NBS TECHNICAL NOTE 1159

U.S. DEPARTMENT OF COMMERCE/National Bureau of Standards

A Fast Response, Low-Frequency Sampling Voltmeter

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NBS technical note

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A FAST RESPONSE, LOW-FREQUENCY SAMPLING VOLTMETER

Barry A. Bell, Bruce F. Field,* and Thomas H. Kibalo**

A low-frequency voltmeter utilizing a sampling technique implemented with microprocessor-based electronics has been developed to perform as a true rms ac voltmeter and distortion analyzer. The instrument makes measurements accurate to ±0.1 percent (of reading) of the fundamental frequency, total harmonic distortion, and true rms voltage of approximately sinusoidal inputs from 2 mV to 10 V and frequencies from 0.1 to 120 Hz. A major feature of this instrument is the special window crossing and error function algorithms which provide a software means for completing a measurement within two signal periods at frequencies below 10 Hz.

Key words: algorithm; converter; distortion; microcomputer; rms value; sampling; signal period.

1. INTRODUCTION

1.1 Background

For many years the accurate measurement of the true rms value of ac voltage has been of considerable interest to the National Bureau of Standards. With the use of high accuracy thermal voltage converters, whose ac-dc difference characteristics have been well established, absolute measurement of ac voltage can be determined from 0.5 to 1000 V to an accuracy of 50 ppm or better over a frequency range from 20 Hz to 100 kHz [1-3].[1]

More recent investigations by the Electrosystems Division at NBS have emphasized the extension of this capability to lower frequency and voltage levels of interest in order to support the design, calibration, and maintenance of vibration transducers (accelerometer-amplifier systems). These transducers typically operate with approximately sinusoidal signals having levels from a few millivolts up to several volts and frequencies from tenths to tens of hertz. An accurate ac voltmeter/calibrator, employing a multi-junction thermal converter device for rms-to-dc conversion, has been developed at NBS which is capable of making rms ac voltage measurements from 0.1 to 50 Hz at 2 mV to 10 V levels with ±0.1 percent accuracy (percent

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[1] Numbers in brackets refer to the literature references listed at the end of this report.

of reading) above 0.5 Hz and 5 mV and ±0.2 percent below these values [4]. Using the built-in calibrator, ac transfer measurements (comparing ac input voltages with the ac calibrator) can be made to an accuracy of ±0.02 percent. This instrument serves as the standard in a calibration service now provided by NBS for ac voltmeters and calibrators at frequencies from 0.1 to 10 Hz [5].

1.2 Low-Frequency Sampling Voltmeter Specifications

A new instrument has been developed using sampling techniques which is capable of measuring the true rms voltage of approximately sinusoidal waveforms from 2 mV to 10 V with frequencies between 0.1 to 120 Hz [6]. Although similar in its coverage of voltage and frequency ranges as the earlier thermal converter-based standard, this instrument is designed to have a very short response time as well as providing measurements of the rms value of the ac component (only) of the input signal, the fundamental frequency, and the harmonic distortion of the signal, which are features not available on the thermal converter-based instrument.

A research and development program at NBS has been concerned with developing and testing improved vibration accelerometers [7,8]. The major motivation for improved accelerometers has been due to interest in accurately predicting the mean-time-between-failure and failure modes of equipment subjected to mechanical vibration. At present, the largest source of uncertainty in calibrating these accelerometers has been in the low-frequency voltage measurements. Also, studies performed to evaluate new accelerometer designs have produced inconclusive results because of the voltage measurement uncertainties. The instrument described herein was developed to fulfill the following requirements of vibration accelerometer calibration systems:

(1) Frequency range - 0.1 to 120 Hz (fundamental).

(2) Input voltage range - 2 mV to 10 V (4 1/2 decades).

(3) Accuracy - ±0.1 percent of reading (2σ).

(4) The instrument should be an easy-to-use portable instrument requiring minimal calibration.

(5) Fast response time is desired, i.e., the measurement should require no more than two periods at the lowest frequency.

(6) Capability of measuring total harmonic distortion of the waveform for at least the first 10 harmonics.

It should also be noted that in many cases vibration accelerometers produce slowly drifting dc offsets which are added to the ac component. It is awkward to manually compensate for a changing offset; thus, it is necessary that the instrument provide for measuring the ac component (only) of the signal.

2. THEORY OF OPERATION

2.1 Instrument Architecture

Traditional measurement methods for ac waveforms all rely on some square law device (thermal converter, log amplifier, etc.) to provide a true rms indication. For this application squaring modules based on log amplifiers do not provide sufficient accuracies at low frequencies. Thermal converters generally have large time constants (of order 1-5 sec) due to the thermal mass of the heater and thermoelement. RMS to dc converters require large time constants for ripple filtering, causing slow response to input voltage level changes. Although improvements in thermal converter time constants can be made, increased ac ripple on the dc output results, requiring additional filtering or smoothing. Thus, if optimized for low-frequency measurements, the response time of such an instrument becomes excessively large. In addition, these methods do not provide any means for measuring the distortion in the waveform.

The general approach taken for this low-frequency voltmeter is to utilize a microcomputer for implementing as many functions as possible with software, including timing and control of the input sampling hardware. A simplified block diagram of the entire instrument is shown in figure 1. A conventional sample-and-hold (S/H) amplifier and successive approximation type of analog-to-digital (A/D) converter are used for digitizing the input signal. The digital values are then transferred to a microcomputer for calculation of the desired results.

The preamplifier (NBS 722 79184A) is a differential input amplifier with programmable gain provided by external gain setting resistors that can be changed by the microcomputer. Four-decade gain settings are available to accommodate the wide range of possible input signals. A data acquisition channel consisting of a S/H amplifier, a second instrumentation amplifier, and an A/D converter are physically contained in one hybrid module (NBS 722 79141A). This second amplifier is normally hardwired to provide only one of three input voltage ranges for the A/D converter: 2.5, 5, or 10 V. However, external relays have been added to allow the range to be set by the microcomputer. Thus, a total of 12 gain settings is possible using combinations of preamplifier gain and A/D range settings. All these settings are necessary in order to obtain adequate resolution for all specified inputs. The A/D converter has 12-bit (11 bits plus sign bit) resolution with a conversion time of 12 μsec.

The memory, timing, and control section (NBS 722 79094B) has two functions. The first is a hardware clock that may be programmed by the microcomputer to output a pulse at regular intervals. This pulse is used to trigger the A/D converter to make a conversion; thus, a series of A/D readings uniformly spaced in time are generated irrespective of the operation of the microcomputer which may be busy with other calculations. The second function of this section is to

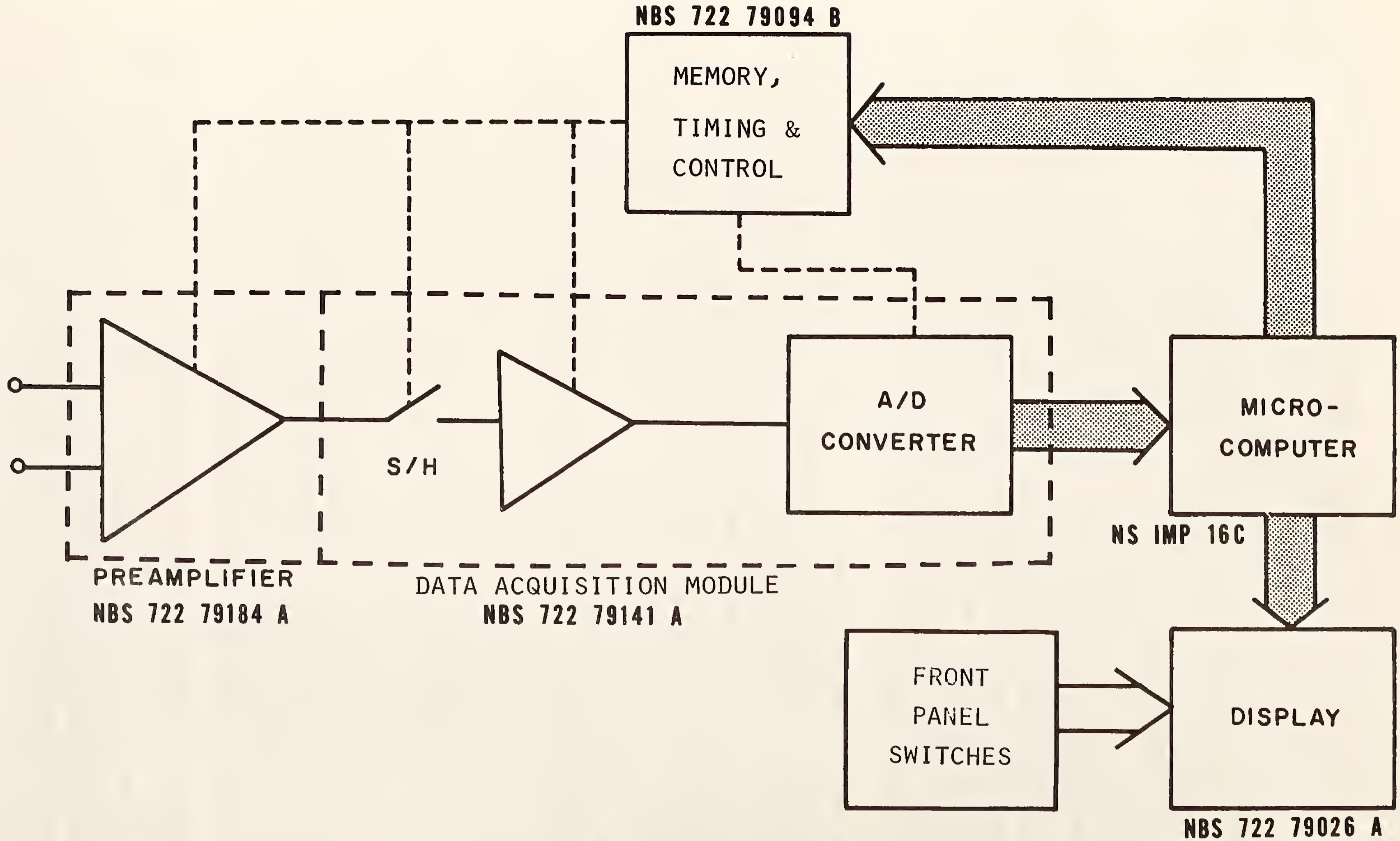

Figure 1. Simplified diagram of the voltmeter.

route control signals generated by the microcomputer to the gain controlling relays for the preamplifier and the data acquisition module.

The microcomputer used is a National Semiconductor IMP-16C.[2] This is a 16-bit word length multi-chip microcomputer. It was chosen mainly for the powerful arithmetic instructions in its instruction set. All peripheral devices (display, A/D converter, etc.) are treated as memory locations and the instruction program (2048 16-bit words) is stored in ultraviolet-erasable programmable read-only-memories (EPROMS). A small amount of random access memory (RAM) is provided on the microcomputer board for temporary storage of input values and program variables.

2.2 Measurement Algorithms

Several kinds of algorithms are used to implement the voltmeter's rms value, frequency (or period), and harmonic distortion functions. Beyond the discussion given here, a more complete description of the signal model for vibration accelerometer systems and associated measurement algorithms is provided in [11].

Because the algorithms for calculating rms value require that the value for the period be accurately determined, it is desirable to have an initial sampling sequence which will adjust the signal level so that it is within the range of the data acquisition preamplifier system and, then, to determine the approximate period. Figure 2 shows the time sequence of the measurement process used within this instrument.

During the first cycle (T_1), sampling is begun at any arbitrary point after the preamplifier gain has possibly been adjusted for off-scale readings. Samples of the waveform are taken and an approximate period is established. In order to measure the period more exactly, and to obtain a set of samples with which to determine the rms value, the sample rate clock (used to trigger the A/D converter) is then set at 128 times the approximate frequency f_0 (the reciprocal of the approximate period). Any necessary adjustments to the gain settings in the sampling hardware are made, a set of 147 samples is taken during the second cycle (T_2), and the corresponding A/D converter readings stored. Depending on the setting of the function switches on the front panel of the voltmeter, the rms value, frequency, or harmonic distortion is calculated during time T_3, and the display is updated with the new value. The whole process is then immediately repeated. Figure 3 shows an overall flowchart of the measurement software that corresponds to the measurement process just described.

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[2]Certain commercial equipment, instruments, or materials are identified in this paper in order to adequately specify the experimental results. In no case does such identification imply recommendation or endorsement by the National Bureau of Standards, nor does it imply that the material or equipment identified is necessarily the best available for the purpose.

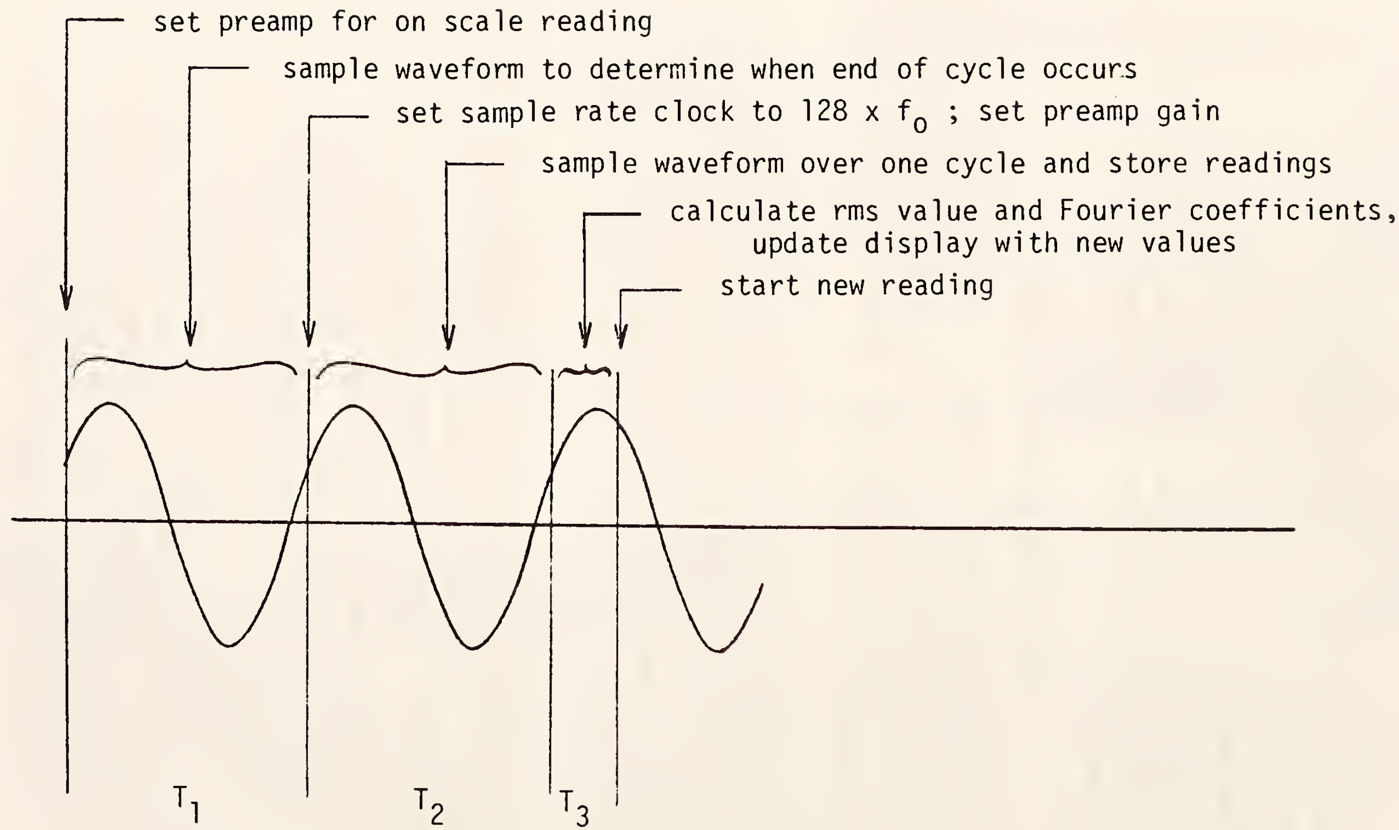


Figure 2. Time sequence of the measurement process.

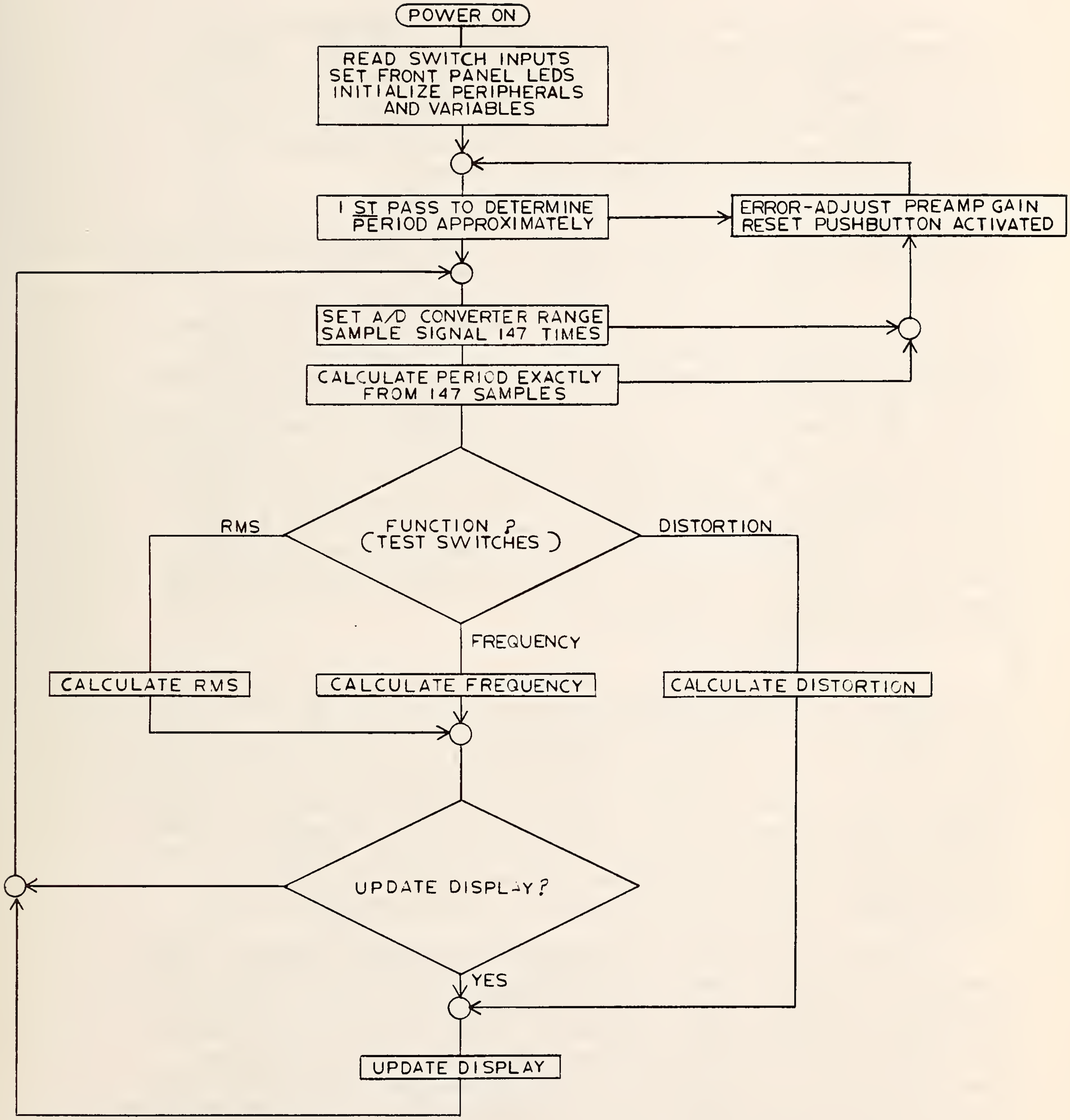


Figure 3. Flowchart of measurement software.

2.2.1 Approximate Period

When starting a measurement of an unknown input waveform, the frequency may be as low as 0.1 Hz or as high as 120 Hz. Sampling at a fast enough rate to capture a 120 Hz signal would produce an excessive number of samples (to store in memory) of a 0.1 Hz signal. A preliminary decision, based on the slope of the signal at the start of the measurement sequence (T_1), determines whether a fast (8 kHz) or slow (500 Hz) sampling rate is used.

With the sampling rate set at 500 Hz (see fig. 4, approximate period detection), an initial reading is taken and temporarily stored in the IMP-16C's RAM. A higher and lower threshold value is calculated at ±0.5 V with respect to this initial reading (upper bound of ±10 V). Thus, an initial "window" of sample values is established as shown in figure 4. Sampling continues with each value tested against the higher and lower threshold until either of these window limits is exceeded. A count is made of the number of samples, N, taken within the window. If the number of samples between the initial value and a threshold is greater than five ($N > 5$), it is assumed that the frequency of the input signal is 1 Hz or less. In this case, sampling continues at 500 Hz until a second window crossing occurs. However, if the number of samples between the initial value and a threshold is less than five ($N < 5$), it is assumed that the input frequency is 10 Hz or greater. To obtain adequate resolution of the period in this case, the sample rate clock is set to 8 kHz and the measurement sequence restarted.

For frequencies less than 1 Hz, the first set of measurements is thus used as an initial voltage window which defines the beginning of one signal period. Sampling continues until a new sample falls within the window values. The next N samples are then saved and tested to see if at least N/4 of them are within the window. If this condition is met, it is assumed that the waveform has passed through the first window crossing (see fig. 4). After the waveform is sampled further and the second window crossing is found, one signal period has then elapsed. The sampling rate clock is then set for 128 times the approximate frequency (the reciprocal of the approximate period).

One problem which must be accounted for with this method is where the window includes a peak on the waveform (see fig. 5). Since the window of values will occur only once more in the period, waiting for two window crossings would include two signal periods. To eliminate this problem, if the initial window does occur over a signal peak, the initial value is modified to start at the peak sample value with a corresponding adjustment of the threshold value. Thus, the beginning of the second window crossing is, again, the peak value which gives an approximate measure of one signal period.

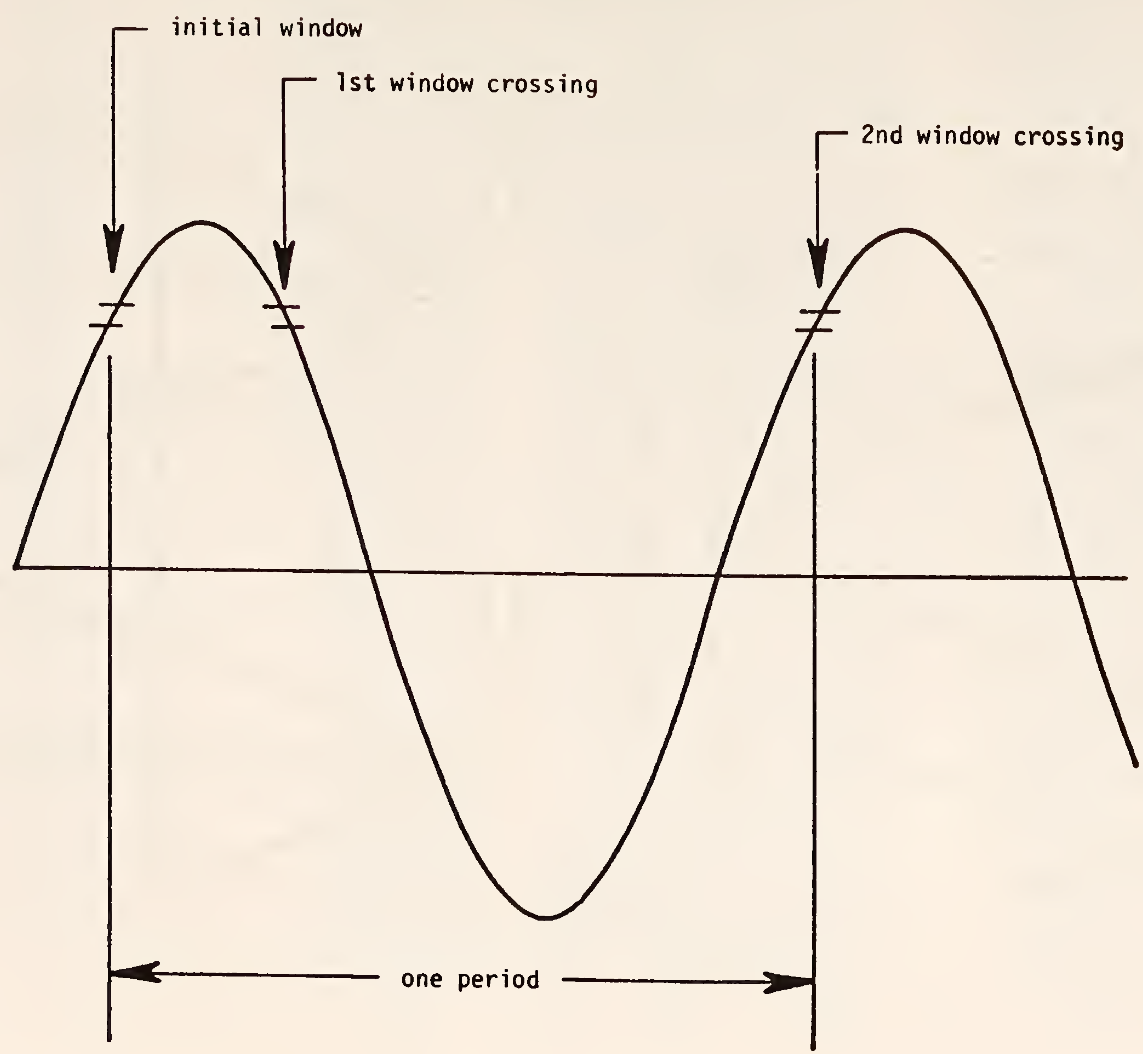
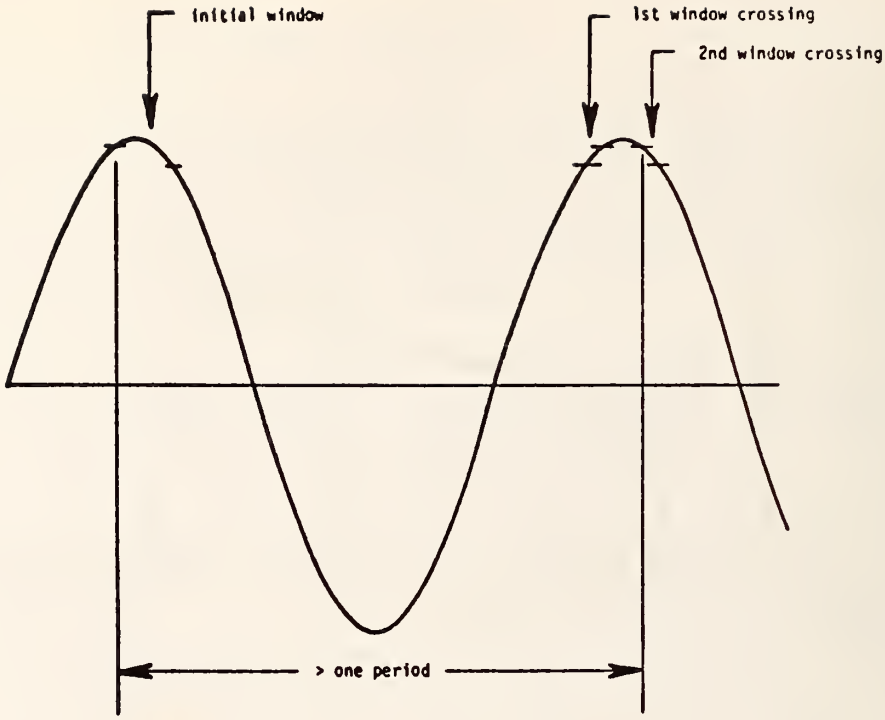
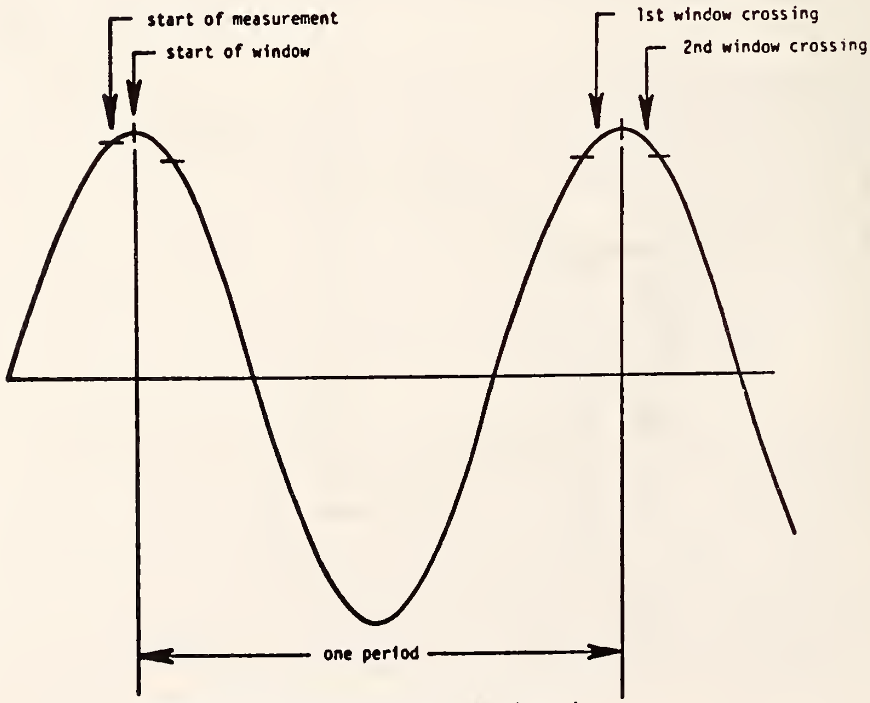


Figure 4. Approximate period detection.



Problem: Initial window starts near a peak



Solution: Modify window to start at the peak

Figure 5. Approximate period detection at the signal peak.

If during the sampling process of finding the initial window of data values the upper bound of ±10 V is exceeded, overrange has occurred and the preamplifier gain is immediately reduced. Once the search is on for two more window crossings, however, autoranging occurs differently, depending on the sampling rate. For 500 Hz sampling, it is possible to check each reading for an overrange condition and reset the preamplifier and A/D converter gains in real-time, i.e., between individual samples. When sampling at 8 kHz, overrange tests are deferred until the end of one signal period. If at this time it is determined that the input was overrange during the measurement sequence, the gains are adjusted and the measurement restarted.

2.2.2 Exact Period

As shown in figure 2, once the approximate frequency f_o has been determined and the sample rate clock set to 128 x f_o, the input waveform is then sampled over a second cycle and the readings stored for subsequent use in calculating the exact period, rms value, or harmonic distortion (refer also to fig. 3). In order to calculate the rms value of the input signal, the period must be known with approximately the same accuracy as is desired in the rms value [11]. As indicated by time T_3 in figure 2 and the main program flowchart in figure 3, 147 samples are taken in order to calculate the exact period. The reason for storing more than the nominal 128 sample values is to take the effects of noise and possible frequency modulation (FM) into account.

An autocorrelation technique is used to determine the maximum point of correlation with respect to the true signal period and the beginning and endpoint samples taken of the waveform. This correlation is accomplished by evaluating an error function that consists of summing the difference in values of samples taken at approximately corresponding times in the cycle:

$$\text{error function} = \sum_{i=1}^{n} \left| E_i - E_{i+N} \right|$$

where N = variable number of samples in period (125, 126, 127, 128, 129, 130, and 131)

E_i = sample values at beginning of cycle

and E_{i+N} = corresponding sample values at beginning of next cycle

n = number of difference values summed; 16 points chosen, based on empirical testing.

If the period was exactly an integral number of samples and the input signal was free of noise or FM, this sum would go to zero for some value of N. However, since this probably is not the case, N is varied as indicated above until a minimum value for the error function is found. The set of seven error functions calculated is then

$$\text{error function 1} = \sum_{i=1}^{16} \left| E_i - E_{i+125} \right|$$

$$\text{error function 2} = \sum_{i=1}^{16} \left| E_i - E_{i+126} \right|$$

$$\text{error function 3} = \sum_{i=1}^{16} \left| E_i - E_{i+127} \right|$$

$$\text{error function 4} = \sum_{i=1}^{16} \left| E_i - E_{i+128} \right|$$

$$\text{error function 5} = \sum_{i=1}^{16} \left| E_i - E_{i+129} \right|$$

$$\text{error function 6} = \sum_{i=1}^{16} \left| E_i - E_{i+130} \right|$$

$$\text{error function 7} = \sum_{i=1}^{16} \left| E_i - E_{i+131} \right|$$

Figure 6 shows a plot of a typical error function E_f,

$$E_f = \sum_{i=1}^{n} \left| E_i - E_{i+\Delta} \right|$$

where Δ = correlation parameter, delta .

Only a 10-point summation was used ($n = 10$) with sampling begun for $\Delta = 0$. For this example, the period contained exactly 360 samples. As is apparent from figure 6, a second unwanted minimum occurs for this error function at approximately $\Delta = 170$. The exact location of this second minimum depends on the starting angle, however, as indicated in figure 7. As the starting angle approaches 90°, the second minimum shifts very close to the true minimum, and it becomes difficult to distinguish between the two (see fig. 8).

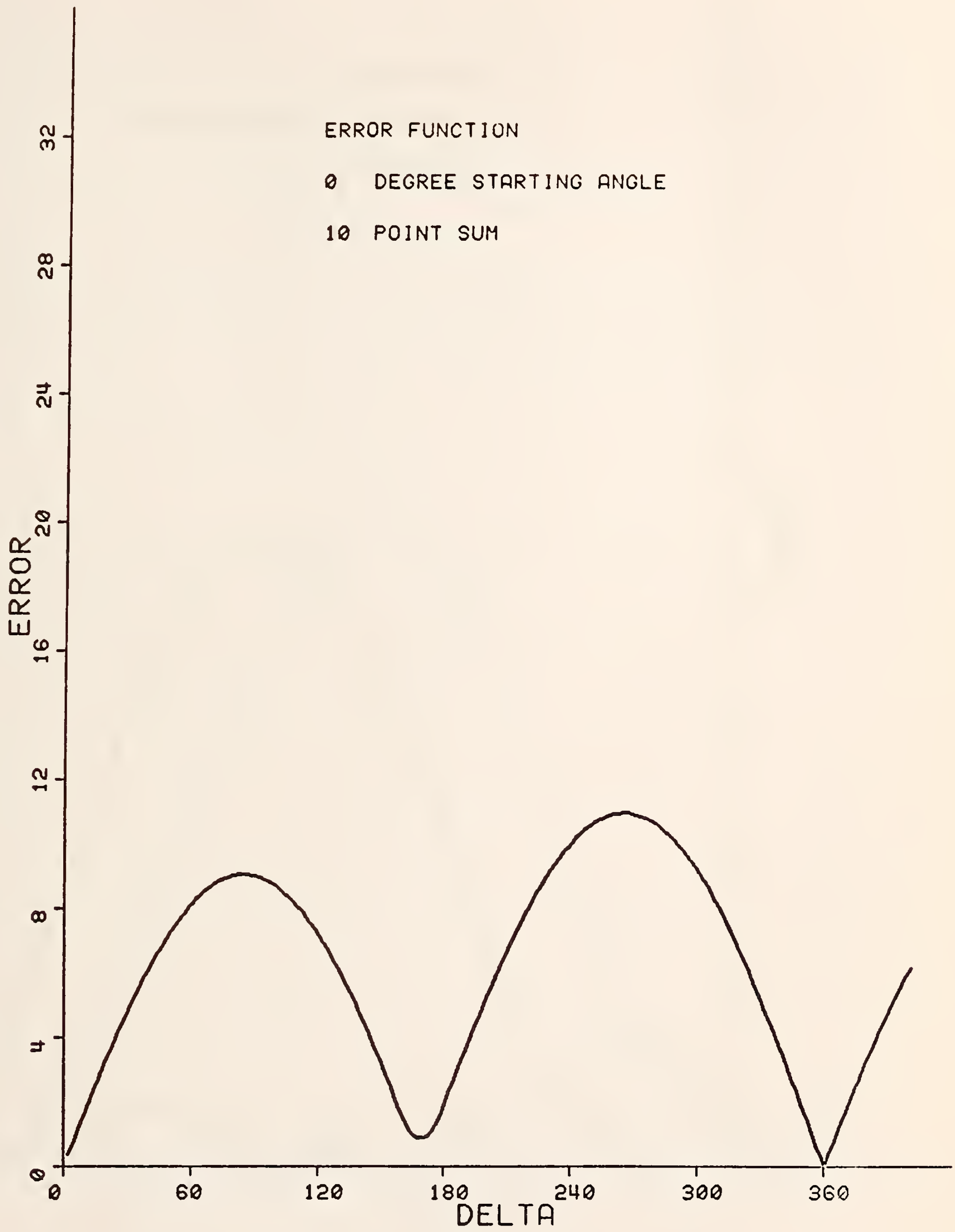


Figure 6. Plot of the error function.

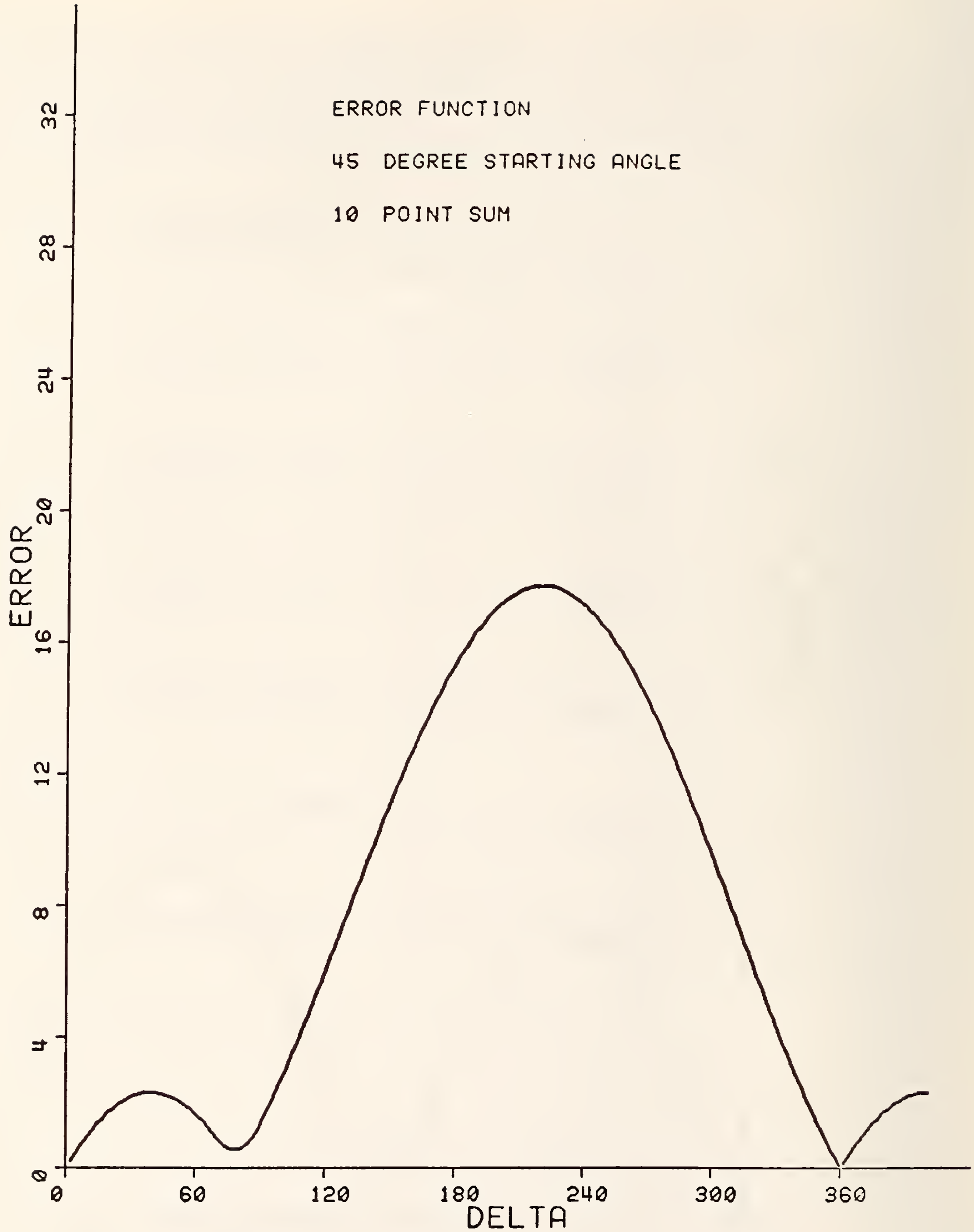


Figure 7. Plot of the error function showing the shift of the undesired minimum due to phase shift of the signal.

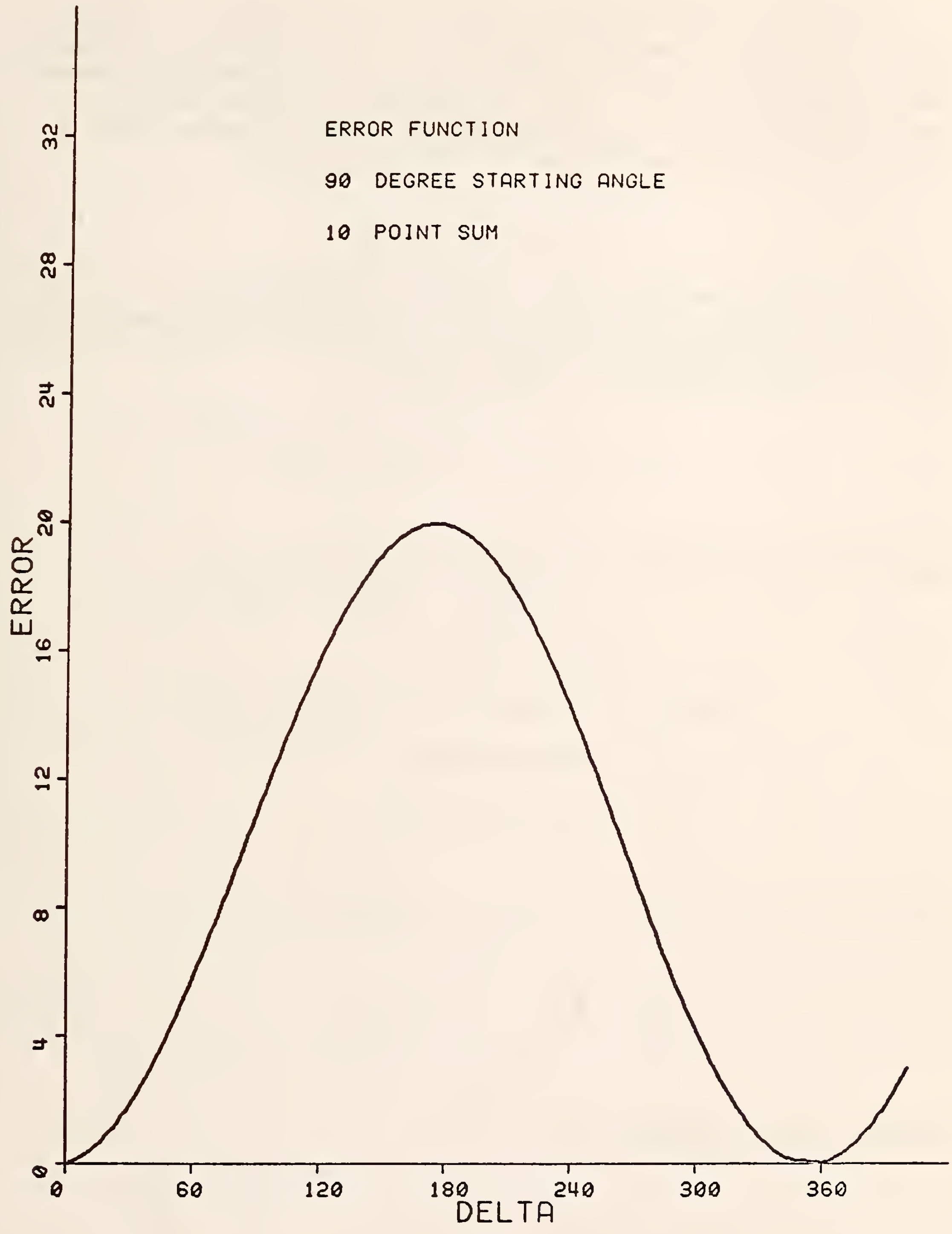


Figure 8. Plot of the error function showing the shift of the undesired minimum as indistinguishable from the true minimum.

Therefore, to provide more resolution than integral values of N will allow, the two smallest sums of the error functions 1 through 7, above, are used to linearly interpolate to the minimum point between them (corresponding to $\Delta = 360$). Experiment has shown that this two-point interpolation provides the required ±0.1 percent accuracy in the resultant value obtained for the exact period.

2.2.3 RMS Value

As stated in section 2.2.2, in order to calculate the rms value of the input signal, the period must be known with approximately the same accuracy as is desired in the rms value. To demonstrate this, let the input signal E(t) be represented as

$$E(t) = \cos t \quad (\text{rms value} \equiv \frac{1}{\sqrt{2}}) \quad .$$

Assume that the period T is determined incorrectly, i.e.,

$$T' = T\,(1-\varepsilon)$$

where T = true period (2π in this case)

T' = measured period

ε = fractional error .

Calculating the rms value directly as defined,

$$E_{rms} = \sqrt{\frac{1}{T'} \int_0^{T'} \cos^2 t \, dt} \qquad .$$

Evaluating the integral Int,

$$\text{Int} = \int_0^{T'} \cos^2 t \, dt = \frac{T'}{2} + \left. \frac{\sin 2t}{4} \right|_0^{T'}$$

or, substituting for T',

$$\text{Int} = \frac{T(1-\varepsilon)}{2} + \frac{\sin 2T\ (1-\varepsilon)}{4} \ .$$

For small ε ,

$$\sin 2T(1-\varepsilon) \cong -\ 2T\varepsilon$$

and

$$\text{Int} \cong \frac{T(1-\varepsilon)}{2} - \frac{T\varepsilon}{2} = \frac{T(1-2\varepsilon)}{2} \ .$$

Then ,

$$E^2_{rms} = \frac{\text{Int}}{T'} \approx \frac{1}{T(1-\varepsilon)} \cdot \frac{T(1-2\varepsilon)}{2}$$

$$\cong \frac{1}{2}\frac{(1-2\varepsilon)}{(1-\varepsilon)} \ .$$

Again, for small ε ,

$$\frac{1}{1-\varepsilon} \cong 1+\varepsilon .$$

Therefore,

$$E^2_{rms} \cong \frac{1}{2}\ (1-2\varepsilon)(1+\varepsilon) = \frac{1}{2}\ (1-2\varepsilon+\varepsilon-2\varepsilon^2)$$

$$\cong \frac{1}{2}\ (1-\varepsilon)$$

or ,

$$E_{rms} \cong \sqrt{\frac{1}{2}\ (1-\varepsilon)} = \sqrt{\frac{1}{2}} \quad \sqrt{1-\varepsilon} \quad .$$

But, for small ε,

$$\sqrt{1-\varepsilon} = (1-\varepsilon)^{1/2} \cong 1-\varepsilon/2$$

and, therefore,

$$E_{rms} \cong \sqrt{\frac{1}{2}}\sqrt{1-\varepsilon} \cong \frac{1}{\sqrt{2}} \quad (1-\varepsilon/2) \quad .$$

Assuming no other errors, the accuracy required for a period measurement is one half the accuracy required for the rms value.

By definition, the rms value of a periodic waveform E(t) is given by

$$E_{rms} = \sqrt{\frac{1}{T}\int_0^T [E(t)]^2 \, dt} \quad , \quad T \equiv \text{period} \; .$$

Although other methods are possible, calculating the rms value directly by means of numerical integration does not require an exact integral number of samples per period, and the calculation is unaffected by frequency modulation. Computer simulations have shown that numerical integration by the midpoint method provides good reproducibility for the value of the integral when measuring a signal with 100 samples/period and a signal-to-noise ratio of 30dB [11]. For midpoint integration,

$$\int_0^T f(t) \, dt \cong \sum_{k=1}^{N} f(t_k) \, \Delta t = \frac{b-a}{N} \sum_{k=1}^{N} f(t_k)$$

where a, b = limits of sample points

N = number of sample points

$$t_k = k\,\frac{(b-a)}{N} + a \qquad k = 1, \ldots, N \ .$$

For this particular implementation, 128 samples are nominally involved, and

$$\int_0^T [E(t)]^2\, dt \cong \frac{128-0}{128} \sum_{k=1}^{128} [E(t_k)]^2$$

or,

$$E_{rms} \cong \sqrt{\frac{1}{T} \sum_{k=1}^{128} [E(t_k)]^2} \ .$$

Calculation of E_{rms} involves finding the square root of a 32-bit binary number. An efficient algorithm for this purpose was used [12], and developed into an assembly language program. The execution time is approximately 27 ms. Also, when the exact period does not contain an integral number of samples, a small end correction is applied.

Because of the need to provide for measuring the ac component only of the input signal, let

$$E(t) = E_0 + A \cos t \qquad (\omega = 2\pi f = 1)$$

where E_0 = dc offset = $\int_0^T E(t)dt$

A = peak amplitude of ac input with period, $T = 2\pi$.

Then,

$$[E(t)]^2 = E_0{}^2 + 2E_0A \cos t + A^2 \cos^2 t$$

and

$$\int_o^T [E(t)]^2 dt = \int_o^T E_o^2 dt + \int_o^T 2E_oA \cos t \, dt$$

$$+ \int_o^T A^2 \cos^2 t \, dt$$

$$= E_o^2 [t]_o^T + 2E_oA \int_o^T \cos t \, dt$$

$$+ A^2 \int_o^T \cos^2 t \, dt$$

$$= E_o^2 T + 0 + A^2 \frac{T}{2} \quad .$$

Hence,

$$E_{rms}(\text{ac only}) = \frac{A}{\sqrt{2}} = \sqrt{\frac{1}{T} \int_o^T [E(t)]^2 \, dt - E_o^2}$$

or, by numerical integration,

$$E_{rms}(\text{ac only}) = \sqrt{\frac{1}{T} \sum_{k=1}^{128} [E(t_k)]^2 - \left[\sum_{k=1}^{128} E(t_k)\right]^2}$$

2.2.4 Harmonic Distortion

When the percent of distortion function switch is selected, the total harmonic distortion of the input signal is calculated during T_3 (see fig. 2).

The data samples from the second cycle (T2) are used to calculate the coefficients of the Discrete Fourier Transform (DFT) of the input signals by means of a 64-point "decimation-in-time" Fast Fourier Transform (FFT) [9,11]. Using the FFT algorithm, the dc, fundamental, and 30 harmonic components of the input signal are obtained. Although for some applications the measurement of the individual coefficients is required, usually the total harmonic distortion (THD) as a figure-of-merit is often most desired, or a measure of spectral purity. THD is obtained by squaring and summing the coefficients of the 30 harmonics, and dividing this sum by the sum of the squares of the harmonics plus the fundamental. This result is then multiplied by 100 and displayed as a percentage on the front panel. As specified by the definition of THD, the dc component of the input signal is not included in either of the sums.

In order to determine the coefficients of the fundamental and 30 harmonics of the input signal, the FFT algorithm requires the multiplication of the data samples by a so-called "twiddle" factor,

$$W_N^m = e^{-j\left(\frac{2\pi}{N}\right)m}$$

where N = total number of samples

m = frequency index .

This factor is convenient since the DFT of the input signal x(t) is given by

$$X(mf) = A(mf) = \sum_{n=0}^{N-1} \chi(nT_s)e^{-j2\pi mfnT_s}$$

where $\omega = 2\pi mf = \frac{2\pi m}{NT_s}$

T_s = sampling time interval

$m = 0, 1, \ldots., N-1$.

Substituting for f,

$$X(mf) = \sum_{n=0}^{N-1} x(nT_s) e^{-j\left(\frac{2\pi}{N}\right)mn}$$

and, using the twiddle factor W_N^m,

$$X(mf) = \sum_{n=0}^{N-1} x(nT_s) W_N^{mn} .$$

Thus, there are nominally N multiplications of the data samples $x(nT_s)$ by the twiddle factor W_N^m, and N complex additions involved in the DFT. The X(mf) are the amplitude coefficients of the transform, where

m = 0 corresponds to the dc component,

m = 1 corresponds to the fundamental component,

$\vdots$ $\vdots$

m = N-1 corresponds to the Nth harmonic component.

These coefficients are, in general, complex by virtue of the complex twiddle factors. Of particular interest, we note that

$$Wk = [W(N-k)]^*$$

where * denotes complex conjugate.

The proof for this equality is as follows:

$$
\begin{aligned}
[W^{(N-k)}]^* &= [e^{-j\left(\frac{2\pi}{N}\right)(N-k)}]^* \\
&= e^{j\left(\frac{2\pi}{N}\right)(N-k)} \\
&= e^{+j2\pi} e^{-j\left(\frac{2\pi k}{N}\right)} \\
&= e^{-j\left(\frac{2\pi k}{N}\right)} \\
&= W^k .
\end{aligned}
$$

Hence,

$$
\begin{aligned}
W^{o} &= [W^{N}]* \\
W^{1} &= [W^{(N-1)}]* \\
W^{2} &= [W^{(N-2)}]* \\
&\vdots \\
W^{(N/2)-1} &= [W^{(N/2\,+\,1)}]* \quad .
\end{aligned}
$$

For the DFT then,

$$
\begin{aligned}
X(f) &= X^{*}[(N-1)f] \\
X(2f) &= X^{*}[(N-2)f] \\
&\vdots \\
X[(N/2 - 1)f] &= X^{*}[N/2 + 1)f]
\end{aligned}
$$

provided that N is even. The real frequency components of the DFT are thus symmetrical about N/2, while the imaginary components are reflected about the horizontal axis as shown in figure 9 below.

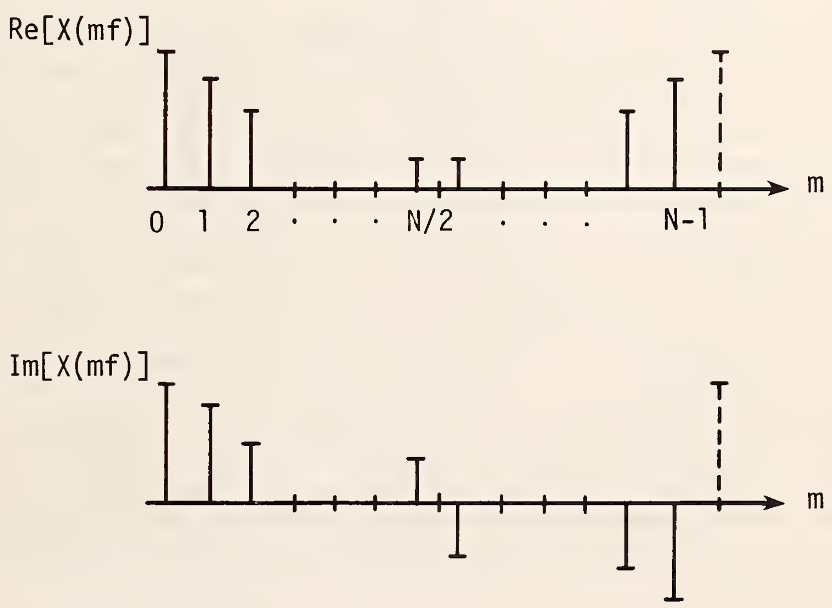


Figure 9. Real and imaginary components of the DFT.

The "negative" frequency terms of the continuous Fourier transform are analogous to the DFT coefficients beyond N/2, but less than or equal to N-1. This property is due to the period extension of the X(mf) spectrum since

$$X(0) = X(Nf)$$

$$X(f) = X[(N+1)f]$$

$$X(2f) = X[(N+2)f]$$

$$\vdots \qquad \qquad \vdots$$

$$X[(N/2 - 1)f] = X[(N + N/2 - 1)f] \quad .$$

Thus it is that with a 64-point FFT (N = 64), we can obtain the N/2 = 32 coefficients or amplitudes for the dc (m = 0), fundamental (m = 1), and 30 harmonics (m = 2, 3,...,31) of the input signal. Also, because of the complex conjugate nature of the 32 twiddle factors, the latter 16 values of W^k are the same as the first 16 with only the sign of the imaginary part changed. A 16 entry look-up table is included in the distortion program with a simple decoding scheme to determine the real and imaginary parts and to set the signs correctly. The total calculation time for the 64-point FFT with 16-bit resolution is typically 400 ms when executed on the IMP-16C.

3. CIRCUIT DESCRIPTION

The basic architecture of this instrument is described in section 2.1. The purpose of this section is to provide a more detailed description of the functional blocks comprising the low-frequency sampling voltmeter as shown in figure 1. In the discussion of these circuits, reference is often made to the various schematic drawings given in appendix C. Photographs, corresponding to each printed circuit board, are given in appendix D. As can be seen in photo 8, which is a back view of the voltmeter with the back panel removed, a card cage is used to contain the IMP-16C microcomputer board, the memory, timing, and control I/O board, and the data acquisition module or A/D board (from top to bottom, respectively). Photo 7 is a front view with the front panel dropped down to show the backplane of the card cage. The display board and the shielded enclosure containing the preamplifier (which are both mounted on the front panel) also can be seen in this photo. Table 1, on the following two pages, provides a back panel wiring diagram which identifies the signal name and how it relates to the pins on the connectors (J1, J3, and J5) used for the CPU, I/O, and A/D boards.

Table 1

Backplane wiring diagram

Name	J1 (CPU)	J3 (I/O)	J5 (A/D)
AD0	10	66	
INTRA	15	56	
AD7	18	80	
AD6	20	78	
AD2	23	68	
AD1	24	60	
AD3	25	91	
BD8	27	24	
AD5	28	74	
AD4	29	76	
MD0	40	128	
MD1	41	130	
MD2	43	132	
MD3	44	126	
MD4	47	124	128
MD5	48	122	126
MD6	50	120	124
MD7	52	118	122
BD5	56	30	
BD1	58	42	
BD4	59	28	
BD3	61	38	
MD14	62	104	108
BD7	63	26	
BD2	64	40	
BD6	65	32	
BD9	67	22	
BD11	68	18	
BD14	69	12	
BD15	70	10	
BD10	73	16	
BD13	74	14	
BD12	75	20	
MD10	76	116	120
MD13	77	106	110

Table 1 (cont.)

Name	J1 (CPU)	J3 (I/O)	J5 (A/D)	
BDO	60	44		
MD12	78	108	112	
MD8	82	112	116	
MD9	87	114	118	
MD11	88	110	114	
AD13	101	96		
MD15	108	102	106	
AD11	109	90		
AD10	110	86		
AD9	112	82		
AD14	115	94		
WRMP	120	54		
CLK	128	48		
JC12	129			GND
RAM SELECT	21	136		
AD15	116	92		
AD12	114	100		
CD15	36			GND
JC14	104			GND
JC15	107			GND
INTCTL, INTEN	85, 131			
CPINT	118			GND
JC13	119			GND
JC12	129			GND
AD8	111	84		
RLYA		133	133	
RLYB		131	131	
RLYC		135	135	
25T		88	88	
27T		129	82	
158, 14T, 17T		136	70	
A/D MVX ADR			74	GND
" " "			76	GND
" " "			78	GND
" " "			80	GND
" " "			68	GND
+15 VDC			11	
-15 VDC			17	
Analog GND			13, 15	
+12 VDC		52		
-12 VDC	31			

3.1 Microcomputer Board

The IMP-16C is a 16-bit parallel processor configured around National Semiconductor GPC/P (General Purpose Controller/Processor) MOS/LSI devices. These MOS/LSI devices are two read-only-memories (ROM) and four register/arithmetic logic units. Since each logic unit handles four bits, 16-bit processing is handled by connecting four units in parallel. Memory on the IMP-16C microcomputer card consists of 256 words of RAM. Data (BD-bus) from/to the peripheral devices and add-on memory (MD-bus) are routed through an input multiplexer on the IMP-16C card to the central processing unit (CPU). Data from/to the on-card memory is also processed in this way. Address data is available via the 16-bit AD-bus.

All the peripheral devices to the IMP-16C microcomputer (sampling rate generator, A/D converter, display, etc.) are memory mapped. The following discussion briefly describes each peripheral and its function, and indicates its associated memory addresses. A linear address scheme is used to address these peripheral devices. Address data from the microcomputer's AD-bus is decoded on the NBS 722 7909A board (IC 31, 32, 33, and 34), and the decoded signals are routed from there to the rest of the system (see schematic 4). Data from the IMP-16C are latched into IC 13, 14, and 15 of NBS 722 7909A for the peripheral devices via the BD-bus. Latching occurs coincident with the write memory pulse (WRMP) which is generated during all write to memory cycles of the IMP-16C microcomputer.

3.2 Sampling Rate Generator (Schematic 4, NBS 722 7909A)

Address 8800 Hex - Clock data for the sampling rate generator is latched from the BD bus, coincident with hex address 8800 and the WRMP pulse. The sampling rate generator, functioning as a programmable interval timer, uses the data word to control the period of the conversion pulses output to the A/D converter.

Clock data appears on the BD bus in the following format:

Bit position	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Data	PS	PS	PS	M	M	M	M	M	M	M	M	M	M	M	M	M

where PS = frequency prescaler bits

M = mantissa .

The 16-bit data word consists of two parts as shown: a 3-bit frequency prescaler and a 13-bit mantissa. The prescaler part is also latched into the frequency prescaler circuitry (IC 9, 10, 11, 12), which divides the clock input (5.76 MHz CPU clock) by 1, 4, 16, 64, 256, 1024, and 4069.

One of the output frequencies selected by the 3-bit prescaler code is sent to a down counter (IC 5, 6, 7, 8) preset by the 13-bit mantissa. Once initialized in this way, the down counter is preset after each zero count to the 13-bit mantissa value. The 13-bit mantissa is stored in latches (IC 1, 2, 3, 4) from the BD latches during the first 8800 hex address. At each zero count of the down counter, a 10 μs pulse is generated (IC 18) and sent to the A/D converter as a convert command (pin 129, J3).

3.3 A/D Converter (Schematic 5, NBS 722 79141A)

The A/D converter chosen for this application has 12-bit resolution (11-bits + sign) with a conversion time of 12 μs. The A/D converter is part of the data acquisition module (Datel MDAS-8D) which contains an analog multiplexer, an instrumentation amplifier, and a S/H amplifier. The instrumentation amplifier is designed with internal feedback resistors to allow three fixed gains (4, 2, 1), providing a full-scale output from the A/D converter for inputs of ±2.5 V, ±5 V, and ±10 V. These ranges are selected via the microcomputer by the switching of relays A, B, and C.

Address 8040 Hex - The control word to set the range for the A/D converter is latched into IC 21 of the NBS 722 7909A board from the BD latches on this board during hex address 8040 and the WRMP pulse. Outputs from IC 21 are used to drive the coils of the range setting relays (A, B, C) on the A/D board.

The following format is used on the BD bus:

Bit position	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Data	x	x	x	x	x	x	x	x	D	D	D	D	x	x	x	x

where x = don't care

DDDD = 0000 = 10 V range

DDDD = 0010 = 5 V range

DDDD = 0011 = 2.5 V range

The data acquisition module has an analog multiplexer on its front end with eight differential input channels. The input channel can be selected by performing the following action:

Address 8200 Hex - The 3-bit multiplexer address code is latched into the data acquisition module, coincident with this address and the WRMP pulse. The only multiplexer channel address presently used in the voltmeter is a code of all zeros. This code is hardwired at the data acquisition module multiplexer address points.

For outputs from the A/D converter, the following action applies:

Address 8400 Hex - Coincident with this address, the tristate outputs of the data acquisition module are enabled and converter data is placed on the MD bus (MDO 4-15).

3.4 Program Memory (Schematic 4, NBS 722 7909A)

The entire program code for controlling the operations of the Low-Frequency (LF) Sampling Voltmeter is contained in EPROM (IC 29, 30). The EPROMs are Intel type 2716s configured as a 2K by 16-bit memory.

Address 3000 - 3800 Hex - The 2716 EPROM select lines are enabled with these addresses. For startup, these addresses are also mapped to F800-FFFF.

3.5 Preamplifier (Schematic 2, NBS 722 79284A)

The preamplifier hardware contains a modular instrumentation amplifier (Analog Devices 606J) which is gain-controlled by relay-selectable resistors. The AD606J amplifier was chosen mainly for its low noise specification. Decade range selection for the voltmeter is accomplished with different preamplifier and data acquisition module gain settings. The preamplifier scales input signals in the range from 10 mv to 10 V rms (full scale) to a level of 7 V rms. For this purpose, four-decade gain steps are provided (by means of the resistors, trimmable potentiometers, and relays) with gains of 700, 70, 7, and 0.7. The input signal may be either single-ended or differential with a common mode voltage of less than 1 V.

To correct for changes in offset of the AD606J amplifier with different gain settings, an auxiliary null adjustment circuit is required. An operational amplifier (Teledyne TP 1319) is used to supply a small positive or negative voltage to correct for the variable bias-current induced offsets when the gain-setting resistors are changed.

Address 8020 Hex - The gain control code for the preamplifier is latched into IC 15 of the NBS 722 79026A display board, coincident with the WRMP pulse and this address. The output of the latch is then buffered to drive the coils of the gain controlling relays.

The following format is used on the BD bus:

Bit position	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Data	x	x	x	x	x	x	x	x	x	x	x	x	D	D	D	D

where x = don't care

DDDD = 0111 = relay A, B, C activated - 10 V range

DDDD = 0100 = relay C activated - 100 mV range

DDDD = 1000 = relay D activated - 10 mV range

DDDD = 0000 = no relays activated - 1 V range .

3.6 Front Panel Numerical Display/LEDs (Schematic 3, NBS 722 79026A)

Address 8080 Hex - Numerical display data is latched into IC 2, 3, 4, 5, and 6, coincident with the WRMP pulse and this address. Selection of the appropriate latch occurs simultaneously, with IC 8 used as a digit decoder. Decimal point data reside in the IC 6 latch while first, second, third, and fourth digit numerical data reside in IC 2, 3, 4, and 5, respectively.

The following format is used on the BD bus:

Bit position	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Data	x	x	x	x	x	x	x	x	D	D	D	D	I	I	I	I

where x = don't care

IIII = numerical data code (0-9 BCD)

DDDD = digit to be selected (1-6 BCD) .

Coincident with all IC 6 updates is a momentary pulsing of LED 13 (display update indicator) through IC 14.

Address 8010 Hex - LED display data is latched into IC 10 and 11 of the display board, coincident with the WRMP pulse and this address. The latch outputs are buffered by IC 12 and 13 which drive the LEDs.

The following format is used on the BD bus:

Bit position	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Data	x	x	x	x	12*	11*	10*	9*	8*	7*	6*	5*	4*	3*	2*	1*

where x = don't care

* - each numerical digit represents an LED position on the display board (see schematic 3). If a logic one or zero exists in any of these positions, there is LED activation or no LED activation, respectively.

3.7 Front Panel Switches (Schematic 6)

Address 8100 Hex - Data representing the front panel switch positions is read from latches IC 26 and 27 of the NBS 722 7909A board, coincident with this address. The switches are connected to these latches via the external 50 line connector and cable from the CPU board to the display board. Switch data is strobed into the latches automatically (by IC 20 on the NBS 722 7909A board) when these latches are addressed by the CPU.

4. OPERATING INSTRUCTIONS

4.1 Installation

The NBS Low-Frequency Sampling Voltmeter mounts onto a standard 19-inch equipment rack. Ensure adequate ventilation. The voltmeter is powered from a nominal 117 V, 60 Hz ac line.

4.2 Controls

4.2.1 Power Switch

The power switch is the toggle switch located on the front panel (see photo 1).

4.2.2 RMS AC + DC, RMS AC, Freq., Dist.

These functions are performed and displayed individually by the voltmeter by pushing the appropriate front panel button (see photo 1). If the signal has already been obtained (that is, a number is displayed on the front panel LED numerical display) and the user desires to see another function, the voltmeter will blank out the display momentarily to perform the new operation once the appropriate pushbutton has been depressed. If either the frequency or amplitude of the incoming signal varies significantly, the display will blank out and take two signal periods before new data is displayed (see section 2, Theory of Operation). The voltmeter averages the last computed rms value with the previous 16 values of the incoming signal stored in RAM on the IMP-16C microcomputer board. If the latest value differs by more than ±1 percent from the running average, this value is then displayed and a new cycle of averaging readings is begun. Consequently, the operator must wait for the voltmeter readings to stabilize if the rms value is changing, particularly at frequencies below 1 Hz. Also, the operator is cautioned to be aware that this averaging feature does not permit simple multiplexing of the input of the voltmeter to several sources.

4.2.3 10 mV, 100 mV, 1 V, 10 V Pushbuttons

The operator selects the range manually by pushing the appropriate button. This action will activate an entire measurement sequence for the voltmeter (see section 2, Theory of Operation) with the internal preamplifier set to the appropriate range value. If the instrument cannot make a measurement on that selected range, it will automatically sequence to the correct range (see section 4.2.4, Autoranging). When the incoming signal level (as indicated by the numerical display) is within the full-scale value of the next lowest range (see section 4.3.6), a more accurate reading can be obtained by manually selecting this lower range.

4.2.4 Autoranging

This feature exists as an internal software control for the instrument to automatically select the appropriate range for measurement purposes and will override an inappropriate range setting the user inadvertently selects. However, this operation takes time, extending the initial measurement sequence, as the instrument must range up over each of the possible ranges from an initial range setting.

4.2.5 Reset

Pushing this button causes an entire measurement cycle to be initiated regardless of anything else.

4.3 Indicators

All controls are indicated active by the voltmeter when the LED corresponding to a given control is turned ON.

4.3.1 Reset

The LED indicator for this function is always ON when the period of the incoming signal has not been determined and sampling has not been successful (see section 2, Theory of Operation). Otherwise, this LED is OFF indicating that the period of the incoming signal is known and samples have been taken.

4.3.2 Display

This LED is momentarily ON during each update to the display. If the Reset LED is OFF, the Display LED will be activated at a fixed rate depending upon the frequency of the incoming signal.

4.3.3 Overload

This LED indicates that too large a signal (>10 V rms) or too large an offset (>10 V) is present at the input terminals.

4.3.4 RMS AC + DC, RMS AC, Freq., Dist.

The instrument will indicate which function is being performed by activating the appropriate LED above the function pushbutton.

4.3.5 Front Panel Numerical Display

This display is a five digit LED numerical display. If the function is frequency, the display will be in hertz. If the function is Dist. (Distortion), the display is in percent.

If the function is rms ac + dc or rms ac, the function is in volts. Volts or millivolts are indicated by the two LEDS located near the front panel LED numerical display.

4.3.6 10 mV, 100 mV, 1 V, 10 V Ranges

The instrument indicates on what range it is currently operating by activating the appropriate LED.

4.4 Signal Input

The incoming signal is applied to the instrument via the front panel banana plugs. These plugs route to a differential amplifier with HIGH indicating the high side of the differential amplifier and LOW indicating the low side. GND indicates instrument ground.

The maximum signal between HIGH and LOW terminals is 10 V rms for a measurement to be obtained. The maximum offset between the LOW and GND terminals is 10 V for a measurement to be obtained.

5. ERROR ANALYSIS

The performance of the Low-Frequency Sampling Voltmeter described herein is dependent upon both the software implementation of the measurement algorithms described in section 2.2 and the hardware implementation described in section 3. The algorithms assume ideal sampling of the input signal and the conversion to digital values. A brief discussion is given here of the measurement errors due to offset and gain in the preamplifier and data acquisition channel, quantization amd nonlinearity of the A/D converter, and timing jitter in the S/H amplifier and sampling rate clock. A more complete analysis of these errors is provided in [11].

5.1 RMS Zero and Gain Error

The preamplifier circuit and data acquisition module (instrumentation amplifier, S/H, and A/D converter) are described in sections 3.3 and 3.5. In this analysis, it is assumed that the transfer function of the data acquisition module is essentially linear for the 2.5, 5, and 10 V ranges provided. Hence, the accuracy of this data acquisition hardware is dependent on how well two points on the transfer function (generally zero and full scale) can be adjusted with the trimpots provided on the module.

The accuracy of the data acquisition module was tested by measuring dc voltages produced by a stable source. The output from the source was also measured by a digital voltmeter, calibrated to an accuracy of better than 50 ppm. Before adjusting the offset (zero) and gain controls, 100 readings were taken from the A/D converter at each of several voltages on the three different ranges and compared to corresponding readings of the DVM. This data is summarized in the plots given in figure 10. After the data were taken, the gain and offset errors were then adjusted to near zero for the three ranges. The resultant gain and offset errors, compiled from averaging 400 readings at each point, are listed in table 2 below. As can be seen, considerable improvement can be made in the unadjusted error by trimming the offset and gain controls, although with only one set of adjustments, all the zero and gain errors cannot be simultaneously adjusted to zero.

Table 2

Final gain and offset errors after adjustment of the A/D converter

	Input voltage (volts)	Error (% of fs)	Gain error (% of fs/fs)	Offset error (% of fs)
5 V range	-4.9963	-0.00098		
	+4.9988	-0.00220		
			-0.00122	-0.00159
2.5 V range	-2.4994	-0.00293		
	+2.4984	-0.06641		
			-0.06347	+0.00347
10 V range	-9.99760	+0.06592		
	+9.97803	+0.05786		
			-0.00806	+0.06189

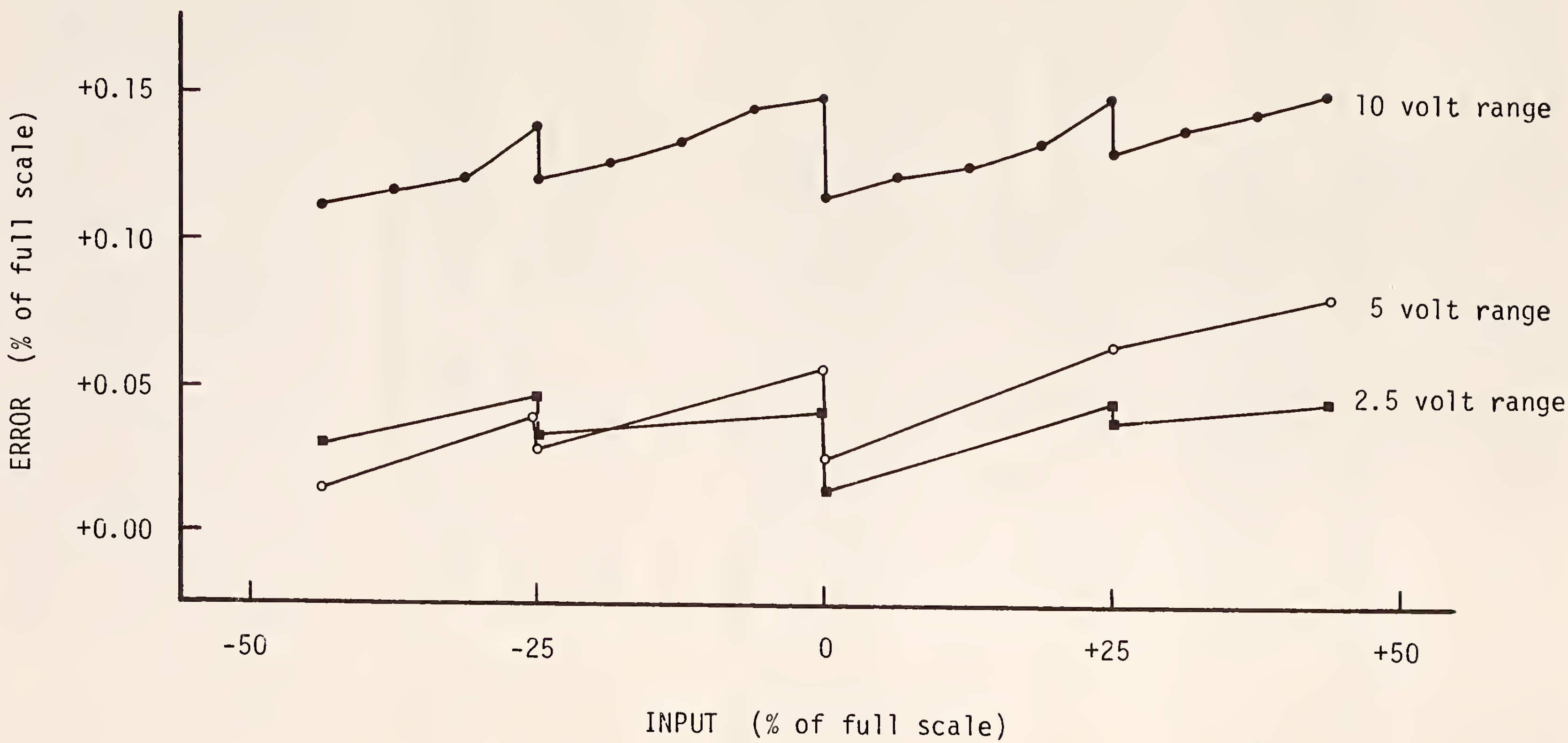


Figure 10. Measured error of the A/D converter before adjustment of the gain and offset errors.

Appendix A provides a calibration procedure (section A.1.2) for adjusting the "BAL" and "NULL" trimpots of the preamplifier to compensate for its offset on all four ranges of the voltmeter. Following this procedure will provide preamplifier offset errors of 0.1 percent of full scale or less.

The residual error at the ± full-scale limits of any of the four voltmeter ranges can thus be translated into an offset (zero) error ε_o and gain error ε_g for the instrument. These errors, in combination with the magnitudes of the ac (M_{ac}) and dc (M_{dc}) components of the signal, produce the following expression for the rms (zero and gain) measurement error:

$$\text{rms error} = \frac{\sqrt{[(1 - \varepsilon_g)M_{dc} + \varepsilon_o]^2 + [(1 - \varepsilon_g)M_{ac}]^2}}{\sqrt{(M_{dc})^2 + (M_{ac})^2}} - 1$$

where

ε_g = gain error in percent of full scale/full scale

ε_o = offset error in percent of full scale

M_{dc} = dc component of signal

M_{ac} = ac component of signal .

Generally, M_{dc} is some fraction of M_{ac} as is the offset ε_o, where M_{ac} is kept near full scale. This expression can thus be evaluated using the gain and offset errors for the data acquisition module and the preamplifier discussed above.

For a signal with $M_{dc} = 0.1\ M_{ac}$ (the maximum allowed by the signal model) and M_{ac} = full scale so that ε_o = offset error in percent of full scale (kM_{ac}), the rms error expression is then

$$\text{rms error} = \frac{\sqrt{[(1-\varepsilon_g)0.1\,M_{ac} + kM_{ac}]^2 + [(1-\varepsilon_g)M_{ac}]^2}}{\sqrt{(0.1\,M_{ac})^2 + M_{ac}^{\;2}}} - 1$$

$$= \frac{\sqrt{[(1-\varepsilon_g)0.1 + k]^2 M_{ac}^2 + (1-\varepsilon_g)^2 M_{ac}^2}}{\sqrt{1.01\,M_{ac}^2}} - 1$$

$$= \frac{\sqrt{[(1-\varepsilon_g)0.1 + k]^2 + (1-\varepsilon_g)^2}}{\sqrt{1.01}} - 1$$

where

$k = \varepsilon_0$ = offset error in percent of full scale

$= \varepsilon_0$ (preamplifier) $+ \varepsilon_0$ (data acquisition module) .

Using the 10 V range gain and offset errors from table 2 (-0.008 percent and 0.062 percent, respectively) and ε_0 (preamplifier) = 0.1 percent, the rms error can be calculated to equal 0.024 percent. Using the 2.5 V range gain and offset errors (-0.063 percent and 0.00347 percent, respectively) and ε_0 (preamplifier) = 0.1 percent, the rms error can be calculated to equal 0.073 percent. Ideally, if the gain error can be assumed to be negligible (see appendix A, section A.1.3), then a further simplification gives

$$\text{rms error} = \frac{\sqrt{(0.1 + \varepsilon_0)^2 + 1}}{\sqrt{1.01}} - 1, \text{ for } \varepsilon_g = 0 \quad .$$

The same total maximum offset error of 0.162 percent of full scale (10 V range) will then cause an rms error of only 0.016 percent.

5.2 Quantization Error

The measurement error of the voltmeter is also due in part to the error attributed to finite quantization of the input signal by the A/D converter. Under certain conditions, this error can be approximated by random noise with zero mean and variance (standard deviation squared) of $q^2/12$, where q is the smallest quantization interval, or one least significant bit (LSB) of the A/D converter [13]. The assumption for this result to be true is that when the signal is located within a particular quantization interval, the difference (or error) between the continuous-time signal and the particular quantization level is uniformly distributed over the quantization interval (i.e., a uniform error probability distribution).

The rms value of the input signal E(t) has been described earlier (section 2.2.3) to be given by

$$E_{rms} = \sqrt{\frac{1}{T} \sum_{k=1}^{N} [E(t_k)]^2}$$

where there are exactly N samples within one period T (nominally 128 in this case). Or, letting the values of E(t) at time values t_k be denoted by E_k,

$$E_{rms} = \sqrt{\frac{1}{T} \sum_{k=1}^{N} (E_k)^2} \quad .$$

With added random noise, R_k, the rms value of the signal plus noise is then

$$E_{rms} = \sqrt{\frac{1}{T} \sum_{k=1}^{N} (E_k + R_k)^2}$$

which, for small R_k, is approximately

$$E_{rms} \cong \sqrt{\frac{1}{T} \sum_{k=1}^{N} E_k^2 \left(1 + \frac{2R_k}{E_k}\right)^2} \quad .$$

It can then be shown that the variance (σ^2) of E_{rms} due to the random noise is $\frac{(q')^2}{12}$ where $q' = 2q$, or $\sigma^2 = \frac{4q^2}{12}$. An initial criterion for determining the A/D converter resolution was that the (random) quantization noise contributes no more than 0.05 percent error (3σ limit) to the noise-free rms value for a reasonable number of samples. Hence, if we assume 100 samples per period, then

$$3\sigma = 0.0005 \sqrt{N}$$

or,

$$\sigma^2 \doteq \left[\frac{0.0005\ \sqrt{N}}{3}\right]^2 = \frac{4q^2}{12} \quad .$$

Thus, q = 0.002887 which requires a converter with 9 bits (plus a sign bit for bipolar operation).

To verify these calculations, however, a series of computer simulation tests were performed. For these tests, 20 sine waves with rms values of unity and with random phase shifts (Start Angle) were generated, quantized to a specific number of bits (8, 9, and 12). The BASIC program and printouts for several sets of simulation calculations are contained in [11]. For each set of 20 calculations (20 different starting angles to randomize the quantization error), the rms value was calculated, based on 10 and 100 samples. The results of these calculations are tabulated below:

Converter resolution (including sign bit)	RMS error	
12 bits	0.0000950	1σ limit for 10 samples
9 bits	0.0006533	
8 bits	0.0013	
8 bits	0.0005013	1σ limit for 100 samples

These values agree rather well with the $\frac{4q^2}{12}$ prediction.

The wide range of input signals (2 mV to 10 V) requires the preamplifier described before, having adjustable gains of 1, 10, and 100 to normalize the signal before it is applied to the data acquisition module. Even so, not all signals will span the range of the instrumentation amplifier, S/H circuit, and A/D converter of the data acquisition module; however, at least 10 bits of resolution are needed for the smallest signal, as discussed above. A 12-bit converter was chosen, therefore, because of the required resolution and accuracy. The gain setting of the instrumentation amplifier in the data acquisition module must then be no larger than four times the next lower setting, so that the signal will always use full scale/4 = $\frac{2^{12}}{2^2} = 2^{10}$ or 10 bits of the converter. In practice, this scaling is accomplished by having gains of 1, 2, and 4 in the instrumentation amplifier (data acquisition module). With a maximum change of 2.5 between ranges, the maximum random rms measurement error due to finite quantization is then

$$\text{random rms error} \leq \frac{3\sigma}{\sqrt{N}} = \frac{3\left(\frac{q}{\sqrt{3}}\right)}{\sqrt{N}} = \sqrt{\frac{3}{N}}\left(\frac{2^{12}}{2.5}\right)^{-1}$$

$$= \sqrt{\frac{3}{10}}\left(\frac{1}{2^{10.7}}\right) = 0.033 \text{ percent .}$$

5.3 A/D Converter Nonlinearity Errors

The A/D converter in the data acquisition module is of the successive approximation type. This type of converter exhibits both linearity and differential linearity errors. Linearity errors indicate that the digital output produced by the converter is not exactly a linear function of the input voltage, whereas differential linearity errors are a measure of the variation in input signal change associated with a one-bit change in the digital output. These errors vary, depending on the particular location along the converter's transfer characteristic.

Figures 11 and 12 show plots from linearity tests made on two different data acquisition modules of the type used in the Low-Frequency Sampling Voltmeter. With the input multiplexer set for one channel and the S/H amplifier operating in the nominal track and hold mode, 1024 evenly spaced output codes of the 10 most significant bits of the A/D converter were tested by measuring their associated transition edge voltages [17]. As described in [17], the transition edge voltages (where adjacent output codes are obtained on a 50 percent basis) accurately define the transfer characteristic.

The center plot is of the measured error data of the data acquisition module, obtained by comparing the transition edge voltages of the module with the output voltages from a 20-bit reference D/A converter (used in the test set), at corresponding digital codes. In the case of figure 11, the maximum linearity errors of approximately ±1 LSB occur at position 777, or $\left(\frac{777-512}{512}\right) \times 10\text{V} \cong 5.2$ V, and at position 241, or $\left(\frac{512-241}{512}\right) \times (-10\text{ V}) \cong -5.3$ V. In figure 12 the linearity of this particular module is seen to be somewhat better. The rms value of the error data is 0.275 LSB as compared to 0.392 LSB for the module of figure 11.

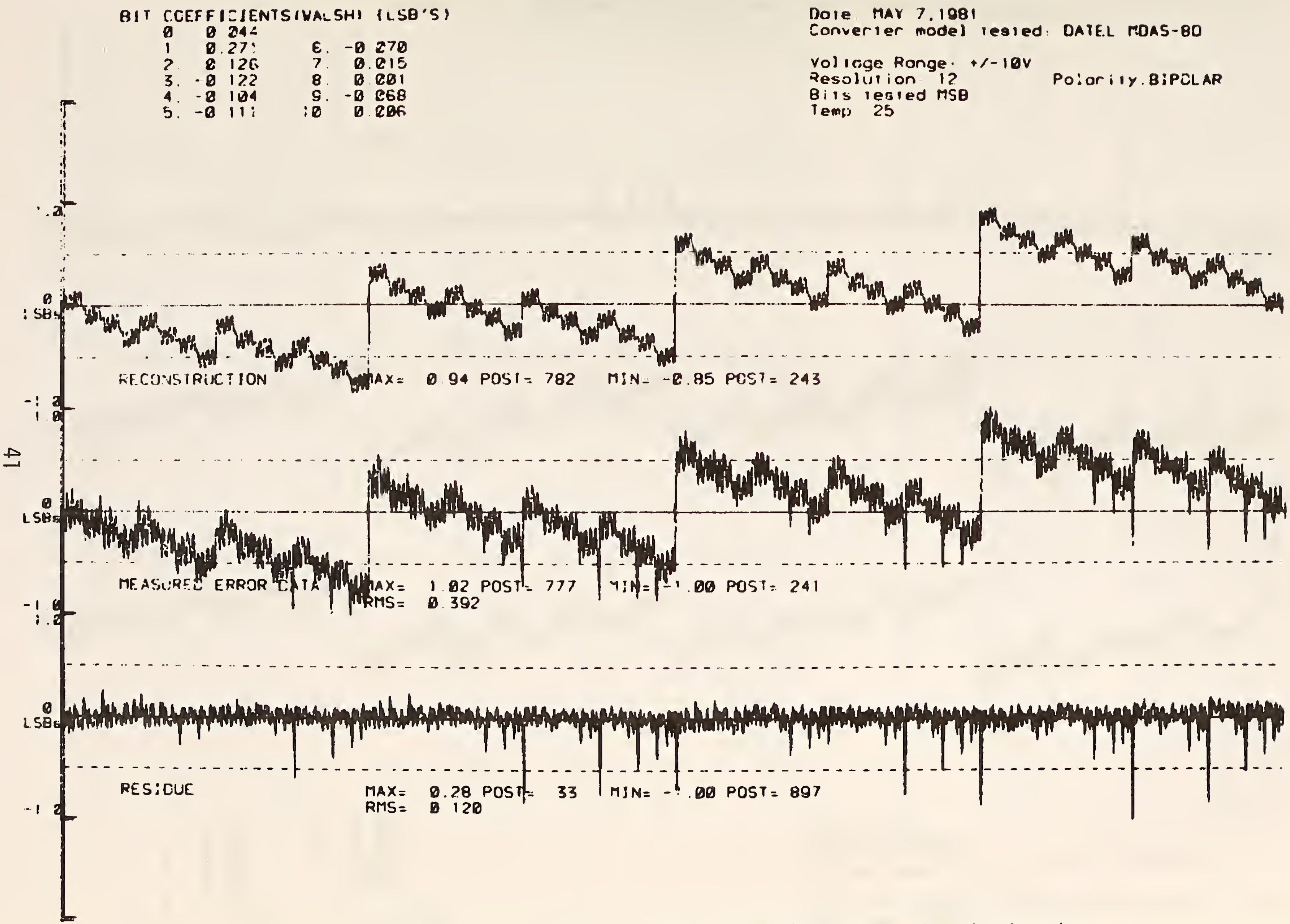


Figure 11. Linearity plots from data converter test set, Module 1.

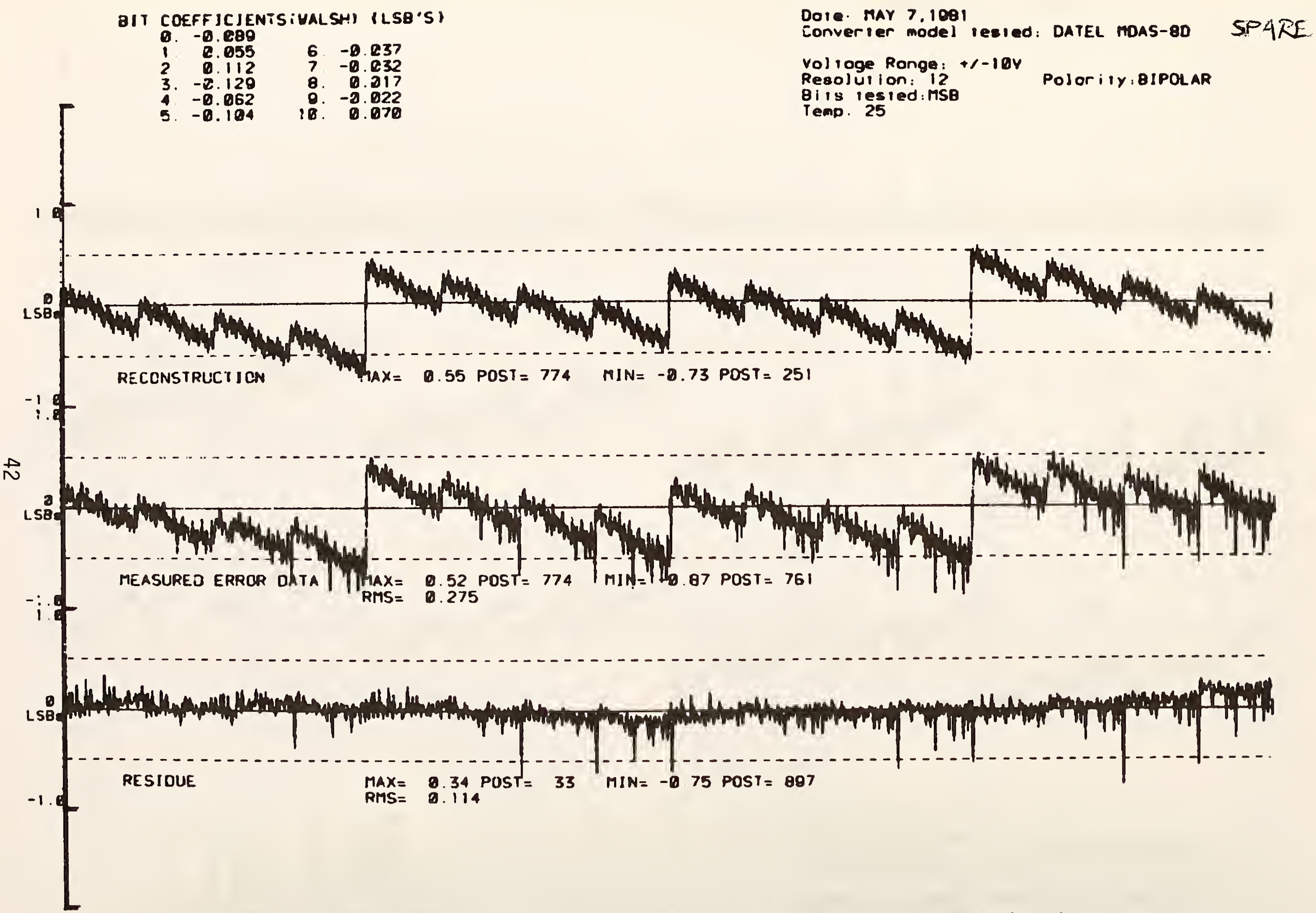


Figure 12. Linearity plots from data converter test set, Module 2.

Significant discontinuities in these plots are seen to occur at half scale and at one-quarter and three-quarters scale, which are major code transition points. Assuming that the linearity error data would not change significantly had all 12 bits of the converter been exercised, an estimate of the worst case differential linearity error can be made from the amplitude of these discontinuities. From the plots for both modules tested, it can be seen that the maximum differential linearity error for these units can be estimated to be 1 LSB.

A good approximation of the measurement error contributed by the nonlinearity of the data acquisition module (primarily, the A/D converter) can be made from the test data in figures 11 and 12. In particular, the worst case error due to either linearity or differential linearity is ~1 LSB or,

$$\text{nonlinearity error (max)} = 1/(2^{12}) \times 100 \text{ percent of full scale}$$

$$= 0.0244 \text{ percent .}$$

For an explanation of the top and bottom plots shown in figures 11 and 12, the reader is referred to [17]. Briefly, these plots are the results of further analysis of the linearity error data. A Walsh transform is made on this error data which is taken to represent a uniformly sampled function having a period of 2^N samples where N, in this case, is 10. Walsh series functions are a set of orthogonal functions analogous to a Fourier series, differing primarily in that Walsh functions are square-like rather than sinusoidal. A reconstruction (inverse transform) of the error data can then be made (top plot) which is a minimum mean squared error representation of the actual measured error data (center plot). Finally, the difference between these two data sets is used to plot a residue function (bottom plot). This function is representative of the effective coupling between bits, or superposition errors. In general, these errors establish a practical limit beyond which no simple adjustments or corrections to the converter are useful.

5.4 Timing Errors

Besides the types of direct amplitude errors described in sections 5.1, 5.2, and 5.3 above, the timing (sampling) jitter in the S/H amplifier and in the sampling rate clock can be translated into their equivalent amplitude errors. Errors associated with the S/H amplifier that can contribute errors when measuring the rms value of a signal are aperture time jitter, acquisition time, and droop rate [14,15]. This same type of sampling time uncertainty also applies to the frequency instability of the sample rate clock.

For this sampling voltmeter application, the aperture time itself is unimportant so long as it remains constant. The aperture time jitter, however, will appear as a random noise voltage superimposed on the input signal samples. The maximum magnitude of this noise can be estimated from the aperture time jitter and the maximum first derivative of the input signal [16]. Letting $E(t) = A \sin \omega t$, then

$$E'(t) = \omega A \cos \omega t$$

or,

$$E'(t)_{max} = \omega A = \frac{\text{amplitude jitter}}{\text{aperture time jitter}}$$

Representing amplitude jitter as kA, then

$$\omega A = \frac{kA}{\text{aperture time jitter}}$$

or,

$$k = \omega \text{ (aperture time jitter)} .$$

Since the S/H amplifier used in the voltmeter has an aperture time jitter of 50 ns less,

$$k_{(max)} \leq 2\pi \text{ (120 Hz)}(50 \times 10^{-9})$$

$$\leq 0.000377 .$$

Hence, the maximum noise jitter is less than 0.038 percent of the peak input amplitude which corresponds to an rms noise error of 0.027 percent.

Acquisition time error occurs when the S/H amplifier is not given sufficient time during the sample mode to allow the output delay and settling time to occur when acquiring the input signal again. The droop rate error of the S/H amplifier is a function of acquisition time, as well as of present input magnitude, previously held input magnitude, and length of required hold time. The output signal of the S/H amplifier must be held within ±1/2 LSB for the duration of the A/D conversion process in order to introduce zero (or negligible) error. For the 12-bit A/D converter used in the data acquisition module, this requirement translates into maintaining the held value to an accuracy of ±0.01 percent for approximately 12 μs.

The frequency instability of the sample rate clock contributes an amplitude error in the same way as aperture time jitter, discussed above. That is, the conversion signal from the sample rate clock may have a random jitter superimposed on an otherwise uniform timing interval that manifests itself as a random noise voltage on the input signal samples. The maximum magnitude of the rms voltage noise can be estimated from the rms value of the clock jitter and the maximum value of the first derivative of the input signal [13]. That is,

$$\text{rms error (clock jitter)} \leq \sigma \text{ (max first derivative)}$$

where σ = rms value of the clock jitter

or, rms error (clock jitter) $\leq \sigma$ (ωA) which is equal to $\sigma\omega$ as a percent of the input signal amplitude A. Because the sampling clock rate is a fixed multiple of the input signal frequency (128x), the rms error due to clock jitter is constant. For a 10 Hz signal the rms value of the clock jitter (σ) was measured as 140 ns, which translates to

$$\text{rms error (clock jitter)} \leq 140 \times 10^{-9}(2\pi\ 10)$$

$$\leq 0.00088 \text{ percent .}$$

Because of the integrated construction of the data acquisition module used, it is not possible to directly evaluate the dynamic characteristics of the internal S/H amplifier. However, a test was performed on the instrumentation amplifier, S/H amplifier, and A/D converter combination by applying a low distortion sine wave source (<0.02 percent THD) to the module, and least squares fitting the digital output data to a pure sinusoid. After correcting for zero and gain errors, the standard deviation for the fit was 0.02 percent. Since this value is comparable to the distortion in the source itself, the previous error estimates appear to be reasonable.

6. INSTRUMENT PERFORMANCE

The resulting performance obtained with the initial LF Sampling Voltmeter prototype and the seven subsequent models (one NBS unit and six for use in the DoD Metrology Centers) will be described in this section. These evaluation tests show typical performance data and do not represent the results of a set of statistically based performance tests.

6.1 Frequency (or, Period) Tests

A commercially available frequency counter with an accuracy of better than 0.01 percent (Tektronix Model DC505) was used to check the accuracy of the frequency (or, period) measuring function. The data shown in table 3 below are from readings taken with the initial LF Sampling Voltmeter prototype [11]:

Table 3

Calibration data for the frequency function of the LF Sampling Voltmeter using an 0.01 percent accurate Tektronix 505A frequency counter (in the period measuring mode)

505A frequency (Hz)	LF Sampling Voltmeter (Hz)	Error (% of reading)
0.10121	0.101	-0.21[a]
1.0123	1.012	-0.03
10.125	10.12	-0.05
14.030	14.03	0.00
30.000	30.01	+0.03
44.503	44.50	-0.01
48.026	48.07	+0.09
55.026	55.09	+0.12
56.363	56.40	+0.07
57.741	57.81	+0.12
60.68	60.70	+0.03
68.743	68.76	+0.03
78.380	78.42	+0.05
102.04	102.1	+0.06
119.00	119.0	0.00

[a]Resolution of 1 percent limited the accuracy of the reading.

These data seem to indicate that the target accuracy of ±0.1 percent of reading is achievable for most frequencies. Limited display resolution prevents a good comparison at the 0.101 Hz value. The random error is most likely due to slight variations in the period as determined by the exact period error function described in section 2.2.2. The small systematic error is probably due to the design of the sample rate clock. A finite time is required to preset the counters in the sample rate clock circuit, after each output pulse, and this reset time adds a larger percentage error at higher frequencies. This effect was born out in tests made on the subsequent units, as shown in table 4 below:

Table 4

Frequency test on engineering model

505A frequency (Hz)	LF Sampling Voltmeter (Hz)	Error (% of reading)
0.0999	0.099[a]	-0.9[a]
1.0984	1.098	-0.04
3.594	3.594	0.00
11.017	11.02	+0.03
20.013	20.02	+0.03
30.050	30.070	+0.067
40.034	40.08	+0.11
68.070	68.18	+0.16
119.19	119.4	+0.18

[a]Resolution of 1 percent limited the accuracy of the reading.

Here we can see the greater effect of the reset time such that at higher frequencies (>40 Hz) the frequency displayed by the LF Sampling Voltmeter is slightly higher than the actual frequency.

6.2 RMS Voltage Tests

As part of the final evaluation of the initial LF Sampling Voltmeter prototype, this instrument was tested in comparison with a commercial 5-1/2 digit ac average responding digital voltmeter (Data Precision 2540A1). These tests were conducted using a high stability sine wave generator specially constructed for this purpose [11]. The requirements for the generator were good cycle-to-cycle amplitude and frequency stability, a frequency range of 0.1 Hz to 120 Hz, and an output adjustable in steps from zero to 7 V rms. Therefore, a digitally synthesized waveform was generated by sending a series of digital values from a read-only-memory (ROM) containing the digital values of a sine wave to a digital-to-analog converter (DAC). Another reason for using this method was the ability to generate a sine wave with a predetermined distortion so that the distortion measurement algorithm of the LF Sampling Voltmeter could also be checked. The generator design, detailed in [11], provided a sinusoidal source whose average ac voltage at 20 Hz was measured to be only 0.01 percent different than at 100 Hz. An active two-pole low-pass Butterworth filter (flat to within 0.1 percent from dc to 106 Hz) was utilized in the output design of the generator. This circuit effectively filters out "glitch" transients in the output of the 12-bit DAC (used to produce the stepped analog sine wave) and to smooth this output into the final sinusoidal form. Then, the calibrated sine wave generator was used to verify the performance of the 2540A1 at 20 Hz. Table 5, on the next page, shows the measurement data for comparison between the 2540A1 digital voltmeter and the prototype sampling voltmeter.

Table 5

Calibration of the rms ac + dc function of the voltmeter by the Data Precision 2540A1 voltmeter at 20 Hz using the waveform generator.

Amplitude setting	2540A1 (volts)	LF Sampling Voltmeter (volts)	Difference[a] (% of reading)
11	7.0700	overrange	-
10	6.4281	6.425	-0.01
9	5.7865	5.783	-0.03
8	5.1425	5.142	+0.02
7	4.5002	4.498	-0.01
6	3.8569	3.854	-0.04
5	3.2133	3.212	-0.01
4	2.5713	2.571	+0.02
3	1.9283	1.928	+0.02
2	1.2860	1.285	-0.04
1	0.6430	0.643	+0.03

[a]The calculated output of the waveform generator at 20 Hz is 7.0676 V for setting 11. A correction of -0.034 percent (calculated minus measured difference for 2540A1 at setting 11) is applied to all the 2540A1 readings before the differences between the two instruments are calculated.

As a further test of the ac rms measurement accuracy of the initial prototype, it was tested with the multi-junction thermal converter-based NBS AC Voltmeter/Calibrator developed at NBS [4,5]. This instrument contains a built-in, ROM-based ac voltage calibrator with ±0.02 percent of reading accuracy from 2 mV to 7 V rms for the 0.1 to 50 Hz frequency range. Table 6 shows test data on the LF Sampling Voltmeter prototype using the NBS AC Voltage Calibrator [11].

Table 6

AC rms tests of prototype LF Sampling Voltmeter

Frequency (Hz)	NBS AC Voltage Calibrator (volts)	LF Sampling Voltmeter (volts)	Error (% of reading)
50	6.384	6.386	+0.03
	5.070	5.070	0.00
	3.159	3.160	+0.03
10	3.159	3.161	+0.06
	5.072	5.074	+0.04

The data in tables 5 and 6 above indicate that the performance of the initial LF Sampling Voltmeter met the rms voltage measurement accuracy goal of ±0.1 percent of reading. However, an important additional performance verification test was subsequently made on the engineering prototype which contains the (four-decade gain setting) preamplifier. The description of this preamplifier is given in section 3.5. These seven engineering models have four input ranges of 10 mV, 100 mV, 1 V, and 10 V (full scale). Table 7, on the next page, shows test data on the engineering prototype of the LF Sampling Voltmeter with readings taken by the NBS AC Voltmeter/Calibrator.

Table 7

Tests of the LF Sampling Voltmeter (with preamplifier), using the NBS AC Voltage Calibrator Standard

	NBS AC Voltage Calibrator (volts)	LF Sampling Voltmeter (volts)	Error % of reading
50 Hz	6.384	6.383	-0.015
	3.159	3.159	-0.00
	2.242	2.243	+0.04
10 Hz	6.384	6.382	-0.03
	3.159	3.157	-0.06
	2.242	2.242	0.00
1 Hz	6.384	6.382	-0.03
	3.159	3.158	-0.03
	2.242	2.243	-0.04
0.1 Hz	6.384	6.384	0.00
	3.158	3.158	0.00
	2.242	2.242	0.00
50 Hz	0.6384	0.6381	-0.05
	0.3159	0.3158	-0.03
	0.2242	0.2242	0.00
10 Hz	0.6384	0.6384	-0.00
	0.3159	0.3158	-0.03
	0.2242	0.2241	-0.04
1 Hz	0.6384	0.6382	-0.03
	0.3159	0.3158	-0.03
	0.2242	0.2242	0.00
0.1 Hz	0.6384	0.6384	0.00
	0.3159	0.3158	-0.03
	0.2242	0.2242	0.00

Table 7 (cont.)

	NBS AC Voltage Calibrator (millivolts)	LF Sampling Voltmeter (millivolts)	Error % of reading
50 Hz	63.840	63.85	+0.016
	31.590	31.58	-0.03
	22.420	22.42	0.00
10 Hz	63.840	63.83	-0.015
	31.590	31.58	-0.03
	22.420	22.43	+0.04
1 Hz	63.840	63.84	0.00
	31.590	31.58	-0.03
	22.420	22.42	0.00
0.1 Hz	63.840	63.84	0.00
	31.590	31.60	+0.03
	22.420	22.44	+0.089
50 Hz	6.381	6.378	-0.05
	3.159	3.158	-0.03
	2.242	2.243	+0.04
10 Hz	6.390	6.388	-0.03
	3.162	3.160	-0.06
	2.241	2.241	0.00
1 Hz	6.391	6.389	-0.03
	3.162	3.160	-0.06
	2.241	2.241	0.00
0.1 Hz	6.384	6.383	-0.015
	3.159	3.158	-0.03
	2.241	2.241	0.00

Figures 13, 14, 15, and 16 are plots of the above error data for the 10 V, 1 V, 100 mV, and 10 mV ranges, respectively. These data show evidence that the preamplifier does not add significant rms measurement errors to the basic design and that these instruments can meet the ±0.1 percent of reading accuracy specification.

6.3 Harmonic Distortion Tests

The high stability sine wave generator discussed in section 6.2 was used to check the total harmonic distortion (THD) measurement algorithm. This function and the associated algorithm is described in section 2.2.4 under Theory of Operation. By reprogramming the contents of the ROM used in the aforementioned sine wave generator, a distorted sine waveform with known ac offset and 4.7 percent of third harmonic distortion was used to test the distortion function (see fig. 17). The THD was measured for many values of voltage and frequency and was found to be within 4.7 ±0.1 percent, except at frequencies >80 Hz where the filter on the waveform generator began to attenuate the third harmonic. Compensating for the reduced third harmonic produced by the generator above 80 Hz yielded results in agreement with the measurements.

7. CONCLUSIONS

The means for the accurate measurement of accelerometer-amplifier outputs and other low-frequency vibration transducers using a sampling technique have been developed. These methods have been embodied in a true rms ac voltmeter and distortion analyzer that has been shown to have a measurement accuracy to ±0.1 percent (of reading) of the fundamental frequency, total harmonic distortion, and true rms voltage of approximately sinusoidal inputs from 2 mV to 10 V and frequencies from 0.1 to 120 Hz.

A microcomputer is utilized for implementing special measurement algorithms and timing and control functions, together with a conventional S/H amplifier and successive approximation type A/D converter (data acquisition module) for digitizing the input signal. By obtaining first an approximate (~1 percent) measure of the signal period with a purely software window crossing technique, a more exact determination of period can then be calculated with an autocorrelation function that consists of summing the difference in the values of samples taken at approximately corresponding times on the waveform cycle. It is shown that, assuming there are no other errors, the accuracy required for the period measurement is one half the accuracy required for the rms value. In the prototype voltmeter, the approximate period is determined in the first cycle and the exact period is then determined from the data samples taken during the second cycle of the input signal.

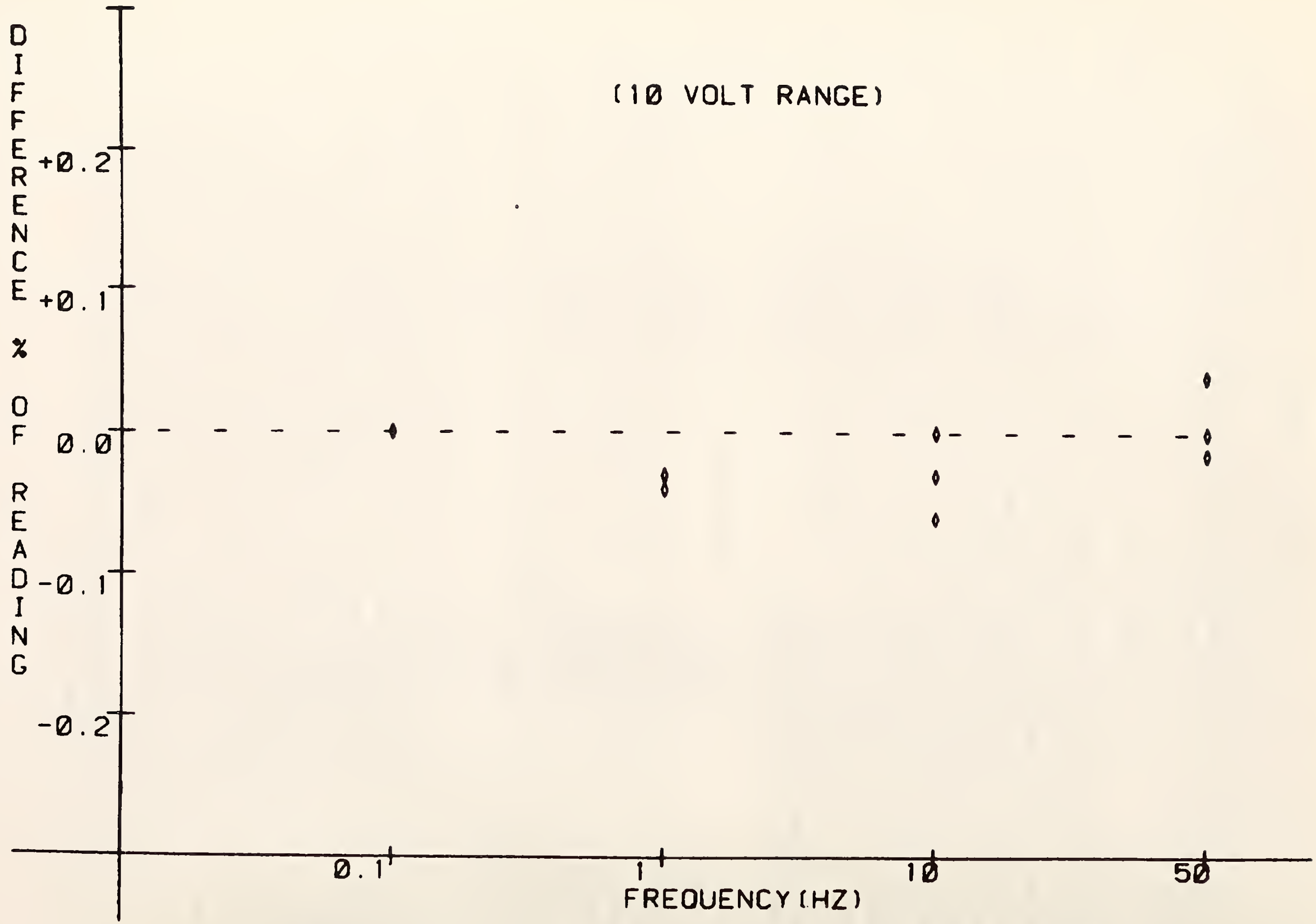


Figure 13. Plot of error data comparing the LF Sampling Voltmeter to the NBS AC Voltmeter/Calibrator (10 V range).

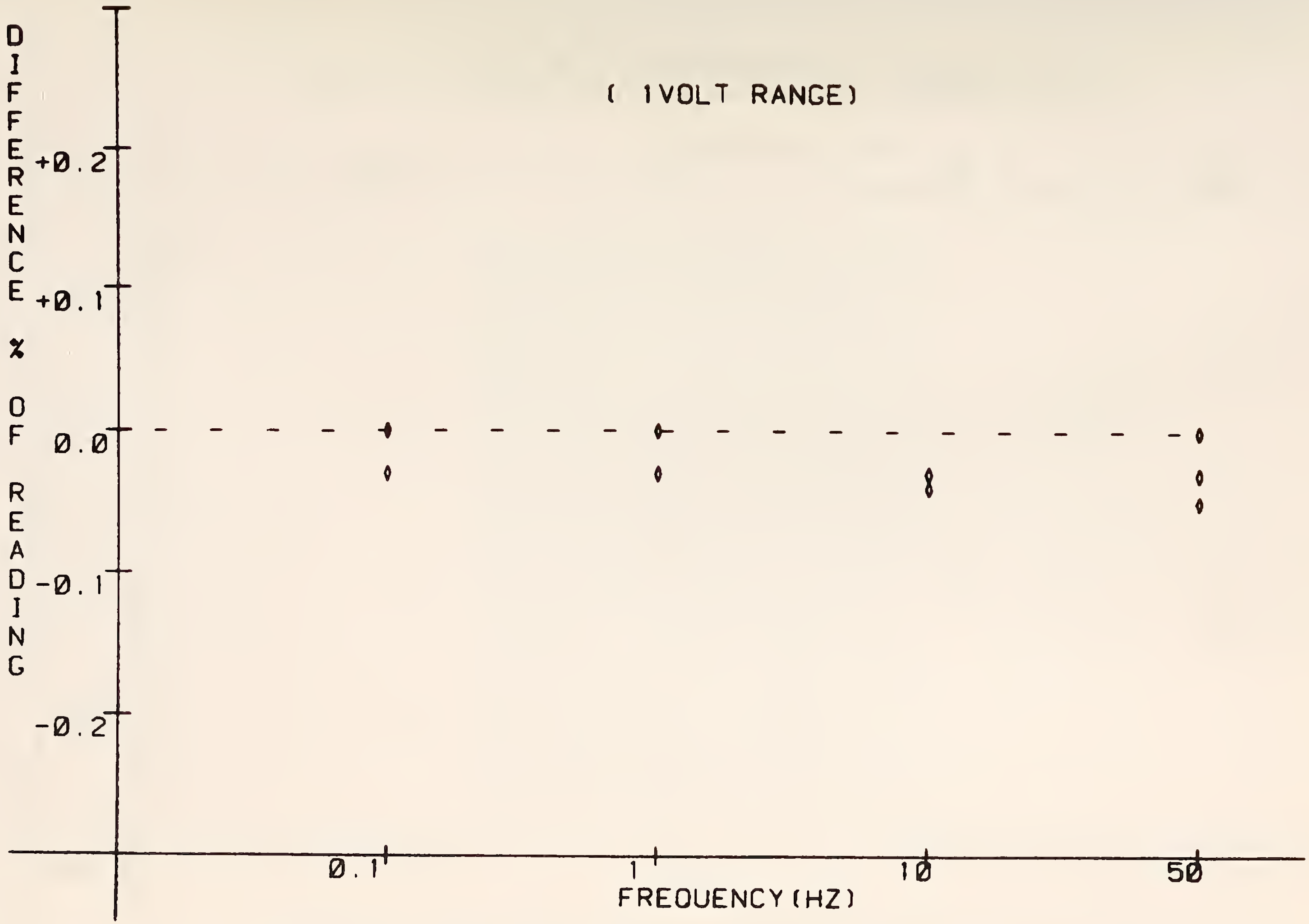


Figure 14. Plot of error data comparing the LF Sampling Voltmeter to the NBS AC Voltmeter/Calibrator (1 V range).

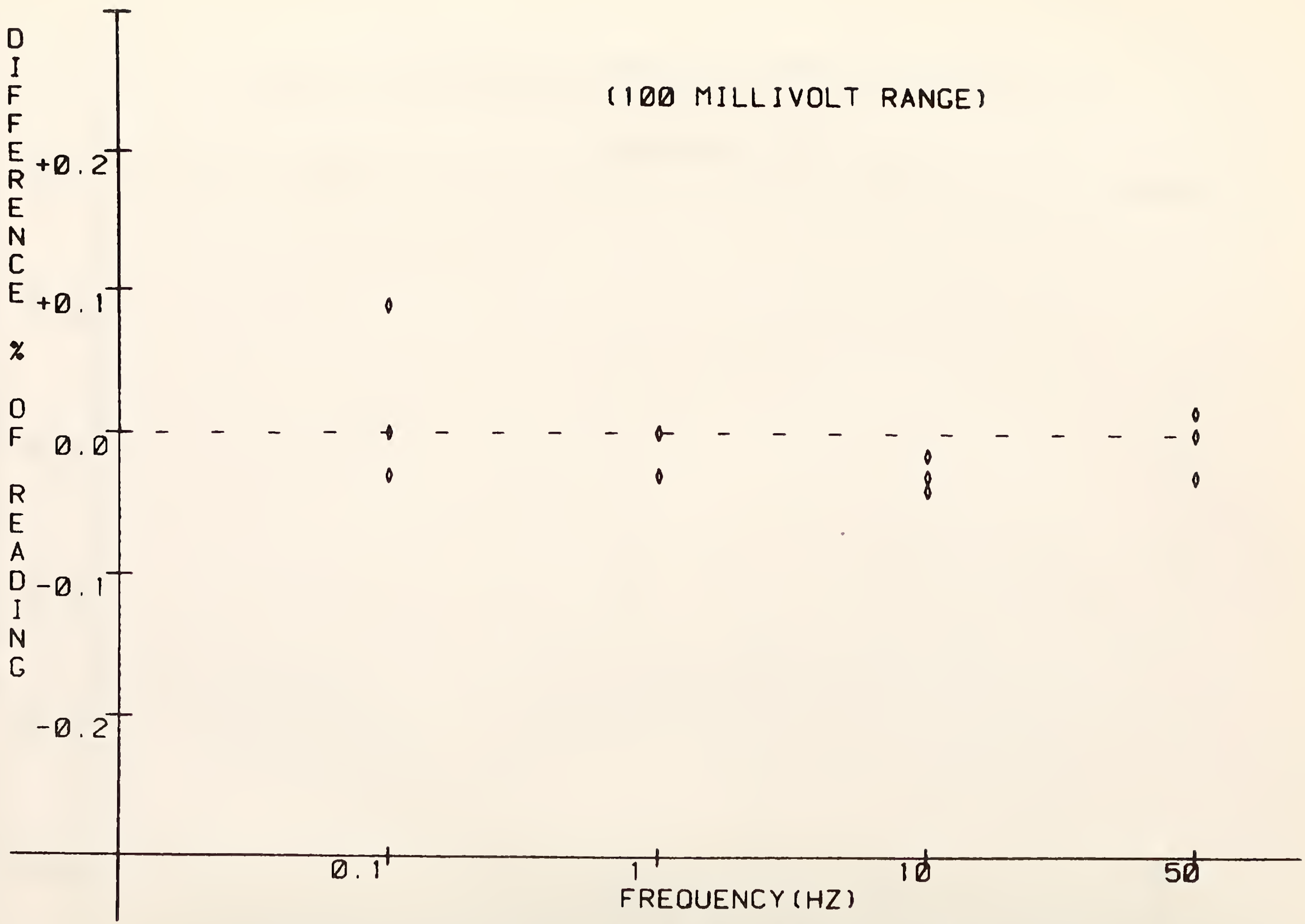


Figure 15. Plot of error data comparing the LF Sampling Voltmeter to the NBS AC Voltmeter/Calibrator (100 mV range).

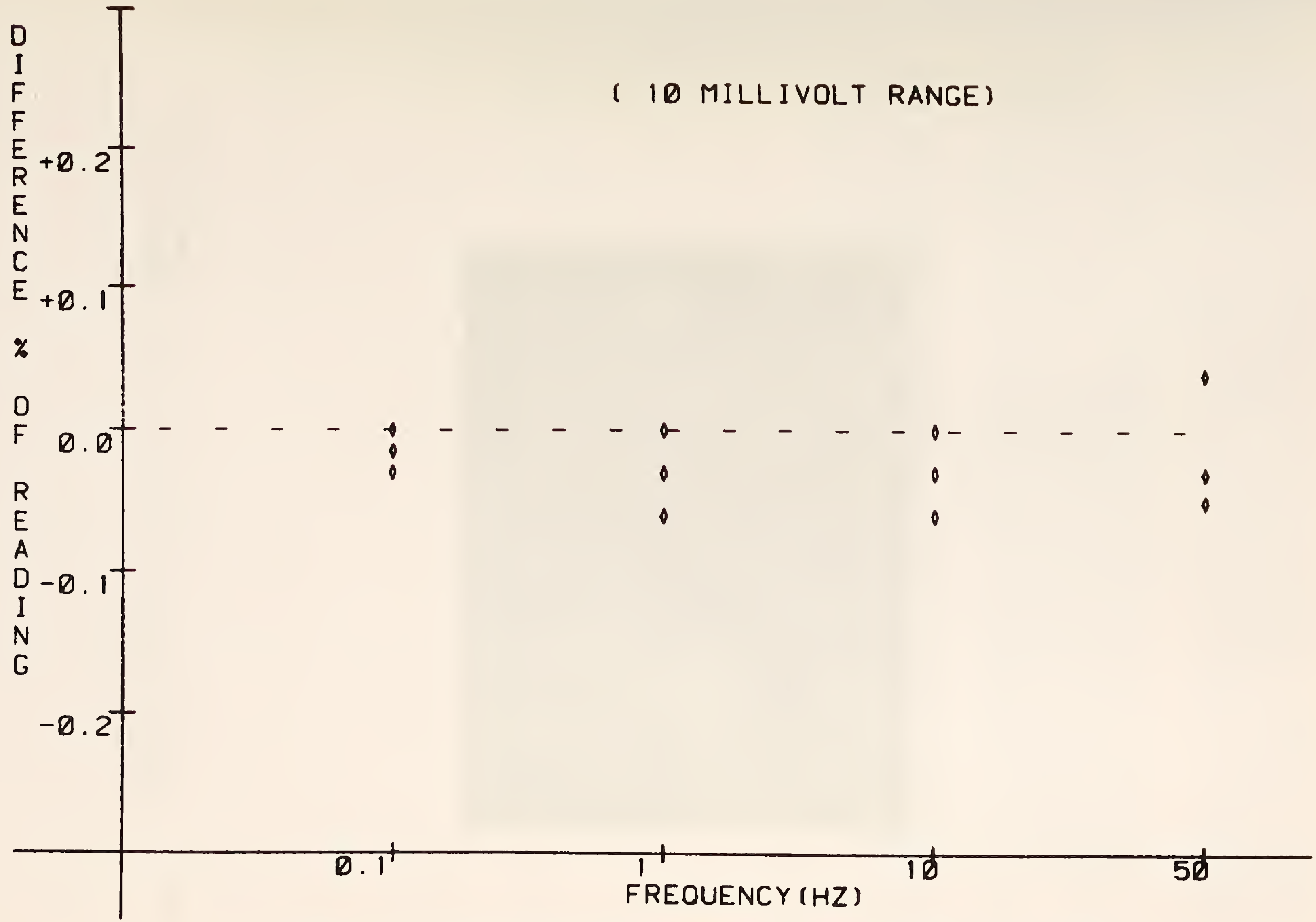


Figure 16. Plot of error data comparing the LF Sampling Voltmeter to the NBS AC Voltmeter/Calibrator (10 mV range).

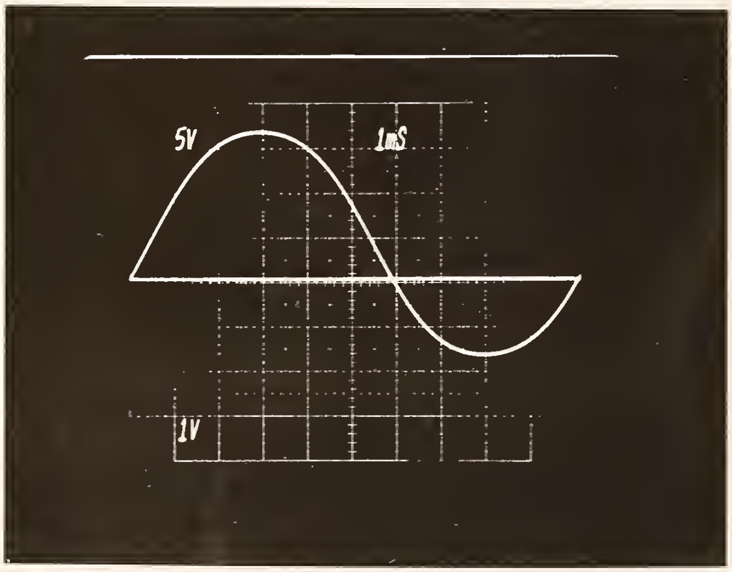


Figure 17. An example of a distorted sine wave (4.7 percent third harmonic distortion) with a dc offset produced by the waveform generator.

Using the mathematical definition of the rms value of a periodic waveform, the rms value (of both the ac plus dc components and the ac only component) of the signal is obtained from the sample set of the second cycle using numerical (midpoint) integration. When the exact period does not contain an integral number of samples, a small end correction is applied.

With nominally 128 samples taken over one period, a 64-point decimation-in-time FFT is computed to obtain the dc, fundamental, and 30 harmonic component coefficients of the input signal. The THD figure-of-merit is then calculated by squaring and summing the coefficients of the 30 harmonics and dividing this sum by the sum of the squares of the harmonic plus the fundamental.

An analysis of the sources of measurement errors and tests of the frequency (period), rms value, and harmonic distortion functions of the voltmeter have shown that the principles developed and employed in the LF Sampling Voltmeter described herein are a significant improvement over previously available means.

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APPENDIX A - CALIBRATION PROCEDURES

A.1.1 Setup

Calibration of the voltmeter should be performed in the order given here and in the following sections.[1] Equipment needed will be a microvolt null detector (with at least a 3 mV full-scale deflection) and an accurate (at least 0.1 percent) ac rms source with the capability of outputting between 10 mV-10 V rms at a frequency of 5 Hz-10 Hz. First, remove the front panel retaining screws and, secondly, situate the front panel so as to have access to both the preamp backplate (where all trim pots are located) and the front panel pushbuttons. Be careful not to short the display board against either the instrument case or the backplane of the card cage. The most convenient position for the front panel is to swing it out from the preamp side as far as connection wires will allow, and tilt the top of the front panel about 30° from the vertical. A portable vise is handy to hold it in this position. Also, the bottom plate should be removed to have access to the preamp coaxial cable end. Once this is done, proceed to sections A.1.2 and A.1.3 on balance and gain adjustments. If offset has already been adjusted, proceed directly to section A.1.3 regarding gain adjustment.

A.1.2 Balance Adjustment

This procedure simultaneously adjusts for preamplifier offset, bias current, and power supply effects. The first step in the procedure is to short the HIGH-LOW input terminals to the GND terminal. Then connect the BNC output of the preamp to the null detector with the null detector set to a less sensitive range. Turn on the voltmeter power switch. Depress and hold the front panel 10 mV pushbutton if the instrument is not on that range setting. Adjust the output offset using the pot label "BAL" on the preamp backplate until a null is obtained. It might be appropriate at this time to increase the null detector sensitivity to be certain that a null has been reached. Next, once this null has been established, decrease the null detector sensitivity, release the 10 mV pushbutton, and depress and hold the 1 V pushbutton. Adjust the pot labeled "NULL" to achieve a new null. Again, increase null meter sensitivity to verify that a null has been reached, then decrease the sensitivity. Release the 1 V pushbutton and depress the 10 mV pushbutton. Readjust BAL for a new null. Depress and hold the 1 V pushbutton. Readjust NULL for a new null. Repeat this process until the desired null (less than a millivolt) is achieved.

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[1] This procedure assumes that the gain adjustments of the data acquisition channel (S/H amplifier and A/D converter) are properly set to make accurate dc voltage measurements on its 2.5, 5, and 10 V ranges.

A.1.3 Gain Adjustment

The source referred to here is the accurate ac rms source mentioned in section A.1.1. A frequency of approximately 5-10 Hz should be available. The actual value is not important and these numbers are chosen to facilitate good low-frequency calibration and to speed up the calibration process. With this instrument the 1 V range should always be adjusted first since it impacts all other ranges. Connect the ac source to the HIGH and LOW terminals, and adjust the source to 500 mV rms. If the instrument is working correctly, a number should appear on the front panel numerical display. Depress and hold, if necessary, the 1 V pushbutton. Adjust "1 V" pot on the preamp backplate until 500 mV appears on the front panel display. It is important to note at this time that since the instrument is averaging data before displaying the results, the immediate effects of turning any gain pot will not be noticed right away. Release the 1 V pushbutton. Adjust the source to 50 mV rms. Depress and hold, if necessary, the 100 mV pushbutton. Trim the pot labeled "100 mV" on the preamp backplate until 50 mV is indicated on the front panel display. Release the 100 mV pushbutton. Adjust the source to 5 mV rms. Depress and hold, if necessary, the 10 mV pushbutton. Trim the pot labeled "10 mV" on the preamp backplate until 5 mV is indicated on the front panel display. Release the 10 mV pushbutton. Adjust the source to 5 V rms. Trim the pot labeled "10 V" on the preamp backplate until 5 V is indicated on the front panel display. Calibration of the instrument is now complete.

APPENDIX A - CALIBRATION PROCEDURES

A.1.1 Setup

Calibration of the voltmeter should be performed in the order given here and in the following sections.[1] Equipment needed will be a microvolt null detector (with at least a 3 mV full-scale deflection) and an accurate (at least 0.1 percent) ac rms source with the capability of outputting between 10 mV-10 V rms at a frequency of 5 Hz-10 Hz. First, remove the front panel retaining screws and, secondly, situate the front panel so as to have access to both the preamp backplate (where all trim pots are located) and the front panel pushbuttons. Be careful not to short the display board against either the instrument case or the backplane of the card cage. The most convenient position for the front panel is to swing it out from the preamp side as far as connection wires will allow, and tilt the top of the front panel about 30° from the vertical. A portable vise is handy to hold it in this position. Also, the bottom plate should be removed to have access to the preamp coaxial cable end. Once this is done, proceed to sections A.1.2 and A.1.3 on balance and gain adjustments. If offset has already been adjusted, proceed directly to section A.1.3 regarding gain adjustment.

A.1.2 Balance Adjustment

This procedure simultaneously adjusts for preamplifier offset, bias current, and power supply effects. The first step in the procedure is to short the HIGH-LOW input terminals to the GND terminal. Then connect the BNC output of the preamp to the null detector with the null detector set to a less sensitive range. Turn on the voltmeter power switch. Depress and hold the front panel 10 mV pushbutton if the instrument is not on that range setting. Adjust the output offset using the pot label "BAL" on the preamp backplate until a null is obtained. It might be appropriate at this time to increase the null detector sensitivity to be certain that a null has been reached. Next, once this null has been established, decrease the null detector sensitivity, release the 10 mV pushbutton, and depress and hold the 1 V pushbutton. Adjust the pot labeled "NULL" to achieve a new null. Again, increase null meter sensitivity to verify that a null has been reached, then decrease the sensitivity. Release the 1 V pushbutton and depress the 10 mV pushbutton. Readjust BAL for a new null. Depress and hold the 1 V pushbutton. Readjust NULL for a new null. Repeat this process until the desired null (less than a millivolt) is achieved.

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[1]This procedure assumes that the gain adjustments of the data acquisition channel (S/H amplifier and A/D converter) are properly set to make accurate dc voltage measurements on its 2.5, 5, and 10 V ranges.

A.1.3 Gain Adjustment

The source referred to here is the accurate ac rms source mentioned in section A.1.1. A frequency of approximately 5-10 Hz should be available. The actual value is not important and these numbers are chosen to facilitate good low-frequency calibration and to speed up the calibration process. With this instrument the 1 V range should always be adjusted first since it impacts all other ranges. Connect the ac source to the HIGH and LOW terminals, and adjust the source to 500 mV rms. If the instrument is working correctly, a number should appear on the front panel numerical display. Depress and hold, if necessary, the 1 V pushbutton. Adjust "1 V" pot on the preamp backplate until 500 mV appears on the front panel display. It is important to note at this time that since the instrument is averaging data before displaying the results, the immediate effects of turning any gain pot will not be noticed right away. Release the 1 V pushbutton. Adjust the source to 50 mV rms. Depress and hold, if necessary, the 100 mV pushbutton. Trim the pot labeled "100 mV" on the preamp backplate until 50 mV is indicated on the front panel display. Release the 100 mV pushbutton. Adjust the source to 5 mV rms. Depress and hold, if necessary, the 10 mV pushbutton. Trim the pot labeled "10 mV" on the preamp backplate until 5 mV is indicated on the front panel display. Release the 10 mV pushbutton. Adjust the source to 5 V rms. Trim the pot labeled "10 V" on the preamp backplate until 5 V is indicated on the front panel display. Calibration of the instrument is now complete.

APPENDIX B - MAINTENANCE AND TROUBLESHOOTING

A.2 Maintenance

The following portion of this manual contains information and procedures designed to aid in the troubleshooting and repair of the Low-Frequency Sampling Voltmeter. Each section is titled by a specific fault possibility with information given on the probable cures.

A.2.1 No power

In the event that the front panel LEDs, numerical display, and back panel fan are not operating, turn off power, check the fuse in the fuse holder on the back panel and replace with a 125 V, 5 A fuse. Before turning power back on, check all the internal power lines in the instrument (yellow, orange, red) for possible shorts. All the power supply outputs are currently protected against shorts, but power failure has been known to occur if different supply outputs are connected together.

A.2.2 Front panel numerical display is functioning; no LEDs are turning on

Turn off power. Check the +12 V line from the Kepco power supply (orange lead). This line runs from the power supply to the backplane connector J3 pin 52, to pin 9 of the 50 pin ribbon connector, to the front panel display board. Make sure the 50 pin connector is secure at both ends. Check the integrity of the +12 V path and, if okay, remove the front panel display board from the front panel with all connectors to this board intact (connector to preamp, switches, and 50 pin connector). Be careful not to short the display board against the case, the backplane, or the front panel. Turn on the power. The LEDs should now work, indicating that a short existed between the display board and the front panel. Turn off the power. Reassemble the instrument, and correct the shorting condition. If the LEDs still do not work, the +12 V supply has probably failed.

A.2.3 Erratic behavior of the LEDs

Push the reset button. If there are still improper results, turn the power off momentarily, then switch power back on. If this procedure fails, remove the front panel display board connector from the pushbutton switches, and examine for loose connections. Also, check to see that the 50 pin connector is secure on both ends, and that all printed circuit boards are secure in their respective card cage connectors.

A.2.4 Reset not functioning

Refer to section A.2.3. If this procedure fails, check the wirewrap connection on the card cage backplane - pin 56 of J3 to pin 15 of J1. If this path is okay, then an integrated circuit failure exists on the memory, timing, and control I/O board (probably IC20).

A.2.5 LEDs functioning; no front panel numerical display

Check the +15 V, -15 V and analog ground lines (red, yellow, green) on the terminal strip located on the backplane of the card cage on the side furthest from the Kepco power supply. Push reset and make sure that the overload indicator is not ON. If it is, then the incoming signal exceeds the limits of the instrument's measuring capability. Check the BNC connection to the A/D converter board from the preamp. Make sure that the A/D converter board is secure in card cage connector J3. If the problem persists after this check, the trouble is probably in the preamp; go to appendix A for calibration.

A.2.6 Erratic detection of the incoming signal

Make sure all printed circuit boards are secure in their respective card cage connector, and check that the 50-pin ribbon connector is secure on both ends. Perform the checks designated in section A.2.5. If the problem still exists after these checks, it is probably an integrated circuit failure problem on the memory, timing, and control I/O board. Replace IC's 8, 7, 6, and 5 (sampling rate generator) on the memory, timing, and control I/O board.

A.2.7 No response to pushbutton selections; functioning otherwise

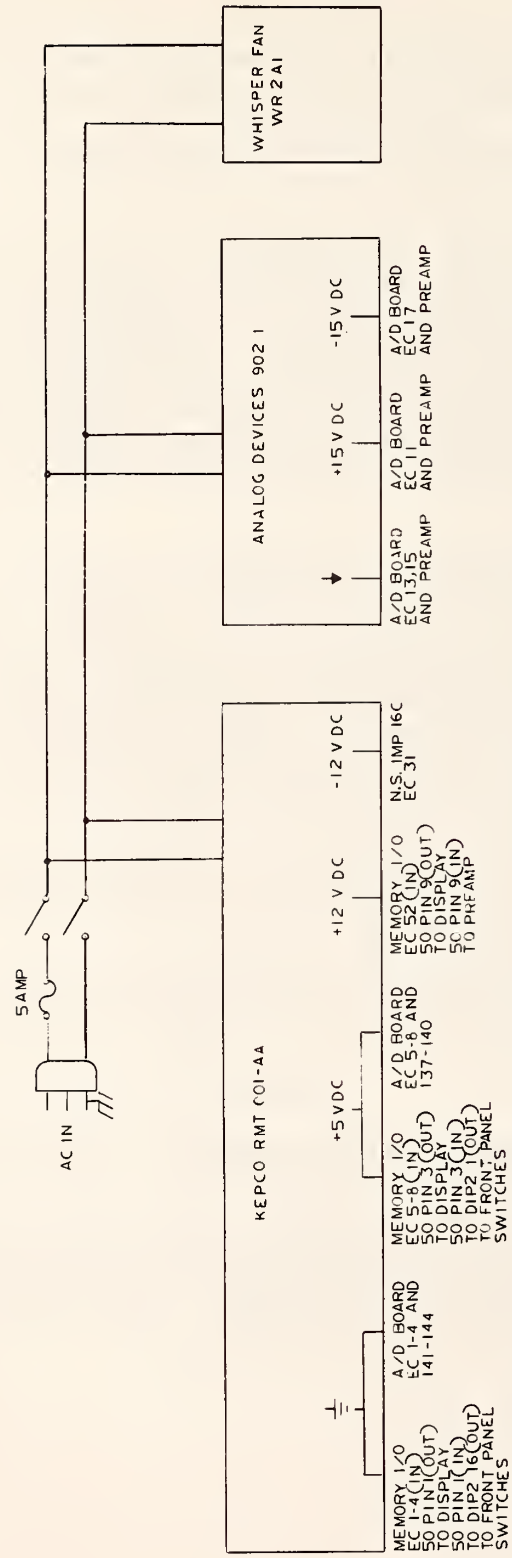
Perform the checks designated in section A.2.3.

A.2.8 No range, mV, or volt LEDs activated, but instrument is working

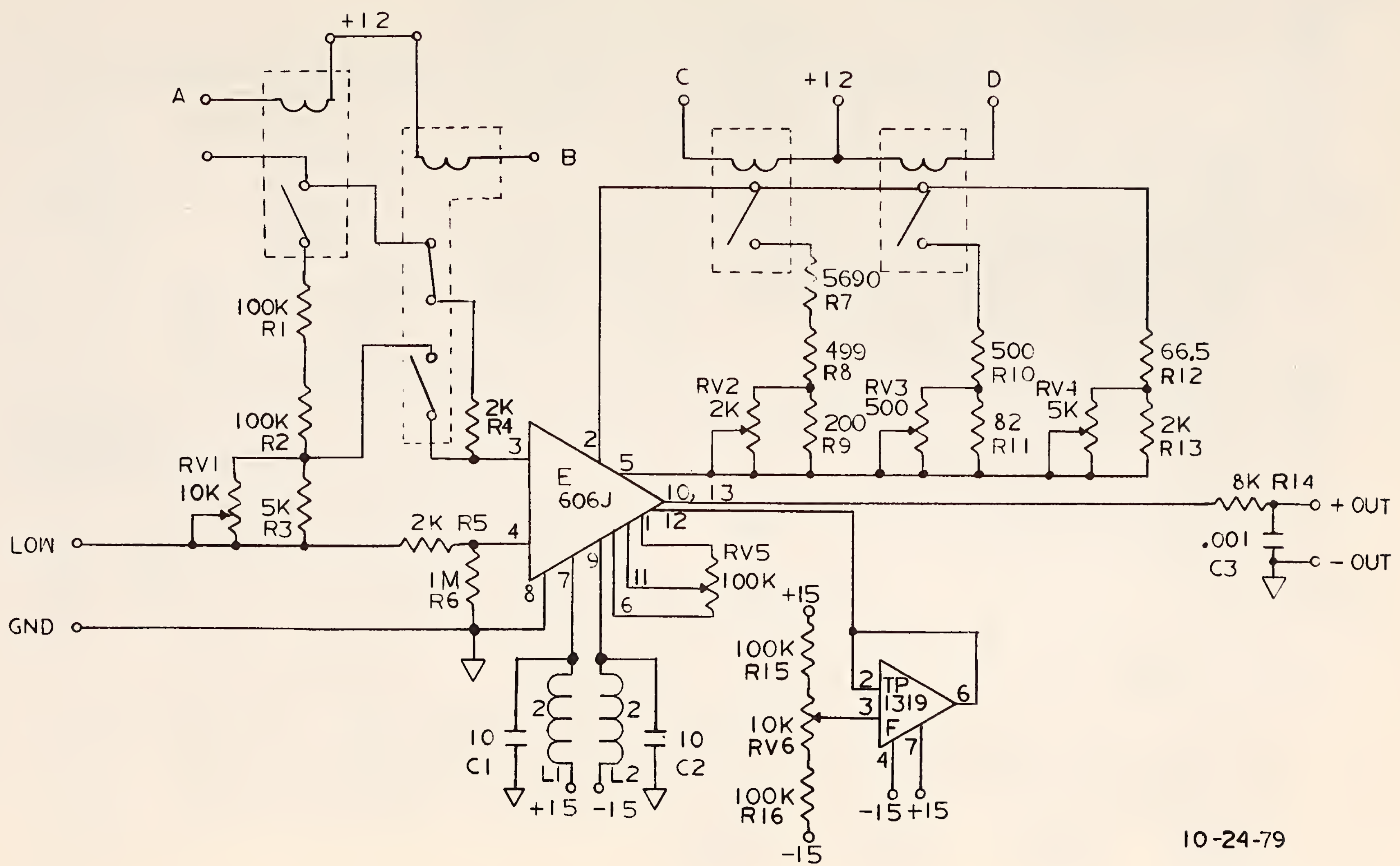
The instrument has just recovered from an overload; push 10 mV range button to restore LEDs.

A.2.9 Improper values being displayed on numerical display; otherwise instrument working

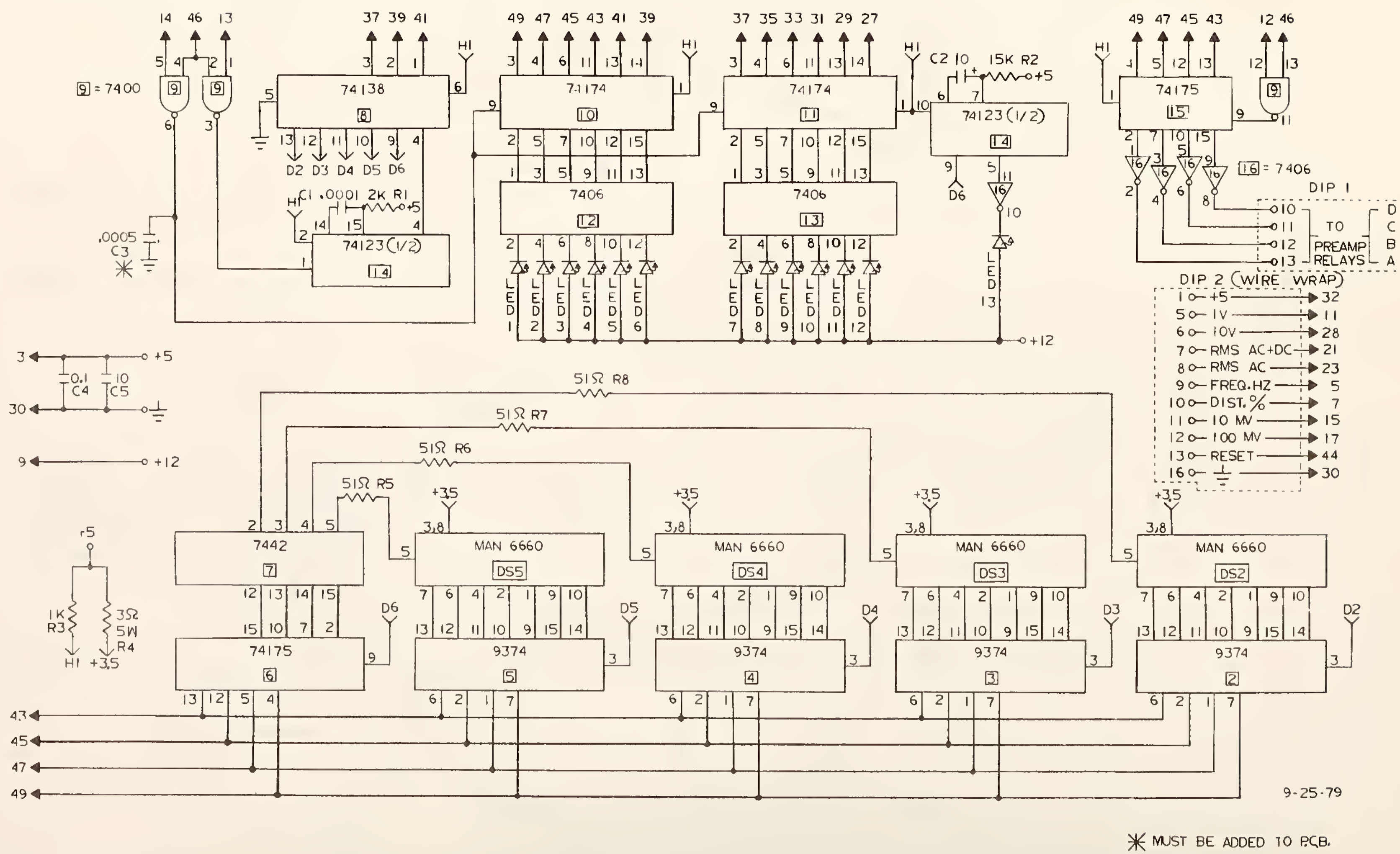
Turn to calibration, appendix A. Perform the necessary procedure. If the problem persists, the number displayed is an overload on that particular range. To correct, push next higher range button. If the problem still exists, refer to section A.2.6 above.



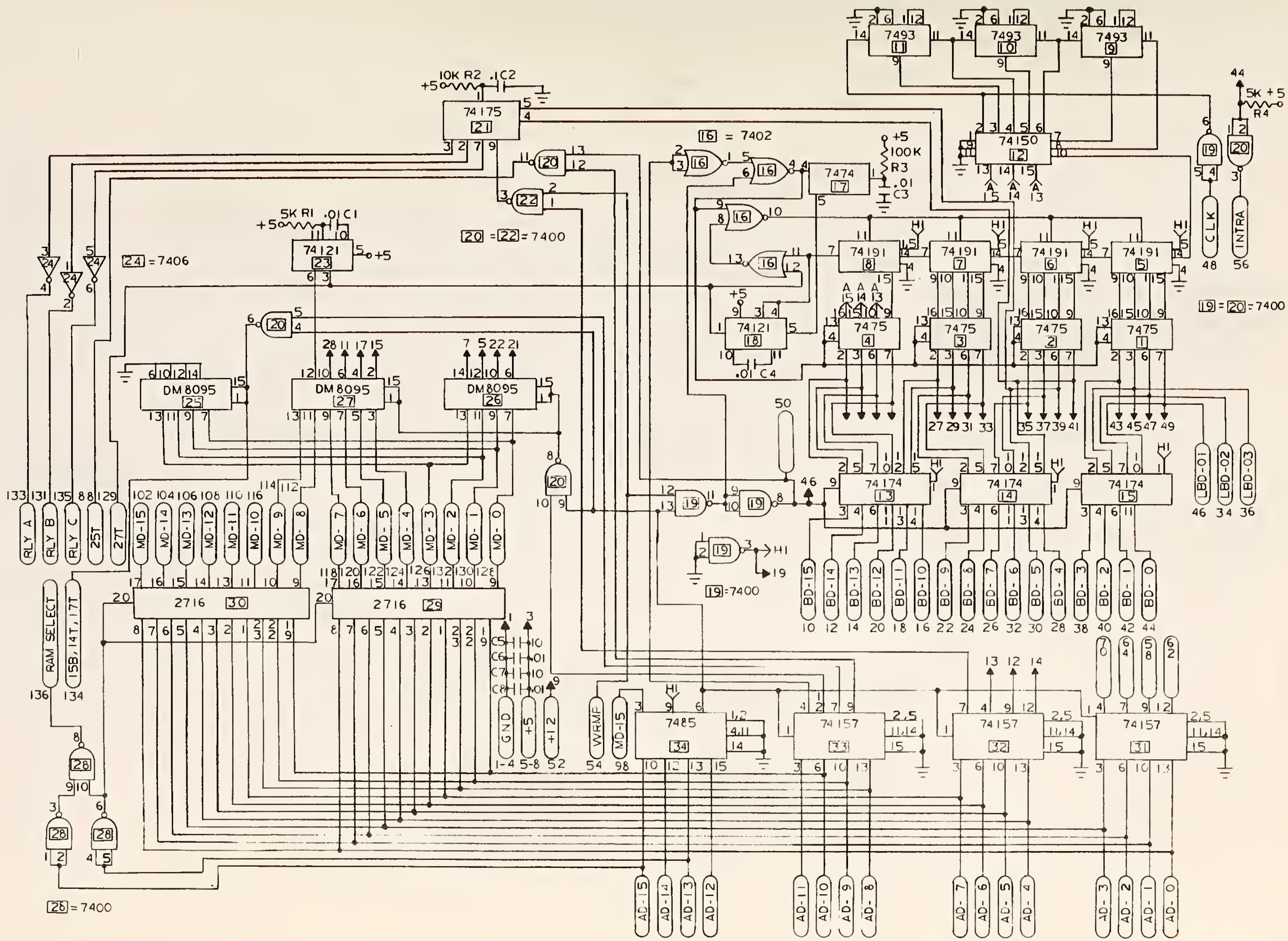
SCHEMATIC 1 POWER SUPPLY DIAGRAM



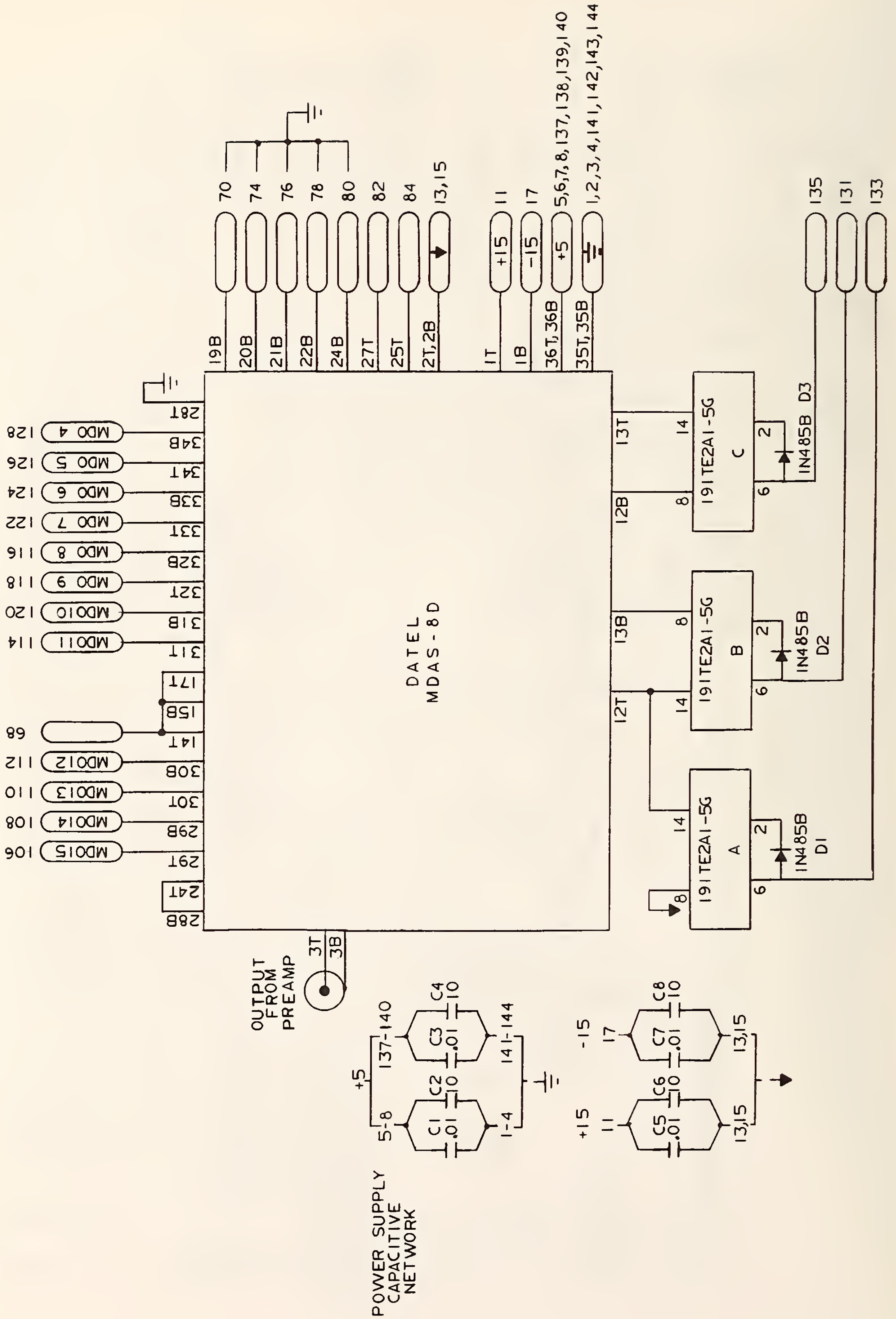
SCHEMATIC 2 PREAMP NBS 722 79284A



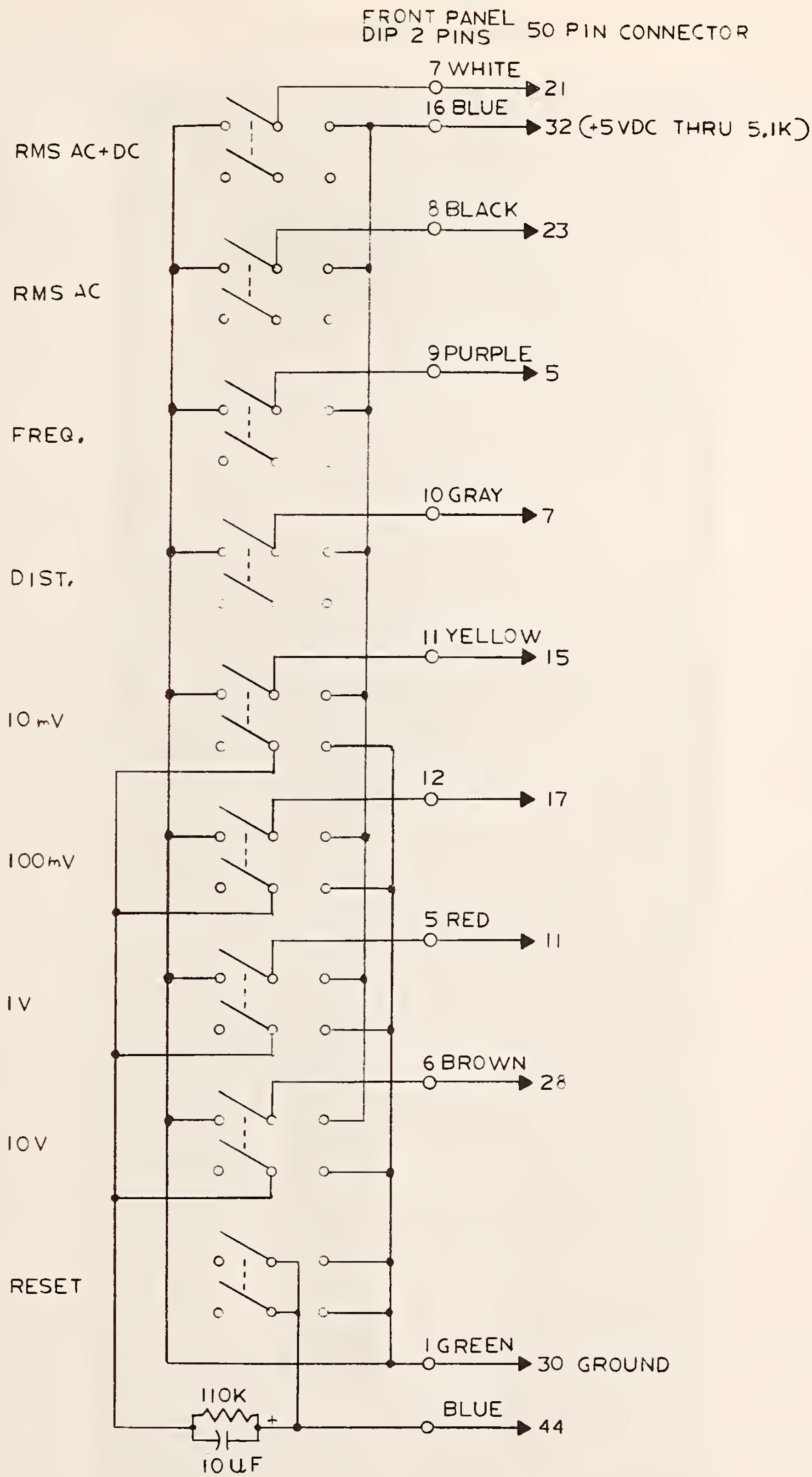
SCHEMATIC 3 DISPLAY BOARD NBS 722 79026A



SCHEMATIC 4   NBS 722 7909A   MEMORY, TIMING, AND CONTROL I/O BOARD



SCHEMATIC 5 A/D BOARD NBS 722 79141A



SCHEMATIC 6 FRONT PANEL SWITCHES WIRING DIAGRAM

PHOTO 1 LOW-FREQUENCY AC VOLTMETER

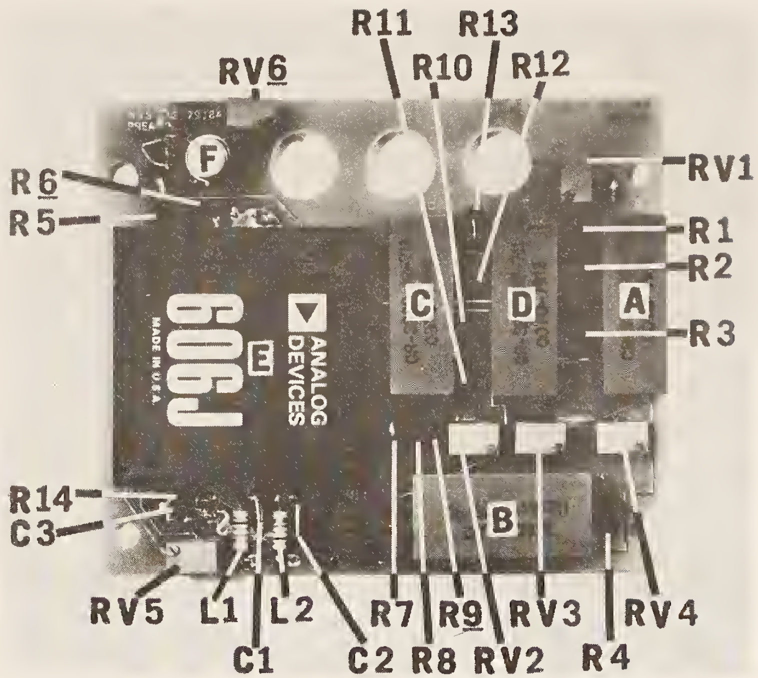


PHOTO 2 NBS 722 79184A PREAMPLIFIER

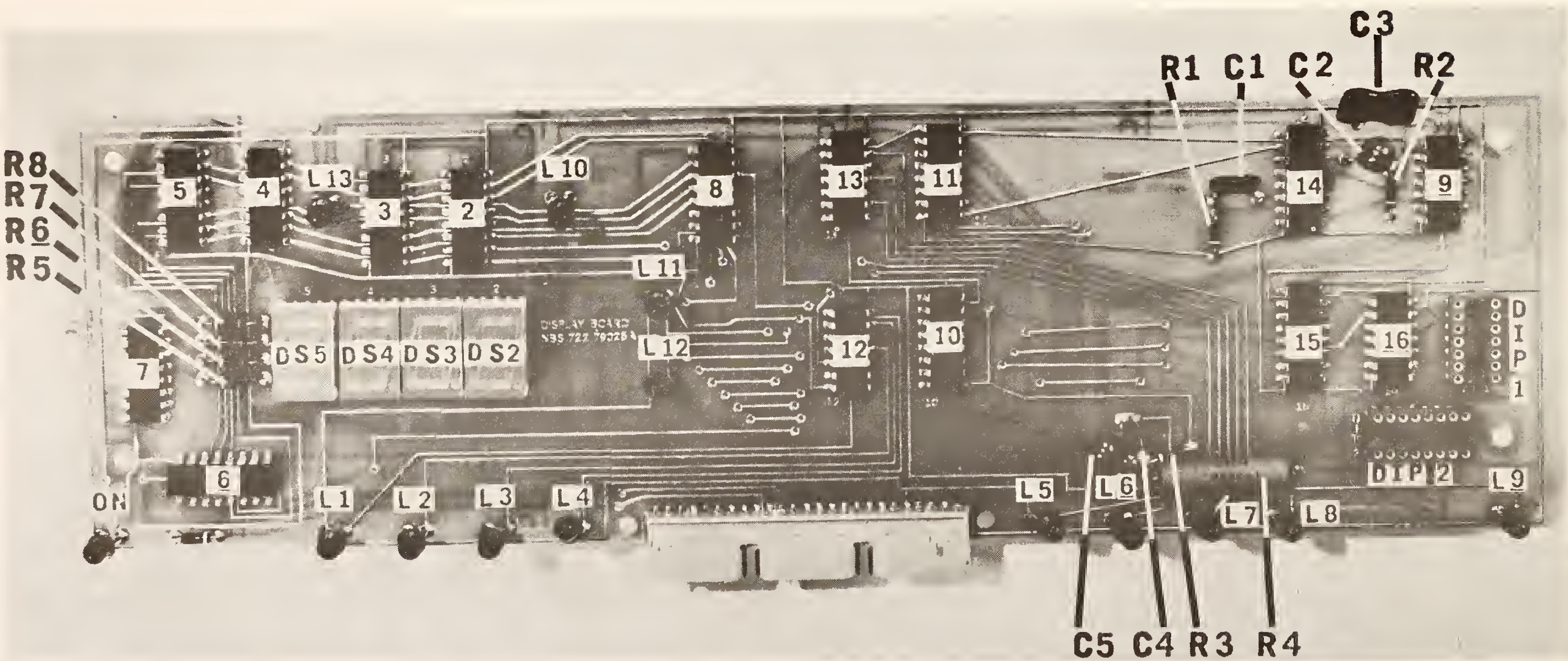


PHOTO 3 NBS 722 79026A DISPLAY BOARD

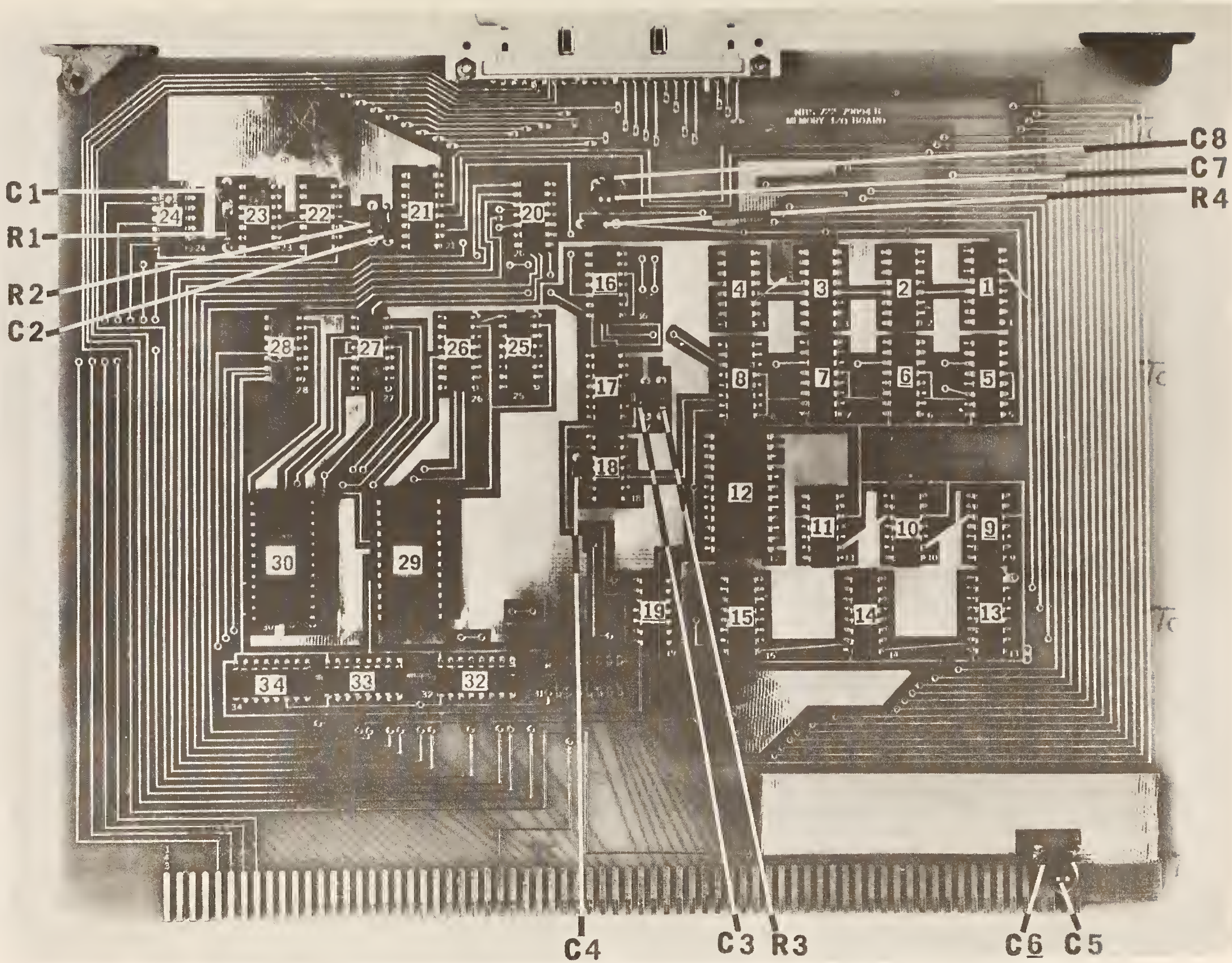


PHOTO 4   NBS 722 79094A MEMORY, TIMING, and CONTROL I/O BOARD

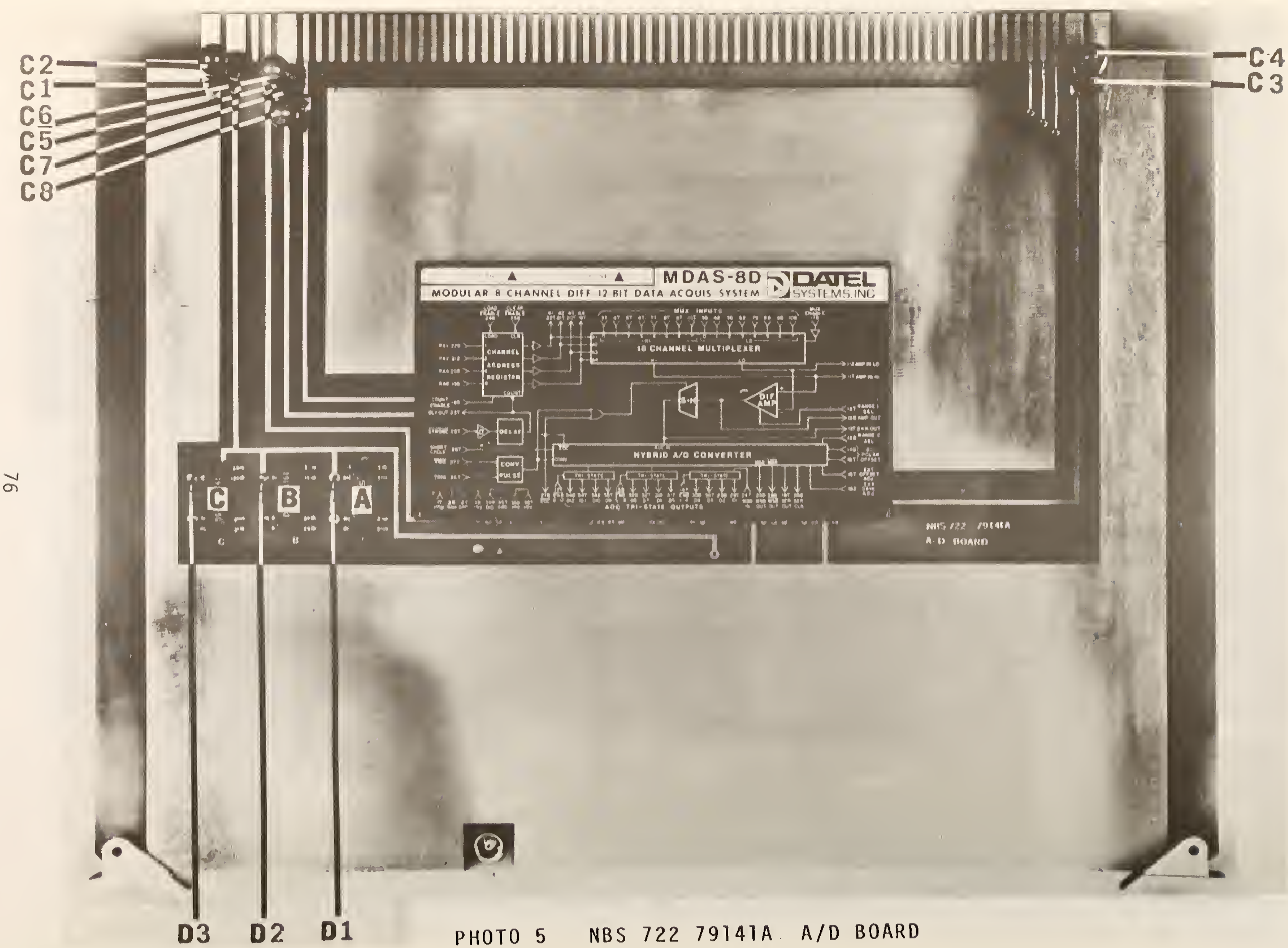


PHOTO 5 NBS 722 79141A A/D BOARD

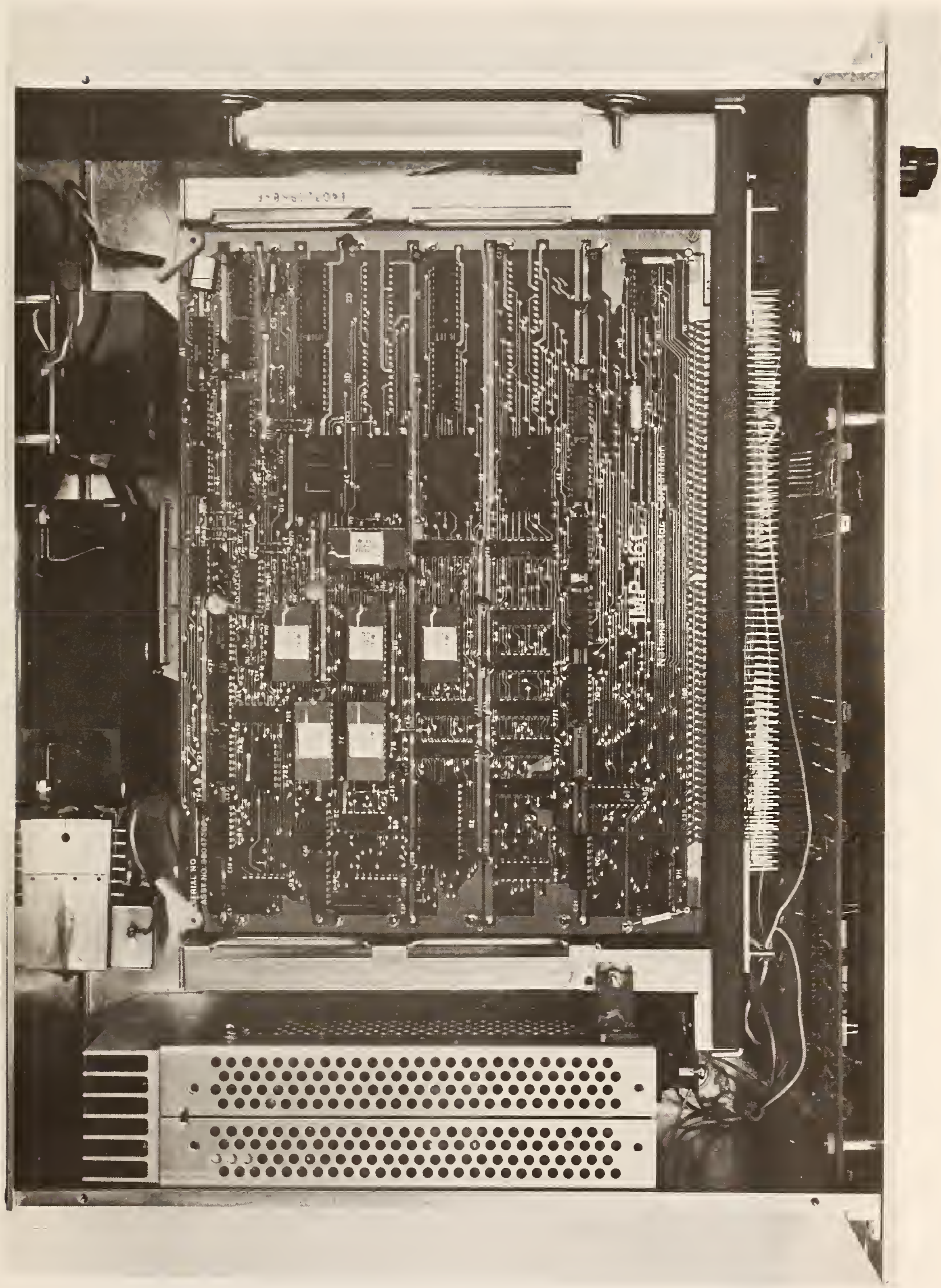


PHOTO 6   TOP VIEW OF INSTRUMENT

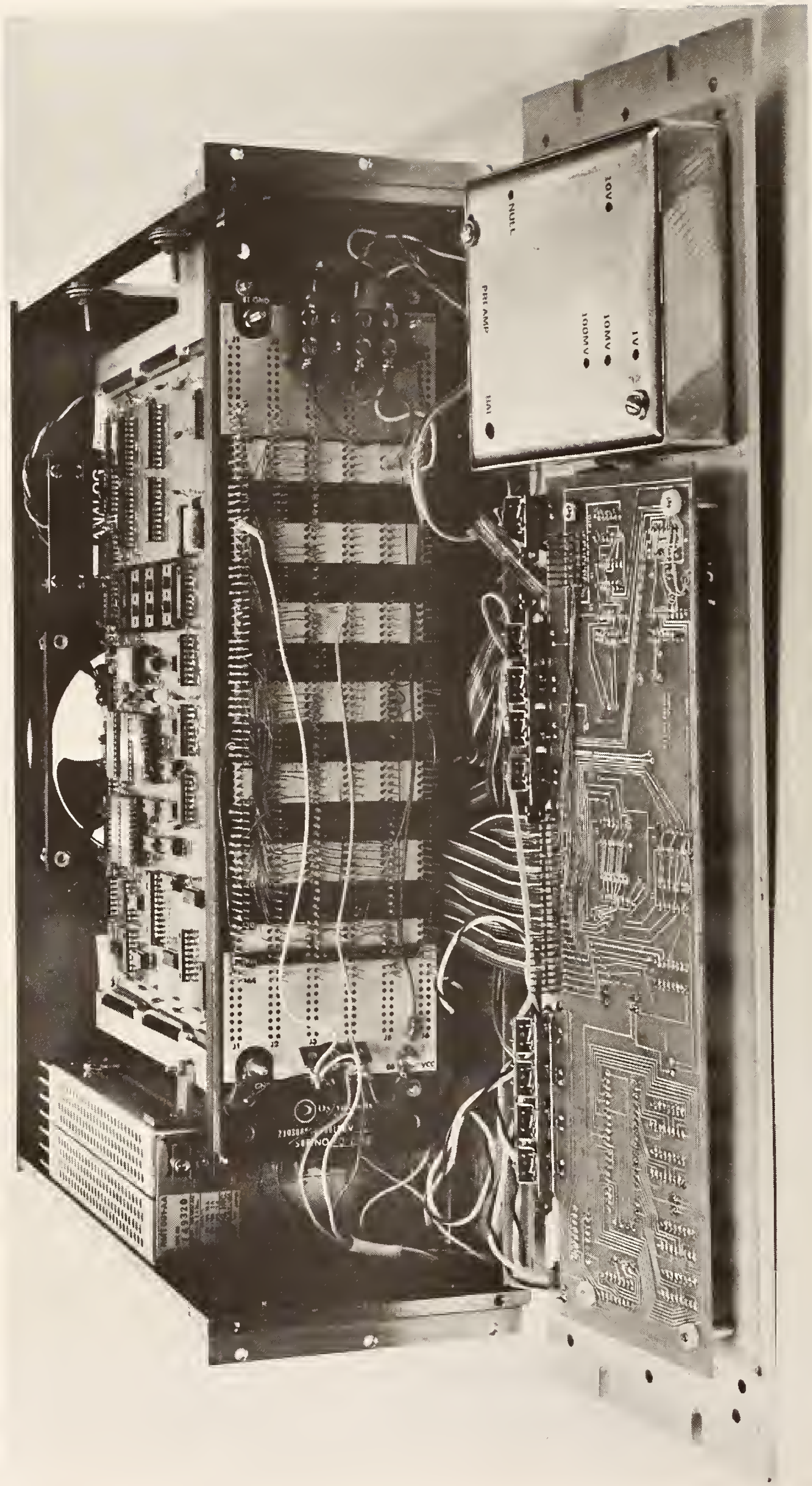


PHOTO 7 FRONT VIEW WITH PANEL REMOVED

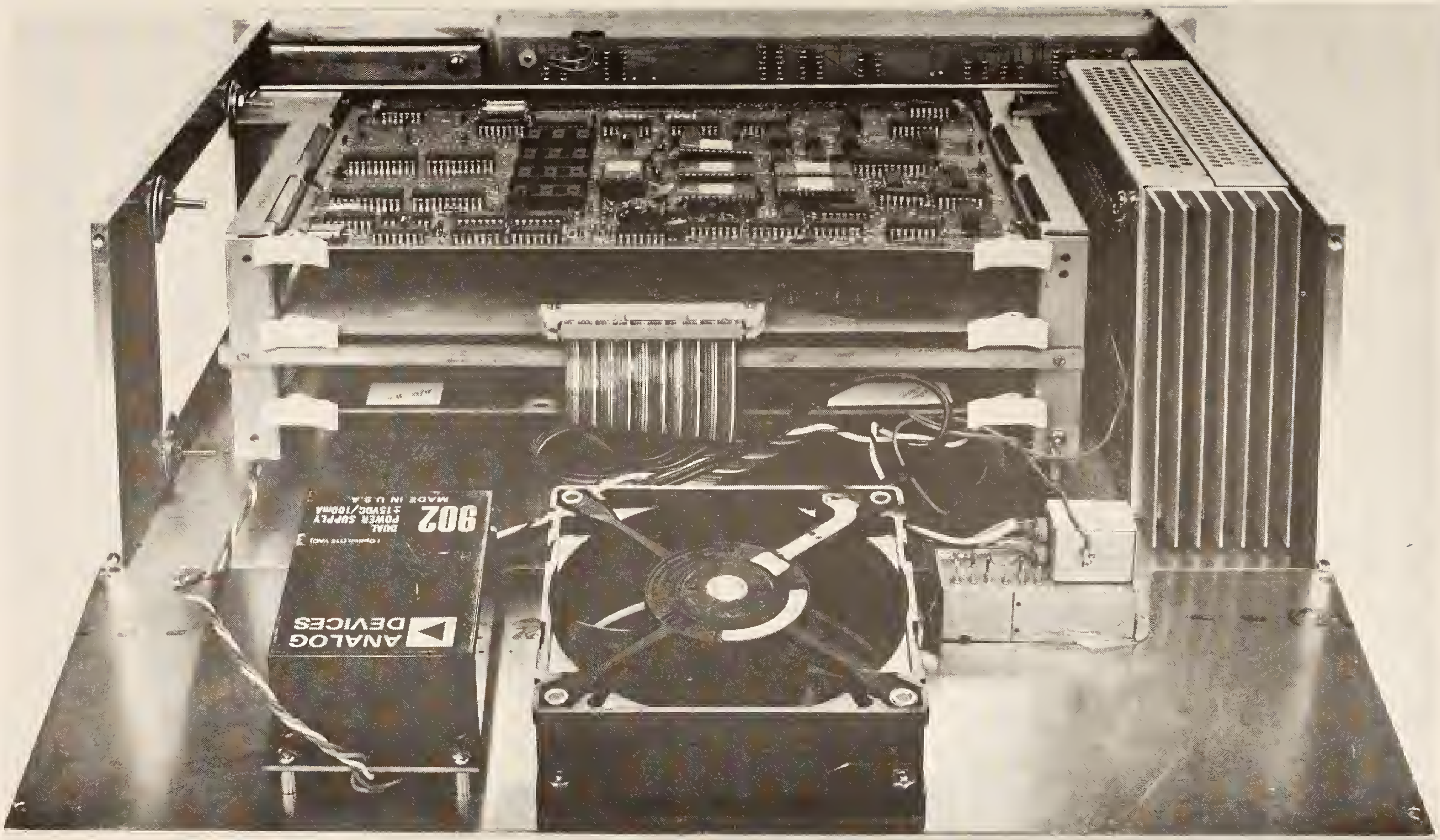


PHOTO 8   BACK VIEW WITH PANEL REMOVED

APPENDIX E - PARTS LIST

NBS 722 79184A Preamplifier Board

	Description	Quantity	Identification No(s)
1.	Analog Devices 606J amplifier	1	E
2.	Teledyne TP319 operational amplifier	1	F
3.	1/4 Watt Vishay resistors		
	66.5 kΩ	1	R12
	5690 Ω	1	R7
	100 kΩ	2	R1, R2
	500 Ω	1	R10
	2 kΩ	3	R4, R5, R13
	499 Ω	1	R8
	82 Ω	1	R11
4.	Helitrim pots (pc board type)		
	5 k	1	RV4
	500 Ω	1	RV3
	2 k	1	RV2
	100 k	1	RV5
	10 k	2	RV1, RV6
5.	Miniature chokes 2 μh, 1/4 watt	2	L1, L2
6.	10 μf, 35 V (polarized) Tantalum capacitor	2	C1, C2
7.	Resistors 10%, 1/4 watt, Tin oxide		
	200 Ω	1	R9
	5.1 k	1	R3
	8 k	1	R14

PARTS LIST

NBS 722 79184A Preamplifier Board (cont.)

Description	Quantity	Identification No(s)
8. Resistors 1%, 1/4 watt		
1 MΩ	1	R6
100 kΩ	2	R15, R16
9. Coto-Coil Relays, CR-3702-12	3	A, C, D
" " " CR-3206-12	1	B

PARTS LIST

Preamp Board (cont.)

Miscellaneous List

	Description	Quantity
1.	3M No. 3406 connector	2
2.	4" flat 14 conductor ribbon cable	1
3.	1/8" MU-METAL box for preamp housing	
	Body 4 7/8" x 4 1/8" x 1 1/8"	1
	Lid 4 3/4" x 4" x 1 1/8"	1
4.	Bendix 3022-12 BNC connector	1
5.	14" RF-174/U shield cable	1
6.	Red & Black dual female banana receptacle chassis mount	
7.	Black single banana receptacle chassis mount	1
8.	14 pin low profile dip socket	1
9.	1/2" 6-32 threaded steel spacer	2
10.	1/4" 6-32 threaded steel spacer	4
11.	6-32 steel nuts	4
12.	1/2" flat head steel screw	2
13.	3/4" flat head steel screw	2
14.	Nylon washer 0.196 ID	6
15.	Nylon nut 6-32	2

PARTS LIST

NBS 7279026A Display Board

Description	Quantity	Identification No(s)
I. INTEGRATED CIRCUITS		
1. 9374 LED digit driver	4	2, 3, 4, 5
2. 7442 BCD to DECIMAL decoder	1	7
3. 74175 (or 74LS175) Quad D-type flip flop	2	6, 15
4. 74138 (or 74LS138) 3- to -8 Line decoder	1	8
5. 74174 (or 74LS174) Hex D-type flip flop	2	10, 11
6. 74123 Dual One-Shot with clear	1	14
7. 7406 30 V Hex inverter/driver	3	12, 13, 16
8. 7400 (or 74LS00) Quad 2 input NAND gate	1	9
II. LEDS		
1. MAN 6660 LED Numerical display	4	DS2 - DS5
2. LITRONIX RED-LIT-C201 (or equivalent)	14	LED1 - LED 13
III. RESISTORS and CAPACITORS		
1. 2 kΩ, 10%, 1/4 W Composition resistor	1	R1
2. 15 kΩ, 10%, 1/4 W Composition resistor	1	R2
3. 1 kΩ, 10%, 1/4 W Composition resistor	1	R3
4. 3Ω , 1%, 5 W Composition resistor	1	R4
5. 51 Ω, 10%, 1/4 Composition resistor	4	R5, R6, R7, R8
6. 100 pF, 100 V, Dipped mica capacitor	1	C1
7. 10 μf, 35 V (polarized) Tantalum capacitor	2	C2, C5
8. 500 pF, 100 V, Dipped mica capacitor	1	C3
9. 0.1 μf, 50 V, Monolithic ceramic capacitor	1	C4
IV. MISCELLANEOUS		
1. 14 pin Low profile dip socket	1	DIP 1
2. 14 pin connector & ribbon cable, 3M #3406 (or equiv.)	1	
3. 50 pin 90° ribbon connector (3M or equiv.)	1	
4. 16 pin (wire wrap) Low profile dip socket	1	DIP 2

PARTS LIST

NBS 722 79094A Memory I/O Board

Description	Quantity	Identification No(s)
I. INTEGRATED CIRCUITS		
1. 2k x 8 2716 EPROMS	2	29, 30
2. 74150 Data selector/multiplexer	1	12
3. 74157 Quad 2 to 1 data selector	2	32, 33
4. 74157 Quad 2 to 1 data selector (optical)	(1)	31
5. 74174 Hex D-type flip-flop	3	13, 14, 15
6. DM8095 Tri-state buffer and latch	3	25, 26, 27
7. 7475 (or 74LS75) Quad bistable latch	4	1, 2, 3, 4
8. 74191 (or 74LQ191) up/down binary counter	4	5, 6, 7, 8
9. 7493 (or 74LS93) 4-bit binary counter	3	9, 10, 11
10. 7474 (or 74LS74) dual D-type flip flop	1	17
11. 74175 (or 74LS175) Quad D-type flip flop	1	21
12. 7400 (or 74LS00) Quad 2 input NAND gate	4	19, 20, 22, 28
13. 7402 (or 74LS02) Quad 2 input NOR gate	1	16
14. 7485 (or 74LS85) 4-bit magnitude comparator	1	34
15. 7406 30 V Hex inverter/driver	1	24
16. 74121 Monostable multivibrator	2	18, 23
II. RESISTORS and CAPACITORS		
1A. 5.1 kΩ, 10%, 1/4 W, composition resistor	1	R4
1B. 5.1 kΩ, 1%, 1/4 W, composition resistor	1	R1
2. 10 kΩ, 10%, 1/4 W, composition resistor	1	R2
3. 100 kΩ, 10%, 1/4 W, composition resistor	1	R3
4. 10 or 22 µf, 35 V (polarized) Tantalum capacitor	2	C5, C7
5. 0.1 µf, 50 V, monolithic ceramic capacitor	1	C2
6. 0.01 µf, 50 V, monolithic ceramic capacitor	5	C1, C3, C4, C6, C8
III. MISCELLANEOUS		
1. 24 pin I.C. socket (Cambion or equiv.) for item I.1.	2	
2. 50 pin 90° ribbon connector (3M or equiv.)	1	
3. Card ejector clip (Cambion or equiv.)	2	
4. #2-56, 3/4" machine screw	2	

PARTS LIST

NBS 722 79141A A/D Board

	Description	Quantity	Identification No(s)
1.	Data Model MDAS-8D data acquisition module and 3-36063-2 72 pin AMP connector	1	
2.	1N485B switching diodes (or equiv.)	3	D1, D2, D3
3.	191TE2A1-5G SIGMA reed relays (or equiv.)	3	A, B, C
4.	0.01 μf, 50 V, monolithic ceramic capacitor	4	C1, 3, 5, 7
5.	10 μf, 35 V (polarized) Tantalum capacitor	4	C2, 4, 6, 8
6.	LEMO size 0 chassis mount coax connector	1	
7.	RG178 coax cable (~3" long)	1	
8.	Card ejector clip (Cambion or equiv.)	2	
9.	1/4" 4-40 screws	2	

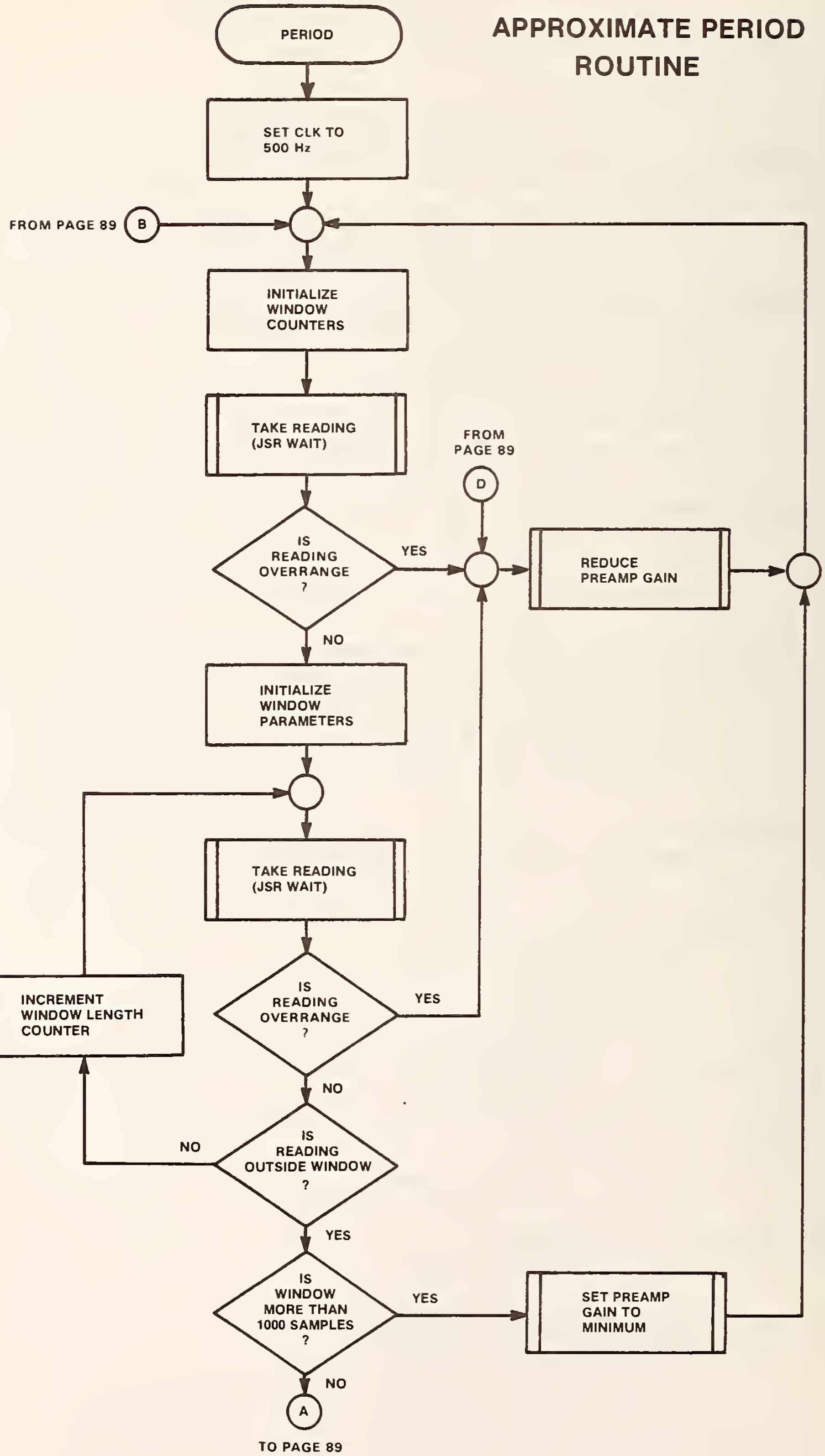
INSTRUMENT PARTS LIST

		Quantity
1.	CPU Board National Semiconductor IMP-16C/3000A	1
2.	I/O Memory Board NBS 722 79094B (see detailed parts list)	1
3.	A/D Board NBS 722 79141A "	1
4.	Display Board NBS 722 79026A "	1
5.	Preamplifier Board NBS 722 79184A "	1
6.	Card Cage National Semiconductor IMP-00H/880	1
7.	Kepco Power Supply RMT-001-AA +12 V at 1 A -12 V at 1 A +5 V at 10 A	1
8.	Analog Devices Dual Power Supply 902I ±15 V DC/100 mA (115 V option)	1
9.	24" 50 conductor flat ribbon cable with 3M No. 3425 50 pin female connectors at each end	1
10.	ROTRON whisper fan WR2A1, 115 V	1
11.	ROTRON #551147 finger guard	1
12.	CORCOM line filter and fuse holder	1
13.	Fuse, 125 V, 5 A	1
14.	Switchcraft all lock 65043-206 switches	1
15.	Switchcraft interlock 65041K-206 switches	1
16.	Plexiglas, dark ruby, 4" x 1 1/2"	1
17.	Wiring set from Kepco supply to card cage (see schematic)	1
18.	Wiring set from Analog supply to card cage "	1
19.	Cable from Analog supply to preamp "	1
20.	Terminal strips, 4 conductor type	2
21.	NBS LF AC Voltmeter, 19" x 4 3/4" front panel for #19 rack	1
22.	Side panel, 13" x 3 1/8"	2
23.	Back panel, 17" x 5 1/4"	1
24.	Top lid	1
25.	Bottom lid	1
26.	Socket for item #8 Analog Devices AC-1040	1
27.	Switch wiring cable to display board (see schematic)	1
28.	Supporting bracket from side panels to front panel	2
29.	Fan power cord for item #10	1
30.	Double pole single throw switch (5 A at 125 V)	1

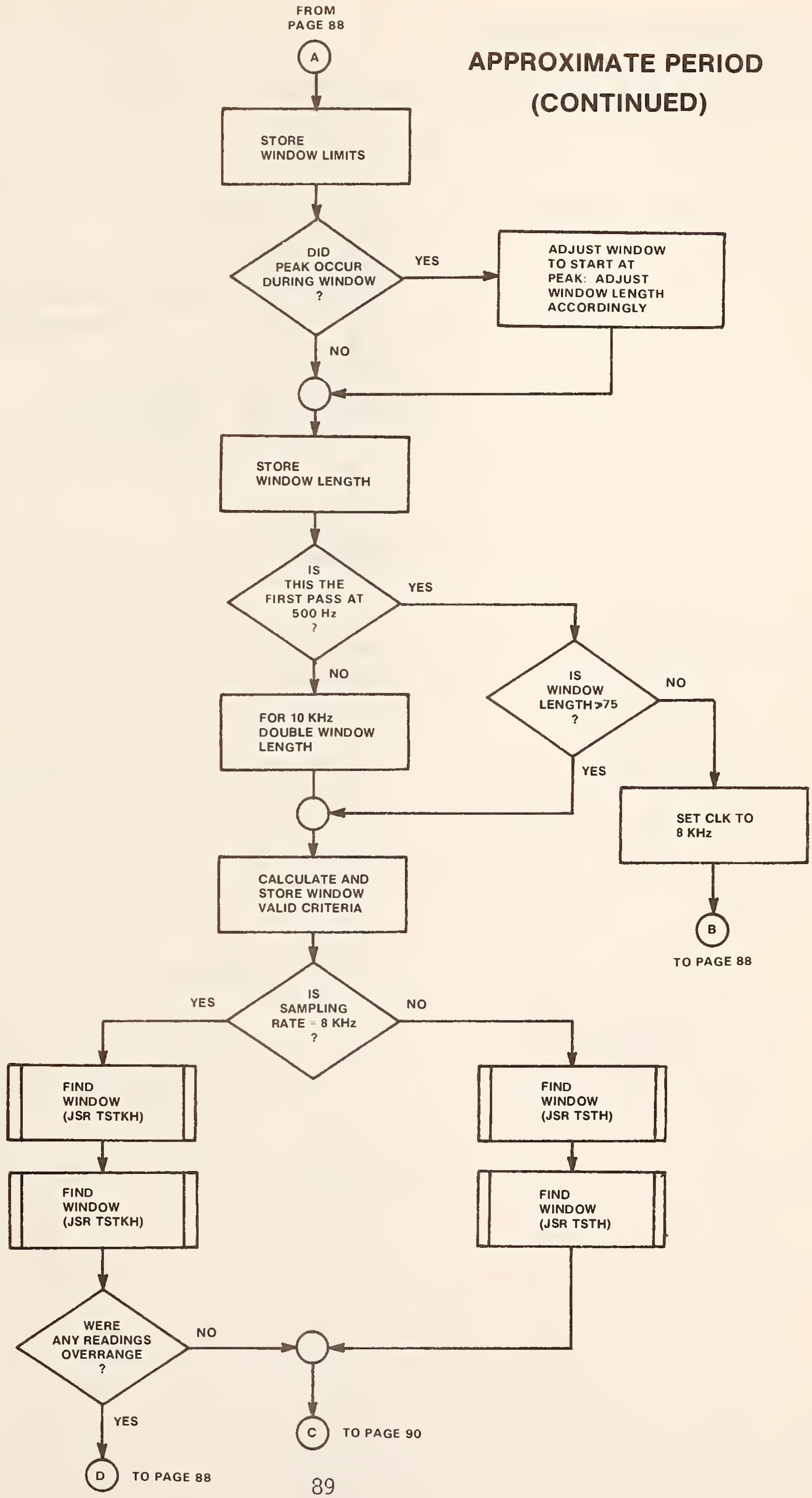
Miscellaneous Hardware

1.	Mounting hardware for fan	
	flat head 3/4" steel 6-32 screw	4
	6-32 steel nut	4
2.	Mounting hardware for 902I power supply	
	6-32 1/4" steel screw	4
	6-32 1/2" steel screw	4
	1/2" thread 6-32 steel spacer	4
3.	Mounting hardware for card cage	
	angle iron 1", 1/2", 1/8" thickness	2
	8/32 steel screws 1/2"	2
	6/32 steel screws 3/4"	6
	6/32 steel nut	9
	steel washer 0.19610	2
4.	Iron mesh, 7" x 5"	1
	6/32 1" steel screw	4
	6/32 steel nut	4
	retaining washer	4
5.	Mounting hardware for front panel	
	6/32 1" flat head steel screw	4
	steel spacer 1/4" 6-32 threaded	8
	insulating washer 0.196 ID	8
	nylon nut, 6-32	8
6.	Side panel to front panel	
	8-32 flat head 1/2" steel screw	4
7.	Bracket hardware (item #28 on instrument parts list)	
	8-32 1/4" steel screw	2
	8-32 1/2" steel screw	3
	8-32 steel nut	3
8.	Side panel to back panel	
	8-32 1/2" steel screw	2
9.	Top lid hardware	
	6-32 1/4" steel screw	4
10.	Bottom lid hardware	
	6-32 1/4" steel screw	4
11.	Mounting hardware for switches	
	4-40 1/4" steel screw	5

APPROXIMATE PERIOD ROUTINE



APPROXIMATE PERIOD (CONTINUED)



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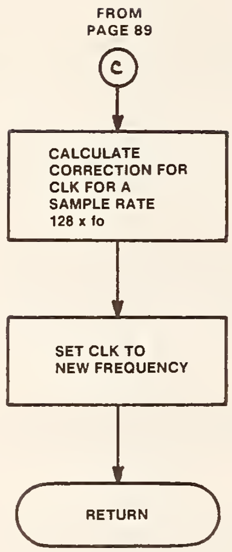
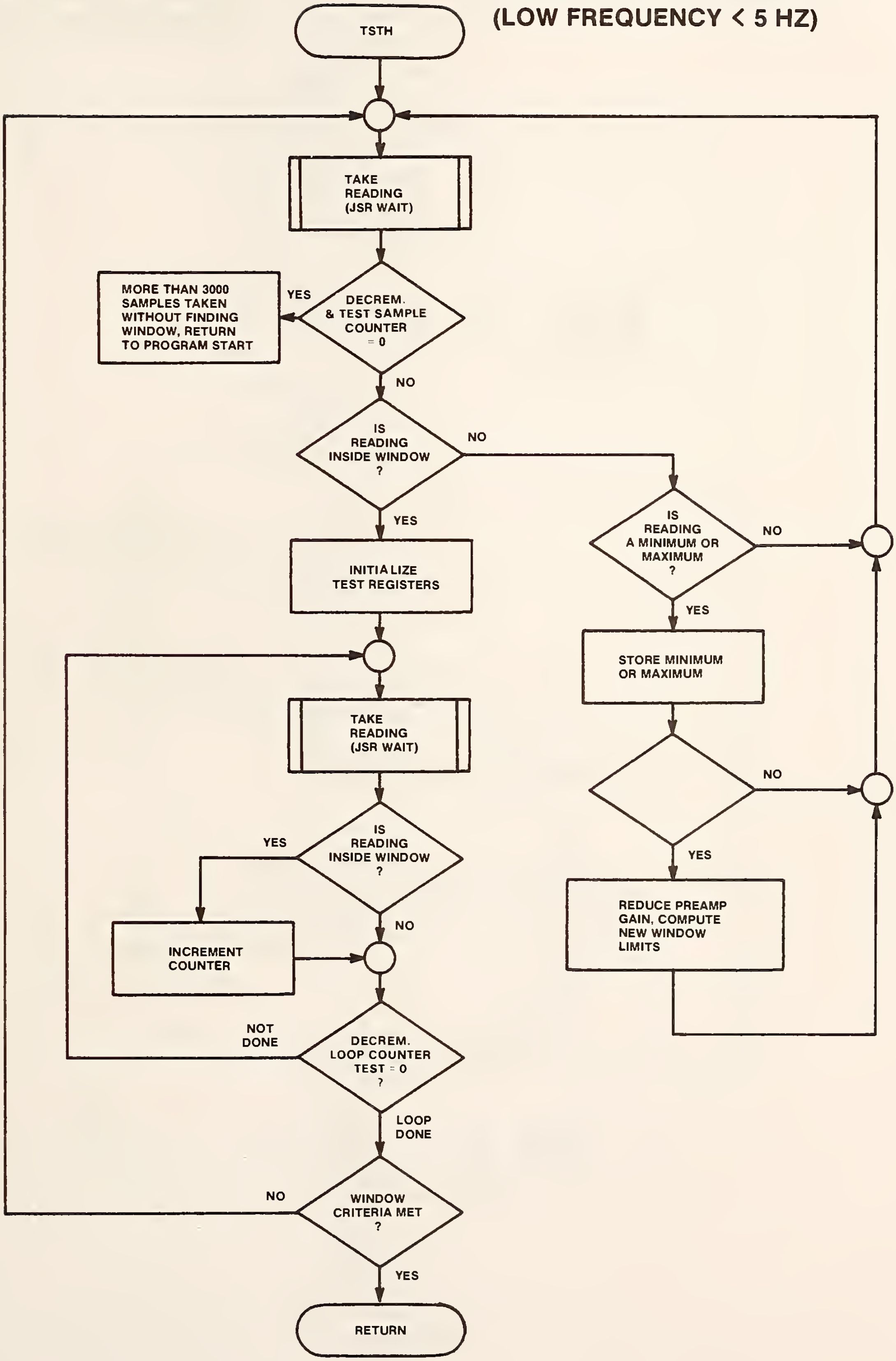
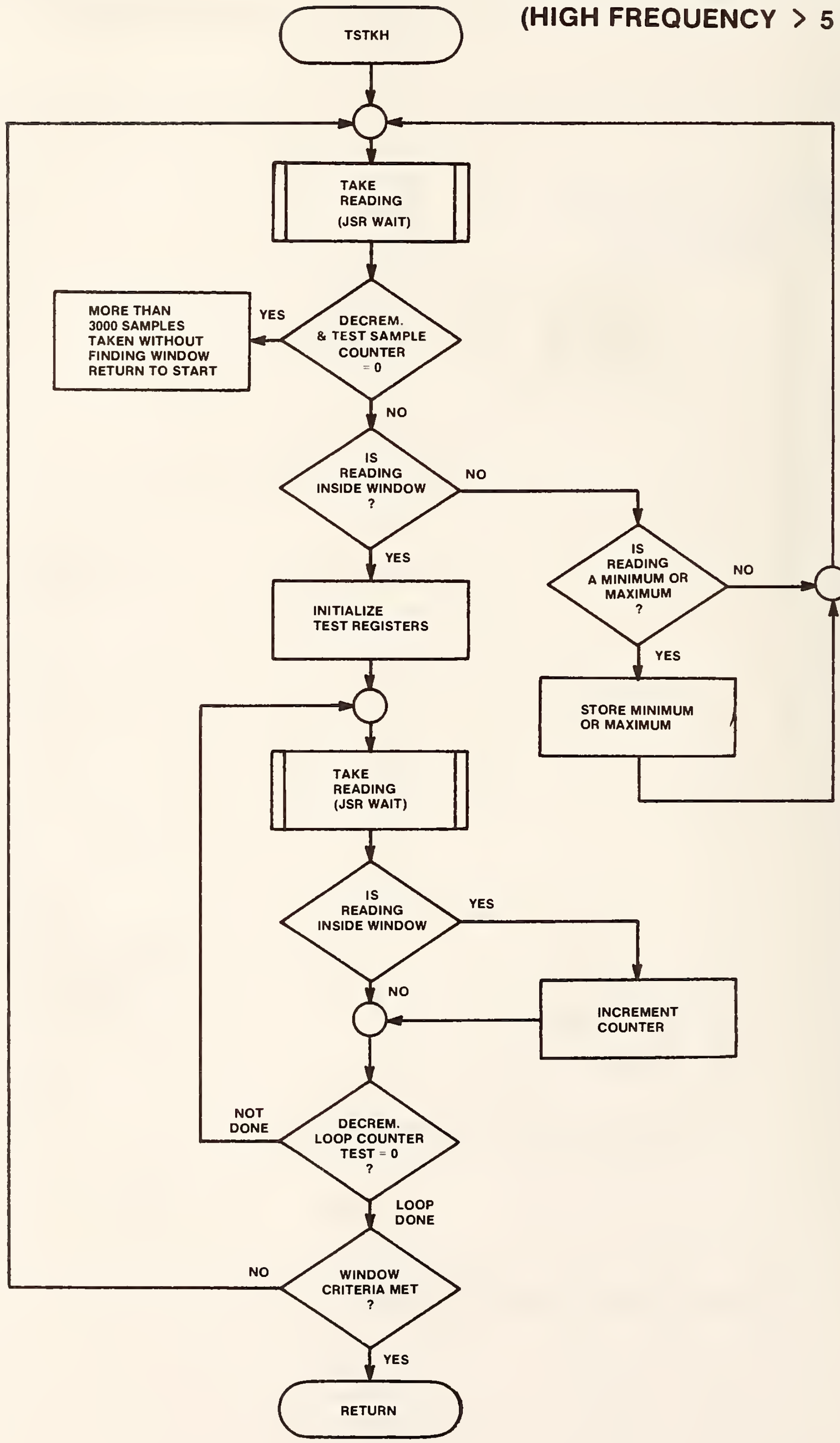


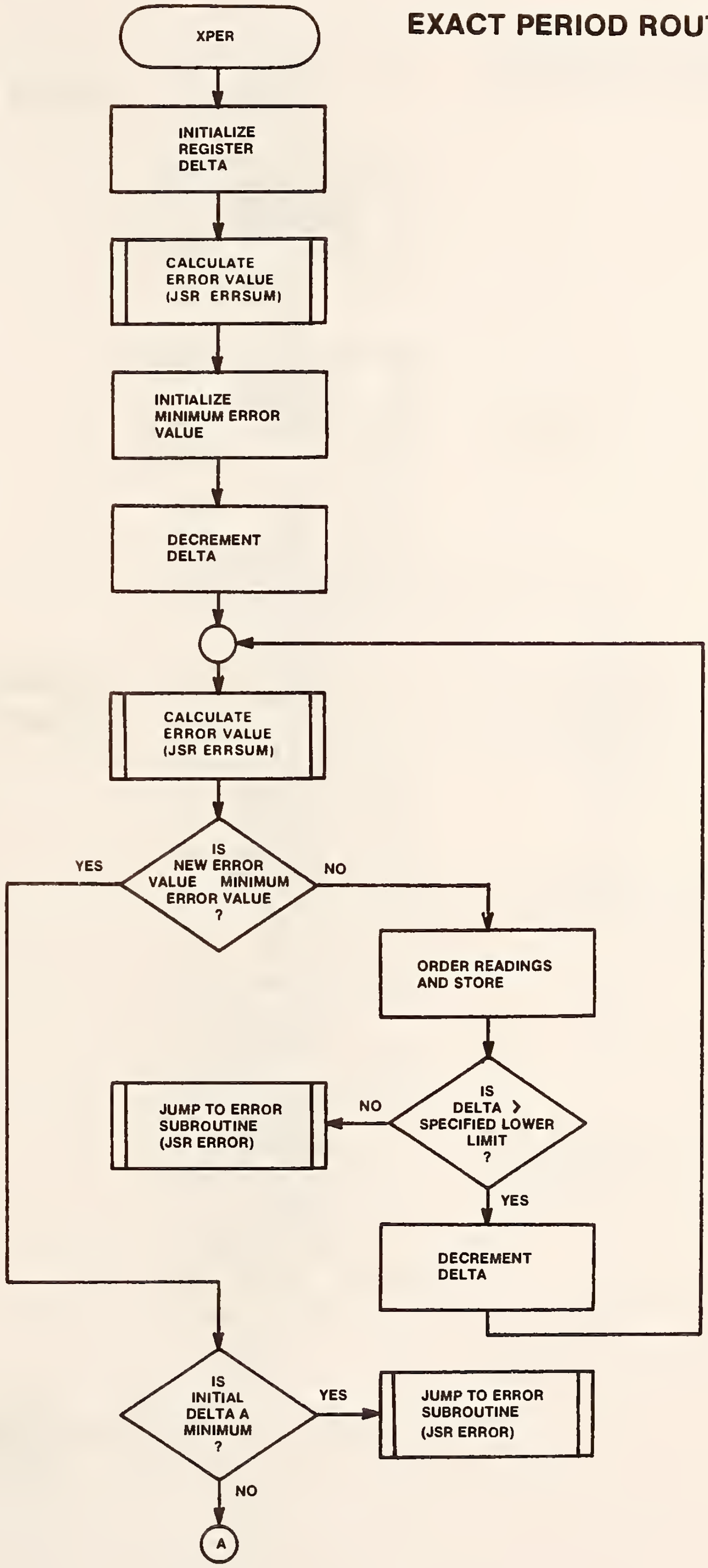
SUBROUTINE IN APPROXIMATE PERIOD TO FIND WINDOW (LOW FREQUENCY < 5 HZ)



SUBROUTINE IN APPROXIMATE PERIOD TO FIND WINDOW (HIGH FREQUENCY > 5 HZ)



EXACT PERIOD ROUTINE



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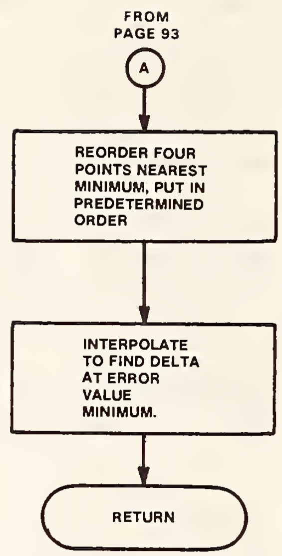
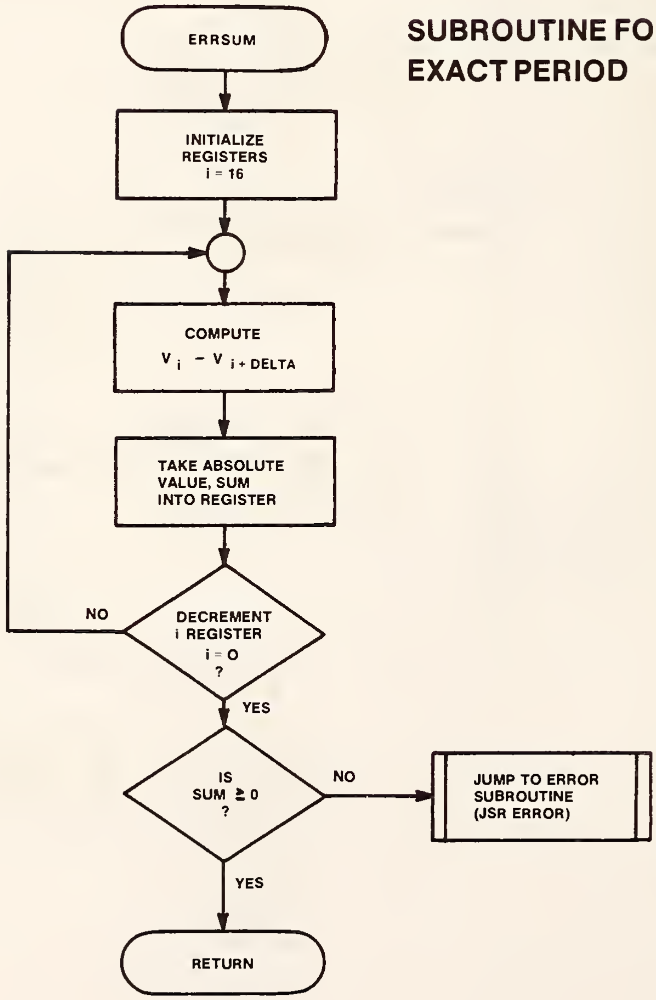
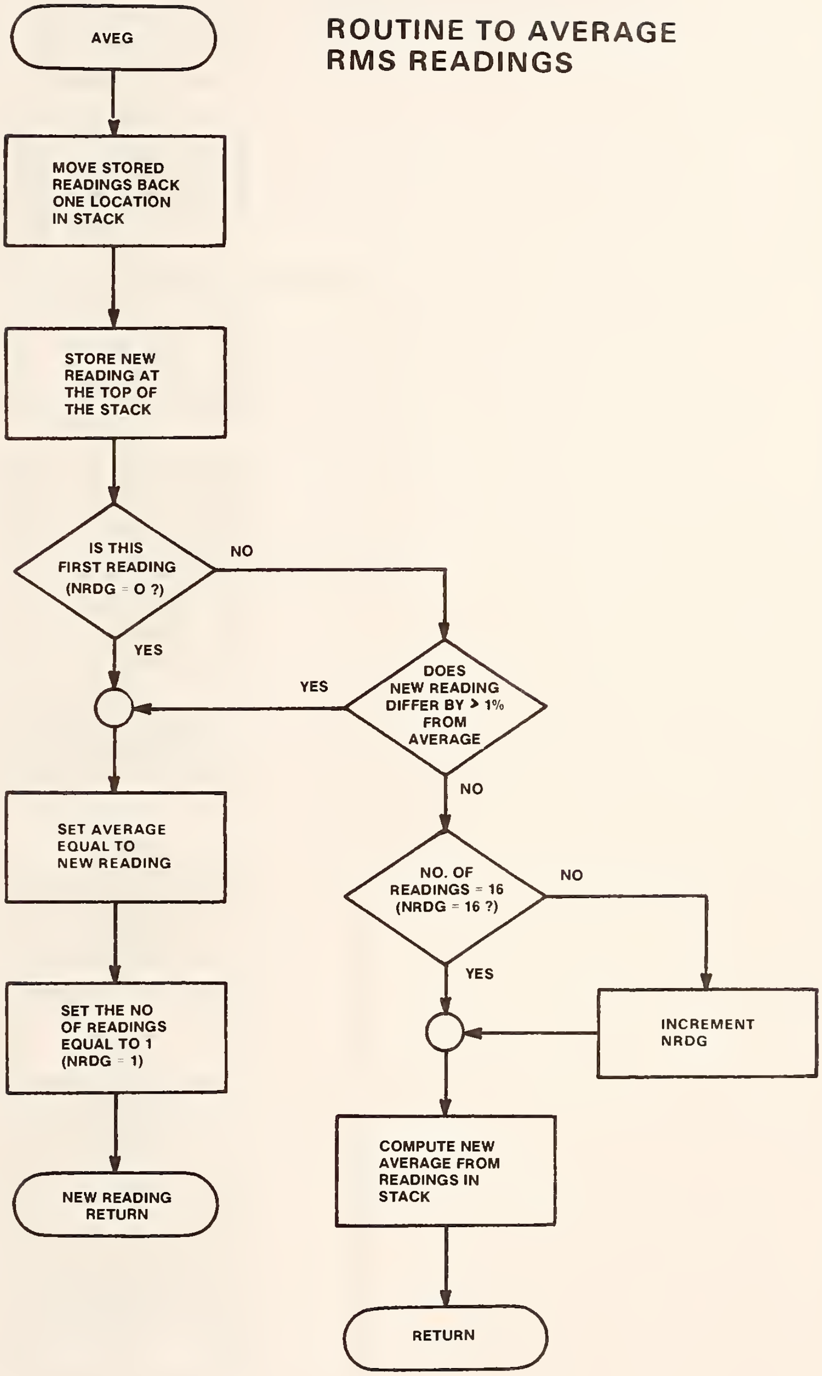
EXACT PERIOD (CONTINUED)



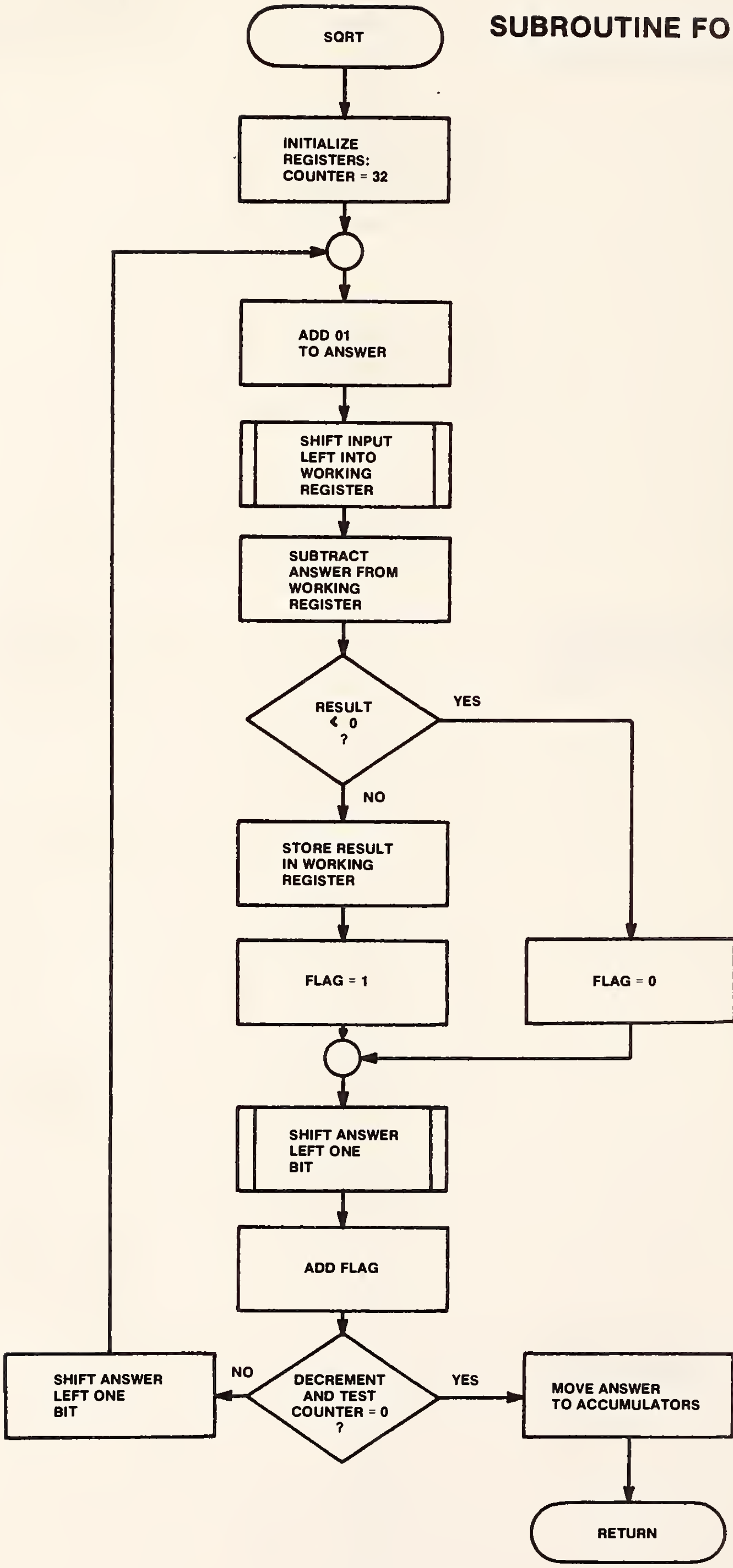
SUBROUTINE FOR EXACT PERIOD



ROUTINE TO AVERAGE RMS READINGS



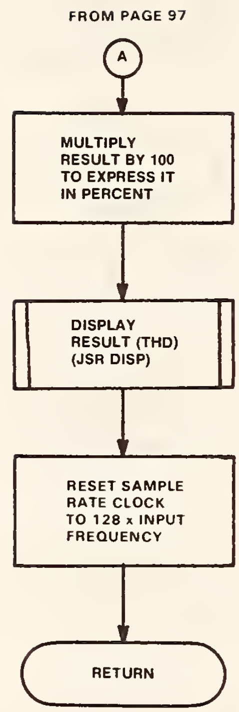
SUBROUTINE FOR SQUARE ROOT



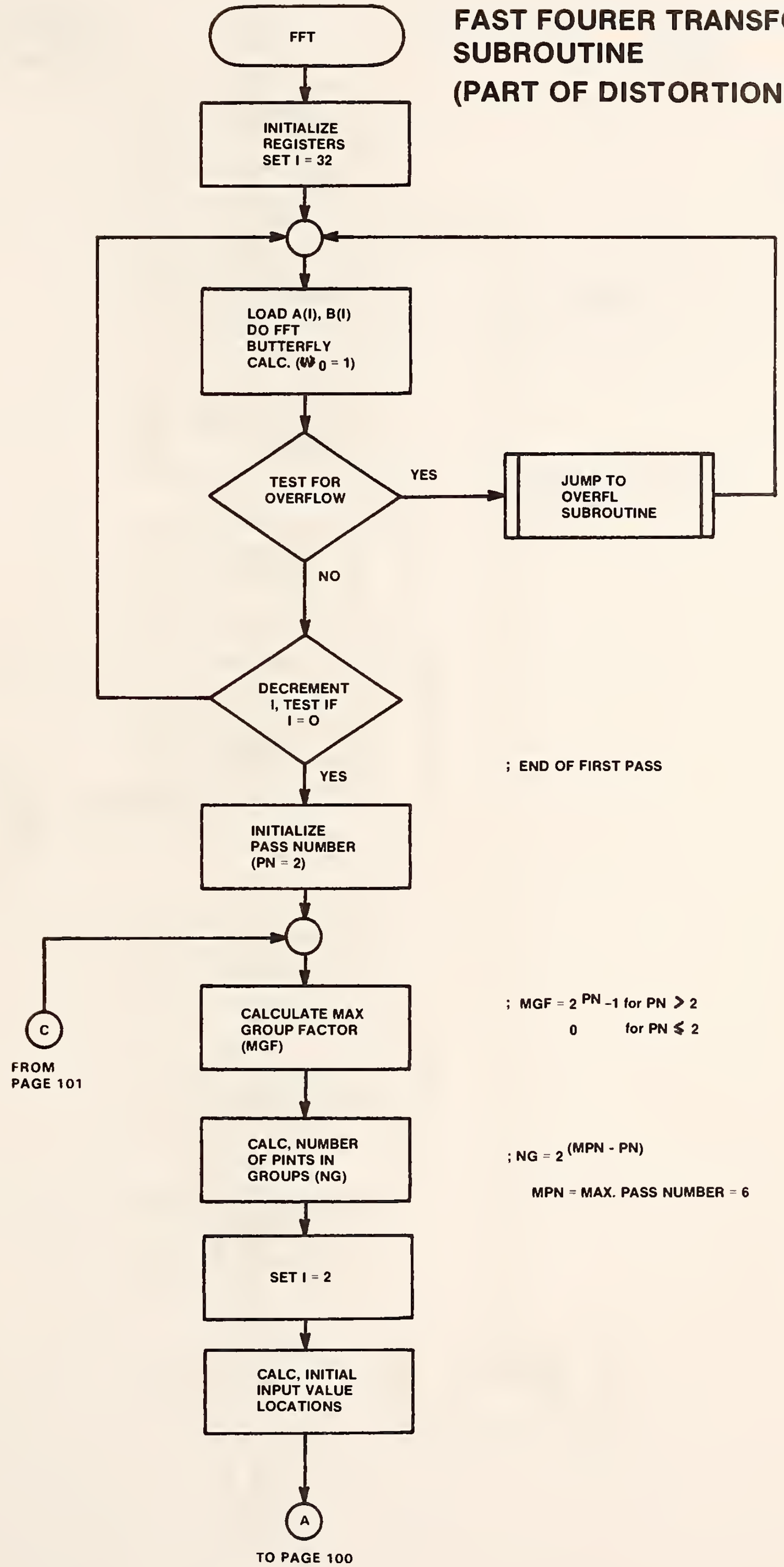
DISTORTION ROUTINE

TO PAGE 98

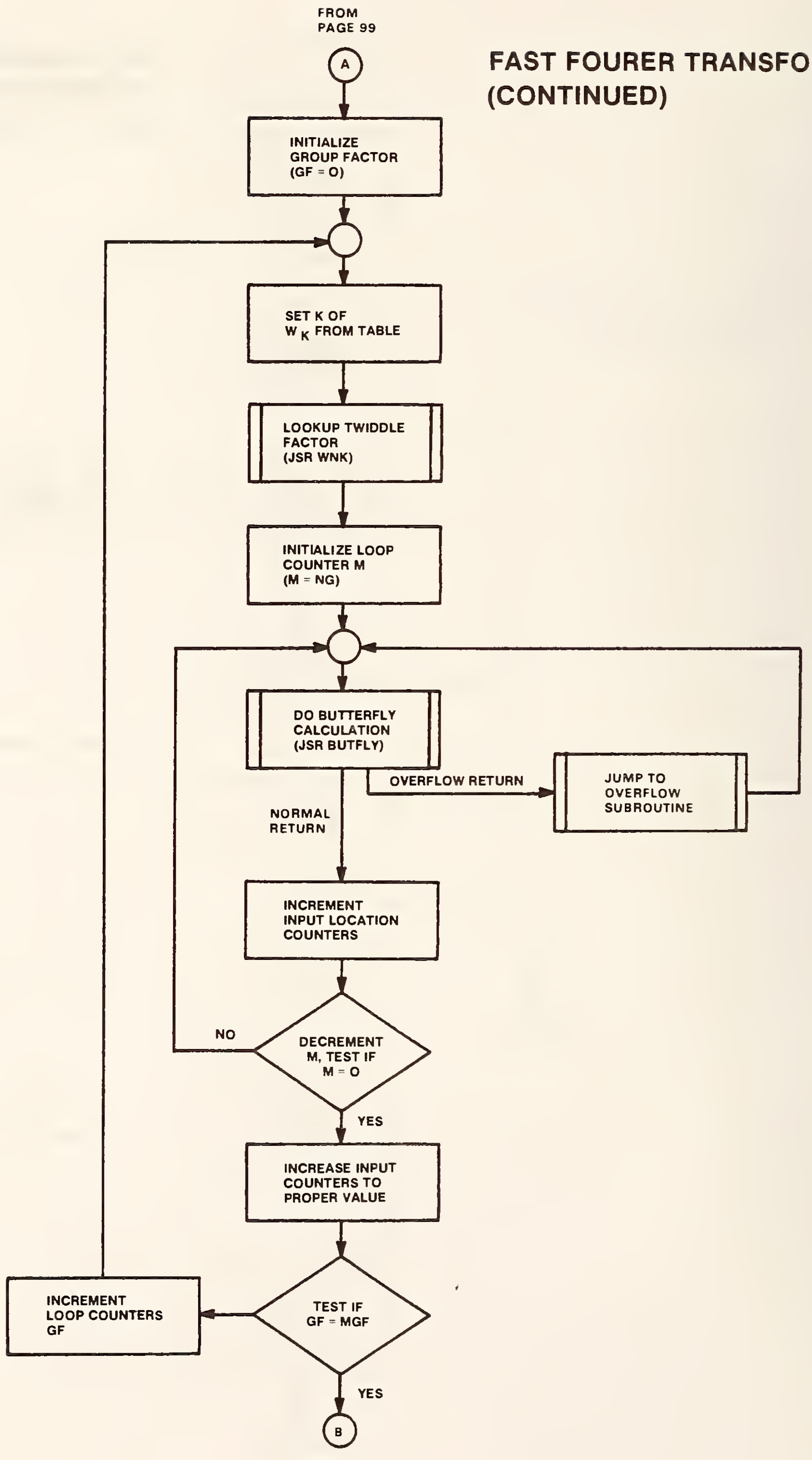
DISTORTION ROUTINE
(CONTINUED)



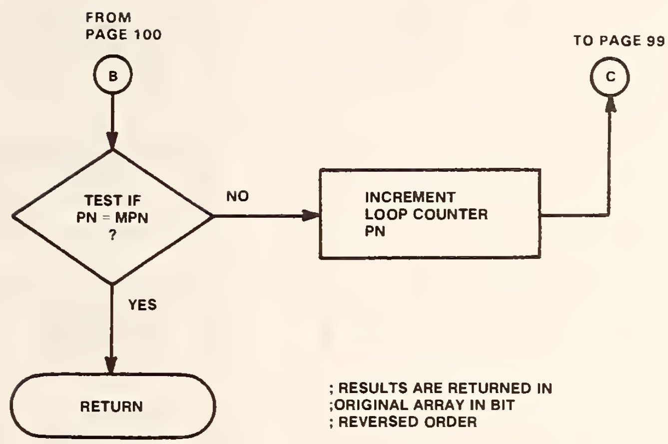
FAST FOURER TRANSFORM SUBROUTINE
(PART OF DISTORTION ROUTINE)



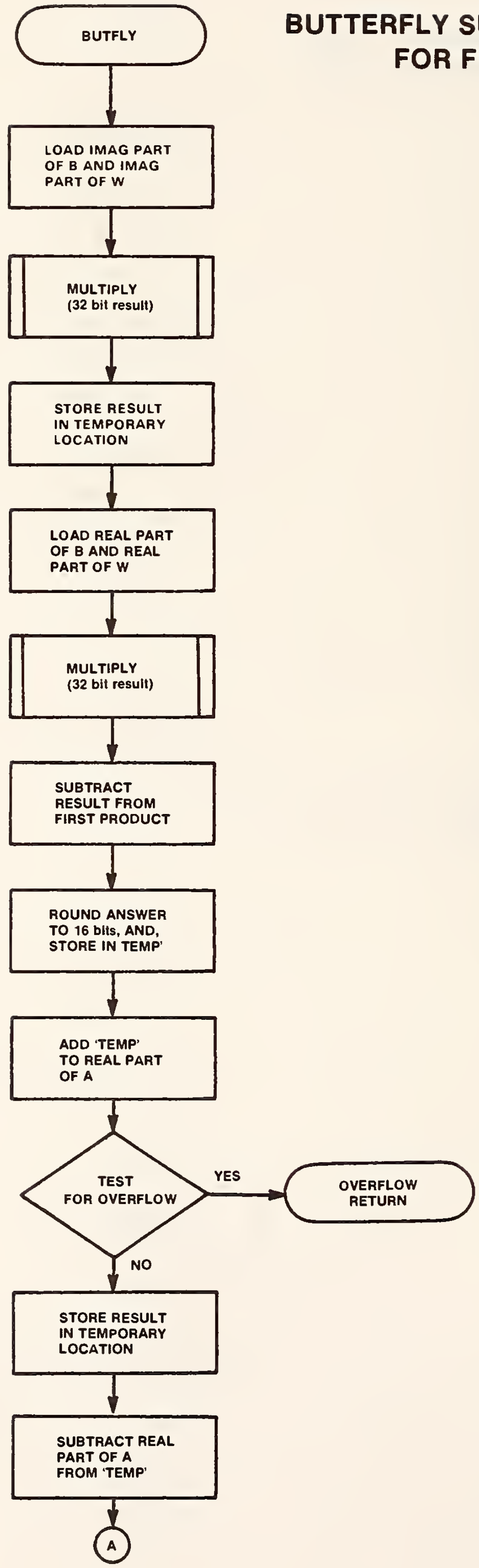
FAST FOURER TRANSFORM (CONTINUED)



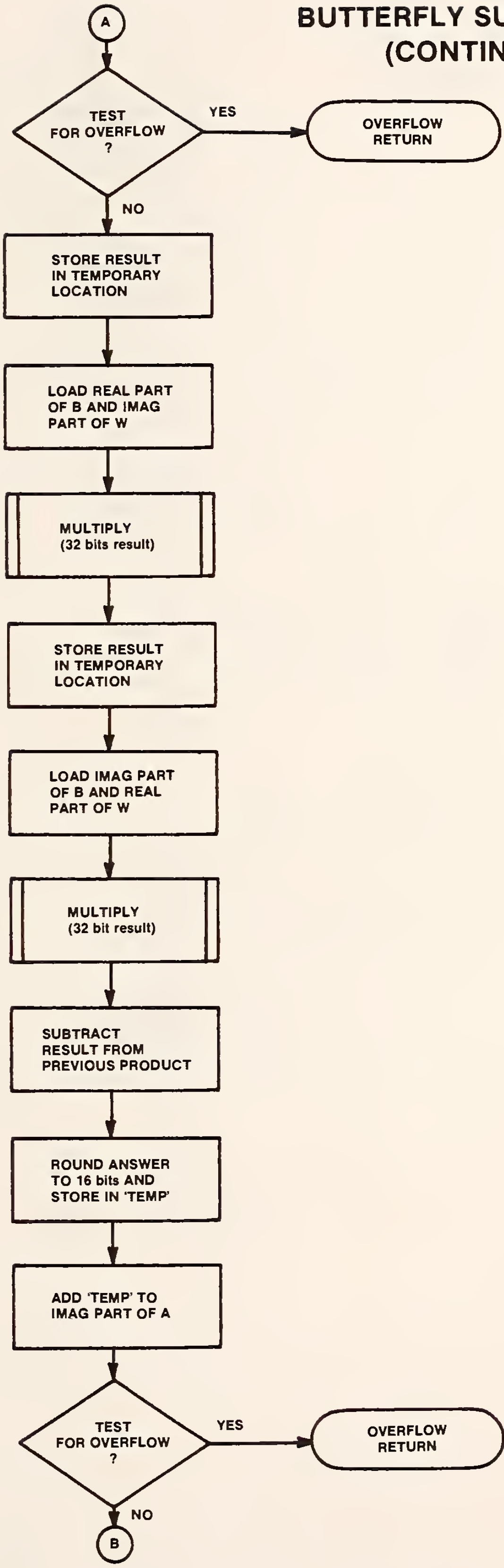
FAST FOURIER TRANSFORM (CONTINUED)



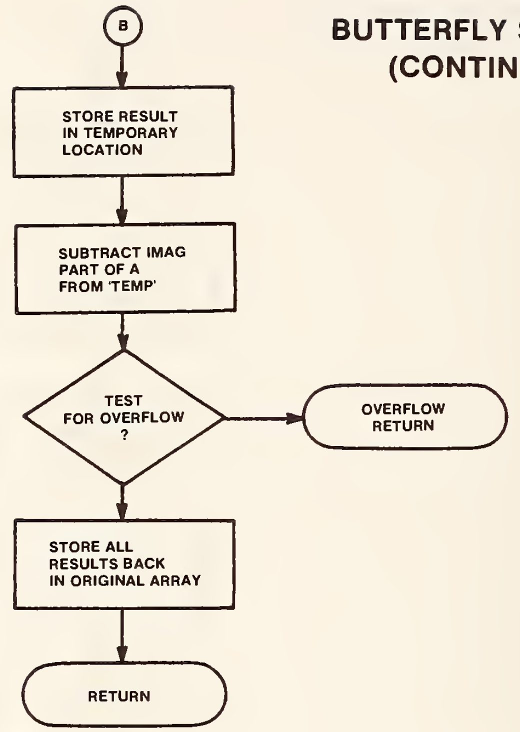
BUTTERFLY SUBROUTINE FOR FFT

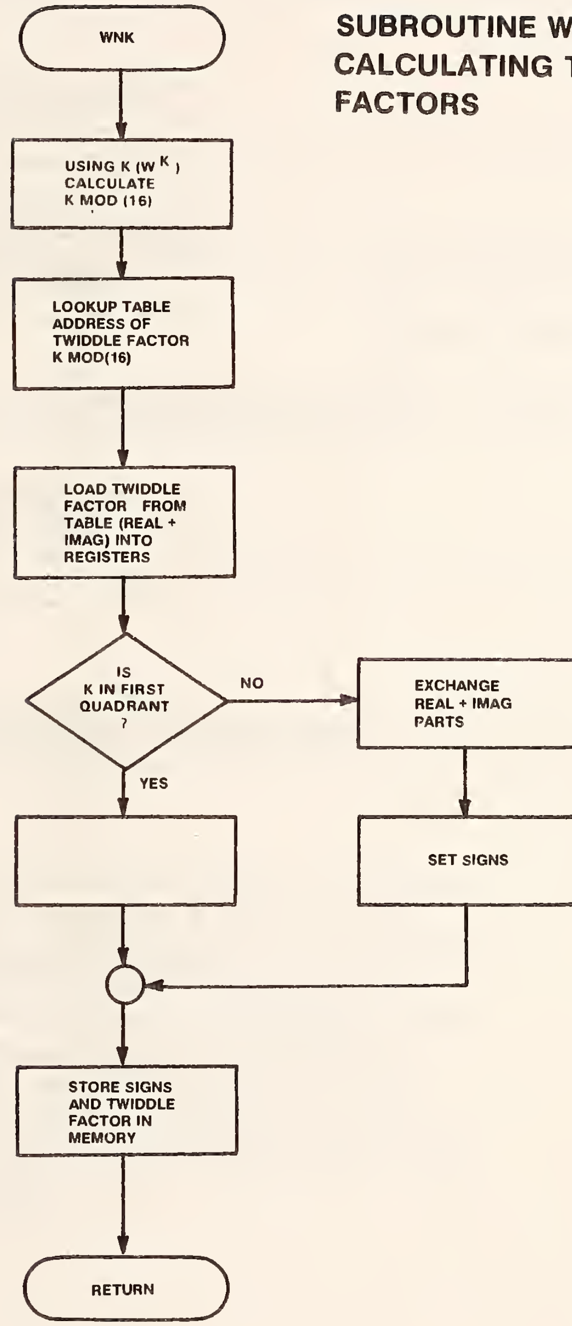


BUTTERFLY SUBROUTINE (CONTINUED)



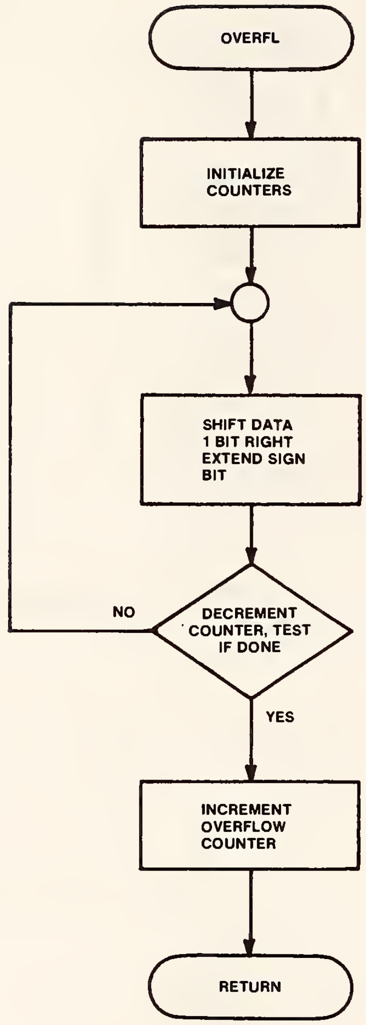
BUTTERFLY SUBROUTINE (CONTINUED)





SUBROUTINE WNK FOR CALCULATING TWIDDLE FACTORS

OVERFLOW SUBROUTINE



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4. TITLE AND SUBTITLE

A FAST RESPONSE, LOW-FREQUENCY SAMPLING VOLTMETER

5. AUTHOR(S)

Barry A. Bell, Bruce F. Field, and Thomas H. Kibalo

6. PERFORMING ORGANIZATION (If joint or other than NBS, see instructions)

NATIONAL BUREAU OF STANDARDS
DEPARTMENT OF COMMERCE
WASHINGTON, D.C. 20234

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DC and Low Frequency Engineering Subgroup of the Calibration Coordination Group (CCG) of the Department of Defense.

10. SUPPLEMENTARY NOTES

[ ] Document describes a computer program; SF-185, FIPS Software Summary, is attached.

11. ABSTRACT (A 200-word or less factual summary of most significant information. If document includes a significant bibliography or literature survey, mention it here)

A low-frequency voltmeter utilizing a sampling technique implemented with microprocessor-based electronics has been developed to perform as a true rms ac voltmeter and distortion analyzer. The instrument makes measurements accurate to $\pm$0.1 percent (of reading) of the fundamental frequency, total harmonic distortion, and true rms voltage of approximately sinusoidal inputs from 2 mV to 10 V and frequencies from 0.1 to 120 Hz. A major feature of this instrument is the special window crossing and error function algorithms which provide a software means for completing a measurement within two signal periods at frequencies below 10 Hz.

12. KEY WORDS (Six to twelve entries; alphabetical order; capitalize only proper names; and separate key words by semicolons)

algorithm; converter; distortion; microcomputer; rms value; sampling; signal period.

13. AVAILABILITY

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