NBS TECHNICAL NOTE 658

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

Development of Electric and Magnetic Near-Field Probes

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Development of Electric and Magnetic Near-Field Probes

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Frank M. Greene

ABSTRACT

This publication describes the development and design of small electric and magnetic near-field probes for measuring hazard-level fields up to 20,000 V/m and 100 A/m, respectively. They were originally designed to be used over the frequency range from 10 to 30 MHz, and consist of short dipole antennas and small, single-turn, balanced, loop antennas to measure the electric- and magnetic-field components, respectively. The probes are intended for use by various researchers in their electromagnetic, radiation-exposure programs for determining the effects of hazard-level, non-ionizing, EM fields on living tissue, electro-explosive devices, and volatile fuels.

In order to later extend the use of the probes to frequencies above 30 MHz, a detailed analysis was made of several types of measurement errors likely to be encountered. The principal errors result from a variation with frequency in: the effective length and impedance of the dipoles; and the electric-dipole response and partial resonance of the loops. Corresponding corrections are given for each type of error as a function of the operating frequency from 10 to 1000 MHz, and as a function of the physical and electrical sizes of the probes.

As a result of the analysis, the dipoles can now be used for measurements at frequencies up to 750 MHz, and the loops to 75 MHz with an estimated uncertainty of 0.5 dB. Applying recommended corrections will provide a substantial further increase in the usable frequency range.

Keywords: Electric near-field probe; electromagnetic-field hazard; field-strength measurements; magnetic near-field probe; near-field measurements; r-f hazard measurements; semiconducting transmission line.

1. INTRODUCTION

This publication describes the development of portable, electric and magnetic near-field probes by the National Bureau of Standards for other U.S. Government Agencies over the past five years.[1] These probes are designed for use in measuring hazard-level, electric and magnetic near fields up to 20,000 volts per meter and 100 amperes per meter, respectively, over the frequency range from 10 to 30 MHz and higher as will be discussed in detail. The probes are intended for use by various researchers in their electromagnetic radiation-exposure programs for determining the effects of hazard-level, non-ionizing, EM fields on living tissue, electro-explosive devices, and volatile fuels.

The probes consist of short, dipole antennas (which measure the component of the electric field parallel to the dipole axis), and small, single-turn, balanced loop antennas (which measure the component of the magnetic field parallel to the loop axis). Each probe has a silicon-junction, semiconductor diode connected internally across its gap to rectify the induced r-f voltage. The d-c output of each probe is transmitted over a special non-metallic, high-resistance, transmission line (developed at NBS) to a remote d-c electrometer voltmeter, located up to 30 feet away. Such a resistance line is essentially "transparent" to the field being measured and substantially reduces the measurement error that would otherwise result from distortion of the field and r-f pickup if a metallic line were used.

It is known that both the amplitude and phase of an electromagnetic field have a much more rapid spatial variation

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[1]This work was sponsored through interagency agreements with the USAF School of Aerospace Medicine, Brooks Air Force Base, Texas 78235, and the National Institute for Occupational Safety and Health of HEW, Cincinnati, Ohio 45202.

in the near zone (close to the source) than they do in the far-zone (at a greater distance from the source). In order to minimize the averaging effect introduced by any finite-sized probe, it is therefore necessary to make its dimensions electrically small when it is used for near-field measurements. A probe cannot be used to study a field with a structure much finer than its own dimensions. A small probe also introduces less perturbation of the field than a larger one, since its scattering cross-section is smaller.

In order to extend the use of the probes to frequencies above 30 MHz, a detailed analysis was made of several types of errors likely to be encountered. The performance of both the electric dipoles and the loop antennas was investigated as a function of their operating frequency (from 10 to 1000 MHz) and as a function of their physical and electrical sizes. Measurement errors, resulting from several causes, have been analyzed as a function of frequency, and corresponding corrections are given in each case to be applied to the measurements.

The errors to be treated result from the following causes:

(a) the variation of the effective-length and internal impedance of a short, electric dipole as a function of its electrical length;

(b) the variation with frequency of the electric-dipole response of a small loop antenna to the electric field present;

(c) the increase in the response of a loop antenna due to partial resonance;

(d) the harmonic frequencies present in the fields being measured;

(e) the use of metallic lines associated with field-strength meters.

The practical mks International System of Units (SI) is used in the principal formulas in this paper.

2. THE ELECTRIC NEAR-FIELD PROBES

2.1 General. The electric-field probes consist of two electrically-short, dipole antennas 10-cm and 5-cm in overall length and are designed to be used for measuring hazard-level, electric fields in the ranges from 200 to 2000 and from 2000 to 20,000 volts per meter, respectively, over the frequency range from approximately 10 to 30 MHz and higher. In addition, a third dipole antenna, having a length of 2.5-cm, was included in the theoretical analysis. This size of dipole appears to be necessary for electric-field measurements at frequencies above about 750 MHz. It also appears that printed-circuit techniques will probably be required in the design of this dipole and its associated r-f filter, in order to minimize extraneous responses.

The general construction of the 5-cm and 10-cm dipole antennas is shown in figure 1 and the dimensions are given in Table I.

Table I

Dimensions of the 5-cm and 10-cm Dipole Antennas

Lengths			Gap Width	Diameters		
L-1	L-2	L-3	G	D-1	D-2	D-3
(cm)	(inches)		(inches)	(inches)		
5.00	0.5625	0.2500	0.004	0.1250	0.4375	0.3750
10.00	0.7500	0.3750	0.031	0.1875	0.5000	0.3750

Notes: (a) see figure 1 for definitions of symbols;
(b) Teflon dielectric is used in the gap having the same thickness as indicated above for the gap width, G.

The dipoles are fabricated in two identical halves as shown in figure 1. The dipole is formed of two brass rods, each of which fits into one half of a split, brass center post

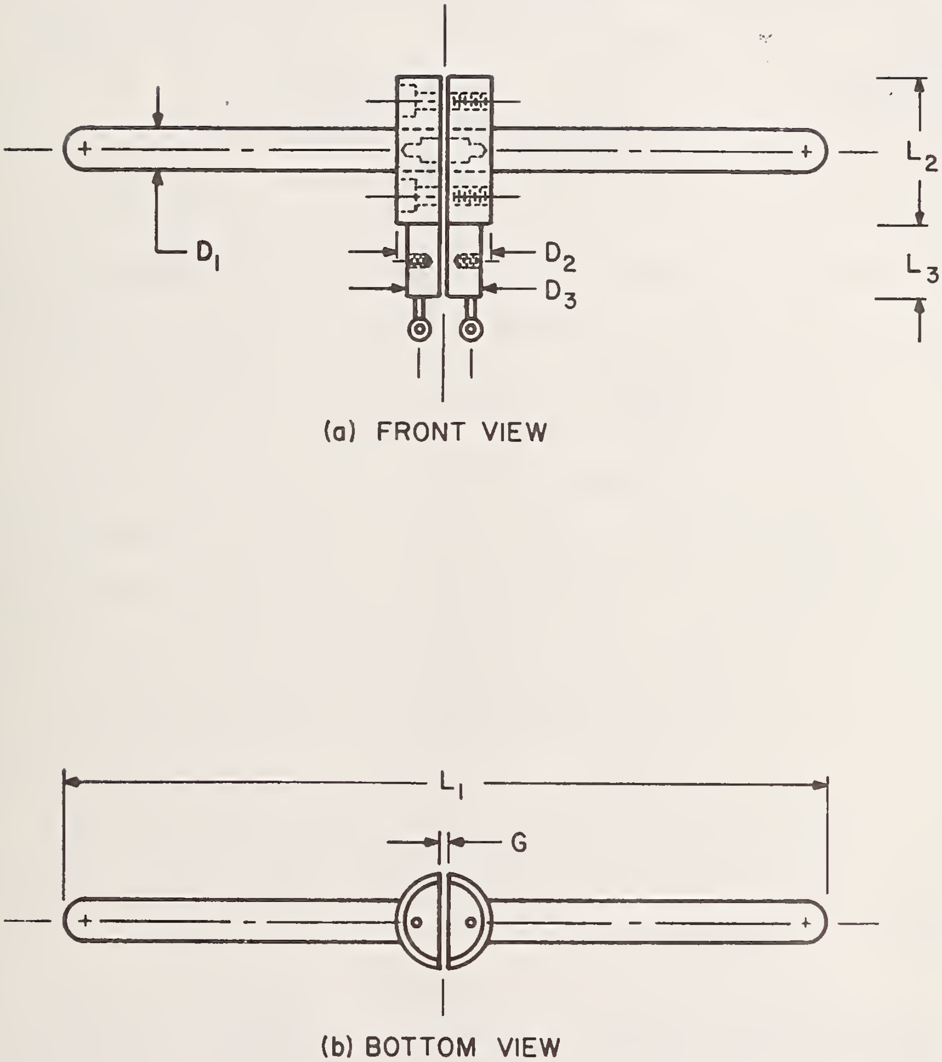

Figure 1. Construction details of the 5-cm and 10-cm dipole antennas. See Table I for dimensions.

and is secured by silver soldering. The inner surfaces of the gap are then milled to the required dimensions and lapped flat and parallel to form a built-in gap capacitor. Teflon dielectric is used in the gap, having the same thickness as indicated in Table I for the gap width, G. A small semiconductor diode approximately 0.076 inch in diameter by 0.170 inch long is mounted inside of a polystyrene sleeve, which mounts internally at the center of the dipole. The diode is connected across the gap by means of spring contacts formed from the diode leads. The two halves of the dipole are fastened together with two 4-40 fillister-head (recessed), nylon machine screws through the center post as indicated. Solder lugs are provided, as shown, for connecting a filter network.

Each dipole is mounted at the end of a 36-inch-long, tubular, fiber-glass handle approximately 1/2-inch outside diameter having a 1/16-inch wall thickness. The 3/8-inch-diameter end of the dipole center post slides into one end of the handle and is secured with two 2-56 flat-head, brass machine screws. Two twisted pieces of the conducting-teflon monofilament, described in Section 7, connect the balanced, dipole-filter output at the probe end of the tubular handle to a military-type PJ-291 miniature, twin plug secured internally at the other end.

2.2 Electrical Characteristics. A schematic diagram of the portable dipole probe is shown in figure 2. Each probe originally had a type 1N4148 silicon-junction, semiconductor diode connected internally across its gap, as shown, to rectify the induced r-f voltage. Type 1N5711 Schottky-barrier diodes were used in the later models because of their improved response at frequencies to 1000 MHz. A balanced r-c, r-f

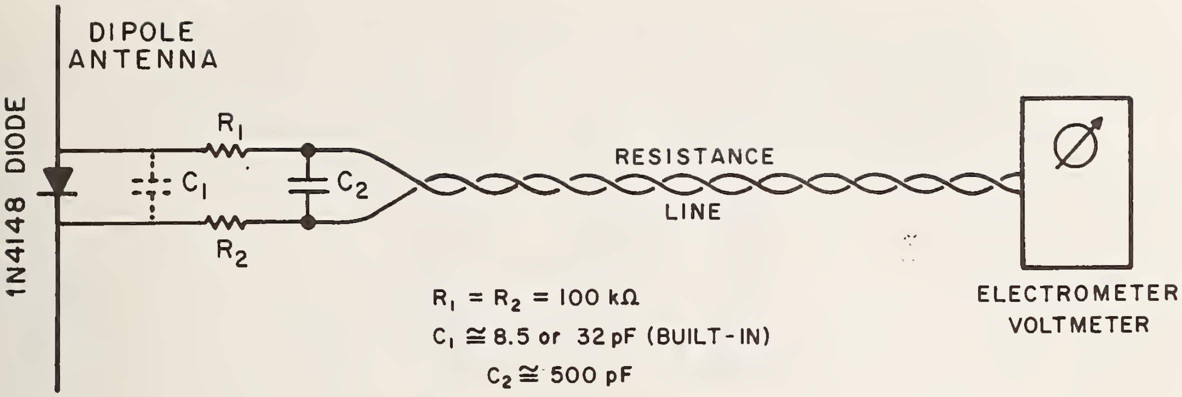

Figure 2. Schematic diagram of the portable dipole probe.

filter, consisting of R-1, R-2, C-1, and C-2, is built into the tubular handle next to the probe to prevent any r-f voltage, picked up on the resistance line, from being rectified by the diode and causing an error in indication. Capacitor C-1 is built into the gap at the center of the dipole, as mentioned in Section 2.1. The resulting value of capacitance is determined by the width of the gap and the type of dielectric material used. The design value selected for C-1 largely determines the sensitivity of the probe as will be discussed further in Section 2.3.

A 30-foot length of the non-metallic, high-resistance line, described in Section 7, is used to connect the probe with a remote, high-resistance, d-c electrometer voltmeter. This meter has an output reading roughly proportional to electric-field strength over a voltage range from approximately 1 to 10 volts. The line has a resistance of approximately 40,000 ohms per foot, giving a total loop resistance of 1.2 megohms. The electrometer is used with an input-resistance setting of 10^8 or 10^9 ohms, so that the error due to the i-r drop on the line is less than 1 percent. The d-c voltage level used is sufficient to override any noise present on the line, as discussed in Section 7.

The detector time constant is such that the probes measure essentially the instantaneous peak value of a steady-state cw field over the measuring range involved here. This has been multiplied by 0.707 to convert the indication to rms (root-mean-square) field strength. For the case of pulsed fields having a duty cycle of 0.05 or greater,[2] the difference between the cw and the pulse indication is small. The probes indicate the rms value of the field during the pulse, i.e., 0.707 times the instantaneous peak value of the pulse.

The equivalent circuit of a short, balanced dipole antenna is shown in figure 3. The self-impedance of a short

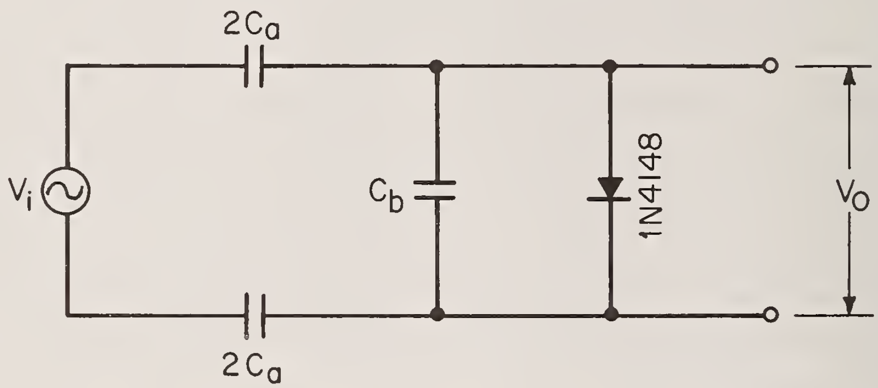

Figure 3. Equivalent circuit of a short dipole antenna.

dipole is essentially capacitive as represented by C_a. The gap capacitance across the center terminals is represented by C_b, which includes the diode capacitance. V_i is the r-f voltage induced in the dipole by the electric field, E. V_o is the output r-f voltage appearing across the gap and diode at the center of the dipole as shown. The radiation and ohmic resistances are considered negligible. The approximate values of C_a and C_b for the 5-cm and 10-cm dipoles are shown in Table II.

[2]The duty cycle of a periodically recurring pulse is the ratio of the pulse width to the period of the pulse.

Table II

Approximate Values of the 5-cm and 10-cm Dipole Capacitances

Dipole Length (cm)	Capacitance C_a (pF)	C_b (pF)
5.00	0.5	32
10.00	0.7	8.5

2.3 The Response of a Short Dipole Antenna. The voltage, V_i, induced in a short electric dipole by the component of the electric field, E, parallel to the axis of the dipole is given by

$$V_i = E \cdot L_{eff} \tag{1}$$

where L_{eff} is the effective length of the dipole, in meters, and is as defined by equation (1) for the receiving case [1].[3]

The effective length is a function of frequency and increases with the electrical length of the dipole. For a thin dipole having essentially a sinusoidal current distribution, the effective length is given approximately by the following asymptotic expression [2,3], valid for $0 \leq \beta\ell < \pi$:

$$L_{eff} \simeq \frac{L}{2}\left[\frac{\tan\frac{\beta\ell}{2}}{\frac{\beta\ell}{2}}\right], \text{ meters}, \tag{2}$$

where $L = 2\ell$ is the overall physical length of the dipole in meters; $\beta = 2\pi/\lambda$, where λ is the wavelength in meters. For an electrically-short dipole having an overall length, $L < 0.1\lambda$, $L_{eff} \simeq (1/2)L$. For a half-wavelength, self-resonant dipole, $L_{eff} = (2/\pi)L$. Several numerical values derived from equation (2) are shown in Table III.

[3]Figures in brackets indicate literature references at the end of this paper.

Table III

Increase in the Effective Length of a Dipole with Electrical Length

ℓ/λ	$\beta\ell$ (radians)	θ (degrees)	$\left[\frac{\tan\frac{\beta\ell}{2}}{\frac{\beta\ell}{2}}\right]$
.01	.06283	3.60	1.0003
.02	.12566	7.20	1.0013
.05	.31416	18.00	1.0083
.10	.62832	36.00	1.0343
.15	.94248	54.00	1.0812
.20	1.25664	72.00	1.1563
.25	1.57080	90.00	1.2732

The angle θ, in Table III, is the equivalent electrical half-length of the dipole expressed in degrees ($\theta = 360\ \ell/\lambda$). The induced voltage, V_i, given by equation (1), varies with the electrical length of the dipole in accordance with equation (2). In the case of the NBS dipoles, the output voltage, V_o, is also determined by the capacitive voltage-divider ratio comprised of C_a and C_b and is given by

$$V_o = \frac{C_a}{C_a + C_b} V_i. \qquad (3)$$

For lossless capacitors, the divider ratio in equation (3) is independent of frequency, giving V_o the same frequency dependence as V_i, so long as C_a and C_b remain fixed in value. The built-in capacitive voltage divider and the dipole effective length are both made use of, in the NBS dipoles, in fixing their measurement sensitivity. However, the apparent dipole capacitance, C_a', increases with frequency, as will be shown, which will give V_o a further frequency dependence, depending on the value of C_b.

By making use of Schelkunoff's analogy with conventional transmission-line theory [4], it can be readily shown that the dependence of the apparent self-capacitance, C_a', of a thin dipole on its electrical length, will be given approximately by the following asymptotic expression, valid for $0 \leq \beta\ell \leq \pi/2$:

$$C_a' \cong C_a \left[\frac{\tan \beta\ell}{\beta\ell} \right], \tag{4}$$

where C_a is the low-frequency self-capacitance of the dipole. The increase in the quantity C_a'/C_a with electrical length is shown in Table IV.

Table IV

Increase in The Quantity C_a'/C_a of a Thin Dipole with Electrical Length

ℓ/λ	$\beta\ell$ (radians)	θ (degrees)	C_a'/C_a
.01	.06283	3.60	1.0013
.02	.12566	7.20	1.0053
.05	.31416	18.00	1.0343
.10	.62832	36.00	1.1563
.15	.94248	54.00	1.4604
.20	1.25664	72.00	2.4491
.25	1.57080	90.00	∞

It can be seen from Table IV that as $\ell/\lambda \to 0$, $C_a' \to C_a$. For $\ell/\lambda \leq 0.03$, the increase in C_a' over C_a will not exceed 1 percent. However, when $\ell/\lambda = 0.15$, the value of apparent self-capacitance, C_a', will be 46 percent larger than the low-frequency value. When the dipole is one half-wavelength long ($\ell/\lambda = 0.25$), its apparent self-capacitance is infinite, due to resonance with its self-inductance. The radiation resistance of the order of 72 ohms, in this case, is in series with these reactances.

2.4 Dipole Antenna Calibration and Measurement Errors. The 5-cm and 10-cm NBS dipoles are usually calibrated at a relatively low frequency (f = 10 MHz), such that $C_a' \cong C_a$, and $L_{eff} \cong L/2$. The gap capacity, C_b, is large compared to C_a. When these dipoles are used at higher frequencies, the change in C_a' and L_{eff} with frequency will represent a measurement error. The percent error in the measurement will be approximately equal to the sum of the percentage increases in C_a' and L_{eff} from their 10 MHz calibration values. This can be determined from equations (2) and (4), or Tables III and IV (for the case when $C_b >> C_a$).

The worst-case measurement errors for a short dipole, due to the above causes, are shown plotted vs. the electrical half-length, ℓ/λ, in figure 4, and vs. frequency in figures 5 and 6 for the 2.5-cm, 5-cm, and 10-cm dipoles. As can be seen, the total measurement error will be approximately 0.5 dB (6 percent) when the overall length of the dipole is 1/8 wavelength ($\ell = \lambda/16$). This would correspond to a frequency of 375 MHz for the 10-cm dipole, 750 MHz for the 5-cm dipole, and 1500 MHz for the 2.5-cm dipole. When it is practicable to calibrate the dipoles at each frequency used, these errors can be greatly reduced.

Typical calibration curves, made at a frequency of 10 MHz, for the 10-cm dipole probe are shown in figure 7 at the end of this section. The same curves can be used for the 5-cm dipole, except that the values of electric-field strength obtained from the curves should be multiplied by a factor of 10. The calibrations were made at NBS in terms of a high-level standard electric field, generated in a special parallel-plate strip-line. This was supplemented by the use of the standard (receiving) antenna method of calibration [5].

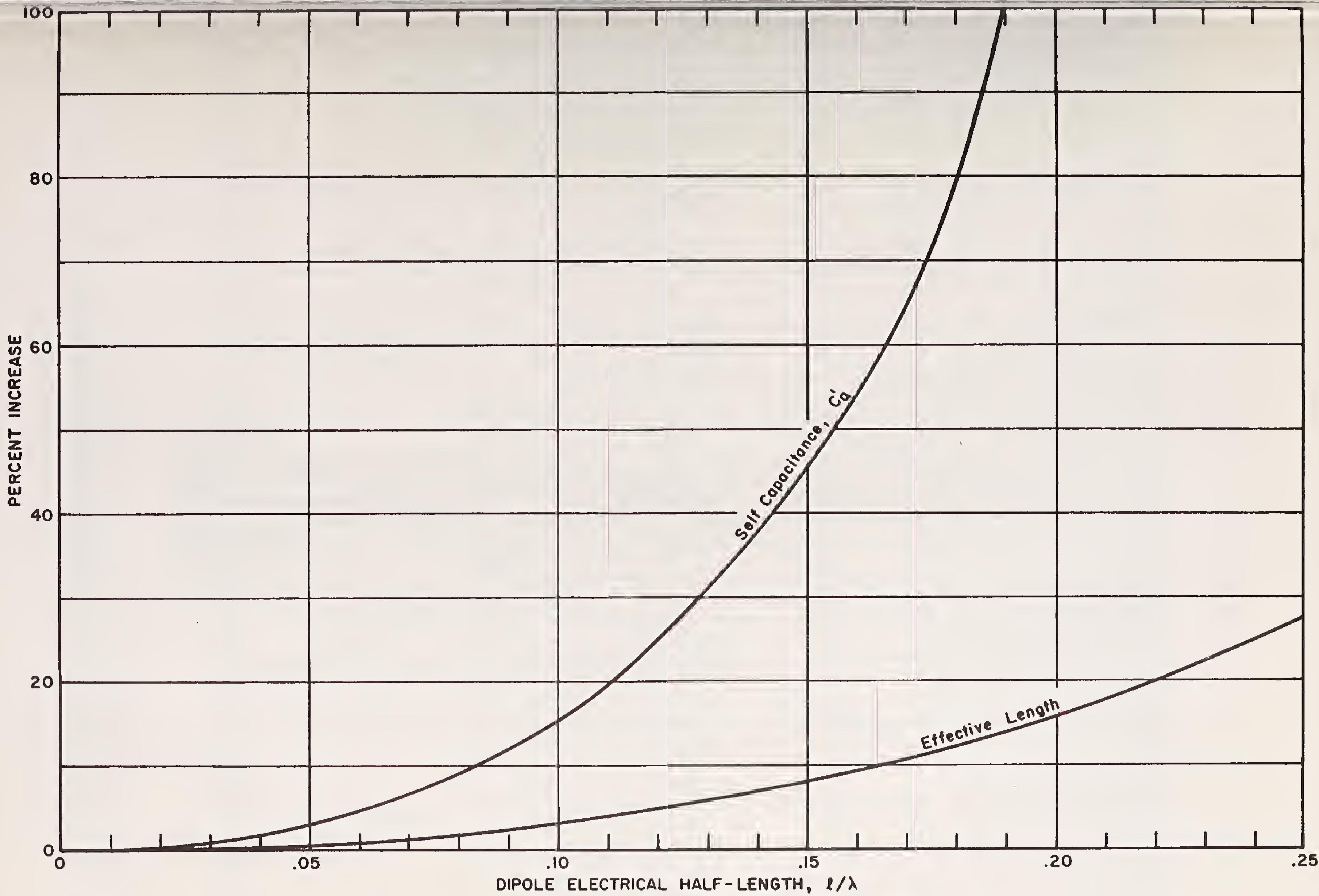

Figure 4. Increase in the effective length and apparent self-capacitance of a short dipole antenna with electrical half-length.

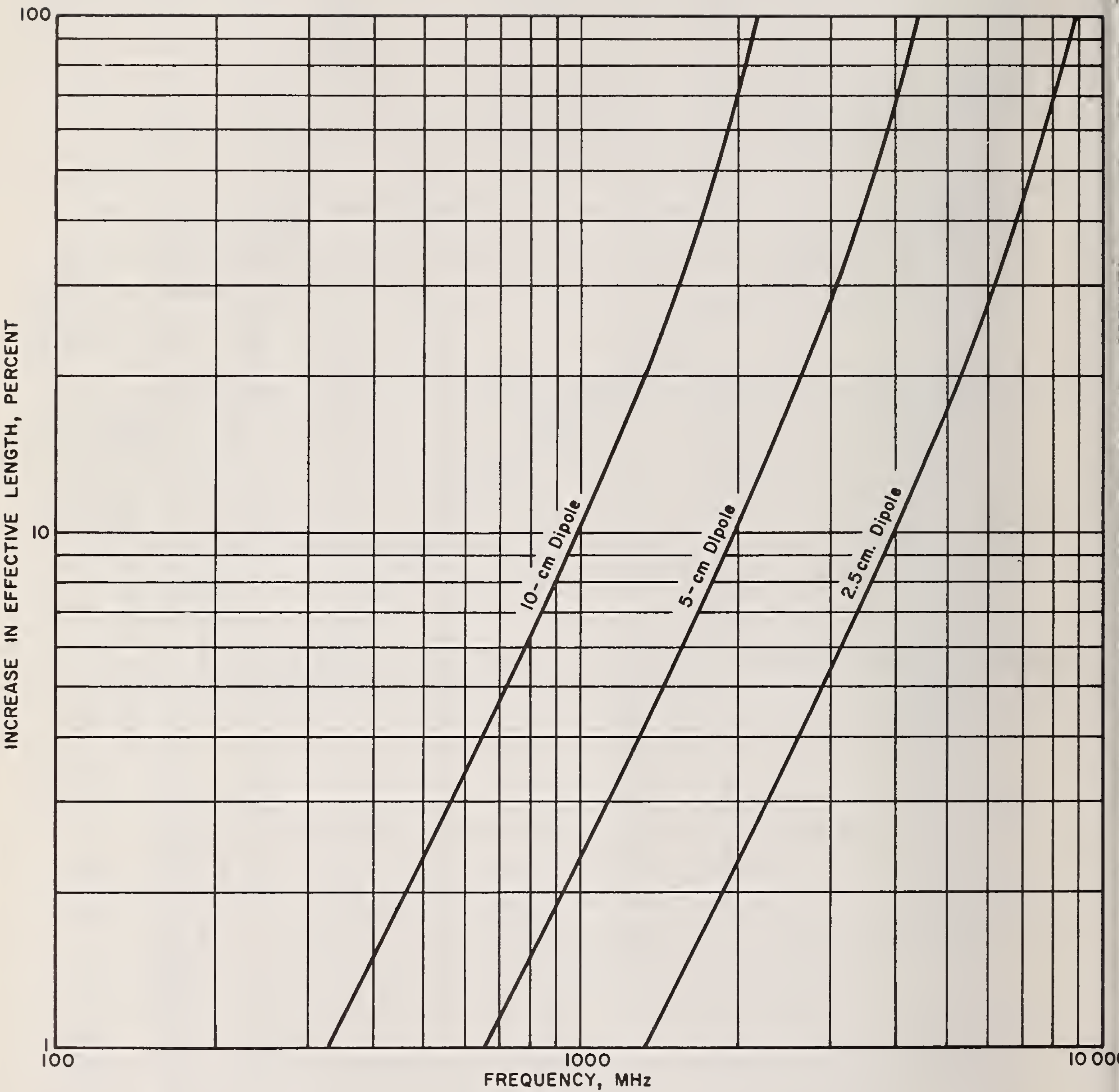

Figure 5. Increase in the effective length of the 2.5-cm, 5-cm and 10-cm dipole antennas with frequency.

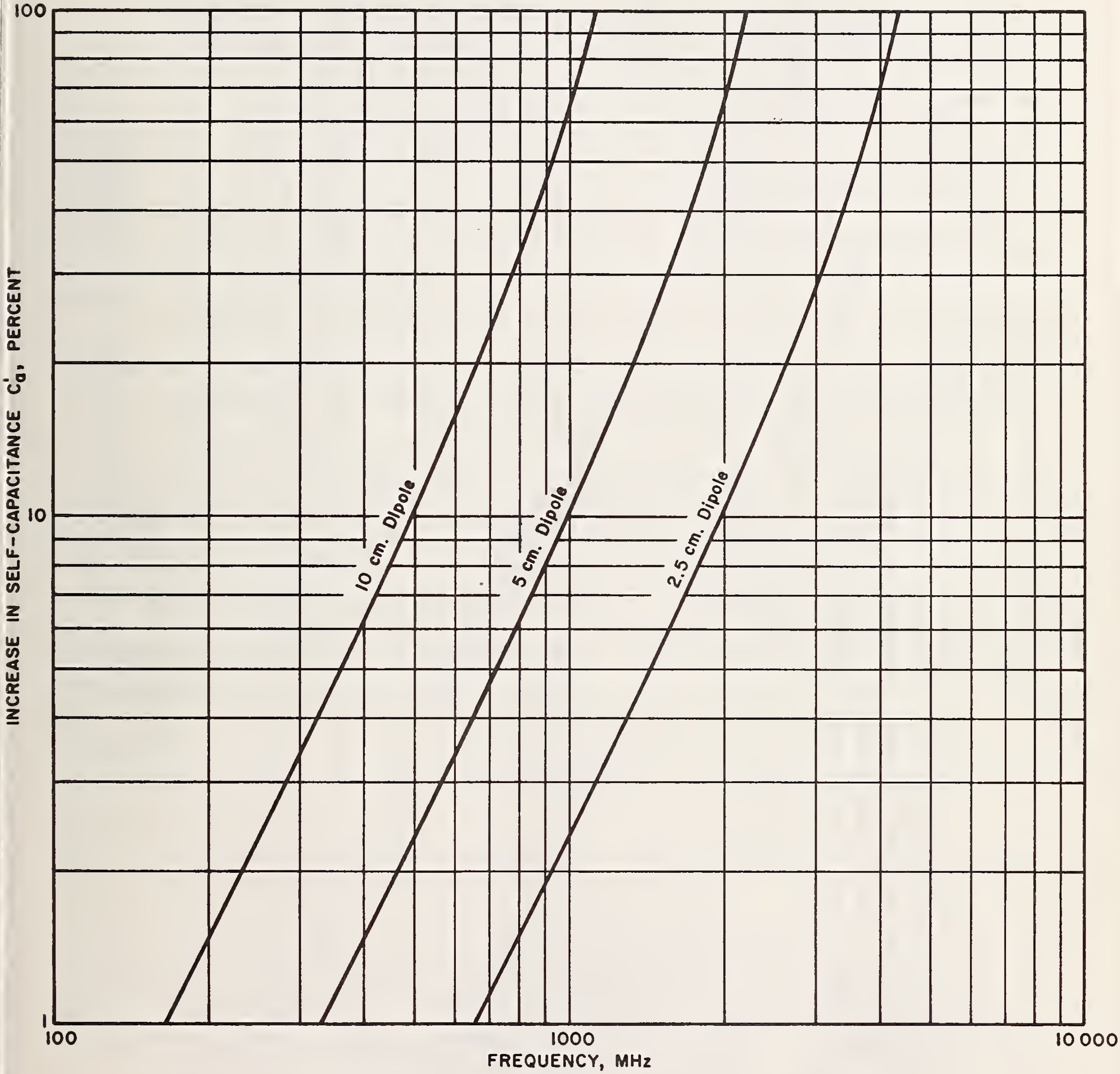

Figure 6. Increase in the apparent self-capacitance of the 2.5-cm, 5-cm and 10-cm dipole antennas with frequency.

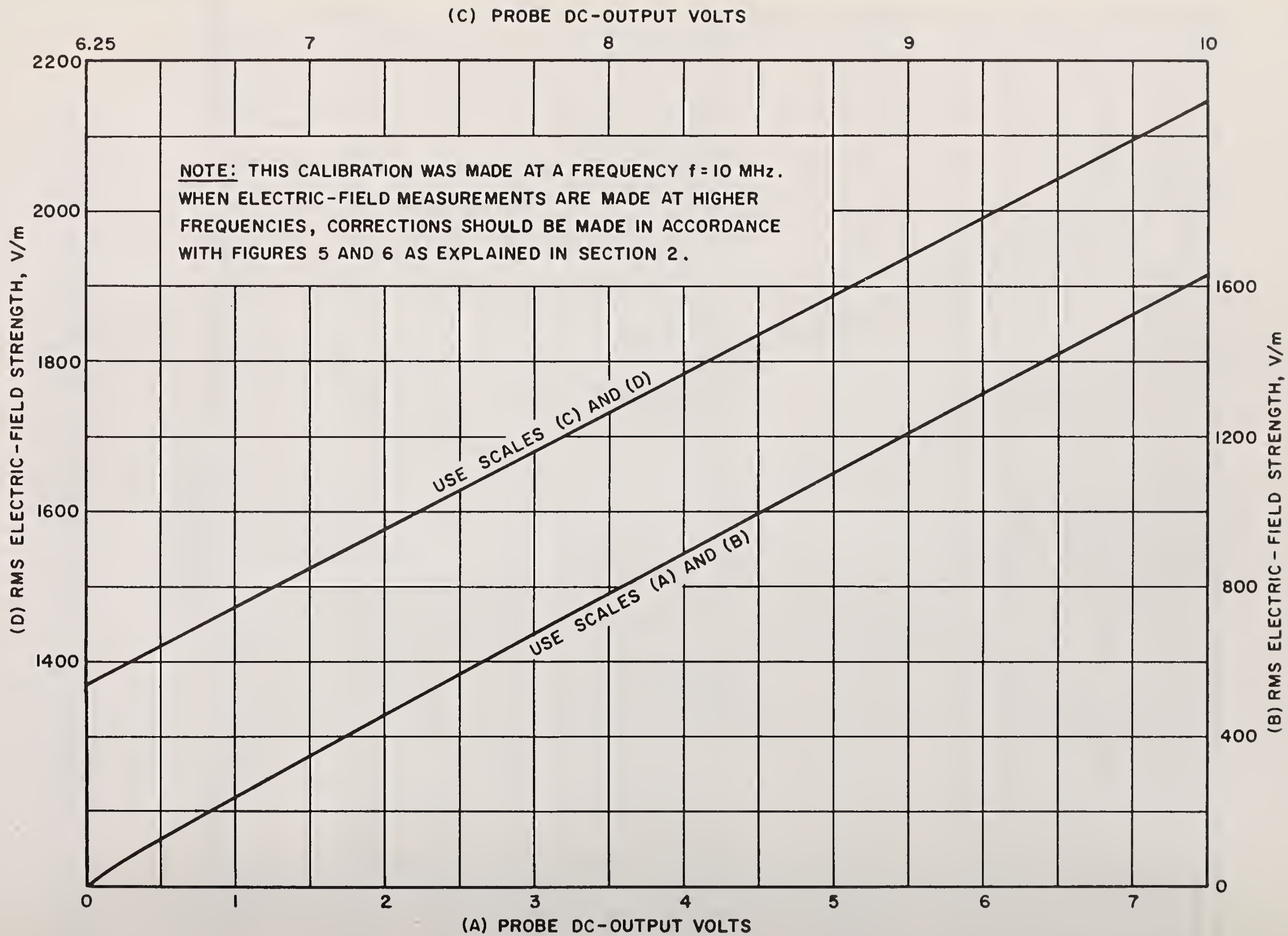

Figure 7. Calibration of the 10-cm dipole probe for USAFSAM, Brooks AF Base (f = 10 MHz).

3. THE MAGNETIC NEAR-FIELD PROBES

3.1 General. The magnetic-field probes consist of two electrically-small, single-turn, balanced loop antennas 10-cm and 3.16-cm in diameter, and are designed to be used for measuring hazard-level, magnetic fields. The response of the loop antennas is directly proportional to frequency, but averages from roughly 0.5 to 5.0 and from 5.0 to 50 amperes per meter, respectively, over the frequency range from approximately 10 to 30 MHz or higher.

In addition, a third loop antenna, having a diameter of 1-cm, was included in the theoretical analysis. This size of loop appears to be necessary for magnetic-field measurements at frequencies above about 75 MHz. It also appears that printed-circuit techniques will probably be required in the design of this loop and its associated r-f filter, in order to minimize extraneous responses.

The general construction of the 3.16-cm and 10-cm loop antennas is shown in figure 8, and the dimensions are given in Table V.

Table V

Dimensions of the 3.16-cm and 10-cm Loop Antennas

Diameters				Gap Widths		Lengths	
D-1	D-2	D-3	D-4	G-1	G-2	L-1	L-2
(cm)	(inches)			(inches)		(inches)	
3.16	0.1250	0.4375	0.375	0.001	0.031	0.5625	0.250
10.00	0.1875	0.6250	0.375	0.001	0.031	0.8750	0.375

Notes: (a) see figure 8 for definitions of symbols;
(b) Mylar dielectric 0.001" thick is used in gap G-1 at the base of each loop.

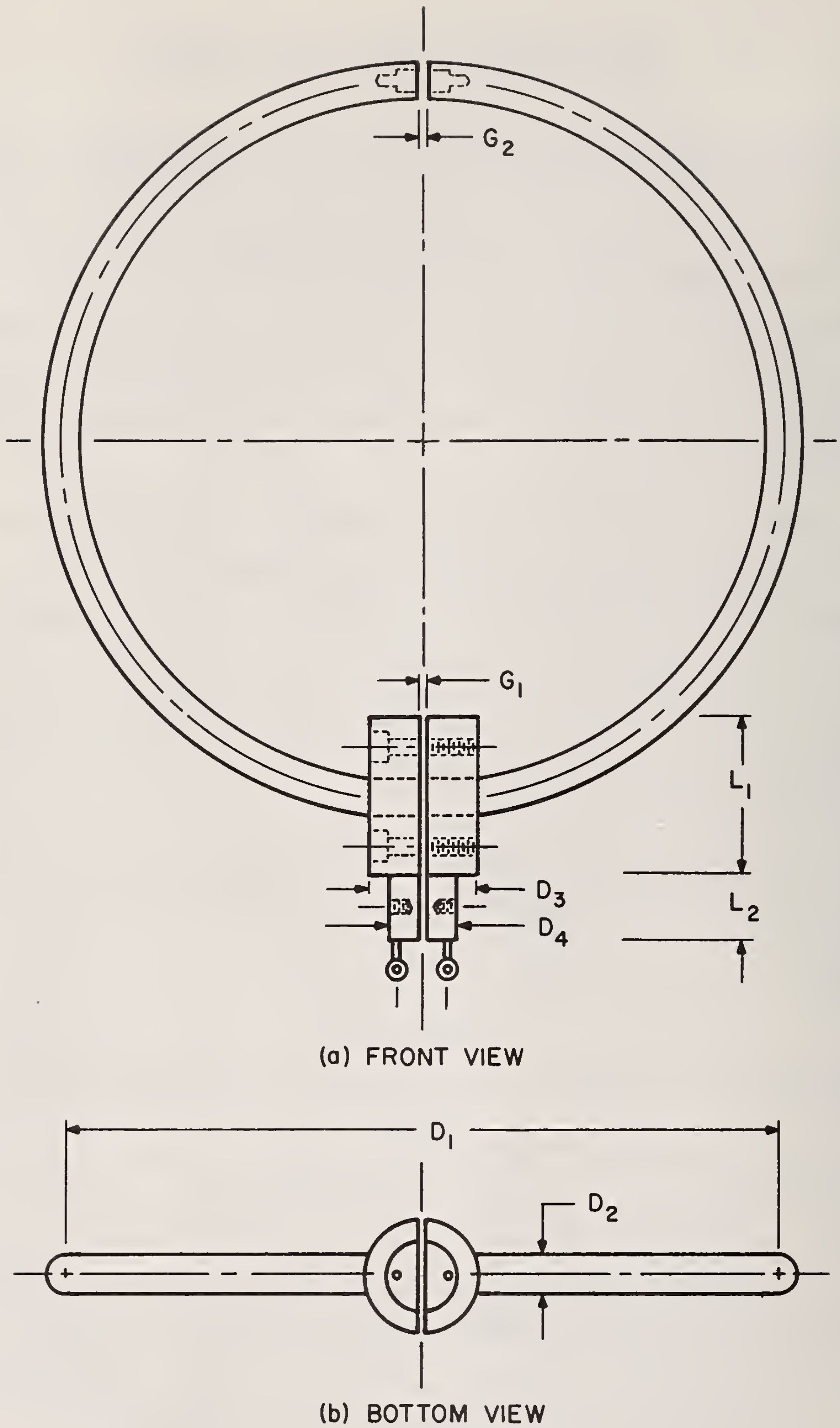

Figure 8. Construction details of the 3.16-cm and 10-cm diameter loop antennas. See Table V for dimensions.

The loop antennas are fabricated in two identical halves as shown in figure 8. The loop is formed of two semicircular brass rings, each of which fits into one half of a split, brass center post and is secured by silver soldering. The inner surfaces of the gap are then milled to the required dimensions and lapped flat and parallel to form a built-in gap capacitor. Mylar dielectric film 0.001 inch thick is used in the gap (G-1), as indicated in Table V. A second insulated gap 0.031 inch wide is provided at the top of each loop as shown. A small semiconductor diode is mounted inside of a flanged polystyrene sleeve, which mounts internally in this gap in a similar manner to that previously described for the dipoles in Section 2.1. The two halves of the loop are fastened together with two 4-40 fillister-head (recessed), nylon machine screws through the centerpost as indicated. Solder lugs are provided, as shown, for connecting a filter network.

Each loop is mounted at the end of a 36-inch-long tubular, fiberglass handle. The loops can be either rigidly fastened to the handle or made to swivel to facilitate aligning the loop axis with the field vector when making measurements. Conducting-teflon monofilaments connect the probe output at one end of the handle to the twin plug at the other, as previously described for the dipoles.

3.2 <u>Electrical Characteristics</u>. A schematic diagram of the portable loop probe is shown in figure 9. Each probe originally had a type 1N4148 silicon-junction, semiconductor diode connected internally across its gap, as shown, to rectify the induced r-f voltage. Type 1N5711 Schottky-barrier diodes were used in the later models because of their improved response at

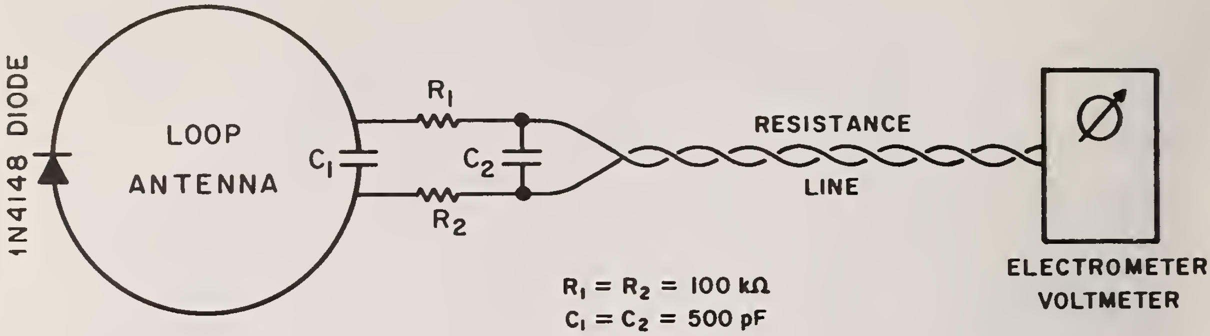

Figure 9. Schematic diagram of the portable loop probe.

the higher frequencies. A balanced r-c, r-f filter, consisting of R-1, R-2, C-1, and C-2, is built into the tubular handle next to the probe to prevent any r-f voltage, picked up on the resistance line, from being rectified by the diode and causing an error in indication. The capacitor, C-1, is built into the base of the loop, as previously discussed, and serves as an r-f by-pass capacitor to complete the loop r-f circuit without short-circuiting the rectified d-c output. A 30-foot length of resistance line connects the probe with the remote, high-resistance, d-c electrometer voltmeter, as in the case of the dipoles.

The equivalent circuit of the small balanced loop antenna is shown in figure 10. L_a is the internal, low-frequency

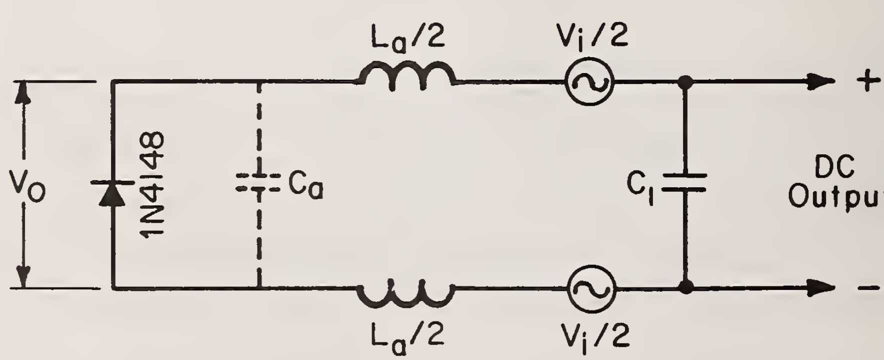

Figure 10. Equivalent circuit of a small loop antenna.

inductance of the loop. C_a represents the combined distributed capacitance of the loop and the gap and diode capacitances, and is what determines the partial resonance effect to be discussed in Section 5. The radiation and ohmic resistances are negligible for small loops in this frequency range and are being ignored in this treatment. V_i is the r-f voltage induced in the loop by the magnetic field, H. V_o is the output r-f voltage appearing across the gap and diode at the center of the loop. At low-frequencies where C_a can be neglected, $V_o \cong V_i$.

3.3 <u>The Response of a Small Loop Antenna</u>. The induced voltage, V_i, can be determined from one of Maxwell's equations relating E and H at any point. For the sinusoidally-time-varying, steady-state case, this is [6,7]

$$\nabla \times \overline{E} = -j\omega\mu\overline{H}. \tag{5}$$

If both sides of equation (5) are integrated over the circular area of the loop, $A = \pi r^2$, and Stokes' theorem is applied, we obtain

$$\oint_0^{2\pi} \overline{E}\cdot\overline{d\ell} = -j\omega\mu HA. \tag{6}$$

ℓ is the circumference of the loop, $\ell = 2\pi r$, meters. From the Maxwell-Faraday law, the left-hand side of equation (6) is the induced voltage, V_i, in the loop. So that

$$\oint_0^{2\pi} \overline{E}\cdot\overline{d\ell} = V_i = -j\omega\mu HA, \text{ volts,} \tag{7}$$

where $\omega = 2\pi f$, $\mu = 4\pi 10^{-7}$, and H is the normal component of the magnetic-field strength in amperes per meter, assumed constant over the area of the loop. It should be noted from equation (7) that the induced voltage, V_i, lags the magnetic field, H, by 90 degrees. This will be made use of later in evaluating errors due to the electric-dipole response of the loop.

From equation (7), the magnitude of the induced voltage can be written

$$|V_i| = 0.2\pi^3 f_{MHz} d^2 H, \text{ volts,} \tag{8}$$

where f_{MHz} is the operating frequency in megahertz, and d is the mean diameter of the loop in meters. As can be seen, the response of the loop (at low-frequencies) is proportional to frequency which somewhat complicates its calibration and use. The response is also proportional to the area of the loop, $A = \pi d^2/4$, in square meters. The relative responses of the 1-cm, 3.16-cm, and 10-cm diameter loop antennas are shown vs. frequency in figure 11.

3.4 Loop Antenna Calibration and Measurement Errors. Measurement errors for a small loop antenna due to its electric-dipole response and partial resonance effects are treated in detail in Sections 4 and 5, respectively. Typical values will be given here for the three sizes of loops under discussion, as determined from figures 15 and 18. The total worst-case error resulting from the above effects is approximately 0.5 dB (6 percent) when the loop diameter is 0.01 wavelength. This would correspond to a frequency of 25 MHz for the 10-cm diameter loop, 75 MHz for the 3.16-cm diameter loop, and 250 MHz for the 1-cm diameter loop. It is possible to greatly reduce these errors by following an experimental procedure outlined near the end of Section 4.

Typical calibration curves, made at a frequency of 10 MHz, for the 10-cm loop probe are shown in figure 12 at the end of this section. The same curves can be used for the 3.16-cm diameter loop, except that the values of magnetic-field strength obtained from the curves should be multiplied by a factor of 10. These curves are valid for use at a frequency of 10 MHz only. When measurements are made at any other frequency, the value of H obtained from the curves should be multiplied by the factor 10/f, where f is the other frequency in megahertz. The calibrations were made at NBS in terms of a high-level standard magnetic field, in a manner somewhat similar to that described for the electric dipoles in Section 2.4.

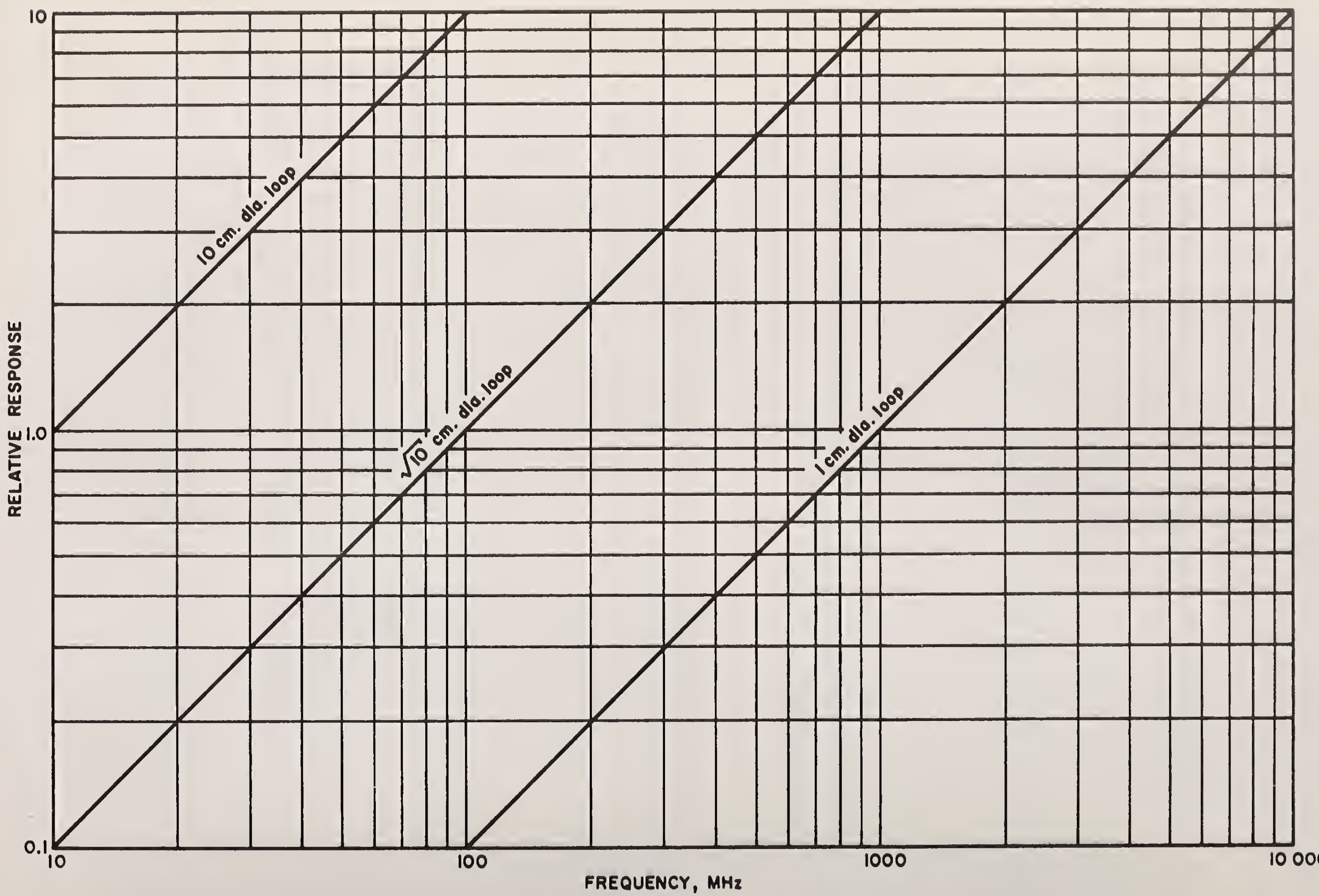

Figure 11. Relative response of the 1-cm, 3.16-cm and 10-cm diameter loop antennas.

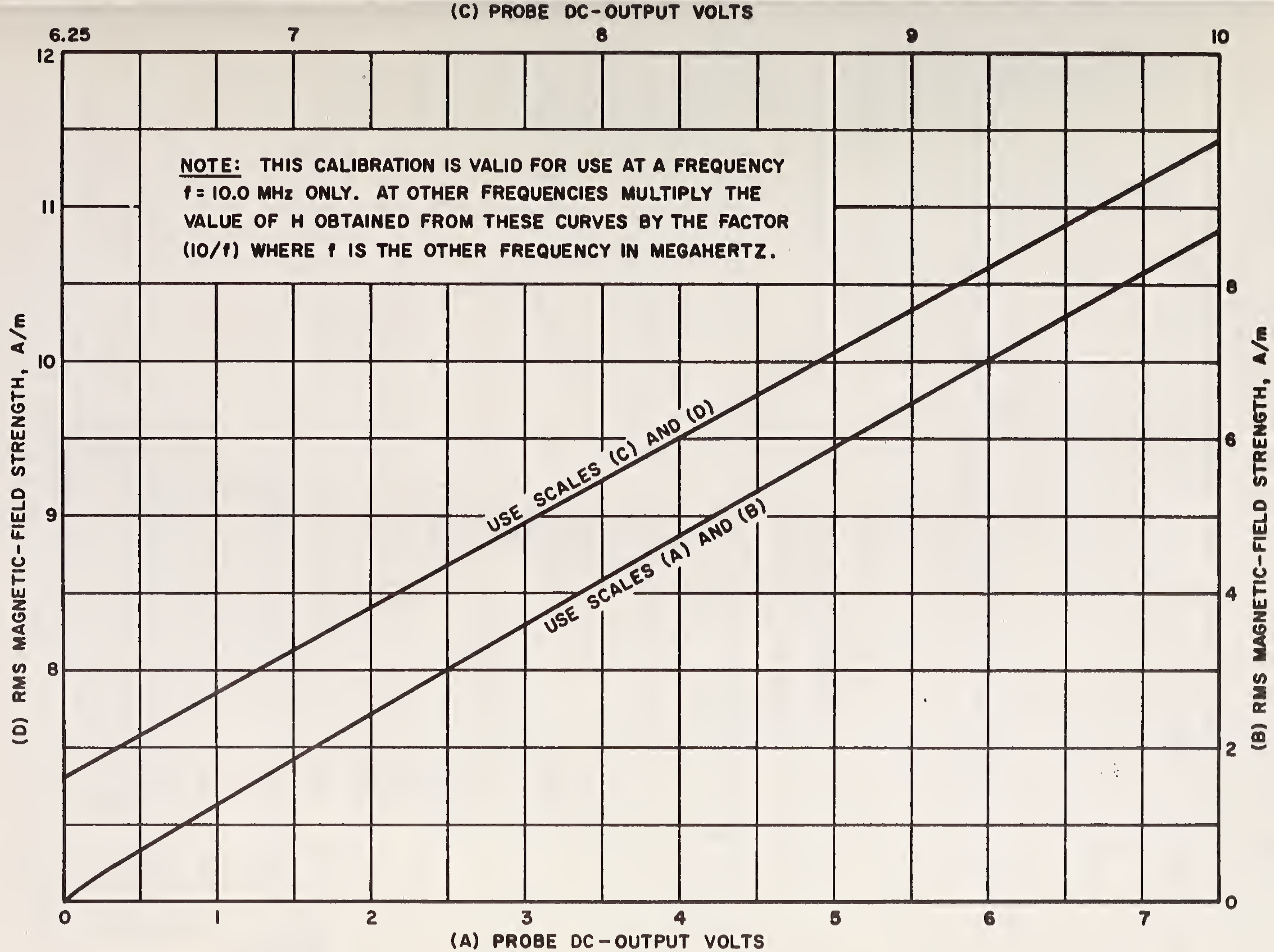

Figure 12. Calibration of the 10-cm diameter loop probe for USAFSAM, Brooks AF Base (f = 10 MHz).

4. THE ELECTRIC-DIPOLE RESPONSE OF A LOOP ANTENNA

This section will treat the electric-dipole response of a loop antenna when measuring H. As is well known, this response will contribute to the H-field measurement error. It can be minimized in any of three ways: (a) by the use of a doubly loaded loop as proposed by Whiteside [8]; (b) by making the linear dimensions of the loop small compared to the wavelength, if a singly loaded loop is used; and (c) by following a procedure, when measuring H, that will average out the electric-dipole response, as will be discussed later.

The equivalent lumped circuit of an electrically small, singly loaded loop antenna showing the electric-dipole and magnetic-dipole responses is given in figure 13. Since the loop is assumed to be small, the radiation resistance has been ignored.

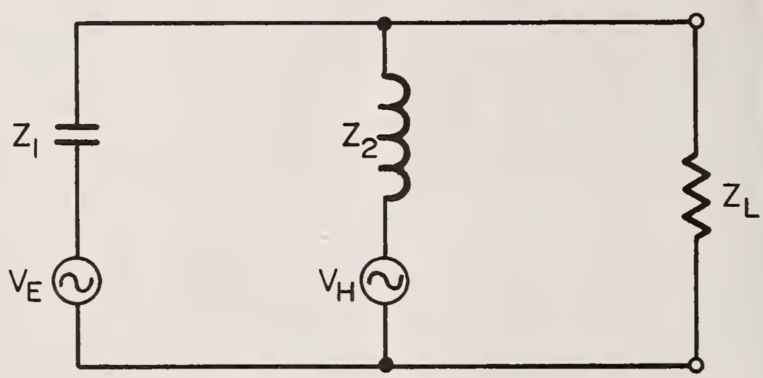

Figure 13. Equivalent lumped circuit of an electrically small, singly loaded loop antenna showing the electric-dipole and magnetic-dipole responses.

V_E is the induced voltage in the loop due to the electric-dipole response to the component of the electric field, E, in the plane of the loop. $V_E = L_{eff} \cdot E$, where L_{eff} is the electrical effective length. V_E is in time phase with E.

V_H is the induced voltage due to the magnetic-dipole response to the component of H normal to the plane of the loop. $V_H = -j\omega\mu H \cdot A$. V_H is in time-phase quadrature with H.

$Z_1 = -j/\omega C$, where C is the low-frequency self-capacitance of the loop. A is the area of the loop. $A = \pi d^2/4$.

$Z_2 = j\omega L$, where L is the low-frequency self-inductance of the loop.

Z_L is the load impedance connected to the loop terminals, i.e., the diode load impedance, the gap capacitance, etc.

The total voltage, V_i, induced in the loop is the complex sum of V_H and V_E, i.e.,

$$V_i = V_H + V_E = V_H\left(1 + \frac{V_E}{V_H}\right). \tag{9}$$

The error in measuring H, caused by the electric-dipole response, can therefore be determined from the ratio of the voltages produced across the load by V_E and V_H. Call these voltages V_{EL} and V_{HL} respectively. The error ratio, which is in general complex, is then given by

$$\frac{V_{EL}}{V_{HL}} = \frac{Z_2}{Z_1}\left(\frac{V_E}{V_H}\right) = -\omega^2 LC\left(\frac{V_E}{V_H}\right). \tag{10}$$

This ratio is independent of the load impedance, Z_L, as can be seen from equation (10). The induced-voltage ratio, V_E/V_H, is given by

$$\frac{V_E}{V_H} = \frac{j3}{4\omega\mu w}\left(\frac{E}{H}\right), \tag{11}$$

where the electrical effective length, L_{eff}, has been assumed to be equal to three-quarters of the length of one side, w, for a square loop [9]. Before equation (11) can be evaluated, both the magnitude and phase of the complex wave-impedance, E/H, must be known.

For the case of a uniform plane wave, where $E = 120\pi H$, equation (11) becomes

$$\frac{V_E}{V_H} = j\left(\frac{3}{4\beta w}\right), \tag{12}$$

where $\beta = 2\pi/\lambda$, and λ = free-space wavelength, meters. When equation (12) is substituted in equation (10) along with Whiteside's expressions for L and C, his formula for the error ratio of a small, singly loaded square loop is obtained

$$\frac{V_{EL}}{V_{HL}} = -j\ 3\pi\left(\frac{w}{\lambda}\right)\left(\frac{\Omega - 4.32}{\Omega - 3.17}\right), \tag{13}$$

where $\Omega = 2 \log_e(4w/a)$, and a is the radius of the conductor forming the loop.

Since the bracketed term involving Ω at the extreme right in equation (13) is approximately unity, if Ω is large, the error ratio for a square loop can be approximated by

$$\frac{V_{EL}}{V_{HL}} \cong -j\ 3\pi\left(\frac{w}{\lambda}\right). \tag{14}$$

For a circular loop of diameter, d, the error ratio is [10]

$$\frac{V_{EL}}{V_{HL}} \cong -j\ 2\pi\left(\frac{d}{\lambda}\right), \tag{15}$$

where λ is the wavelength in meters. As can be seen, the error ratio is directly proportional to the electrical size of the loop, w/λ or d/λ. It can be seen from equations (14) and (15) that the error ratio for a circular loop is approximately 3.5 dB less than for a square loop, when $d = w$.

For the <u>plane-wave</u> case in which E and H are in time phase, the error ratio given by equation (15) is imaginary, as shown. In the <u>near-field</u> case in which E and H may approach a time-phase-quadrature relationship (in the limit as the source is approached), equation (15) becomes real in the limit, and also must be corrected for the new ratio E/H existing in the near zone. In the former plane-wave case, an error ratio of 0.1 will result in an error due to E-field pickup of only 0.5 percent. In the near-zone case the resulting error may be as high as 10 percent or more in the limit for the same error ratio.

If the magnitude and phase angle of the wave impedance are not known, the following experimental technique can be used to minimize the error due to E-field pickup. While making an H-field measurement, the loop is slowly rotated in its own plane, about its axis. If no E-field pickup is present, the output indication of the loop should not change. If a change is noted, the correct value of H can be determined by taking the average of the high and the low readings as the loop is slowly rotated through 360 degrees.

Equation (15) can be used to estimate the worst-case error in the H-field measurements and is shown plotted vs. d/λ in figure 14. For a value of $d/\lambda = 0.1$, for example, this error may be as high as 60 percent. This error is also shown plotted vs. frequency for the 1-cm, 3.16-cm, and 10-cm diameter loops in figure 15.

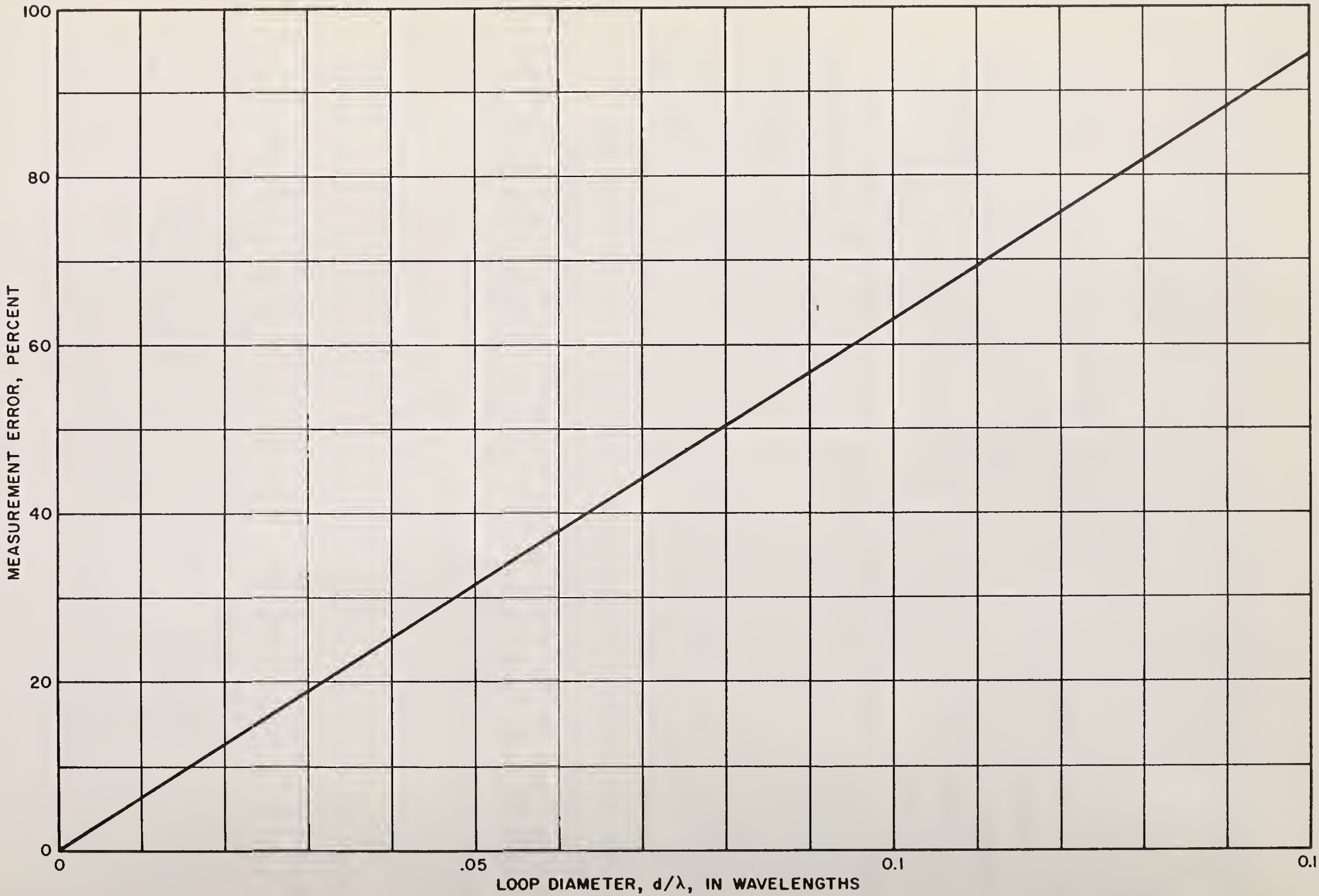

Figure 14. Worst-case measurement error of a circular-loop antenna due to its electric-dipole response.

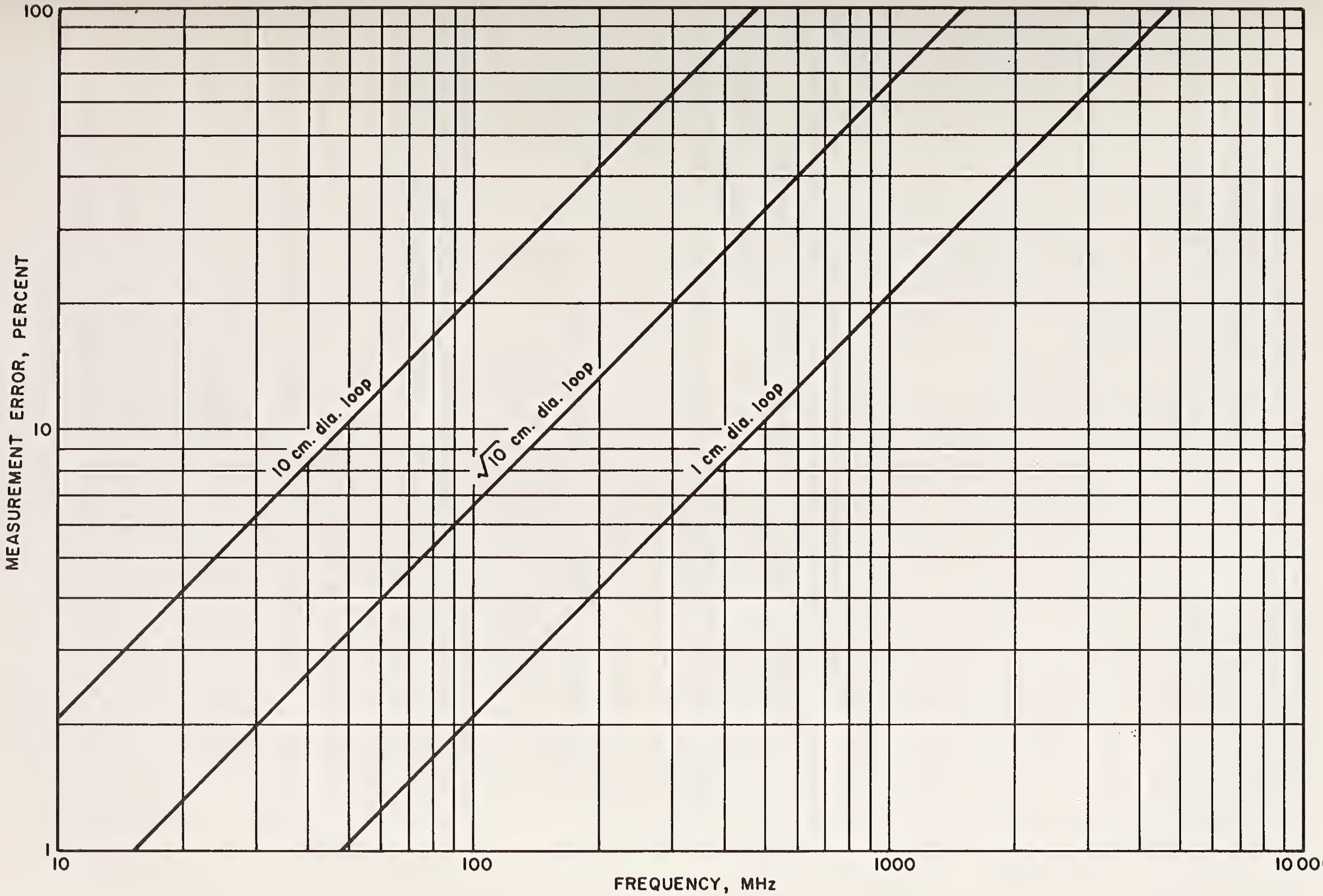

Figure 15. Worst-case measurement error of the 1-cm, 3.16-cm and 10-cm diameter loop antennas due to their electric-dipole response.

5. PARTIAL-RESONANCE EFFECT IN A LOOP ANTENNA

Partial resonance is the result of the combined effect of the distributed capacitance of the loop and the gap and diode capacitances, represented by C_a in figure 10 (Section 3.2).

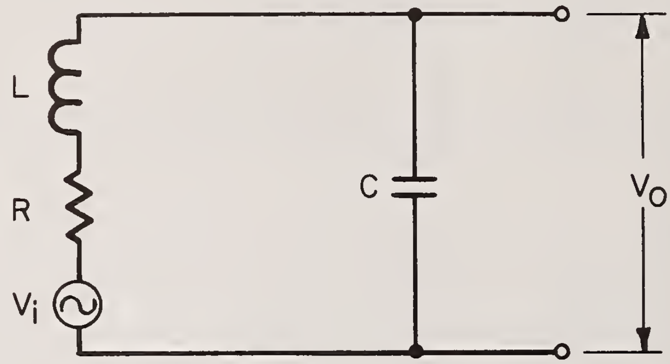

Figure 16. Simplified equivalent circuit of a small loop antenna.

Using the simplified circuit diagram shown in figure 16, the response can be written [11]

$$\frac{V_o}{V_i} = \frac{-j\ \frac{1}{\omega C}}{R + j\left(\omega L - \frac{1}{\omega C}\right)} = \frac{-j\ \frac{\omega_o}{\omega}}{\frac{1}{Q} + j\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)}, \qquad (16)$$

where $Q = \frac{X_o}{R}$, $X_o = 2\pi f_o L = \frac{1}{2\pi f_o C}$, $\omega_o = 2\pi f_o$, and f_o is the self-resonant frequency of the loop. The resistance, R, is assumed to remain constant with frequency, but turns out to be negligible anyway, as will be seen.

The magnitude of equation (16) can be rearranged to give

$$\left|\frac{V_o}{V_i}\right| = \frac{1}{1 - \delta^2}\left[1 + \frac{\delta^2}{Q^2(1-\delta^2)^2}\right]^{-\frac{1}{2}}, \qquad (17)$$

where $\delta = \omega/\omega_o$ is the ratio of the operating frequency to the resonant frequency of the loop. It can be shown that if $Q \geq 10$, and $\delta \leq 0.75$, the second term within the bracket will be negligible, and equation (17) will simplify to

$$\left|\frac{V_o}{V_i}\right| \cong \frac{1}{1 - \delta^2}. \tag{18}$$

Equation (18) can be used to estimate the required correction in the loop measurement for values of δ up to 0.75, and is shown plotted in figure 17. The correction is also shown plotted vs. frequency for the 1-cm, 3.16-cm, and 10-cm diameter loops in figure 18. Under the above conditions these corrections are essentially independent of the Q of the loop (that is, independent of its losses) and depend only on the ratio, ω/ω_o, defined above. The approximate self-resonant frequencies for the three loop probes are given in Table VI.

Table VI

Approximate Self-Resonant Frequencies of the Loop Probes

Loop Diameter	Frequency
10.00 cm	280 MHz (a)
3.16 cm	760 MHz (b)
1.00 cm	2060 MHz (c)

Note: (a) and (b) are measured frequencies. (c) is projected from (a) and (b).

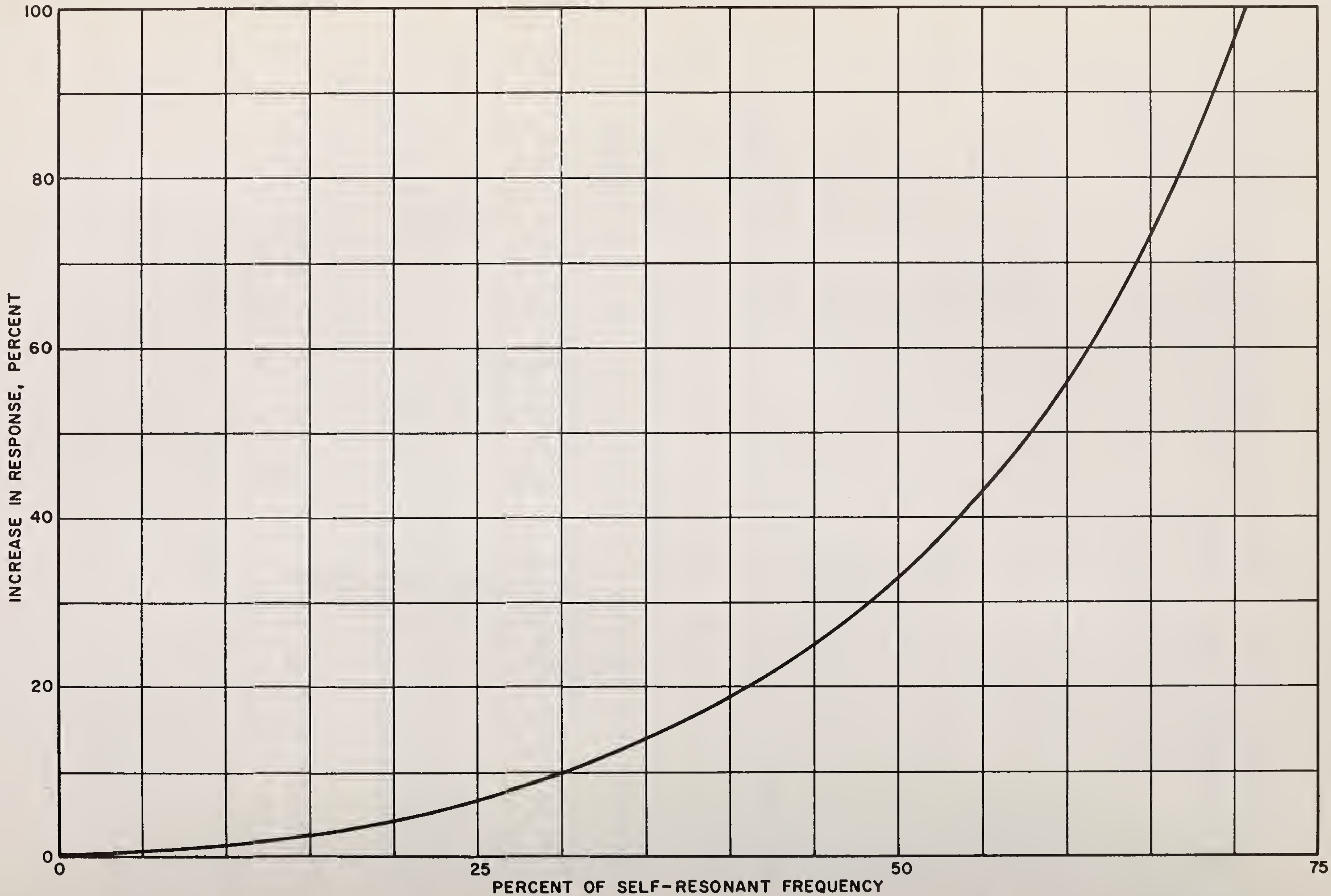

Figure 17. Increase in the response of a loop antenna due to partial resonance.

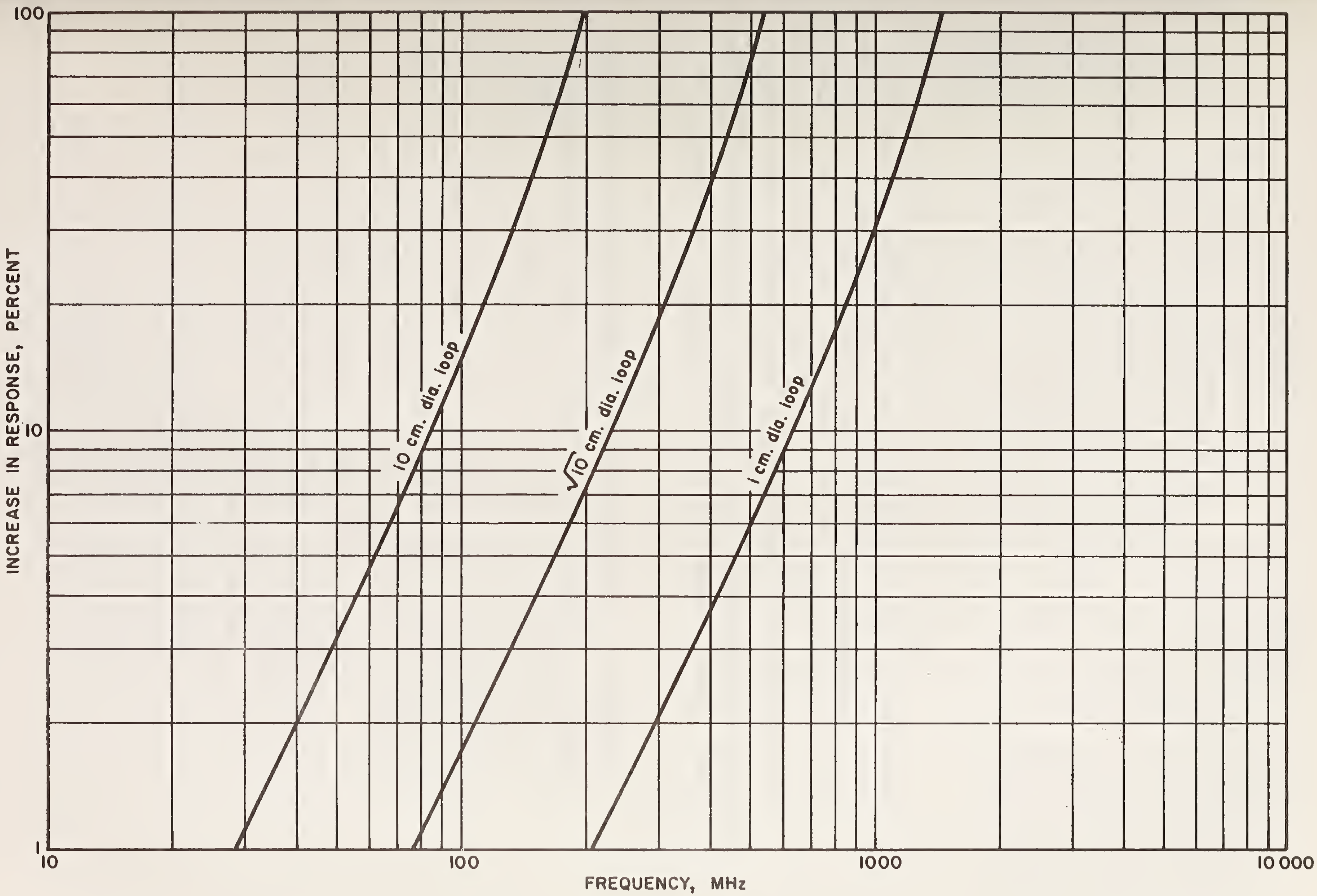

Figure 18. Increase in the response of the 1-cm, 3.16-cm and 10-cm diameter loop antennas due to partial resonance.

6. ERRORS DUE TO FIELD HARMONICS

If harmonics are present in the fields being measured, they can contribute to the resulting measurement error. The effect of field harmonics in the case of electric-field measurements made with short, dipole antennas is not as serious as it is in the case of magnetic-field measurements made with small, loop antennas. This results from the fact that the principal response of a short dipole does not increase as rapidly with frequency as does the response of a loop, as has been shown. The worst-case correction for dipole measurements can be estimated by adding linearly the magnitudes of the various field harmonics present, since the detector circuit is essentially peak reading as mentioned in Section 2.2. The magnitude of each harmonic should be multiplied by the appropriate factor as determined from figures 5 and 6 in Section 2, in order to take into account the changing dipole response with frequency.

As an example, consider a third harmonic field at 120 MHz, the level of which is -30 dB with respect to a 40 MHz fundamental-frequency field being measured. The magnitude of this harmonic is roughly 3 percent of the fundamental. In the case of the dipole antenna, the increase in response at 120 MHz is approximately 0.6 percent for the 10-cm dipole and 0.1 percent for the 5-cm dipole, as can be estimated by extrapolating figures 5 and 6. Assuming that the harmonic signal adds linearly to the fundamental, the resulting worst-case measurement error is 3.6 percent for the 10-cm dipole, and 3.1 percent for the 5-cm dipole due to this cause.

The principal response of a loop antenna is proportional to frequency, as was shown in figure 11 for the 1-cm, 3.16-cm, and 10-cm diameter loops, and as was discussed in Section 3.3. In addition, both the electric-dipole response of the loop, and the partial-resonance effect, previously discussed in Sections 4 and 5, respectively, also increase with frequency, which can further accentuate the harmonic error. The harmonics of many transmitters are at least 30 decibels below the fundamental-frequency output of the transmitter. If the harmonics are stronger than this, it is possible to use filters in the transmitter output to further attenuate them, although many commercial transmitters probably do not take advantage of this.

If the levels of the harmonic fields are known, the resulting contribution to the worst-case measurement error, due to the above three factors, can be estimated by reference to figures 11, 15, and 18. Since the phase relationships between these three factors is, in general, not known, the worst that can happen is for them to add linearly to the principal response of the loop probe at the fundamental frequency.

The apparent magnitude of the harmonic, in the above example, will be enhanced by a factor of three (from 3 percent to 9 percent) due to the loop's principal response. The electric-dipole response and partial resonance of the loop will further increase this to a worst case error due to this one harmonic of 14 percent for the 10-cm loop and 10 percent for the 3.16-cm loop as can be determined from figures 15 and 18. If the effect of only one additional harmonic were taken into account, the worst-case errors could be doubled. This, of course, would be in addition to the other errors discussed. There is no simple way of accurately evaluating these harmonic errors unless the actual harmonic content of the field is known.

7. THE NBS CONDUCTING-PLASTIC TRANSMISSION LINE

7.1 Introduction. In the past, it has been found difficult to accurately measure the absolute strength of electromagnetic fields having a complex spatial distribution over a significant range of magnitudes. Part of the difficulty has been in the use of a metallic r-f transmission line between the dipole measuring antenna and the receiver, or selective voltmeter, located at a point remote from the antenna. Such metallic lines often cause large measurement errors because the line not only perturbs the field being measured, but the unwanted r-f currents induced on the line contribute to the total response of the field-strength meter.

One method of avoiding this difficulty is to use some form of radio, optical, or acoustical telemetering transmitter in place of the metallic line. This transmitter, located at the measuring antenna, can be used to transmit the desired information to a remote readout unit. This technique has been employed in various ways by a number of experimenters in the field. However, because of the nature and complexity of the telemetering instrumentation, the resulting devices have had varying degrees of success, but in general have been seriously limited in the usable range of frequencies and magnitudes over which absolute measurements could actually be made.

The NBS near-zone field-strength meters being described in this paper avoid these telemetering difficulties by making use of a special technique. The complex telemetering instrumentation, mentioned above, is eliminated by transmitting the detected output of the measuring antenna over a completely non-metallic electrical transmission line which does not appreciably interact with the field being measured. If the transmission line is made of sufficiently high-resistance material, it will be essentially "transparent" to the surrounding field. Thus, it will not interact appreciably with the field being measured or have unwanted coupling with the measuring antenna. This is to say that the r-f currents induced in the transmission line by the field will be negligibly small, resulting in insignificant reradiation, or that any r-f energy propagating along the line will be heavily attenuated because of the extremely high loss of the line. However, the detected d-c output voltage of the measuring antenna can be transmitted without appreciable loss if a high-resistance d-c vacuum-tube voltmeter or electrometer is used as the readout device at the receiving end of the line [12].

7.2 Conducting-Plastic Materials. The high resistance, non-metallic transmission lines used by NBS employ conducting-plastic monofilaments in place of the usual copper conductors. In the type of material used, the plastic matrix is made conducting to the desired extent by filling (up to 30 percent) with finely powdered carbon (carbon black) during initial stages of manufacture while the plastic is in a more or less liquid or molten form. The finely-divided carbon is dispersed throughout the plastic to make as homogeneous a mixture as possible. Electrical conduction results from actual physical contact between the carbon particles suspended in the plastic matrix. Not all plastics lend themselves to this application. Only certain types of plastic molecules will properly accept carbon particles in a manner which renders the material electrically conducting, without a deleterious effect on the final mechanical properties. For example, nylon is one of the types which, so far, has been found to be unsuitable by various manufacturers. Several types of conducting-plastic were tested at NBS for this purpose. The most suitable was found to be conducting-Teflon made to NBS specifications.

7.3 Construction of the Conducting-Teflon Transmission Line. The parallel-conductor balanced transmission line was constructed using two 0.03 inch diameter nylon-coated, conducting-Teflon monofilaments, each approximately 30 feet in length. Each filament was in turn pulled through a separate 1/8 inch diameter, vinyl-coated, woven-fiber-glass sleeve for added strength and insulation. These two 30-foot lengths were then pulled together through an equal length of heat-shrinkable irradiated-polyvinyl chloride tubing, which served as the outer jacket of the parallel-conductor line. Conducting silver cement was used to electrically bond the ends of the conducting-Teflon filaments to the special twin-connector plugs at both ends of the line. The oversized outer vinyl jacket was then shrunk in place over the entire length of line, and over the plugs at the ends, as well, so as to make them an integral part of the line. A high-temperature heat gun was used to perform the shrinking.

The completed 30-foot length of line was tested in a high-level electromagnetic field. The interaction with the field was found to be approximately two orders of magnitude less than that found for the copper line used as a reference. A high noise level was found to exist on the line, due to changing contact resistance between the carbon particles (when the line was flexed). This noise was found to average approximately 0.2 volt on open circuit, and approximately 0.02 volt when one end of the line was terminated in a resistance of 10 megohms. However, since the d-c level transmitted over this line was in the range from 1 to 10 volts, the "flexural" noise did not present a problem. The Teflon line was found to be considerably more stable electrically than the other plastic lines tested, and therefore was adopted for use with NBS near-zone field-strength meters.

7.4 Electrical Characteristics of the Conducting-Teflon Line.

Conductor Resistance. The nominal volume resistivity of the conducting-Teflon plastic material used is 3 ohm-cm. The measured d-c resistance of this material in 0.03-inch-diameter-monofilament form is 20,000 ohms per lineal foot, giving a loop resistance for the twin-conductor balanced line of 40,000 ohms per foot.

Mutual Capacitance. The mutual capacitance between the parallel conductors of this line, as measured at a low audio frequency, is approximately 10 picofarads per foot.

Propagation Function. The degree of interaction of the conducting plastic transmission line with the field being measured is directly related to the line loss, or attenuation, of the resulting wave induced on the line. By making the attenuation sufficiently great, this interaction can be reduced to negligible proportions. The propagation function of this transmission line can be derived from the classical transmission line equations to be found in any engineering text on the subject [13]. This theory becomes only approximate in the case of the lossy line, however, and is offered here only to illustrate the degree of attenuation to be expected for such a line. In regard to the electrical parameters of this line, it can be shown, at least for frequencies below 100 MHz, that the series inductive reactance per unit length, ωL, is negligible compared to the series resistance per unit length, R, and that the shunt conductance per unit length, G, is negligible compared to the shunt susceptance per unit length, ωC. That is $\omega L \ll R$, and $G \ll \omega C$. When these conditions are substituted in the complete equations for the attenuation and phase functions of the reflectionless transmission line, the following expressions are obtained:

Attenuation Function $\alpha \cong \left[\frac{\omega CR}{2}\right]^{\frac{1}{2}}$, nepers per unit length. (19)

Phase Function $\beta \cong \left[\frac{\omega CR}{2}\right]^{\frac{1}{2}}$, radians per unit length. (20)

It is interesting to note that for this case, both the attenuation and phase functions are given by the same expression. These quantities both increase as the square root of frequency over the frequency range for which the equations are valid.

7.5 Transmission Line Attenuation. The attenuation per unit length along the line can be determined approximately from equation (19) for both the balanced- and the common-modes of propagation, by substituting in appropriate values of R and C. The use of the values R = 40,000 ohms/foot, and C = 10 pF/foot, given previously, yields a balanced-mode attenuation of approximately 6.1 nepers (53 dB) per foot at a frequency of 30 MHz for the conducting-Teflon line. To determine the common-mode attenuation, the two transmission-line conductors are considered as operating in parallel, making R' = 10,000 ohms/foot. The value of resistance to use in equation (19) is R = 2R' = 20,000 ohms/foot. If the insulated line is lying flat on the ground plane, its capacitance, C', to the plane can be considered to be approximately four times that given previously, or 40 pF/foot (assuming the spacing between conductors to be double their height above the image plane). The value of capacitance to use in equation (19) is then C = C'/2 = 20 pF/foot. This, then, results in the same r-c product, and therefore the same attenuation for the common-mode case as for the balanced-mode of propagation.

Thus, it can be seen that the line loss at 30 MHz for either mode of propagation is extremely high for the conditions assumed. This means that the r-f current flowing at any point along the portion of transmission line, lying on the ground plane, will be attenuated by more than 40 dB (100X) in traveling a distance of only one foot. The common-mode attenuation decreases somewhat for the short portion of line near the probe, since its capacitance to the plane is less. However, this has presented no observable difficulty, as long as this section of line is maintained normal to the probe axis to preserve the electrical symmetry. From actual tests made on this line in an electromagnetic field, the attenuation achieved in this line appears to be sufficient to reduce the interaction with the field to negligible proportions.

The attenuation was measured in the laboratory from d-c to 100 MHz using a conducting-Teflon monofilament as the center conductor of a 5-foot length of rigid coaxial line having a copper outer conductor 1.25 inches inside diameter. The measured attenuation agreed with the theoretical, based on classical transmission-line theory, to within a proportion of 1 dB in 10 over the above frequency range. The results are shown plotted in figure 19.

As can be seen from equations (19) and (20), the phase shift, in radians, along the line is equal to the attenuation in nepers for the same distance. It is a little surprising to find that the 30-foot line used in our tests is one-half wavelength long at approximately 10 kHz, due to the extremely rapid phase shift along the line. Conversely, the velocity of propagation, v, along the line is extremely low, as can be determined from equation (21), which is based on the relation, $v = \omega/\beta$.

$$v \cong \left[\frac{2\omega}{CR}\right]^{\frac{1}{2}}, \text{ unit distance per second.} \qquad (21)$$

The velocity of propagation for the above line is approximately 0.094 v_o, at a frequency of 30 MHz, where v_o = free-space velocity. The velocity is proportional to $(\omega)^{\frac{1}{2}}$ over the frequency range for which equation (21) is valid.

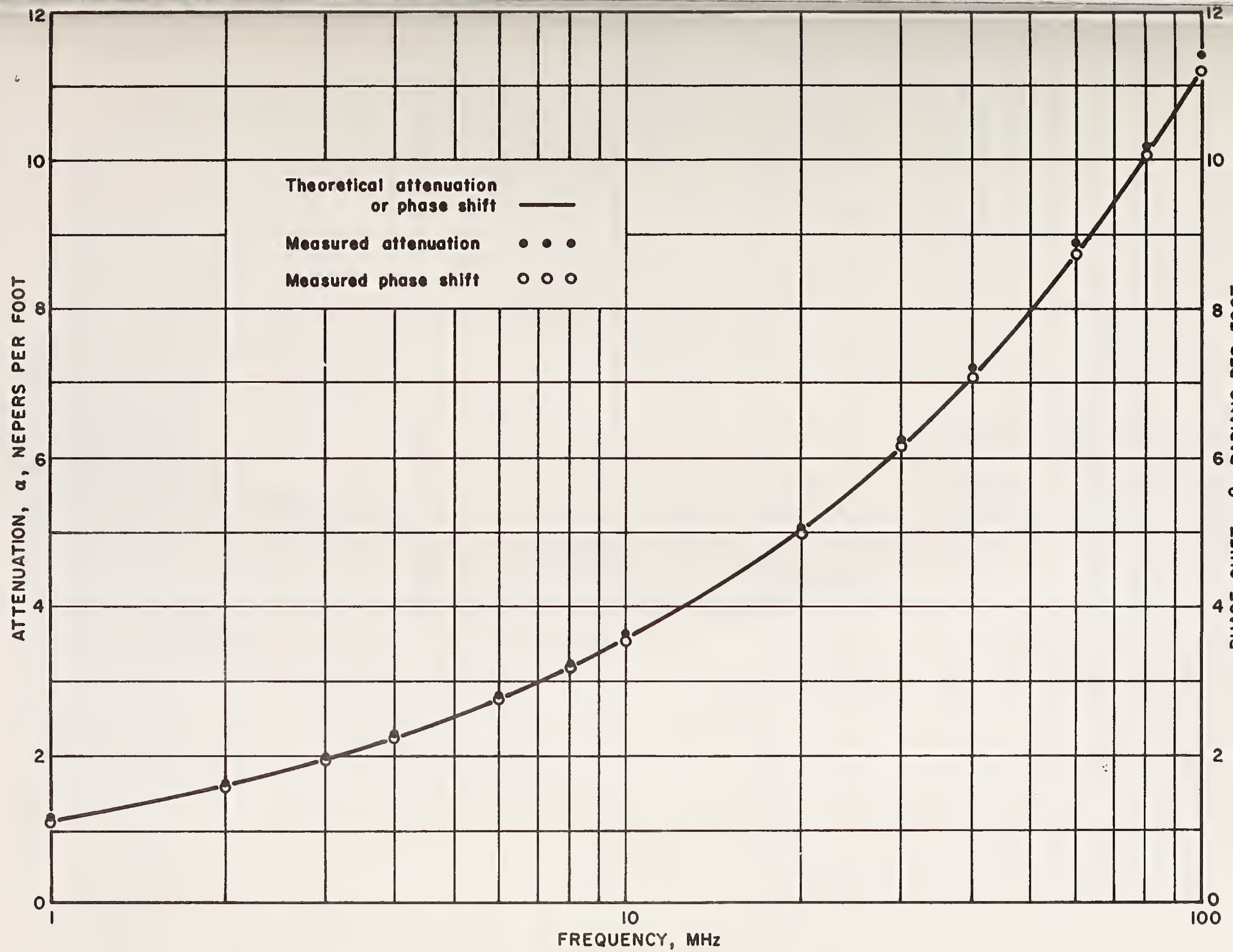

Figure 19. Attenuation and phase-shift of the conducting-plastic transmission line.

8. CONCLUSIONS

The near-field probes described in this publication will enable researchers to accurately measure hazard-level, non-ionizing, electric and magnetic fields in conjunction with their various radiation exposure programs being carried out in the frequency range from 10 to 1000 MHz. These probes in conjunction with an electromagnetic near-field synthesizer, recently developed at NBS [14], will also enable researchers to evaluate any near-field effects encountered in connection with their studies of r-f biological hazards.

The recent discovery at NBS and elsewhere of the leading importance of the role played by the magnetic component of the field in the area of r-f biological hazards to humans in this frequency range, points up the need for accurate magnetic-field probes for further evaluating this effect.

The two sizes of dipole- and loop-antenna probes constructed were originally intended for use at frequencies from 10 to 30 MHz only. If the frequency corrections, resulting from the performance analysis discussed, are applied, the estimated measurement uncertainty will be less than 0.5 dB (6 percent) at those frequencies for which the overall length of the dipole does not exceed $\lambda/8$, or the loop diameter, $\lambda/100$. This will permit electric-field measurements at frequencies up to 750 MHz, and magnetic-field measurements at frequencies up to 75 MHz. The addition of the smaller, third dipole and loop probes, treated theoretically, will extend these frequencies to 1500 MHz and 250 MHz, respectively.

Thus, it is possible to reduce the measurement uncertainty, or to extend the usable frequency range, by following the procedures outlined. The probes have performed reliably and provide an electromagnetic hazard-measurement capability not previously available in this frequency range.

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15. SUPPLEMENTARY NOTES

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

This publication describes the development and design of small electric and magnetic near-field probes for measuring hazard-level fields up to 20,000 V/m and 100 A/m, respectively. They were originally designed to be used over the frequency range from 10 to 30 MHz, and consist of short dipole antennas and small, single-turn balanced, loop antennas to measure the electric-and magnetic-field components, respectively. The probes are intended for use by various researchers in their electromagnetic, radiation-exposure programs for determining the effects of hazard-level, non-ionizing, EM fields on living tissue, electro-explosive devices, and volatile fuels.

In order to later extend the use of the probes to frequencies above 30 MHz, a detailed analysis was made of several types of measurement errors likely to be encountered. The principal errors result from a variation with frequency in: the effective length and impedance of the dipoles; and the electric-dipole response and partial resonance of the loops. Corresponding corrections are given for each type of error as a function of the operating frequency from 10 to 1000 MHz, and as a function of the physical and electrical sizes of the probes.

As a result of the analysis, the dipoles can now be used for measurements at frequencies up to 750 MHz and the loops to 75 MHz with an estimated uncertainty of 0.5 dB. Applying recommended corrections will provide a substantial further increase in the usable frequency range.

17. KEY WORDS *(six to twelve entries; alphabetical order; capitalize only the first letter of the first key word unless a proper name; separated by semicolons)*

Electric near-field probe; electromagnetic-field hazard; field-strength measurements; magnetic near-field probe; near-field measurements; r-f hazard measurements; semiconducting transmission line.

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