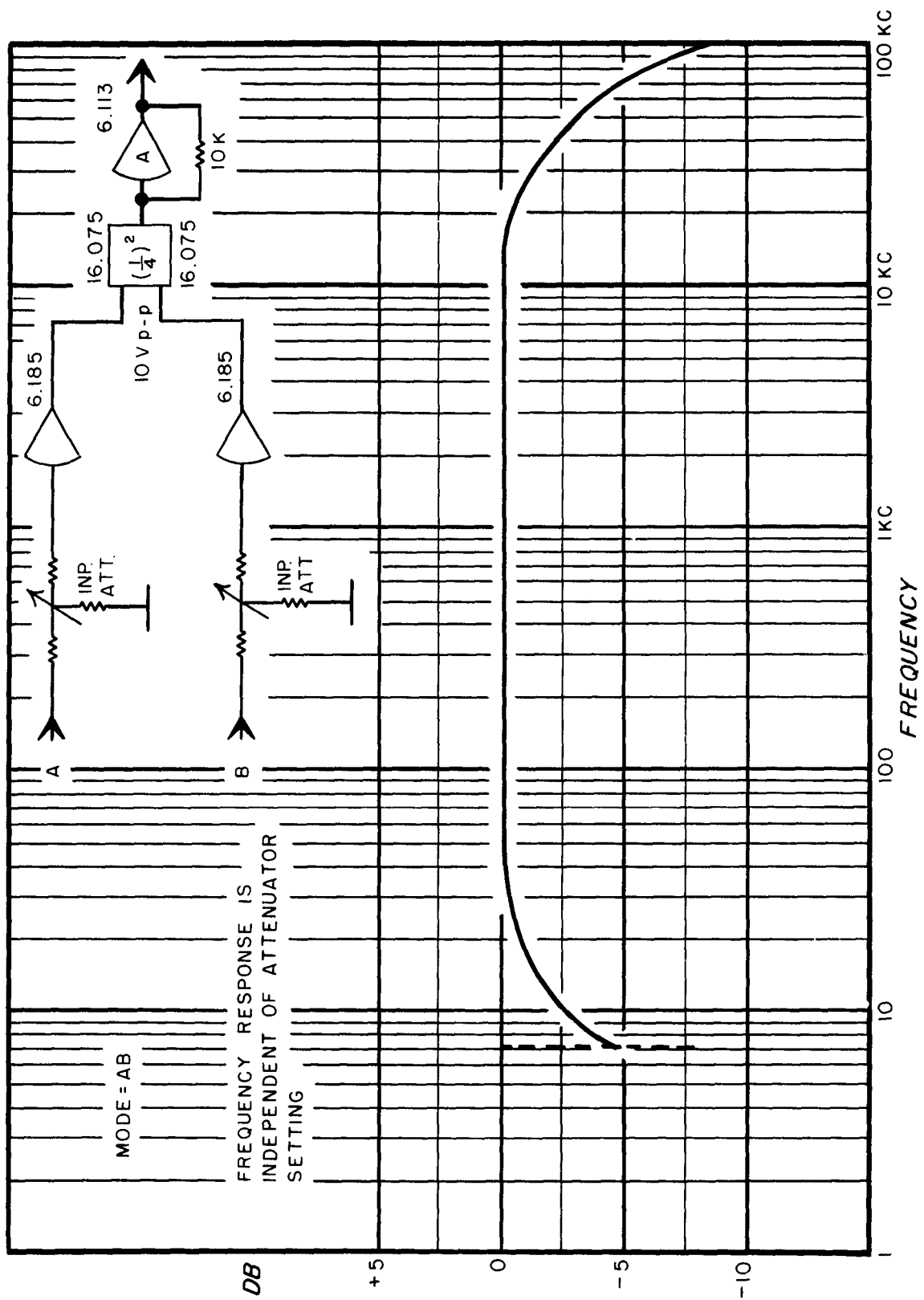


Figure 10 Frequency response of Mode AB.

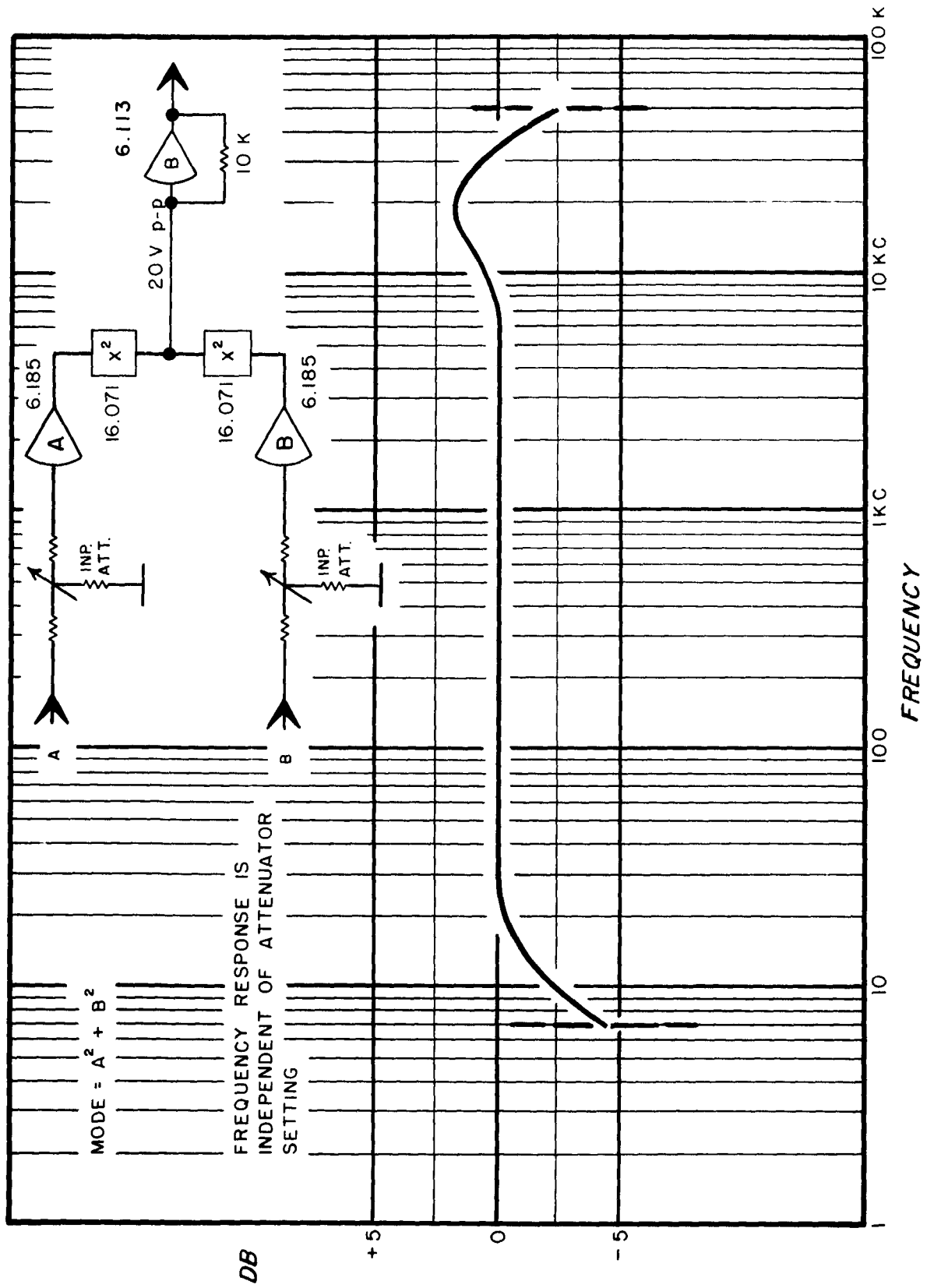


Figure 11 Frequency response of Mode $A^2 + B^2$.

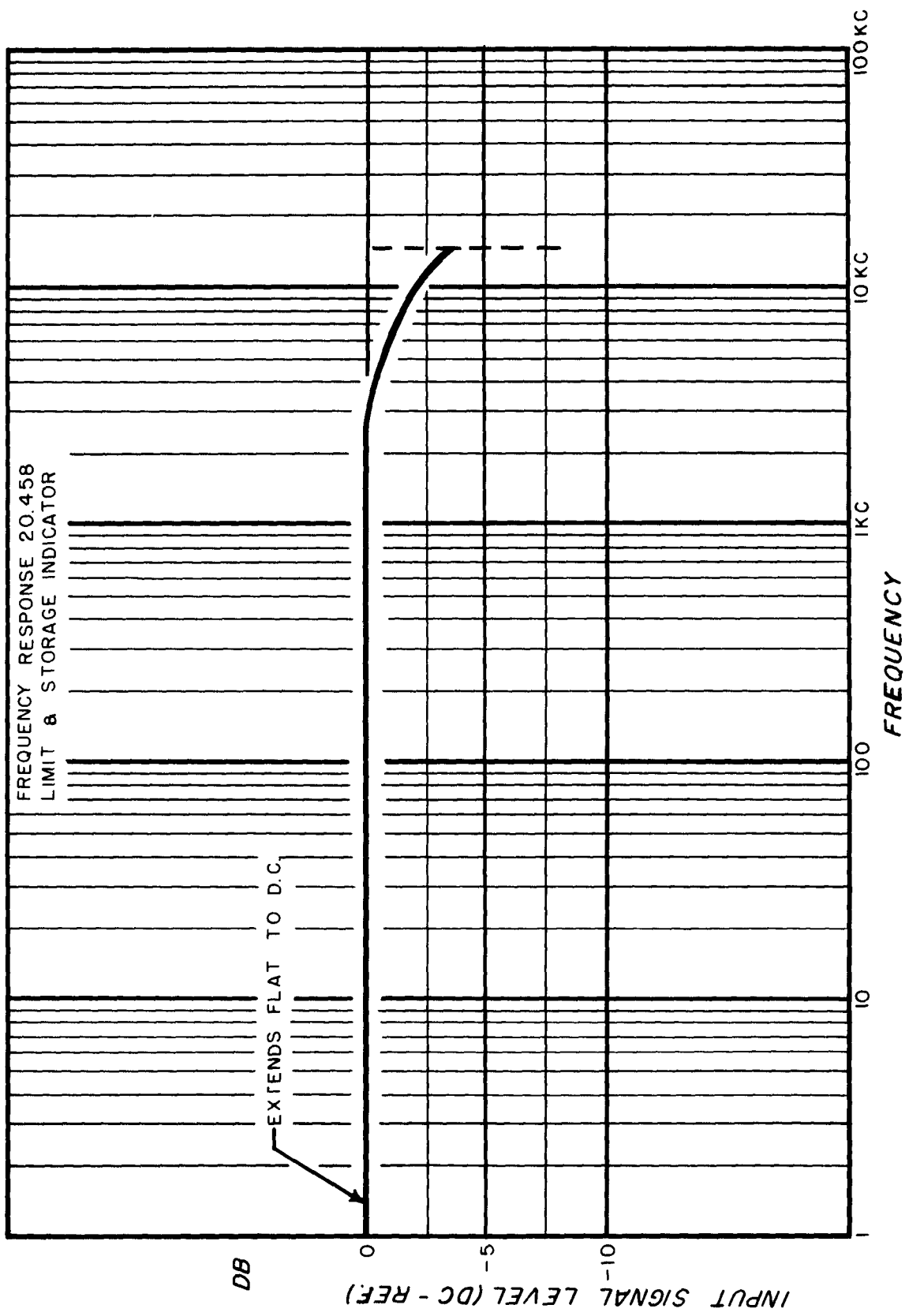


Figure 12 Frequency response of Limit and Storage Indicator.

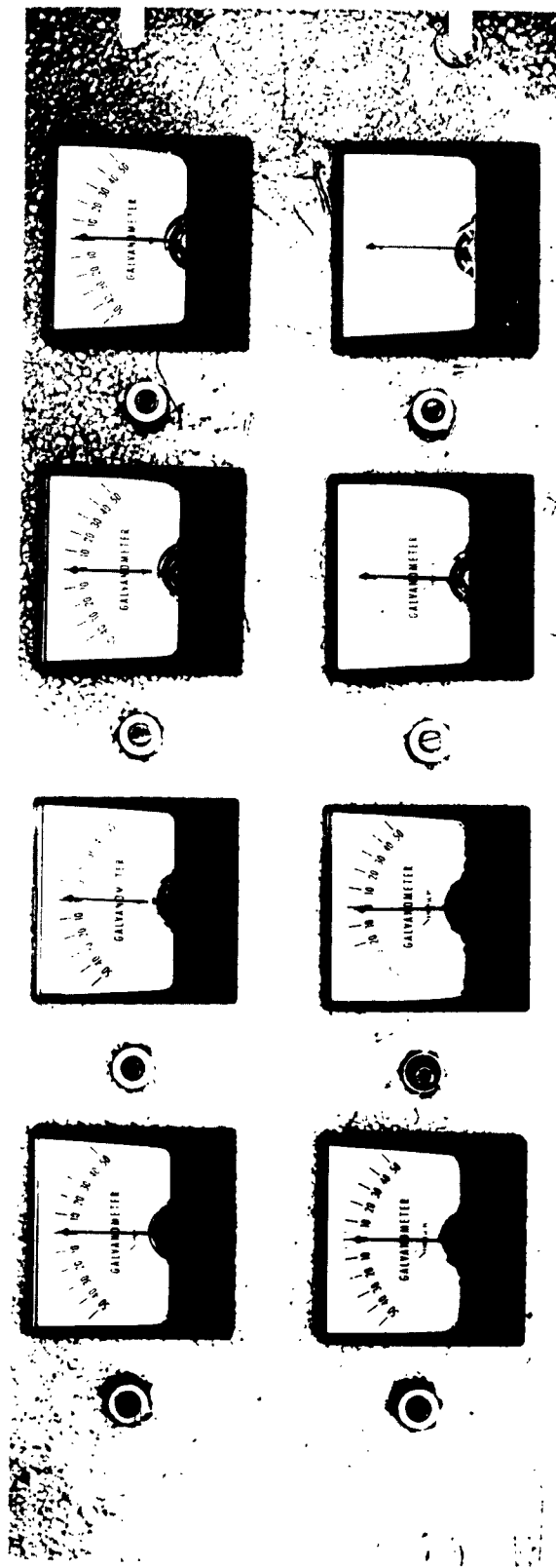


Figure 13 Supplementary readout panel for Oceanographic Computer.

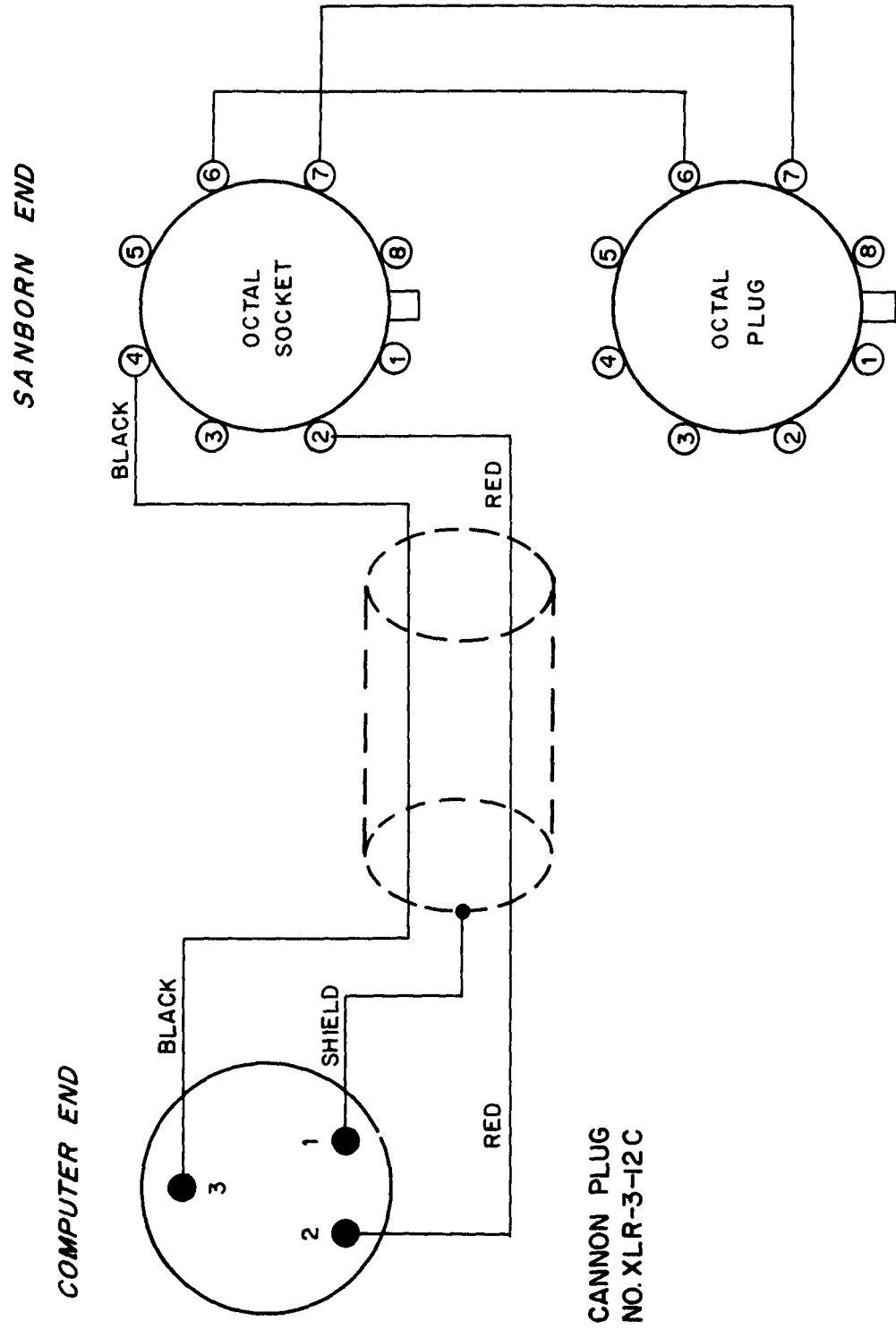


Figure 14 Output cables for Oceanographic Computer.

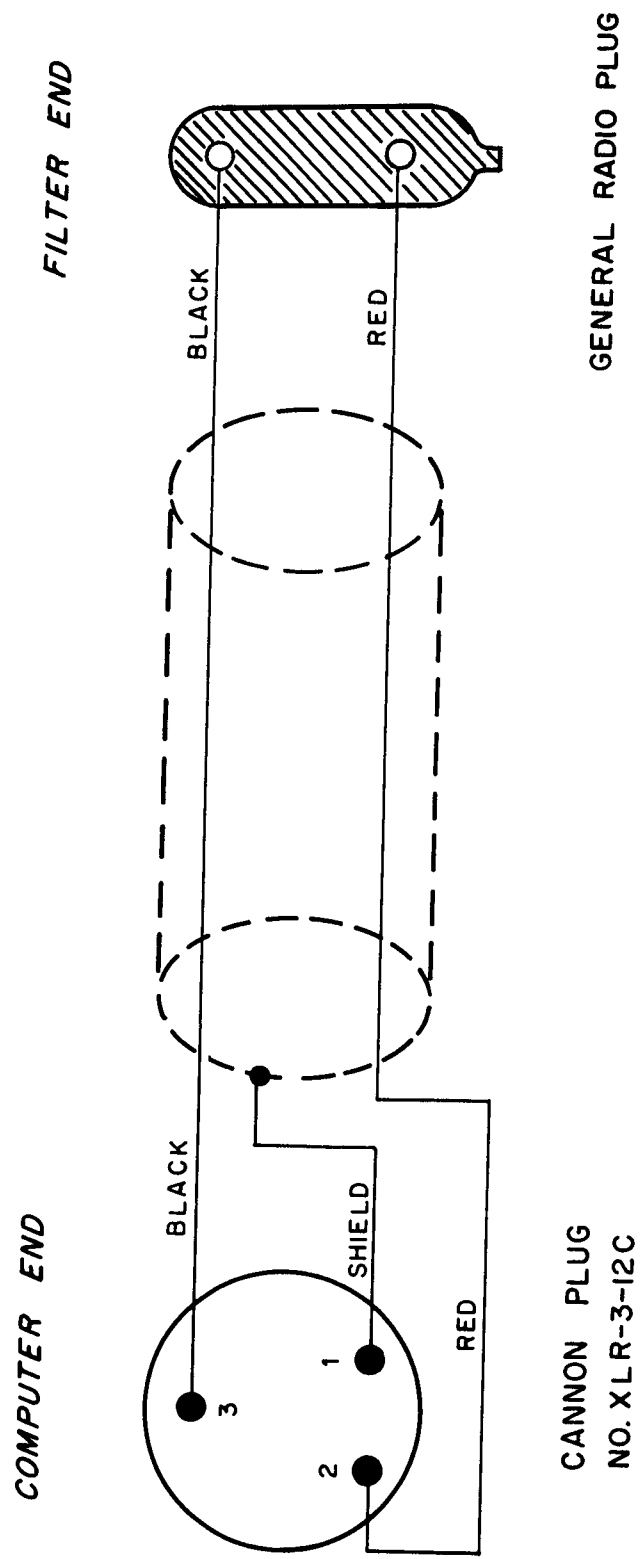


Figure 15 Input cables for Oceanographic Computer.

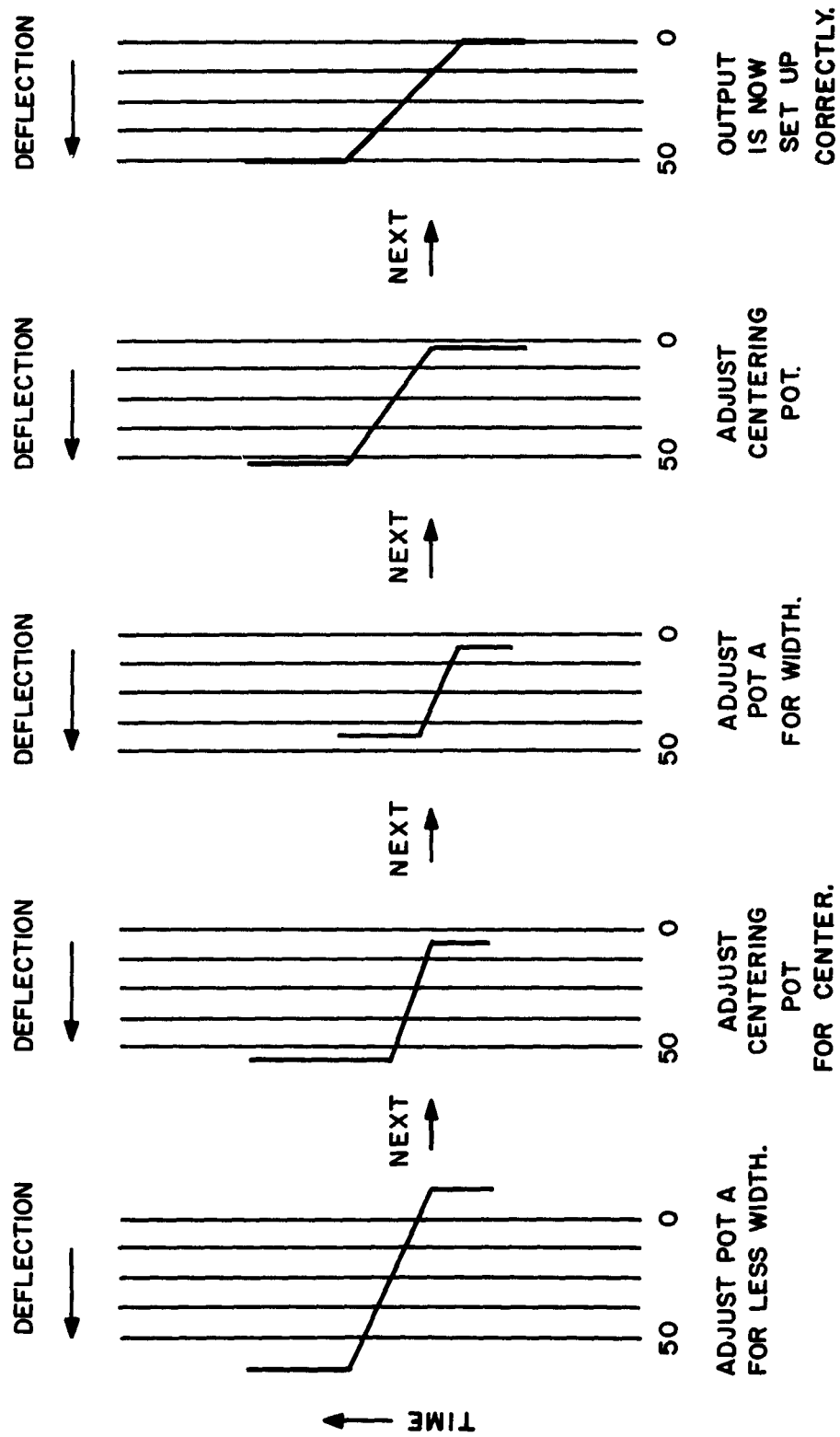


Figure 16 Appearance of Sanborn Trace during adjustment of output scale and centering.

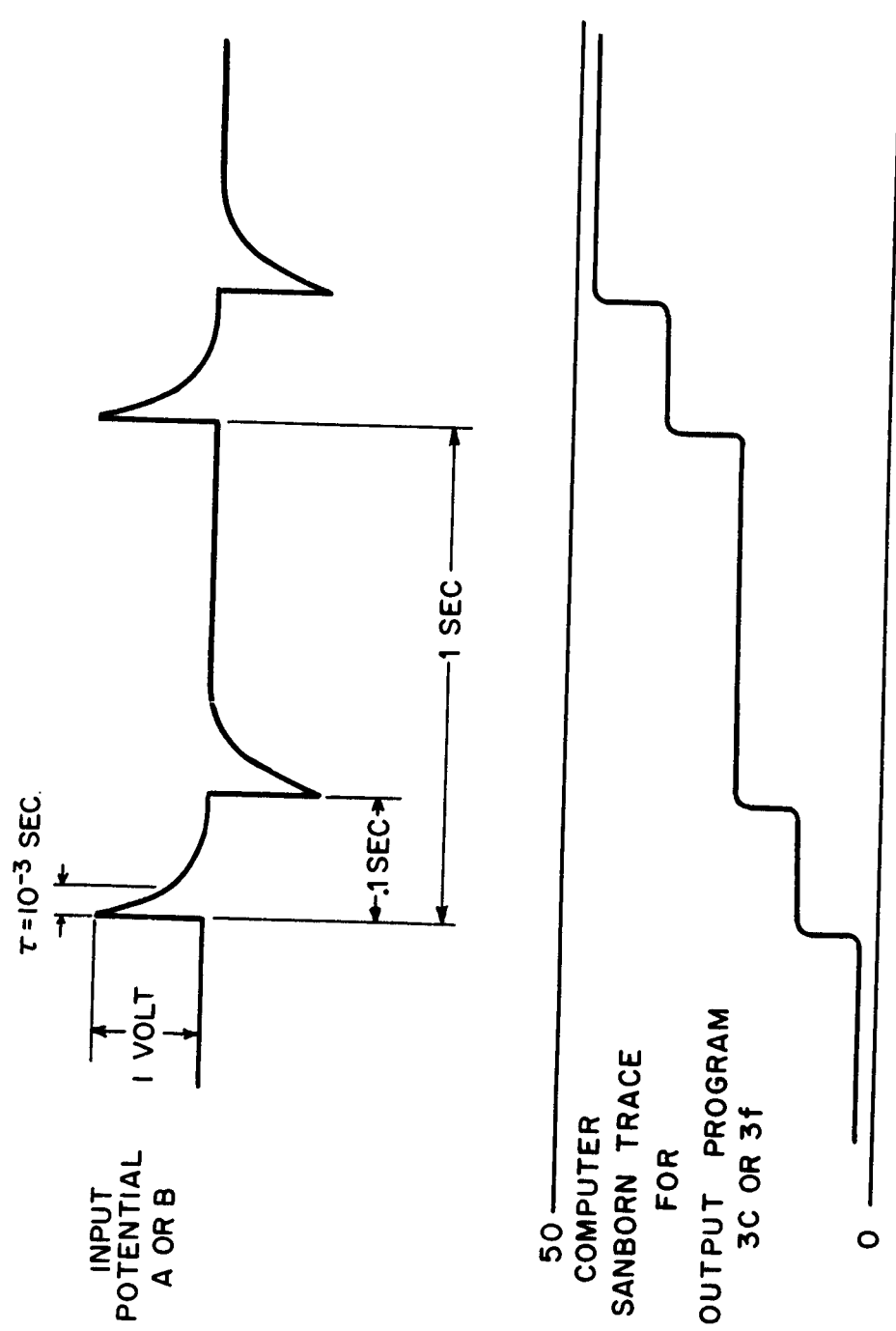


Figure 17 Pulse calibration signal.

APPENDIX A

Phase Shift Circuit for Testing Analog Computer Multipliers

In testing the multiplication of two signals by an analog computer, it is desirable to be able to check the accuracy of the operation with respect to the phase relationship between the input signals. In order to make such a check, a phase-shifting network was designed and built at the WHOI analysis laboratory.

This network is capable of operation at spot frequencies of 30, 90, 600, 6000, and 20,000 cps., and can produce sine waves of any desired phase difference at these frequencies. The output of the phase-adjustable channel varies slightly in amplitude with a change in phase, while the phase fixed (reference) channel has a gain control so that the amplitude can be matched to that of the variable channel.

The input impedance of the phase shifter is high, and any signal generator producing a good sine wave can be used to drive the network. The phase shifter requires about two amperes of 117 volt 60 cycle power. It is preferable to operate this device from a voltage-regulated line.

The two outputs are of low impedance. This is accomplished by the use of cathode follower output stages. Each channel is capable of feeding a 600-ohm input without distortion. If the circuit is loaded more heavily, however, it is possible to overload the output amplifier and thus cause distortion.

The entire phase shifter is built in an 11" x 7" x 2" chassis with an 11" x 5" panel (see Fig. 1A). It is obvious, from the figure, that future models should be built on a larger chassis to prevent the overcrowding of components. Since the analysis laboratory at WHOI needs only one such unit, the prototype is the only one built.

For future models, the following changes are strongly recommended. 1) use of a larger chassis -- probably a 10" x 12" x 3" chassis would not be too large. 2) an electronically regulated power supply is very desirable, as the main "bugs" in the circuit proved to be interaction between stages through the power-supply, and the decoupling shown in the schematic is just barely sufficient for practical operation at the low frequencies.

Appendix A-2

The circuit, Figure 2A, is composed of two channels. The phase fixed (reference) channel is composed of two cathode-follower stages with a gain control placed between them. The phase variable channel provides a cathode-follower input stage, the output of which is fed to two resistive-capacitive phase shifting networks. One of these networks shifts the phase sixty degrees in advance, and the other shifts the phase sixty degrees in arrears. The capacitors of these networks are switched by a ceramic selector switch to change operating frequency.

Following this, there are three split-load phase inverters. One is driven directly from the cathode-follower while the other two are driven by the two phase-shifting networks. The six outputs from these three split-load phase inverters provide sine waves which are 0, 60, 120, 180, 240, and 300 degrees out of phase with the input to the channel. These phases are applied to the contacts of a selector switch across which a potentiometer may be switched. This potentiometer allows one to interpolate anywhere between two consecutive phase taps of the switch. The output of the interpolation pot is applied to the grid of a triode amplifier with negative feedback. From there, a cathode-follower is driven and used as an output stage.

In use, the phases will not be perfect, especially at the very low and very high frequencies. Thus, the phases and the interpolating pot cannot be calibrated. In lieu of dial calibration, an XY oscilloscope may be used to measure the phase difference. When the scope is set up with the reference channel monitored by the X axis and the phase variable channel monitored by the Y axis, the phase difference can be computed from the dimensions of the ellipse. The ellipse must be set up as shown in Figure 3A. $X_1 = -X_2$ and $Y_1 = -Y_2$ is a necessary condition for measurement by this method. Then the dimensions A and B are read off the scale of the scope. The phase difference is given by the equation: $\phi = \arcsin \pm A/B$. When the ellipse tilts to the right, as in Figure 3A, $-90^\circ < \phi < 90^\circ$. When the ellipse tilts to the left, $90^\circ < \phi < 270^\circ$. When the ellipse is a circle, $\phi = 90^\circ$ or $\phi = 270^\circ$.



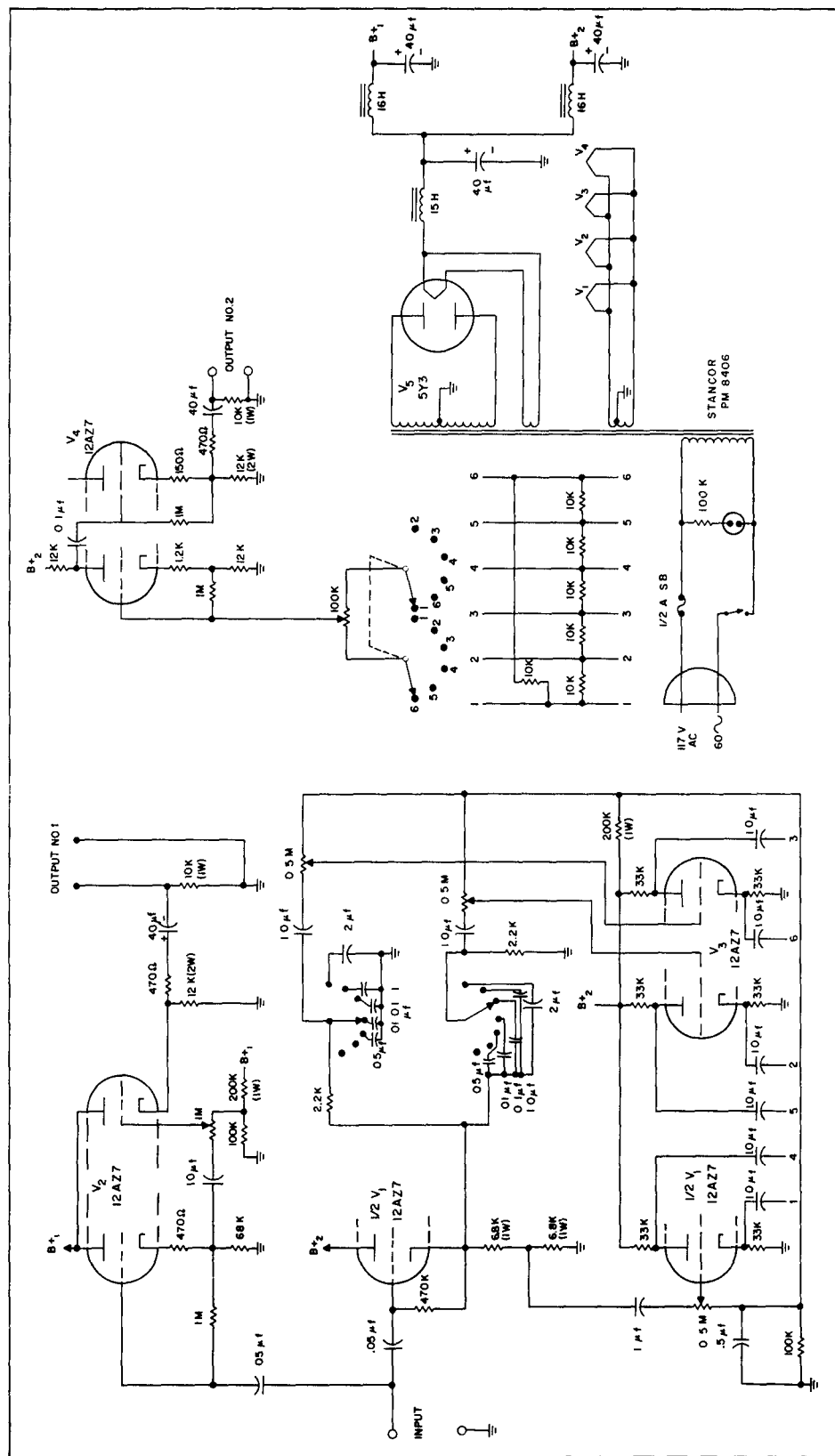
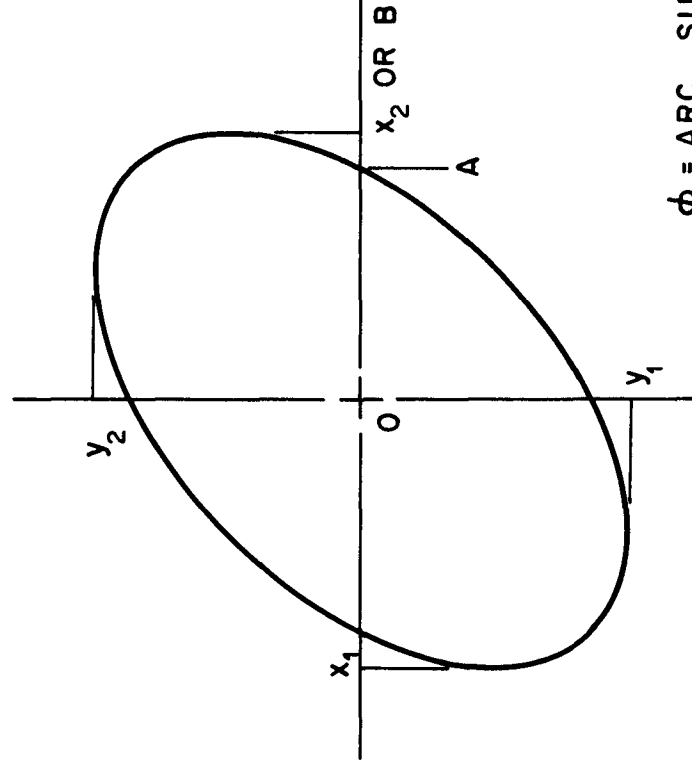


Figure 2 A Circuit diagram of WHOI Phase Shifter.



$$\phi = \text{ARC SINE} \pm \frac{A}{B} \text{ IF } x_1 = -x_2 \text{ AND } y_1 = -y_2$$

Figure 3 A Oscilloscope Trace used for calibrating Phase Shift.

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