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ARMOUR RESEARCH FOUNDATION OF ILLINOIS INSTITUTE OF TECHNOLOGY

ARF Project No. E112

INSTRUMENTATION AND MEASUREMENT TECHNIQUES STUDY

Report No. 12

Twelfth Quarterly Progress Report

1 November 1961 - 31 January 1962

Signal Corps Contract No. DA 36-039 SC-78269

DA Project Nr. 3G89-01-021-01

for

U. S. Army Signal Research and Development Laboratory

Fort Monmouth, New Jersey

25 years of research

ARF 5112-Q12

ARF Project No. E112

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This study is aimed at advancing the state of the art of electrical measurements, and directed toward the examination of current and recent research findings in the physical sciences for exploitation in electrical measurements and instrumentation. The new techniques developed due to the scope of the program are intended for ultimate use in the development of improved military electronic test equipment.

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INSTRUMENTATION AND MEASUREMENT TECHNIQUES STUDY

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I. PURPOSE

This study is aimed at advancing the state of the art of electrical measurements and directed toward the examination of current and recent research findings in the physical sciences, for exploitation in electrical measurements and instrumentation. These new techniques, after optimization by further research, are intended for ultimate use in the development of improved military electronic test equipment, ranging from laboratory-type engineering measurement standards to field-type test sets and indicators intended to provide rapid and simple performance tests on prime tactical electronic equipment. The study records throughout the entire program are compiled in technical reports arranged to provide design data and approach suggestions for use by military test equipment development engineers.

II. ABSTRACT

A current probe for use with printed circuitry which senses branch currents in parallel circuits and operates at a fixed frequency of 159 Kc is described. The sensitivity is such that one milliampere in a conductor produces seven volts output from the associated amplifier. The noise equivalent input is about one microampere. Preliminary testing has proven the device successful.

The location of faults on loaded spiral-four cable by means of a phase measurement technique is evaluated experimentally. Agreement with predicted results is not sufficient to yield accurate results, however, there is evidence that the same procedure in conjunction with empirical data from lines in good operating condition would produce results with adequate accuracy.

The theory and operation of a wideband 90° phase shifter is discussed. This device uses Hall-element product modulators and should be capable of minimizing intermodulation distortion in SSB equipment.

III. PUBLICATIONS, LECTURES, REPORTS AND CONFERENCES

Publications: An instruction manual for the operation of the Visual Circuit Tester was completed and forwarded to the Signal Corps.

Lectures: None

Reports: None

Conferences: A project conference was held at ARF on 20 December 1961, with representatives of ARF and USASRDL present. Work accomplished during the eleventh quarter was discussed, and the current probe was demonstrated. Work being done during the twelfth quarter was outlined. The priority of emphasis on the various phases was outlined.

Trips: A trip was made to USASRDL by ARF personnel to take measurements on loaded spiral-four lines on 16 January 1962. The results of these data are reported in the Factual Data section of this report.

IV. FACTUAL DATA

A. IN CIRCUIT TESTING

1. Visual Test Unit

During this report period the visual test unit was delivered to the sponsor together with an instruction manual. The manual described by means of oscillograms the tests that could be made to locate faults in a simple transistorized pulse amplifier. Four of these amplifiers were supplied to the sponsor.

2. Current Probe for Printed Circuits

The previous quarterly report (report No. 11) described the theory of operation of a current probe as a convenient means of measuring magnitude and phase of currents in parallel connected components on printed circuits.

The early tests made use of conventional magnetic recording heads as current probes. These tests showed the feasibility of such a system but indicated that the magnetic recording heads were undesirable as current probes because of large physical size, limited upper frequency range, low Q windings, and curved front surfaces.

3. Current Probe Design

Several current probe designs were fabricated and tested using both laminated and ferrite magnetic structures. Tests indicated that ferrite magnetic circuits provided the best signal to noise ratio for excitation frequencies greater than about ten kilocycles.

A current probe was fabricated from a two inch length of one quarter inch diameter type H ferrite rod. One end of the rod was ground so that the cross section was approximately a one eighth by one quarter inch rectangle. A one eighth inch hole was drilled through the flattened end of the ferrite rod and a ten mil gap was milled into the end of the rod. This magnetic structure was wound with one hundred turns of number 46 wire and potted in a three eighths inch diameter copper tube to provide shielding and protection against breakage. The inductance of the probe is 300 microhenries and has a Q of 15 at 160 kc.

4. Tuned Preamplifier

In order that the current probe can be used to measure small values of current with good signal noise ratio, it is desirable that the probe be used in conjunction with a tuned preamplifier.

An ac powered, fixed tuned preamplifier has been constructed for use with the current probe previously described. The preamplifier has fixed gain and the only control is an ac power switch. The amplifier is fixed tuned at a frequency of 159 kc and contains a capacitor at the input terminals to resonate the probe inductance at this frequency. This particular frequency was chosen for convenience in calculating reactance values of capacitors and inductors. At this frequency a 1000 μμ fd capacitor and a one millihenry inductor have a reactance of one thousand ohms.

5. System Response

The response of the probe and amplifier is such that one milliampere current in a conductor produces seven volts at the output of the amplifier. The noise level of the system is less than the level produced by one microampere current in the conductor, and the maximum current the system will accept without amplifier overload is about three milliamperes. This combination gives a dynamic range of about 70 decibels.

The frequency response of the probe and amplifier combination exhibits band pass characteristics. The bandwidth of the combination is 2 kc at 1 db, 3 kc at 3 db, 4 kc at 6 db, 8 kc at 20 db, and 20 kc at 40 db.

The circuit diagram of the tuned amplifier is shown schematically in Figure 1. The appearance of the current probe and preamplifier is shown in Figures 2 and 3.

It should be noted that for printed circuit conductors between 1/16 inch and up to 1/4 inch, the sensitivity of the probe remains constant. For the smaller width, the alignment of the probe (along the conductor) is critical, while for conductors greater than 1/4 inch the field on either side of the probe's sensitive area becomes appreciable. Current in adjacent conductors is fairly well shielded from the probe by the copper eddy - current shield surrounding the probe. The field surrounding an inductor, however, is generally strong enough to affect the reading by a few percent.

There is also a certain amount of electrostatic pickup between the probe and the conductor. This electrostatic pickup component is not in phase (in general) with the desired current and can usually be eliminated by suitable grounding of the generator and probe terminals.

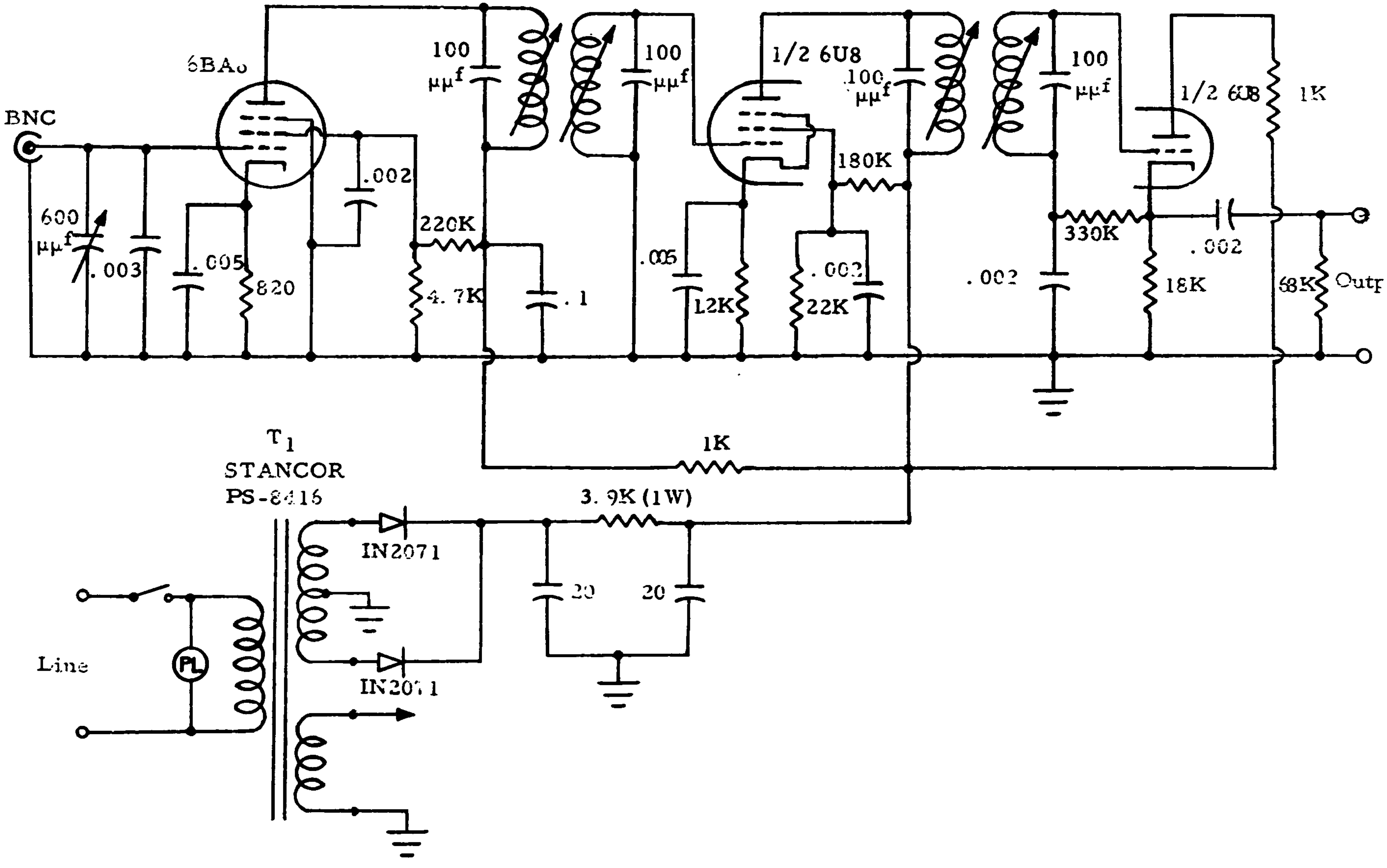

FIG. 1 159 Kc PRE - AMPLIFIER CIRCUIT DIAGRAM

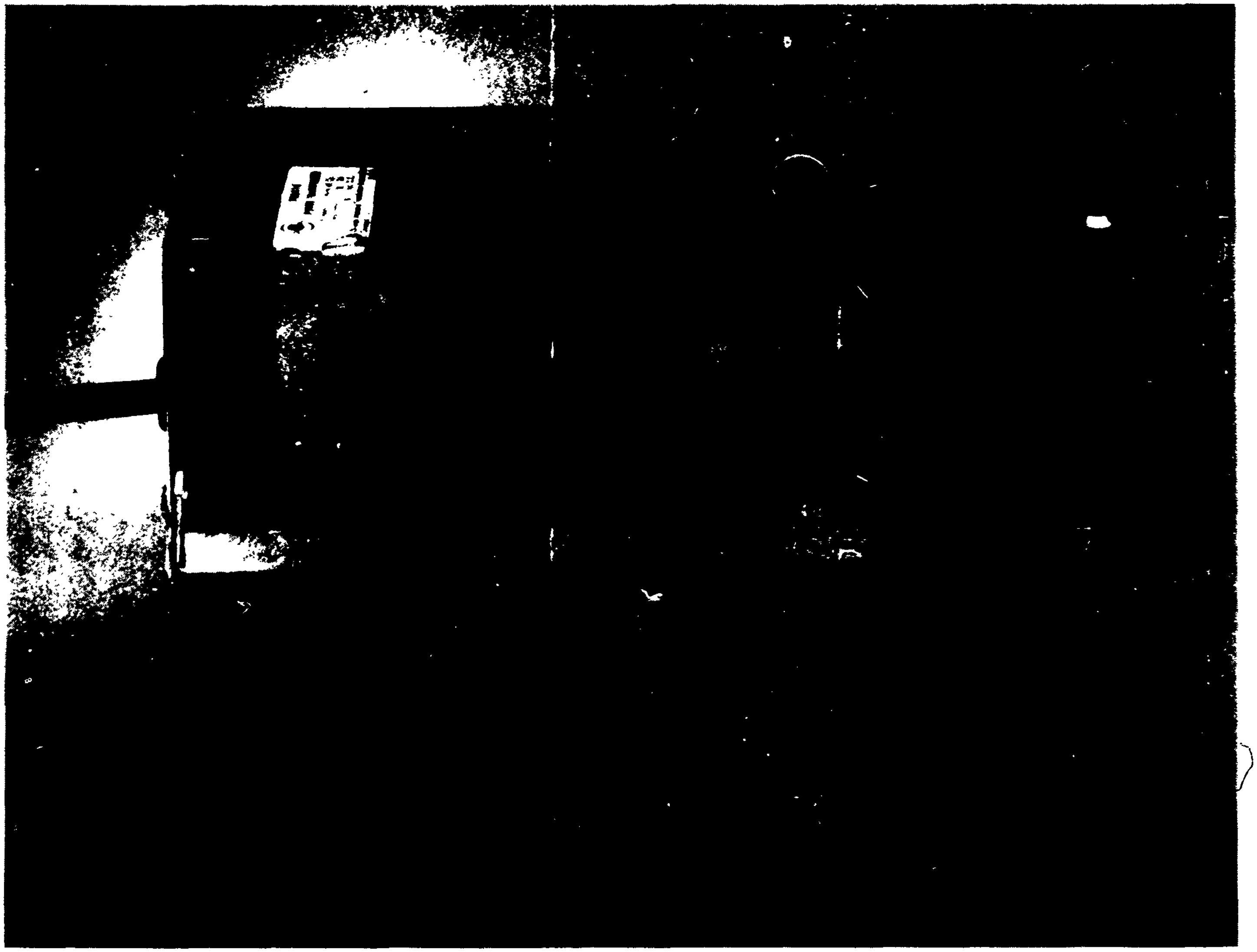

FIG. 2 159 Kc PRE-AMPLIFIER

FIG. 3 CURRENT PROBE AND PRE-AMPLIFIER

In summary, it may be said that the errors are mainly due to extraneous magnetic fields entering through the ferrite end of the probe, and some small electrostatic pickup.

To date the testing of the current probe has been of a preliminary nature, that is, mainly to determine some of the device's limitations. It has not been coupled to a phase meter, for example, to determine the exact precision to which relative phase angles can be measured.

Various relatively simple configurations of parallel passive elements such as shown in Fig. 3 have been tested with success. At 159 Kc, 1000 μμf and one millihenry each have a reactance of 1000 ohms, which probably represents an optimum choice for a single frequency device since at this frequency, excitation voltages and currents should each remain low enough to maintain safe dissipation levels in all components.

During the next period the device will undergo quantitative as well as further qualitative testing. Such a program should determine the success to which precision measurements can be made, the necessity of providing one or more additional frequencies of operation, and what, if any, attenuation needs to be provided.

B. LINE FAULT LOCATION

1. Introduction

In the Ninth Quarterly Report for this contract there is a description of a method of locating a short or open circuit on a transmission line by comparing the phase angle of the input impedance of the defective line to the phase angle of the characteristic impedance. This method will hereafter be called the phase-coincidence method, and the lowest frequency at which these two phase angles are equal will be called the phase-coincidence frequency.

During the past quarter, the study of unloaded spiral-four cable with repeaters was completed, and primary attention was then paid to loaded cable. In the previous (eleventh) quarterly report, the requirement for a 20-mile range for locating faults on unloaded cable with repeaters was discussed, and the accompanying data suggested that this limit could not be reached in the practical case. Measurements have since been made on an actual unloaded line containing repeaters, and the experimental data seems to confirm the impossibility of locating faults at 20 miles with this technique. There was almost no change in phase angle readings as the fault was moved from 14.5 miles to 20 miles. Unfortunately, the measurements were made with the cable shield tied to the test equipment ground, and it was discovered later that this connection alters the readings from the true values. The lack of significant variation between the 14.5 mile and 20 mile readings, however, still indicated confirmation of the original conclusion.

Theoretical and experimental data will be presented in the following paragraphs relative to loaded spiral-four cable without repeaters.

2. Predicted Results for Loaded Spiral-Four

Computer predictions have been made of phase angle versus frequency for spiral-four cable with inductive loading at 1/4 mile intervals. The CU-260/G loading coils used with this cable have a total loop inductance of

6 millihenries per conductor pair. The ac resistance varies with frequency, however, the variation is not known and the resistance is small compared to that of 1/4 mile of cable, so the published value of 3.77 ohms (measured at 20 kc) was used.

Table 1 shows the computed values of the loaded line parameters as a function of frequency. The values of arg Z_0 (the phase angle of Z_0) are plotted in Fig. 4, and the curve is repeated for reference in Fig. 5 through 9.

The predicted values of cable input impedance angle as a function of frequency for fault distances of 2, 10, 20, 30, and 40 miles were computed, and these are shown in Fig. 5 through 9. Both open-circuit and short-circuit faults are shown.

From the computed values of β in Table 1, the length of a quarter-wave section of cable as a function of frequency was determined, and these results are plotted in Fig. 10. This curve determines the distance to an open or short circuit fault when the phase-coincidence frequency is known, therefore Fig. 10 is titled "Fault Distance vs. Phase-Coincidence Frequency." The following procedure will demonstrate the method.

Taking Fig. 5 as an example, the lowest frequency at which each fault curve intersects the arg Z_0 curve is noted. These values are approximately 2730 cps for both the short circuit and open circuit curves. Going now to Fig. 4, the fault distance corresponding to 2730 cps is seen to be 2.0 miles. Exact correlation should not be surprising in this case, since the curves are the results of computer predictions assuming a perfectly

Table I

PARAMETERS OF LOADED SPIRAL - FOUR CABLE (WF-8/G)

FREQ.	$\|Z_0\|$	ARG Z_0	α	β
50	2035	-42.6°	.0355	.0382
70	1717	-41.7	.0412	.0456
100	1440	-40.6	.0481	.0557
150	1189	-38.1	.0567	.0704
200	1041	-36.0	.0633	.0843
300	871.7	-32.0	.0725	.110
500	723.5	-25.8	.0817	.163
700	657.1	-21.2	.0875	.216
1000	612.1	-16.5	.0895	.297
1500	581.7	-11.8	.0923	.436
2000	570.1	- 9.2	.0929	.577
2400	567.2	- 7.7	.0954	.688
2700	561.9	- 7.0	.0944	.776
3000	561.0	- 6.0	.0986	.861
3200	559.5	- 5.6	.0990	.918
3600	557.9	- 5.1	.0971	1.032
4000	556.5	- 4.4	.0989	1.146
4500	554.6	- 3.9	.0979	1.288
5000	553.4	- 3.7	.0945	1.430
5500	550.1	- 3.2	.0962	1.577
6000	548.9	- 3.0	.0944	1.720
6500	544.9	- 2.8	.0927	1.869
7000	542.5	- 2.6	.0938	2.016
7500	541.1	- 2.4	.0951	2.157
8000	537.9	- 2.4	.0907	2.306

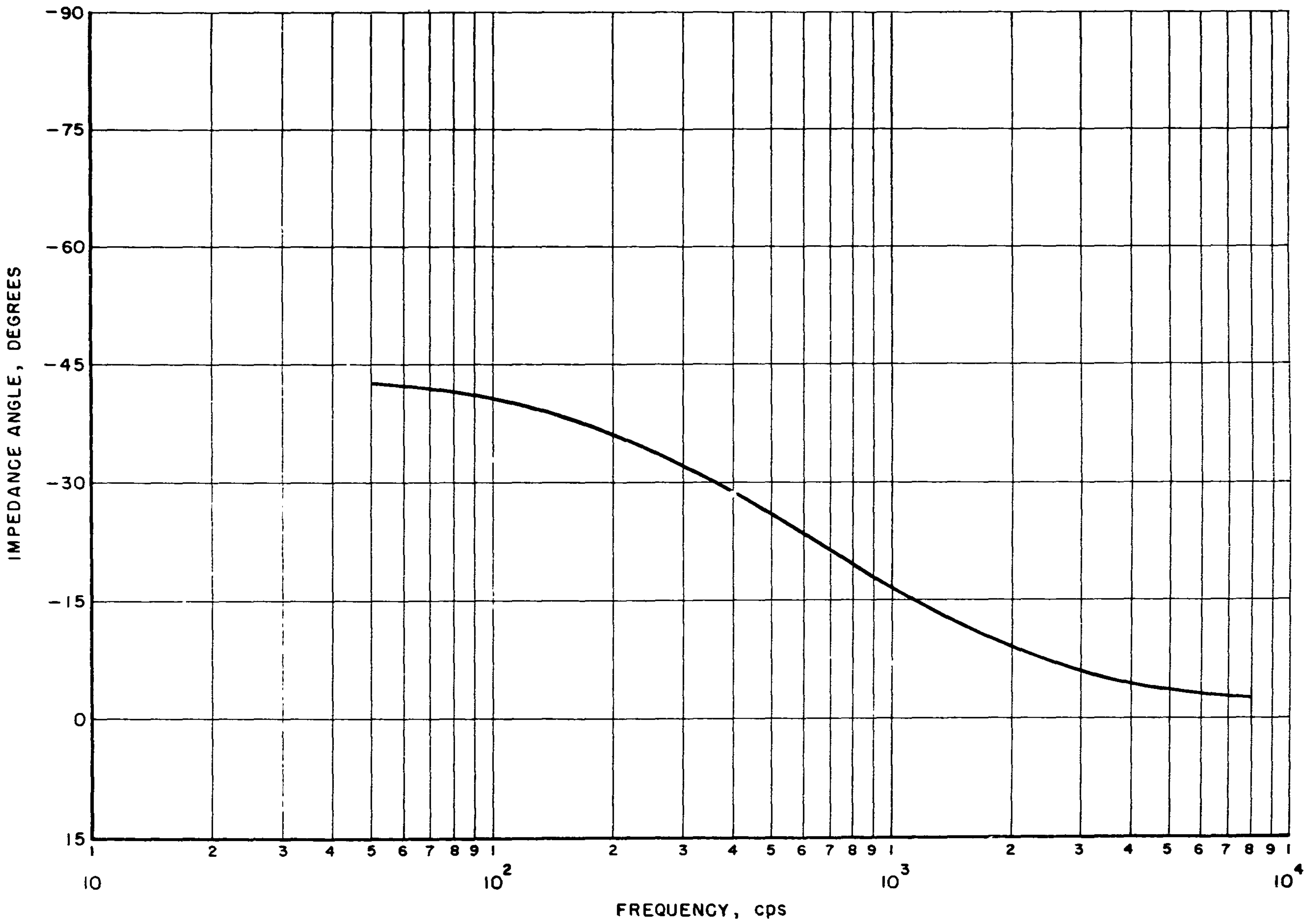

FIG.4 CHARACTERISTIC IMPEDANCE ANGLE VERSUS FREQUENCY FOR LOADED SPIRAL-FOUR CABLE (WF-8/G)

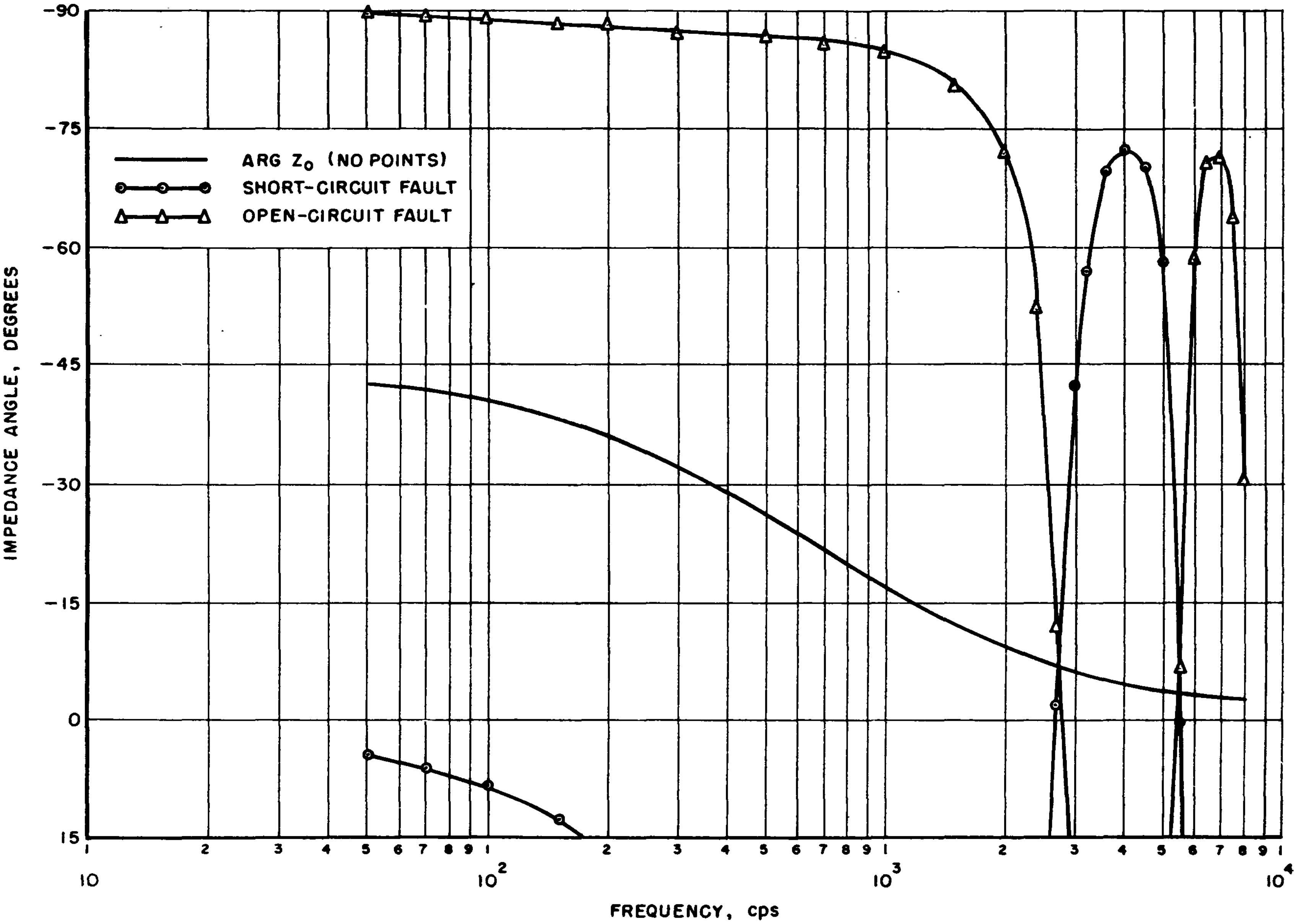

FIG. 5 PREDICTED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 2 MILE FAULT

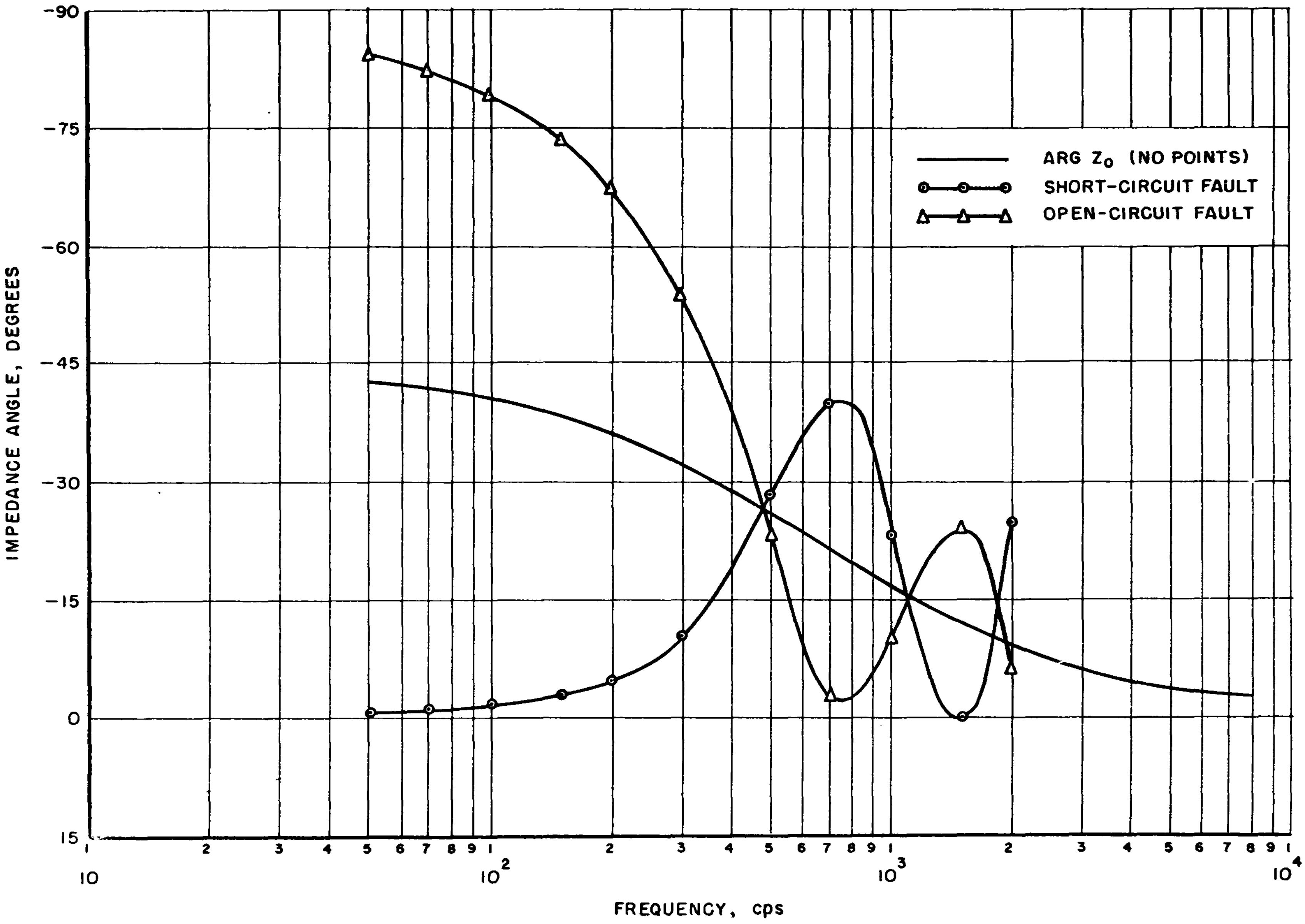

FIG. 6 PREDICTED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 10 MILE FAULT

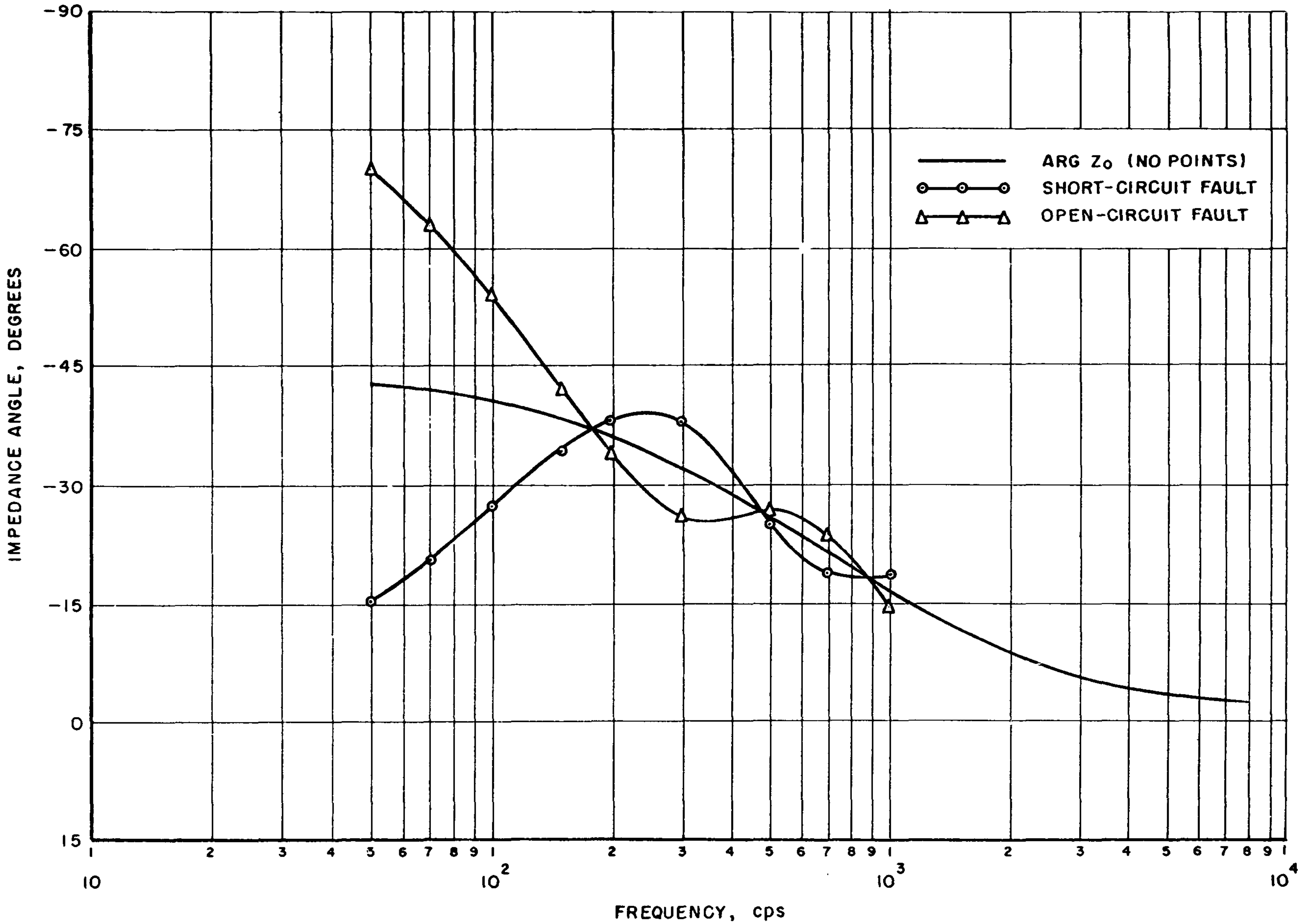

FIG. 7 PREDICTED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 20 MILE FAULT

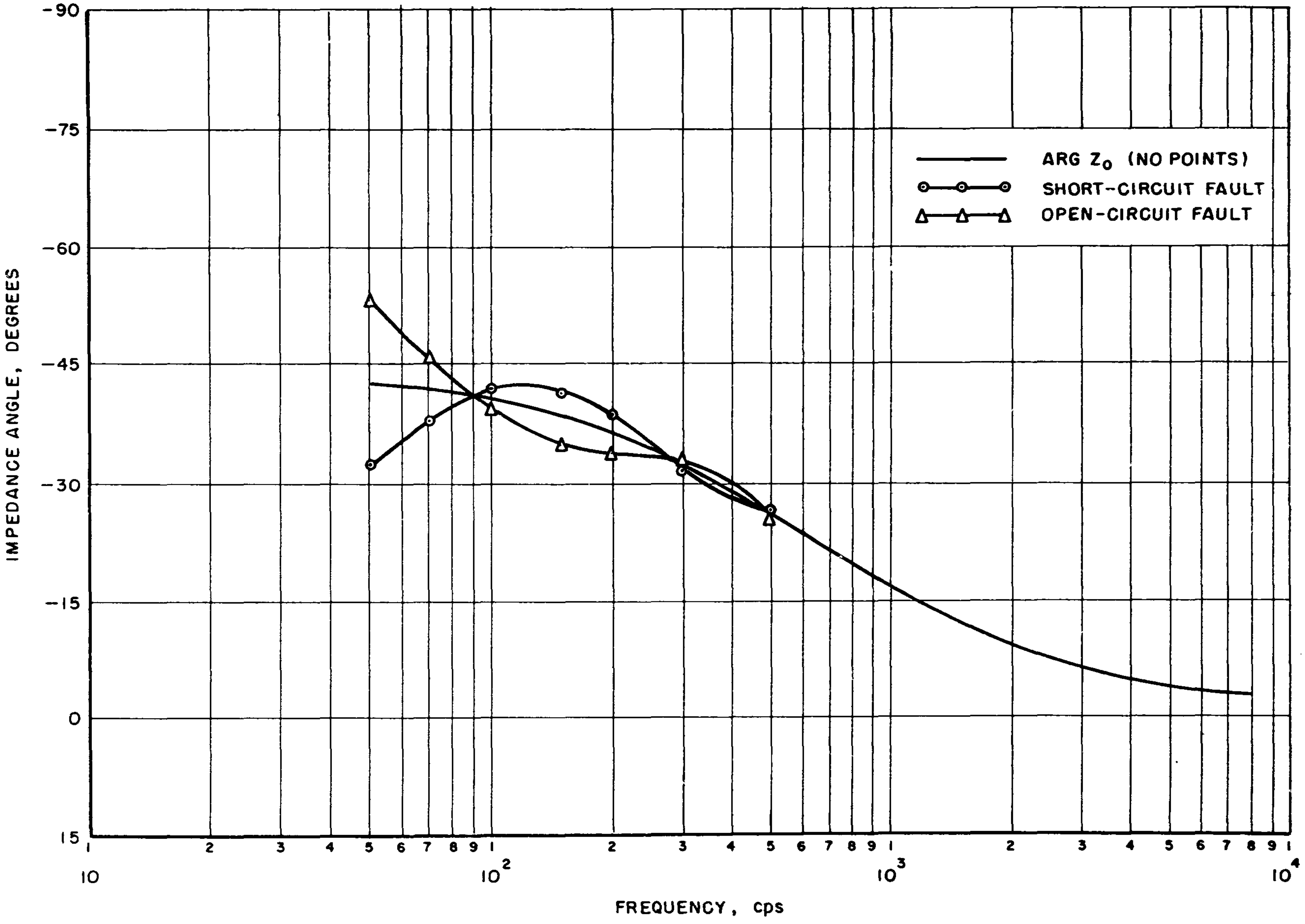

FIG. 8 PREDICTED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 30 MILE FAULT

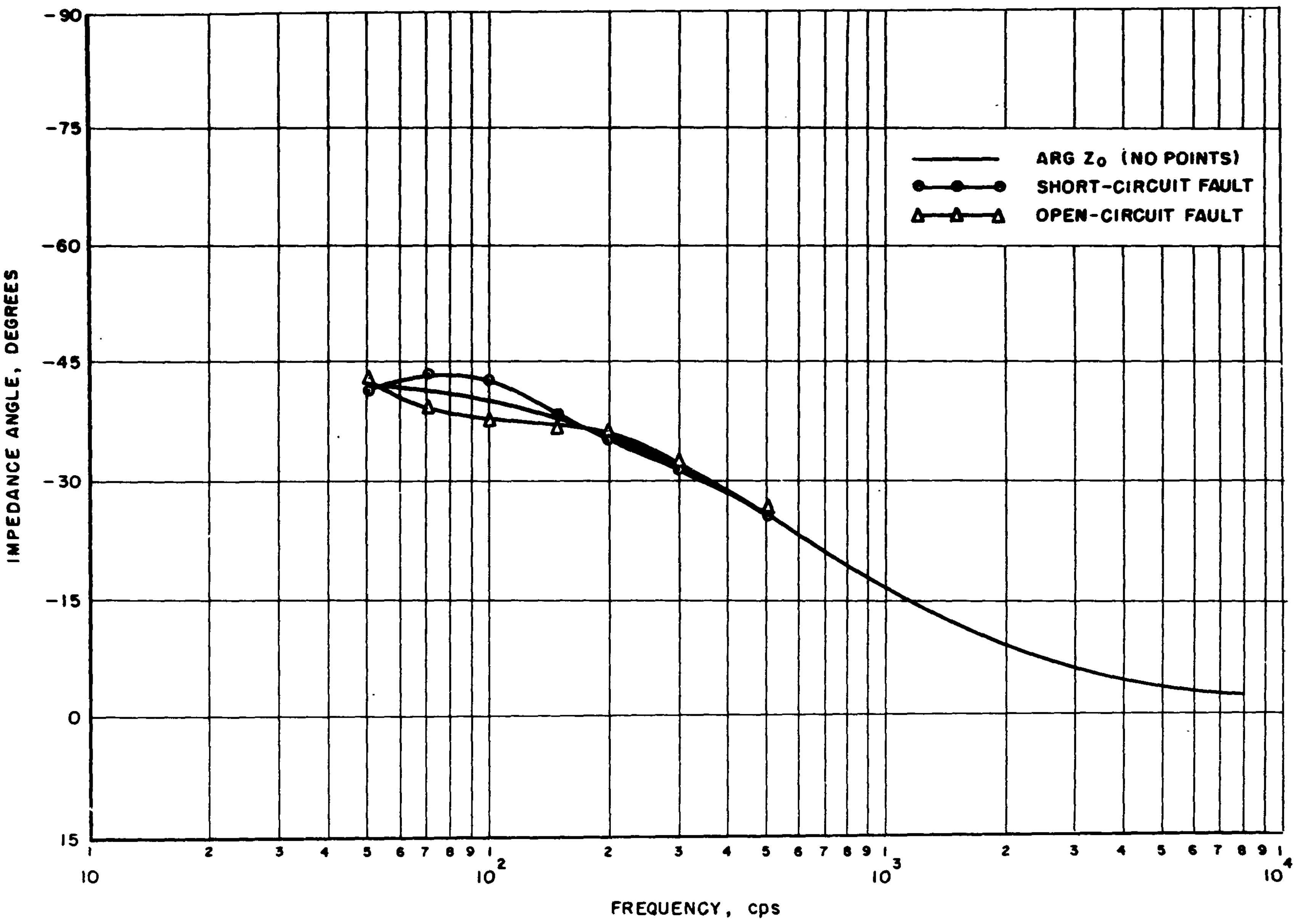

FIG. 9 PREDICTED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 40 MILE FAULT

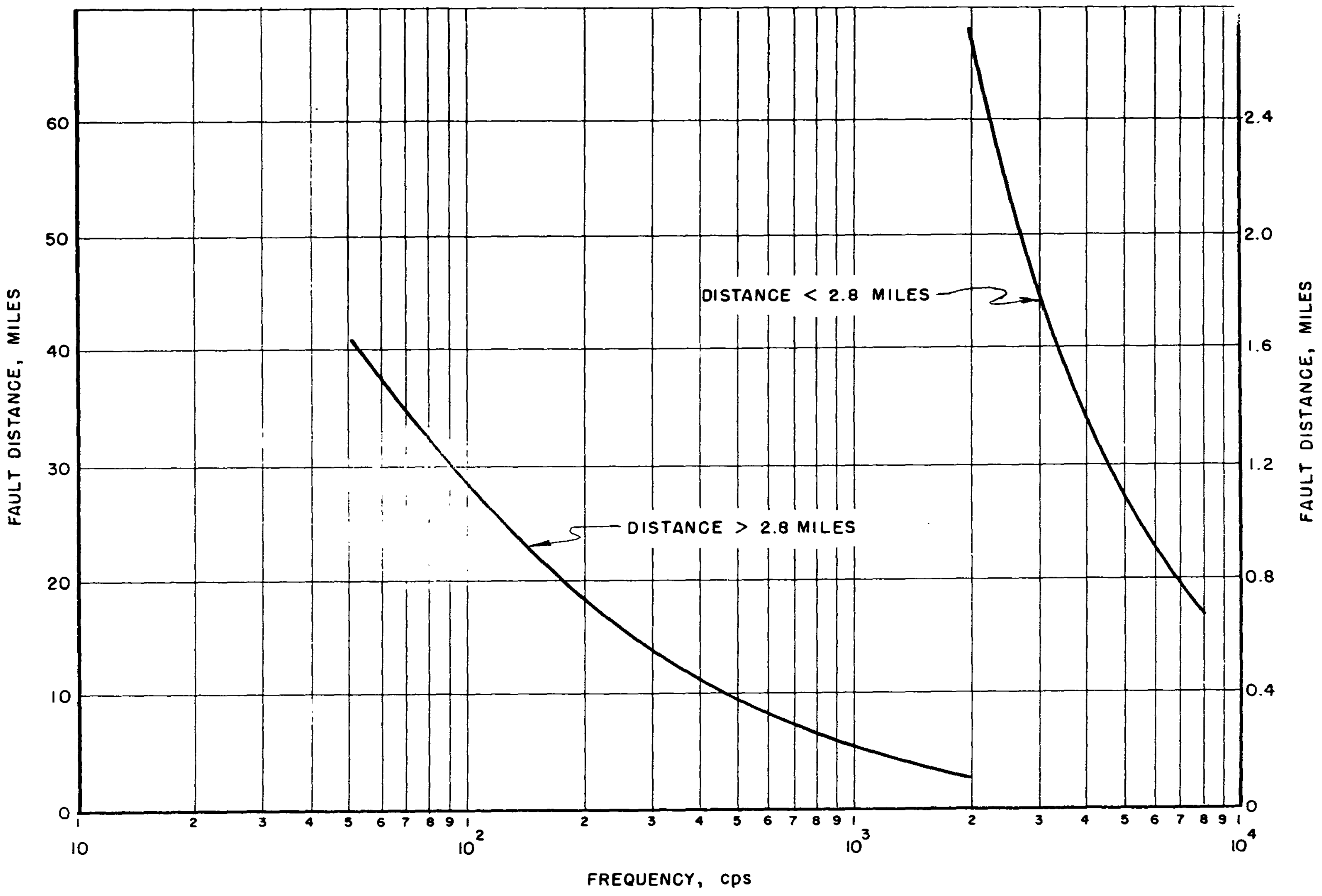

FIG. 10 FAULT DISTANCE VERSUS PHASE COINCIDENCE FREQUENCY

uniform line.

3. Results of Measurements on Loaded Spiral-Four

Measurements were made of input impedance angle on actual lines at USASRDL. At each fault distance, one cable pair was opened and the other short-circuited. A maximum of 15 miles of cable was available.

The data for faults at 2, 10, and 15 miles is plotted in Fig. 11 thru 13, with the arg Z_0 curve superimposed on each graph.

Fig. 11, showing data for a fault at 2 miles, will be examined in detail. The frequencies at which the fault curves intersect the arg Z_0 curve are 2480 cps (short-circuit fault) and 2820 cps (open-circuit fault). These frequencies correspond to 2.21 miles and 1.95 miles, respectively. Theoretically, the fault curves and the arg Z_0 curve should have intersected at a single point. The fact that this does not occur may be due to reading errors and to inaccurate knowledge of the Z_0 of the line. If the latter is the primary reason for error, the frequency at which the two fault curves intersect should correspond closely to the fault distance. In Fig. 11, this frequency is 2600 cps, corresponding to a fault distance of 2.11 miles. Greater discrepancies appear in Figs. 12 and 13.

Table 2 tabulates the results for Figs. 11 thru 13. Intersection frequency no. 1 is that at which the short-circuit fault curve intersects the arg Z_0 curve, no. 2 is that at which the open-circuit fault curve intersects the arg Z_0 curve, and no. 3 is that at which the open-circuit and short-circuit fault curves intersect.

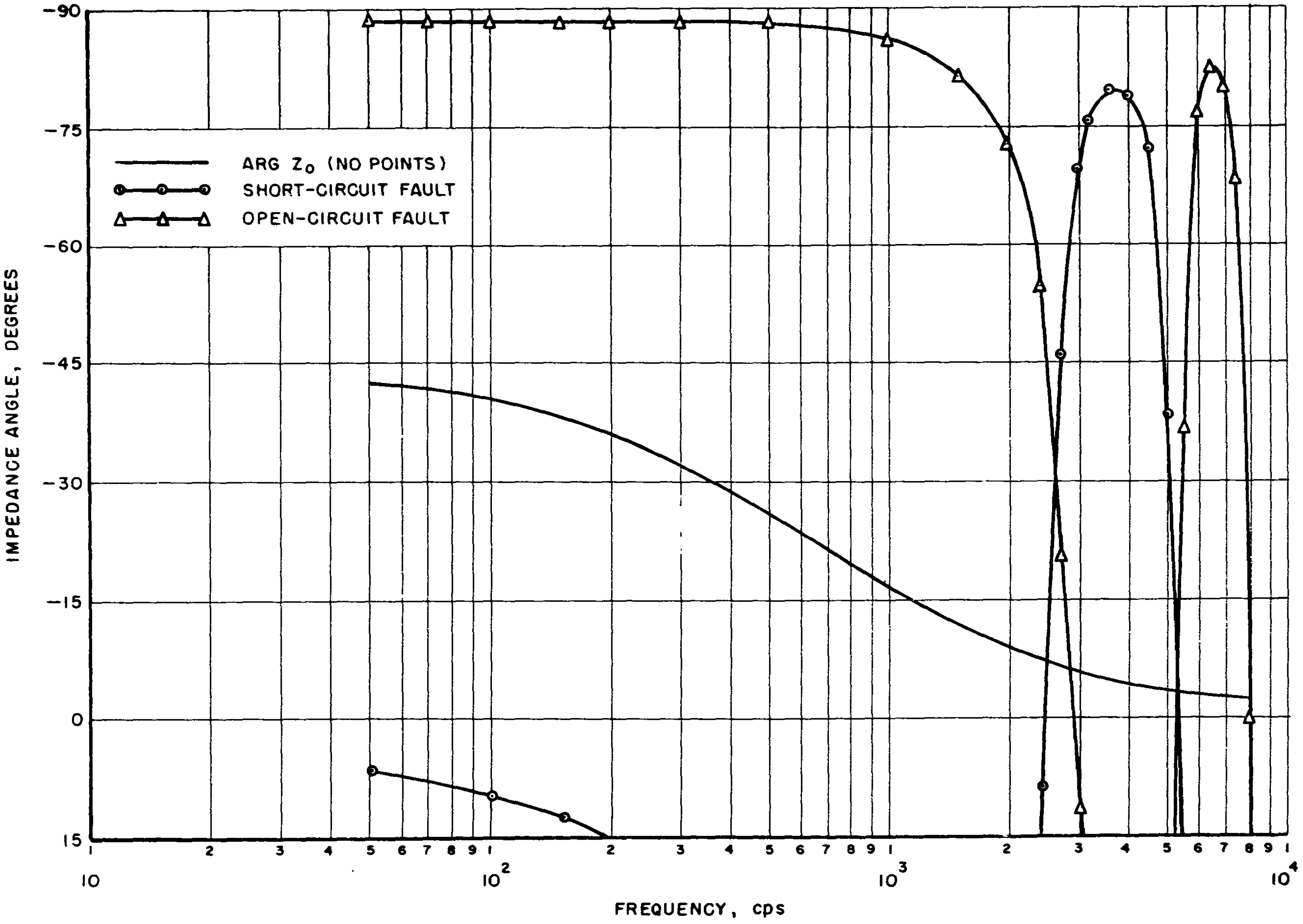

FIG. 11 MEASURED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 2 MILE FAULT

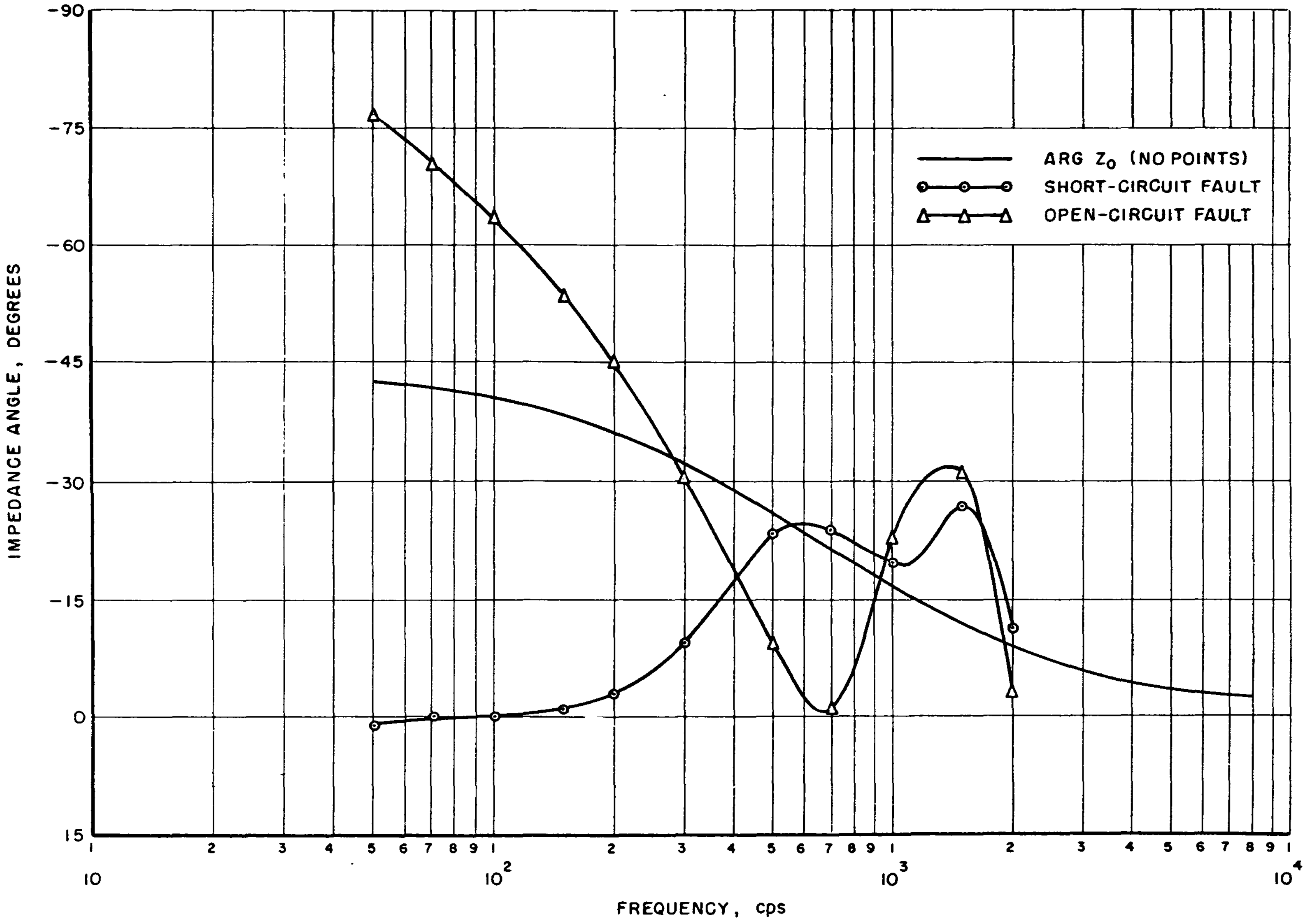

FIG. 12 MEASURED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 10 MILE FAULT

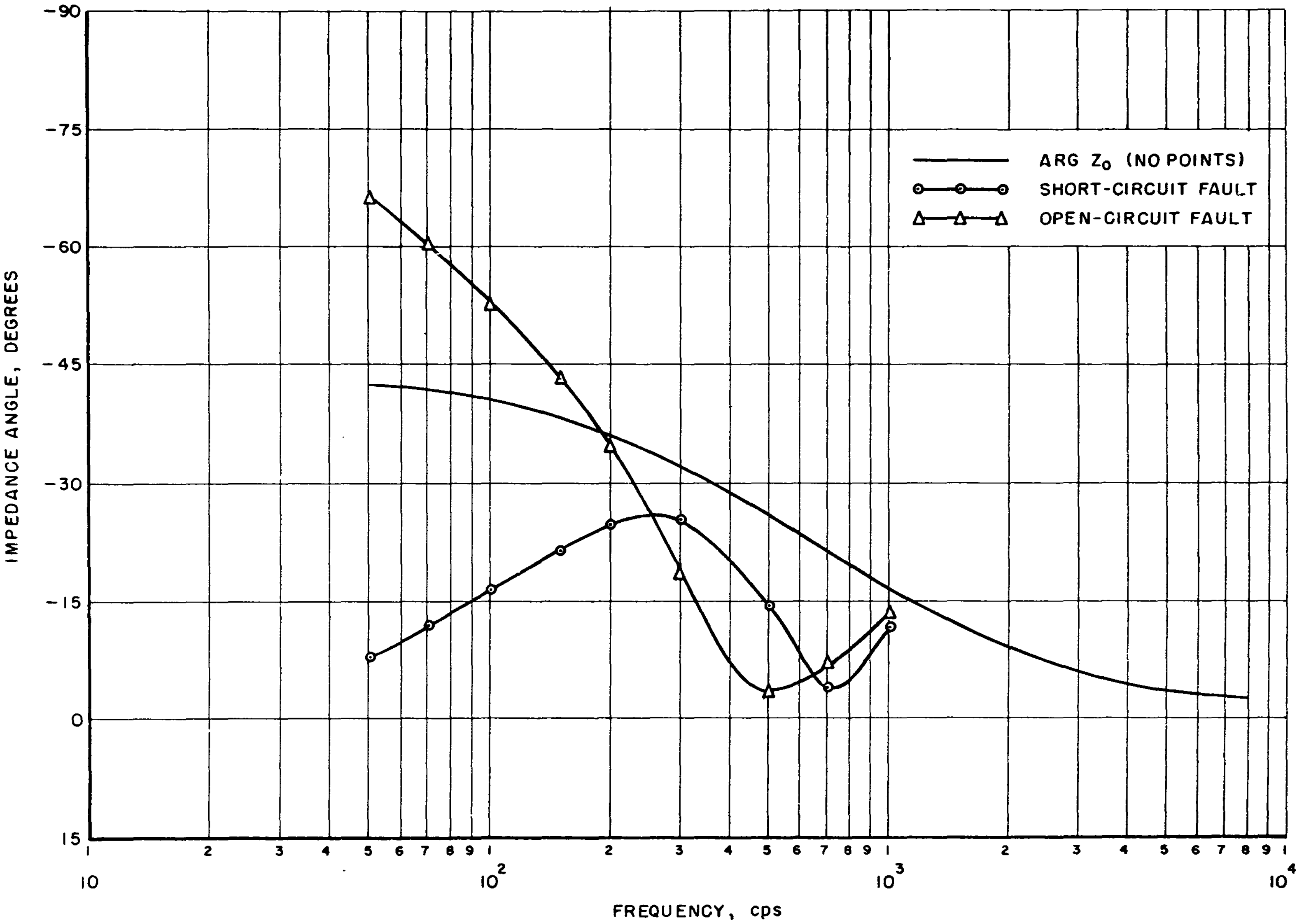

FIG. 13 MEASURED INPUT IMPEDANCE ANGLE OF LOADED SP-4 CABLE — 15 MILE FAULT

Table 2

TABULATION OF FAULT DISTANCE MEASUREMENTS
LOADED SP-4 CABLE

Figure number	11	12	13
Intersection freq. no. 1 (cps)	2480	565	*
Corresponding distance (miles)	2.21	8.7	*
Intersection freq. no. 2 (cps)	2820	283	187
Corresponding distance (miles)	1.95	14.7	19.5
Intersection freq. no. 3 (cps)	2600	408	253
Corresponding distance (miles)	2.11	11.0	15.9
Actual fault distance (miles)	2	10	15

4. Conclusions

The results in Table 2 are obviously not sufficiently accurate for practical fault location purposes. Examination of Figs. 12 and 13 shows that no reasonable curve of arg Z_0 vs. frequency could even closely fit the measured data. The reasons for the discrepancies are not known at this time, but they could include capacity unbalance to the sheath, non-uniformity of cable sections and loading coils, variation of parameters with temperature and aging, and instrumentation errors.

It is believed that a possible basis for a fault location method does exist here. The phase curves do have the general appearance that was

---

* No intersection occurred on the first cycle of variation.

predicted, and the manner in which they change as the fault distance increases is as expected. An empirical method, based on measurements on actual cables, could have adequate accuracy for locating faults. It would require some uniformity between cables, or else measurements would have to be made on individual installations when they were known to be in good working order. We will continue to examine in detail the characteristics of the line, in order to determine the exact reasons for the measurement discrepancies and to evolve an appropriate empirical fault location technique.

C. INTERMODULATION DISTORTION REDUCTION IN SINGLE-SIDEBAND EQUIPMENT

1. Introduction

The Eleventh Quarterly Report outlined a technique of generating quadrature audio signals by means of Hall-effect product modulators and suitable filters. This technique is presently being breadboarded, and preliminary data indicate promising results.

2. Theoretical Considerations

The operation of the quadrature signal generator in theoretical terms is as follows:

Let $$e_i(t) = \sum_{n=1}^{N} E_n \cos(\omega_n t + \phi_n)$$

where E_n is the peak amplitude of the nth term, ω_n is the angular frequency associated with E_n, and ϕ_n is an arbitrary phase angle associated with E_n. Further, let ω_c represent the angular frequency of a carrier of fixed frequency and having unit amplitude.

A product then is made of these terms which has the following form:

$$e_1 = e_i(t) \cos \omega_c t \qquad (1)$$

or

$$e_1 = 1/2 \sum_{n=1}^{N} E_n \cos \left[(\omega_n - \omega_c) t + \phi_n \right] + E_n \cos \left[(\omega_n + \omega_c) t + \phi_n \right] . \qquad (2)$$

Physically, this constitutes a spectrum centered about ω_c and having components differing from ω_c by $\pm \omega_n$. If ω_c is chosen to be greater than the highest ω_n it is the possible to select (by filtering) either the $\omega_c + \omega_n$ or the $\omega_c - \omega_n$ sideband. Assuming that the difference terms are eliminated we are left with:

$$e_2 = 1/2 \sum_{n=1}^{N} E_n \cos \left[(\omega_c + \omega_n) t + \phi_n \right] . \qquad (3)$$

(It is possible that an additional phase term should be included to account for the phase characteristics of the filter, however, the mathematics then becomes obscured in detail.)

The e_2 components are then passed to two additional product forming networks. One network is excited by $\cos(\omega_c t + \phi_c)$ and the other by $\sin(\omega_c t + \phi_c)$. ϕ_c is an arbitrary shift associated with the network which provides these signals and for simplicity can be set equal to zero.

Therefore two signals are generated having the following form,

$$e_3 = e_2 \cos \omega_c t \qquad (4)$$

and

$$e_4 = e_2 \sin \omega_c t \qquad (5)$$

Substituting and expanding gives

$$e_3 = \frac{1}{4}\sum_{n=1}^{N} E_n \cos(\omega_n t + \phi_n) + E_n \cos\left[(2\omega_c + \omega_n)t + \phi_n\right] \qquad (4a)$$

$$e_4 = \frac{1}{4}\sum_{n=1}^{N} E_n \sin(\omega_n t + \phi_n) + E_n \sin\left[(2\omega_c + \omega_n)t + \phi_n\right] \qquad (5a)$$

Equations (4a) and (5a) contain the desired components,

$$e_5 = \frac{1}{4}\sum_{n=1}^{N} E_n \cos(\omega_n t + \phi_n)$$

and

$$e_6 = \frac{1}{4}\sum_{n=1}^{N} E_n \sin(\omega_n t + \phi_n)$$

and in addition unwanted components near $2\omega_c$. These unwanted components are again removed by filtering. In these two filters it is apparent that the phase characteristics can not be neglected. In order for each component of E_n from one filter to be in quadrature with the corresponding E_n component from the other filter, the phase characteristics must be exactly the same. As long as the separation between carrier frequency and highest modulation frequency is reasonable this should present no particular problem.

3. Practical Considerations

In order to implement the technique outlined above it is necessary to consider the available devices for the product forming devices and the filters. As mentioned in the previous quarterly report, several balanced modulator types were investigated, and from these the Hall-effect appeared to most nearly provide the necessary characteristics. The form used is a standard ferrite cup-core with the center post ground down to provide room for the Hall effect crystal. There are three terminals soldered to the crystal, two on the opposite sides, and one on one end of the crystal. The modulation voltage is fed in parallel to the two side connections of the crystal and the third connection is used as the return. The magnetic field is provided by a coil wound on a bobbin that fits around the cup-core center post. The output signal is taken from the two opposite crystal terminals and fed to an impedance matching transformer. The output of the transformer is then passed to the first filter network. Initially, a high pass filter was used since it was assumed that linear (modulation) terms would be present along with the lower sideband components. This, however, has not proved to be the case, and so the lower sideband is being used. The phase and attenuation characteristics of this filter are shown in Fig. 14. The audio filters following the second set of modulators will have similar characteristics but the cut-off frequency will be slightly lower. It can be seen from Figure 14 that the problem of duplicating phase characteristics should not be insurmountable since the phase change as a function of frequency is quite smooth.

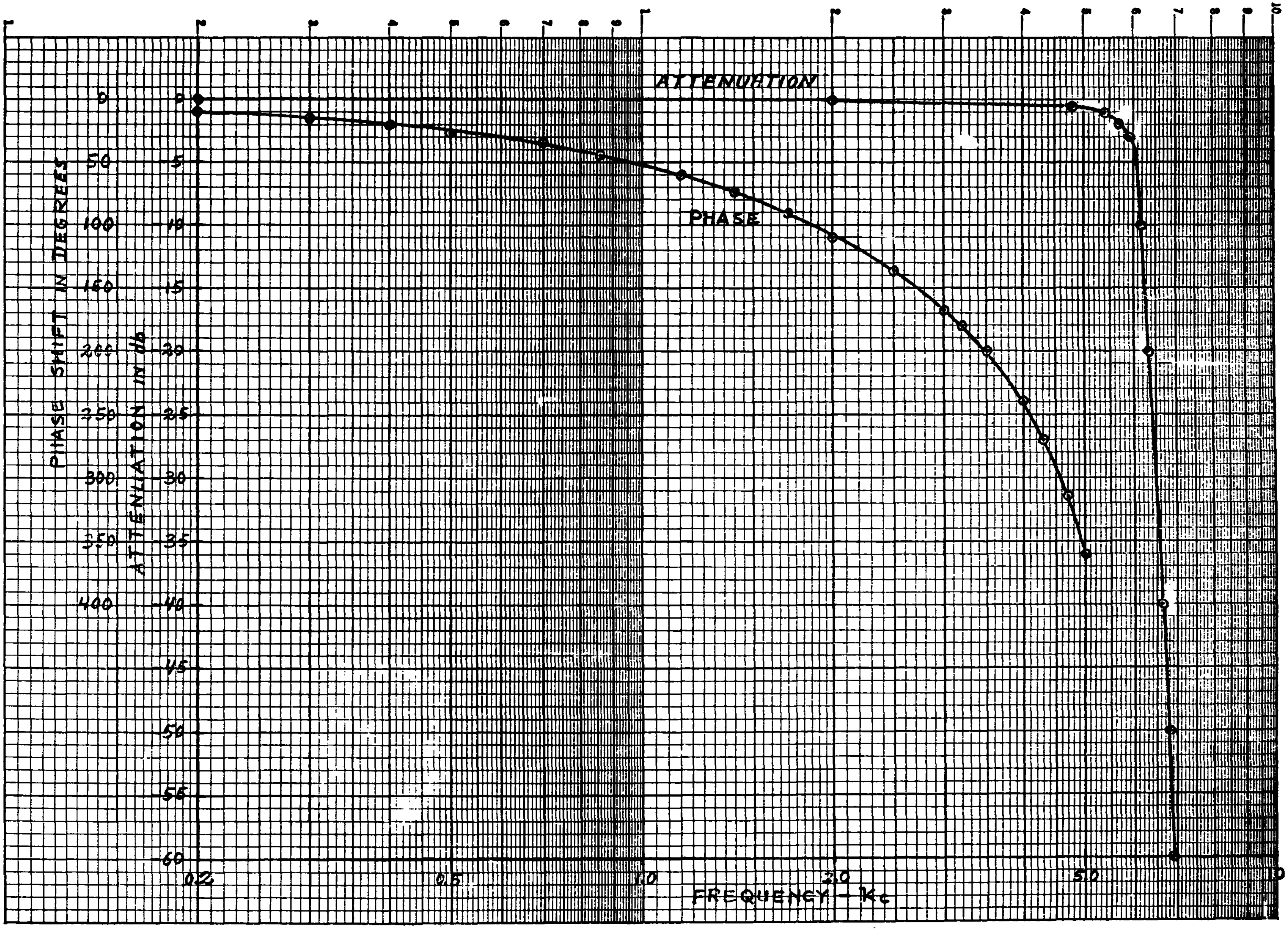

FIG. 14 LOW PASS FILTER ATTENUATION AND PHASE CHARACTERISTICS

At the present time the circuitry up to the second set of modulators is complete, the second modulators have been fabricated, and the audio low pass filters are under construction. It is anticipated that preliminary data on the characteristics of this device should be available during the next quarter.

V. CONCLUSIONS

The printed-circuit current probe has a sensitivity of seven volts for one milliampere in the printed circuit conductor, which should prove adequate for most needs. The equivalent noise input is one microampere. Nearby fields are greatly reduced in the final model over earlier versions, due to better design and shielding. In-circuit testing of many paralleled components with the unit is quite feasible, although the use of the instrument will require somewhat more skill than a volt-ohmmeter.

The instruction manual outlining the use of the visual circuit tester was written and submitted to the Signal Corps.

The general appearance of the experimental data obtained for the phase coincidence method for loaded spiral-four cable was similar to the computed data. Quantitatively, inconsistancies appear which can only be resolved by further line measurements. Manufacturers line specifications may be a limiting factor, along with aging and loading coil parameter changes.

The 90° phase shifter employing Hall-element product modulators should provide accurate quadrature modulation signals for use in a phasing type SSB generator. In the technique employed the problem of providing identical filter networks and Hall-elements does not appear insurmountable.

VI. PROGRAM FOR THE NEXT INTERVAL

For various reasons, the level of effort on this contract has been reduced during the past quarter, and it is expected that this level will continue during the next quarter.

The current probe and its associated amplifier will be further evaluated during the next period, with the aim of determining the ancillary equipment necessary. The additional equipment would provide amplitude and phase readout capability.

In the area of reduction of IM distortion in SSB modulators, work will continue on the breadboard circuitry which will provide the signals needed for a phasing type SSB modulator.

Upon receipt of the TS-1009 Fault Locator from USASRDL, work will be initiated on the location of faults on open-wire lines. In addition, a more complete set of impedance measurements will be made on the loaded spiral-four cable. From this, the analytical and measured discrepancies can be resolved, and possibly other measurement techniques developed if appropriate.

VII. IDENTIFICATION OF KEY PERSONNEL

The following is a list of the key personnel working on this program and the approximate hours worked during the twelfth quarter:

		Man-Hours
A. T. Ashby	Research Engineer	251
T. A. Jackson	Associate Engineer	408
J. N. Van Scoyoc	Engineering Advisor	256

As noted previously, the level of effort has been reduced during this quarter for various reasons.

Respectfully submitted,

ARMOUR RESEARCH FOUNDATION
of Illinois Institute of Technology

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A. T. Ashby, Research Engineer

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by A. T. Ashby. Twelfth Quarterly Progress Report for Nov. 1961 -
Jan. 1962, 34 P. (Report No. ARF 5112-36)
Signal Corps Contract DA 36-039 SC-78269
Unclassified Report

A current probe for use with printed circuitry which senses branch currents in parallel circuits and operates at a fixed frequency of 159 Kc is described. The sensitivity is such that one milliampere in a conductor produces seven volts output from the associated amplifier. The noise equivalent input is about one microampere. Preliminary testing has proven the device successful.

The location of faults on loaded spiral-four cable by means of a phase measurement technique is evaluated experimentally. Agreement with predicted results is not sufficient to yield accurate results, however, there is evidence that the same procedure in conjunction with empirical data from lines in good operating condition would produce results with adequate accuracy.

The theory and operation of a wideband 90° phase shifter is discussed. This device uses Hall-element product modulators and should be capable of minimizing intermodulation distortion in SSB equipment.

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1. Test Equipment
2. Instrumentation
3. Impedance
4. Field Wire Communication Systems
5. Modulators
6. Sidebands
7. Phase Shifters

I. Title: Fault Location
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