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**IEEE Std 291-1991**

(Revision of IEEE Std 291-1969, Revision and  
Redesignation of IEEE Std 284-1968 and IEEE Std 302-1969)

# **IEEE Standard Methods for Measuring Electromagnetic Field Strength of Sinusoidal Continuous Waves, 30 Hz to 30 GHz**

Sponsor

**Wave Propagation Standards Committee  
of the  
IEEE Antennas and Propagation Society**

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**Abstract:** Two standard methods for field-strength measurement are described. The standard-antenna method consists of measuring the received power or open-circuit voltage developed in a standard receiving antenna by the field to be measured and computing the field strength from the measured voltage and the dimensions and form of the standard antenna. The standard-field method consists of comparing voltages produced in an antenna by the field to be measured and by a standard field, the magnitude of which is computed from the dimensions of the transmitting antenna, its current distribution, the distance of separation, and effect of the ground. The measurement procedures are outlined, including calibration of commercial field strength and extension of the methods to microwave frequencies. Methods for measuring power radiated from an antenna under several different conditions are briefly presented, and the important considerations for securing useful and accurate measurements are described.

**Keywords:** Field strength, receiving antenna, transmitting antenna

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## Foreword

(This Foreword is not a part of IEEE Std 291-1991, IEEE Standard Methods for Measuring Electromagnetic Field Strength of Sinusoidal Continuous Waves.)

IEEE Standard 291-1991, on the measurement of electromagnetic field strength, updates, consolidates, and supersedes three previously issued standards: 1) IEEE Std 284-1968, 2) IEEE Std 291-1969, and 3) IEEE Std 302-1969. Within the IEEE, this work has been coordinated with the Antennas and Propagation Society Standards Committee, the Broadcast Technology Society, the Electromagnetic Compatibility Society, and the Instrumentation and Measurement Society. The standard has also been coordinated with the FCC and with the international organizations of CCIR (International Radio Consultative Committee) and URSI (International Union of Radio Sciences).

SI (rationalized MKS) units have been used throughout the document.

This standard was prepared by the Subcommittee on Field Strength Measurement of the Wave Propagation Standards Committee. At the time that this standard was approved, the Subcommittee on Field Strength Measurement had the following membership:

**M. Kanda, *Chair***

G. H. Hagn  
D. A. Hill

E. B. Larsen  
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G. S. Smith

The Wave Propagation Standards Committee (WPSC) of the IEEE Antennas and Propagation Society (APS), which balloted and approved this document for submission to the IEEE Standards Board, had the following membership:

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W. L. Stutzman  
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\*Past Chair

The subcommittee, the WPSC, and the APS appreciate the assistance of all those who contributed to this standard.

The final conditions for approval of this standard were met on April 2, 1991. This standard was conditionally approved by the IEEE Standards Board on March 21, 1991 with the following membership:

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\*Member Emeritus

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# IEEE Standard Methods for Measuring Electromagnetic Field Strength of Sinusoidal Continuous Waves, 30 Hz to 30 GHz

## 1. Introduction

### 1.1 General

Most measurements relating to radio wave propagation involve the measurement of field strength. Standard methods for the measurement of this fundamental quantity are described in this document. The power radiated from an antenna (Section 3) also may be determined by measurement of field strength.

### 1.2 Symbols and Units

The general symbols used are given in the following list, together with units to be used when the symbols appear in equations containing numerical coefficients. When more specific or restrictive meanings are indicated by subscripts, the symbols are further defined after the equation in which they are used.

$a$	= Attenuation, numeric.
$A$	= Amperes.
$A$	= Area of a loop antenna, in m <sup>2</sup> .
$A$	= Norton approximation to Sommerfeld attenuation function, numeric.
$A$	= Cable attenuation, in dB.
$b$	= Effective noise power bandwidth, in Hz.
$b_i$	= Impulse bandwidth, in Hz.
$c$	= Velocity of propagation of an electromagnetic wave in a vacuum = $2.9979246 \cdot 10^8$ m/s.
$C$	= Capacitance, in F.
$C_{inc}$	= Incident coupling factor for 4-port dual directional coupler, numeric.
$C_{refl}$	= Reflected coupling factor for 4-port dual directional coupler, numeric.
$d$	= Distance from transmitter to receiver, in m.
$D$	= Maximum dimension of an antenna, in m.
$E$	= Root-mean-square (RMS) value of electric field strength, in V/m.

$E_o$	= Direct wave electric field strength, in V/m.
$EIRP$	= Effective isotropically radiated power, in W.
$f$	= Frequency, in Hz.
$g$	= Intrinsic power gain of an antenna over that of an isotropic antenna, considering its directive gain minus dissipative losses but not its loss caused by impedance mismatches, numeric. The gain of a transmitting antenna is equal to $4\pi$ times the ratio of the power density in a given direction to the net power delivered to the antenna from the transmitter.
$g_d$	= Gain of a short dipole at the surface of a lossy ground, numeric.
$g_e$	= Equivalent gain relative to an isotropic antenna, numeric.
$g_T$	= Power gain of the transmitting antenna in a given direction, numeric.
$g'$	= Standardized receiver gain, numeric.
$G$	= $10 \log_{10} g$ = Antenna gain, in dB, relative to an isotropic antenna (dBi).
$G_e$	= $10 \log_{10} g_e$ = Equivalent gain, dBi.
$h$	= Height of a grounded vertical monopole antenna, in m.
$h$	= Height of an antenna above ground, in m.
$H$	= Root-mean-square (RMS) magnetic field strength, in A/m.
$H$	= Inductance, in H.
$I$	= Root-mean-square (RMS) antenna current, in A.
$k$	= Antenna factor, in reciprocal meters ( $m^{-1}$ ).
$K$	= Antenna factor, in $Db(m^{-1})$ .
$L_{eff}$	= Effective length of an antenna, in m.
$L$	= Physical length of an antenna, in m.
$N$	= Number of loop turns, numeric.
$p$	= Numerical distance.
$P$	= Power, in W.
$q$	= Impedance mismatch factor, numeric.
$Q$	= $2\pi$ times the ratio of energy stored to energy dissipated per cycle. $Q$ is also expressed as the ratio of reactance to series resistance at resonance.
$r_1$	= Radius of a transmitting loop antenna, in m.
$r_2$	= Radius (or mean radius) of a receiving loop antenna, in m.
$R$	= Resistance, in $\Omega$ ; real part of antenna input impedance, in W.
$R_L$	= Real part of load impedance, in $\Omega$
$S$	= Power flux density of an electromagnetic wave, in $W/m^2$ .
$V$	= Root-mean-square (RMS) voltage, in V.
$VSWR$	= Voltage standing wave ratio, numeric.
$V_{oc}$	= Open-circuit voltage across the center gap of a dipole, in V.
$X$	= Imaginary part of antenna input impedance, in $\Omega$
$X_L$	= Imaginary part of load impedance, in $\Omega$
$Z$	= Antenna input impedance, in $\Omega$
$Z_L$	= Load input impedance, in $\Omega$
$Z_o$	= Characteristic impedance, in $\Omega$
$\Gamma$	= Voltage reflection coefficient.
$\epsilon_o$	= Permittivity of free space = $(\mu_o c^2)^{-1} \cong 8.854 \cdot 10^{-12}$ F/m $\cong (36\pi(10^9))^{-1}$ F/m.
$\epsilon_r$	= Real part of ground permittivity relative to that of vacuum.
$\eta$	= Radiation efficiency of an antenna, in percent.
$\eta_o$	= Intrinsic wave impedance of free space = $\sqrt{\mu_o/e_o}$ = $4\pi c \cdot 10^{-7}$ = $376.7303$ ( $\cong 120\pi$ ) $\Omega$
$\lambda$	= Wavelength, in m.
$\mu_o$	= Permeability of free space = $4\pi \cdot 10^{-7}$ H/m.
$\sigma$	= Real part of ground conductivity, in S/m.



- $\tau$  = Pulse duration, in s.  
 $\theta$  Angle between  $L_{\text{eff}}$ , the (vector) effective antenna length and  $\vec{E}_o$ , the (vector) electric incident field.

Unless otherwise stated, units are assumed to be those of the International System (SI). For convenience, several relationships and prefixes are tabulated below.

1 in = 0.0254 m			
1 mi = 1609.34 m			
1 m = 0.000621371 mi			
1 in <sup>2</sup> = 0.00064516 m <sup>2</sup>			
1 m <sup>2</sup> = 1550.003 in <sup>2</sup>			
pico (p)	= 10 <sup>-12</sup>	deka (da)	= 10
nano (n)	= 10 <sup>-9</sup>	hecto (h)	= 10 <sup>2</sup>
micro ( $\mu$ )	= 10 <sup>-6</sup>	kilo (k)	= 10 <sup>3</sup>
milli (m)	= 10 <sup>-3</sup>	mega (M)	= 10 <sup>6</sup>
centi (c)	= 10 <sup>-2</sup>	giga (G)	= 10 <sup>9</sup>
deci (d)	= 10 <sup>-1</sup>	tera (T)	= 10 <sup>12</sup>

## 1.3 Terms Used in This Document

### 1.3.1 Field Strength of an Electromagnetic Wave

A general term that usually means the magnitude of the electric field vector (in V/m) at a point in the field, but which may mean the magnitude of the magnetic field vector (in A/m). Field strength has sometimes been called field intensity, but this usage is now deprecated since intensity connotes power in optics and in radiation. At frequencies above about 100 MHz, field strength in the far field is sometimes identified with power density,  $S$ . For a linearly polarized plane wave in free space,  $S = E^2/\eta_0 = E^2/(\mu_0/\epsilon_0)^{1/2}$ , where  $E$  is the electric field strength and  $\mu_0$  and  $\epsilon_0$  are the permeability and permittivity of free space, respectively. When  $S$  is expressed in W/m<sup>2</sup> and  $E$  in V/m, the denominator is often rounded off to  $120\pi$  or  $377 \Omega$ .

### 1.3.2 Accuracy

The degree of freedom from error, that is, conformity to the correct value. Accuracy is distinguished from precision, which is defined as the repeatability of measurement data and is usually quantified as the standard deviation of that data. The accuracy of an indicated value is expressed by the ratio of the error of the indicated value to the true value. It is usually expressed in  $\pm \%$  or  $\pm$  dB. Since the true value cannot be determined exactly, the measured or calculated value of highest available accuracy is taken to be the true value or reference value. Hence, when a meter is calibrated in a given echelon, the measurement made on a meter of a higher accuracy echelon usually will be used as the reference value. Comparison of results obtained by independent measurement procedures is often useful in establishing the true value and its precision.

### 1.3.3 Echelon

A specific level of accuracy of calibration in a series of levels, the highest of which is represented by an accepted national standard. There may be one or more auxiliary levels between two successive echelons. The classification of accuracy levels into echelons is shown in Section 4

### 1.3.4 Calibration Factor

The factor that, at a given frequency, expresses the relationship between the field strength of an electromagnetic wave impinging upon the antenna and the indication of the field strength meter. It is a function of the antenna gain and impedance, receiver linearity, etc.

### 1.3.5 Antenna Factor

That portion of the calibration factor of a field strength meter that is determined by the characteristics of the antenna. It is usually defined as the ratio of the field strength to the voltage developed across a specified receiver resistance, such as 50  $\Omega$ . It is often expressed in dB, as in the following equation for the electric-field antenna factor ( $K$ ) for the special case of an antenna connected to a 50  $\Omega$  receiver ( $R_L = 50 \Omega$ ).

$$K = 20 \log_{10} \left( \frac{E \text{ in V/m}}{V \text{ across } 50 \Omega} \right), \text{ in dB (m}^{-1}\text{)} \quad (1)$$

## 1.4 Complications That are Avoided

This report is limited to the art of measuring field strength of steady-state sinusoidal continuous-wave emissions. Signals that must be measured in practice are not usually in this limited class because they have time variations such as modulation or fading. Furthermore, some signals (such as impulsive interference) may occupy a band of frequencies far wider than the bandwidth of the measuring instrument. Nonsteady-state, nonsinusoidal signals will not be discussed here, except to mention that they may often be measured with instruments calibrated under continuous-wave (CW) conditions, taking into account certain additional factors depending on the behavior of the measuring equipment (see [1] and [2])<sup>1</sup>, or with instruments calibrated under special conditions. Care must be exercised when measuring fields that are best described as random variables, for significant errors can occur when such measurements are made using detectors calibrated with CW fields (see 1.5.2).

An electromagnetic field often has a complex structure as a result of the existence of more than one propagation path. Components may arrive at the measuring location from several directions, and they may have different polarizations. The resultant field strength varies with position and may vary in amplitude and phase across the dimensions of the receiving antenna. The indication of a field strength measuring instrument in such a case may not be unique because its calibration is obtained in terms of a plane wave. To avoid confusion, it is necessary to add to the measurement report a statement giving the type, polarization, and size of the receiving antenna, its orientation, and the possible existence of multipath propagation.

Similar considerations apply to measurements made in the near field of a source or reflector of the field. In such regions, the field may be nonuniform within the dimensions of the antenna. The field strength can then be considered to be the indication of a measuring receiver with a stated type and size of antenna when the equipment has been calibrated under conditions equivalent to plane-wave free-space propagation. For measurements related to a particular communication service, it is appropriate to use an antenna typical of those normally used in that service.

## 1.5 Equipment for Measurement of Field Strength

Field strength measurement is accomplished through the use of a combination of the following elements (see Fig 1):

<sup>1</sup>The numbers in brackets correspond to those of the bibliographies in Section 6

- 1) An antenna
- 2) A transmission line, optical link, or other coupling networks
- 3) Frequency-selective circuits
- 4) Amplifying and attenuating circuits
- 5) A detector of known input/output characteristics and sufficient dynamic range
- 6) An indicating device such as a panel meter, chart recorder, cathode ray tube, or digital readout

In addition, unless the amplifying and indicating devices are sufficiently stable in gain and sensitivity to be relied upon throughout the time interval between calibrations, it is necessary to provide a signal source having the required time stability in output level to serve as a reference by which the sensitivity of the equipment may be adjusted to a standard condition. Typically, this signal source is a CW signal generator, an impulse generator, or a random-noise generator.

Measurement of field strength is sometimes performed with an assembly of individual elements such as an antenna, receiver, recorder, and standard signal generator. These measurements, however, are facilitated by the use of a “field strength meter,” a name given to a specialized portable unit containing all the necessary elements, including a reference-level signal source and sometimes a built-in antenna.

## 1.6 Factors Affecting the Units Used for Reporting the Results of Field Strength Measurements

The components of a basic receiver for measuring field strength affect the quantities measured and the units applied to these quantities. The receivers used for field strength meters (FSMs), less properly called field intensity meters (FIMs), in use today employ the following five basic elements:

- 1) Antenna and related circuits (e.g., baluns, transmission lines)
- 2) Receiver (rf and IF)
- 3) Detector
- 4) Metering or data processing device
- 5) Calibrator (which may be internal or external to the receiver, see Fig 1)

The antenna and receiver (through the IF amplifier) select the signal spatially and spectrally, respectively, and provide amplification. In the case of a relatively narrow-band antenna and wide-band receiver, the antenna may also contribute to the spectral filtering process. The antenna polarization, of course, is also important. The units used to describe the results of a given measurement are determined primarily by the calibrator, the detector, and the meter. The antenna and receiver characteristics, however, are often used to normalize the data in a way that directly affects the units used to report the results. These normalizations will be considered first.

The antenna normalization, which relates electromagnetic field quantities and their units to circuit quantities and their units, involves converting a known (measured) circuit quantity (e.g., voltage) to a field quantity (e.g., electric field strength) by means of a known antenna factor (see 1.3.5). Hence, a standard (or calibrated) antenna must be employed in order to be able to properly perform this normalization that relates the received voltage in V to an incident electric field in V/m. Other units for field strength include  $\mu\text{V/m}$  and dB above 1  $\mu\text{V/m}$ , often written as dB (1  $\mu\text{V/m}$ ). If the antenna effective area is used to normalize a receiver power, typical units for the power density would be  $\text{W/m}^2$  and dB above 1  $\text{W/m}^2$ .

The receiver normalization involves the effective bandwidth when the spectral range of the wave whose field strength is being measured exceeds the receiver's bandwidth. The measured voltage (or power) is observed in some bandwidth, and results may be normalized in terms of a unit bandwidth (1 Hz, 1 kHz, or 1 MHz). This approach is relatively straightforward for white noise (i.e., noise with a constant power spectral density over the frequency band of interest) and the normalization of average noise power or rms voltage, but it can lead to problems of misinterpretation or error when used in measuring the field strength of signals with nonwhite power spectra. Normalization with respect to bandwidth can be applied to both electromagnetic field and circuit quantities and leads to units such as  $\text{W}\cdot\text{m}^{-2}\cdot\text{Hz}^{-1}$  and dB above 1  $\mu\text{V/MHz}$ .

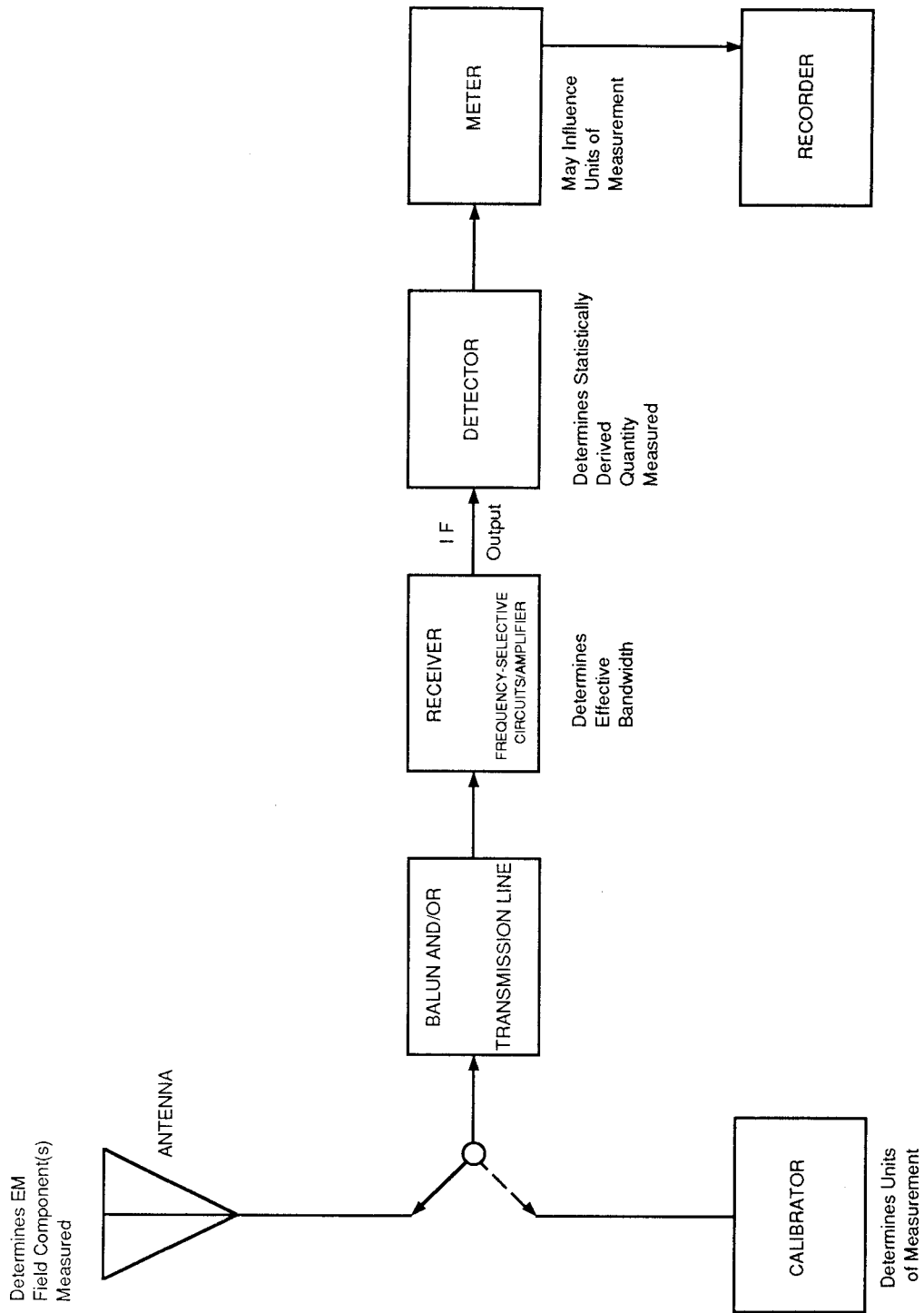


Figure 1—Measurement System Functional Block Diagram

Care must be taken to select the appropriate bandwidth for normalization (e.g., effective noise-power bandwidth,  $b$ , or impulse bandwidth,  $b_i$ ). It is well known that the peak response at the output of a narrowband linear receiver to a noise impulse at the receiver input is proportional to the area under the voltage-time waveform of the impulse. This peak response is independent of the exact shape of the input waveform, provided the duration of the impulse is small relative to the inverse of the impulse bandwidth of the receiver,  $b_i$ . For example, if the output of a pulse generator is a single rectangular pulse of amplitude  $AV$  and duration  $\tau$  s, then the peak of the output waveform of the receiver can be expressed as

$$V(\text{peak}) = (k \cdot A \cdot \tau) b_i, \text{ in V} \quad (2)$$

where

$k$  is the constant of proportionality.

The impulse bandwidth can be computed from

$$\frac{1}{b_i} = \frac{1}{V(\text{peak})} \int_0^{\infty} V(t) dt \quad (3)$$

where

$V(t)$  is the envelope of the receiver's impulse response measured at the receiver IF output.

Notice that an input of twice the amplitude ( $2A$ ) lasting but half the duration ( $\tau/2$ ) would have produced an identical output. The observed quantity,  $V(\text{peak})$ , can give information only about the impulse strength in volt-seconds, but not information about the exact amplitude, duration, or shape of the impulse. For this reason, it is expedient to calibrate peak detectors intended for use in measuring nonoverlapping noise impulses with impulse calibrators. The output of the calibrator is given in volt-seconds, or dB ( $1 \mu\text{V}/\text{MHz}$ ).

The detector in Fig 1 is designed to produce the desired statistic of the noise envelope to be measured. This gives rise to the basic quantity to be metered (e.g., rms voltage, average voltage, quasi-peak voltage, or peak voltage). The units in which the statistics are measured depend upon the calibration source, the detector, and the meter (or data processor or both). As an example, consider rms-voltage, peak-voltage, and average-voltage detectors. The importance of the calibration waveform is illustrated in Table 1 (see [B65] ), which gives the error produced for several selected waveforms when rms, peak, and average detectors are calibrated with the rms value of a sine wave from a CW signal generator.

**Table 1—Examples of Errors When Using Average and Peak Meters Calibrated With the RMS Value From a Sine Generator to Observe Other Waveforms**

Waveform	RMS Meter Indicates	Rectified Average Meter (Calibrated in RMS) Indicates	Peak Meter (Calibrated in RMS) Indicates	Error (dB)	
				Average Meter	Peak Meter
Sine wave	0.707	0.707	0.707	0.00	0.00
Sine wave plus 100% Third Harmonic					
In phase	1.000	0.944	1.09	-0.50	0.75
Out of phase	1.000	0.472	0.382	-6.52	-8.36
Square wave	1.000	1.111	0.707	+0.91	-3.00
Gaussian noise	1.000	0.887	—	-1.05	—
Pulse train					
Duty cycle = 0.1	0.318	0.111	0.707	-9.14	+6.10
Duty cycle = 0.01	0.100	0.011	0.707	-19.17	+16.99

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## 1.7 Division into Frequency Ranges

It is convenient to classify techniques for measurement of field strength into three approximate frequency ranges as follows:

- 1) Frequencies below 30 MHz
- 2) Frequencies from 30 MHz to 1 GHz
- 3) Frequencies from 1 GHz to 30 GHz
- 4) Frequencies above 30 GHz (beyond the scope of this standard)

This division is useful because the optimum technique is determined by the dimensions of practical antennas, the wavelengths of signals to be measured, and the effects of the terrain at and near the measurement location.

Below about 30 MHz, practical antennas are usually small compared to one wavelength. A common measuring antenna is a loop having one or more turns, often electrically shielded, with a diameter of 0.3 to 1 m. Another type is the short monopole (rod or whip) antenna with a length of 0.1 to 3 m. Both of these antenna types are electrically small and have low efficiency. That is, the radiation resistance is low compared with the loss resistance.

Also, below 30 MHz, it is usually necessary to measure field strength at heights that are electrically near the earth. The characteristics of the ground and nearby vegetation, conductors, and structures affect differently the polarization, angle of arrival, and the strength of the electric and magnetic components of the field. They may also influence the antenna input impedance. Measurements using electrically-shielded loop antennas are usually influenced less by nearby objects than those using rod antennas.

From about 30 MHz to 1 GHz, practical antennas have dimensions that are comparable to one wavelength. A common antenna for field-strength measurements in this range is a half-wave ( $\lambda/2$ ) resonant dipole. The dipole is usually connected to the measuring instrument by use of a balanced-to-unbalanced transformer (balun) and a coaxial transmission line. A  $\lambda/2$  dipole differs from the loop and rod antennas in that it is more efficient. Broadband antennas or highly-directional antennas are sometimes used in the upper portion of this frequency range.

Above about 1 GHz, the collecting area of a dipole becomes too small to provide the desired sensitivity. At these frequencies, it is common to employ antennas collecting energy from apertures that are large compared with one wavelength, such as horns. These antennas are usually characterized by high efficiency and appreciable directivity. Coaxial or waveguide transmission lines are usually employed.

## 1.8 Calibration Procedures and Standards

When a field-strength measuring instrument is adjusted to standard sensitivity by using a reference source, the reading of the indicating device is converted to field strength by application of a calibration factor that usually varies with frequency. Typical general-purpose field strength meters have several continuously-tuned frequency ranges and an amplitude range of 80 to 120 dB. The calibration of such an instrument at only a modest sampling of frequencies and signal levels may require a large number of measurements.

Calibration methods may be classified into the following two basic types:

- 1) *Direct Calibration.* The antenna is exposed to an electromagnetic field, the strength of which is accurately known, see [B77] , [B93] . The field strength may be known by computation from the measured current in, or power delivered to, a transmitting antenna of known dimensions and current distribution or gain. In this approach, known as the standard-field method, the effects of ground reflections or reflections from the walls of an anechoic chamber (if present) must be accounted for. This can be achieved by averaging the in-phase and out-of-phase reflections (see [B69] ) as the separation distance between the transmitting and receiving antennas is varied. Alternatively, the field strength may be determined from the voltage induced in (or power available from) a special receiving antenna of known dimensions or gain. This latter approach is called the standard-antenna method. In principle, a field-strength meter may be directly calibrated by the use of known field strengths at all levels and frequencies in its range. For practical reasons, however, such calibrations are usually performed at only one or two levels at each required frequency. These measurements are supplemented by measurements of attenuator step accuracy, dial scale accuracy, etc.
- 2) *Indirect Calibration.* In indirect calibrations, the calibration factor is calculated from the computed or measured characteristics of the receiving antenna and measured instrument characteristics. The antenna is removed from the field-strength measuring equipment and replaced by a calibrated standard-signal generator having an impedance equal to that of the antenna. The equipment is calibrated as an rf voltmeter (or power meter) against the standard signal generator. An antenna factor is computed for each frequency from the dimensions and current distribution of the antenna, from consideration of the antenna as an aperture, or from the measured gain of the antenna. If a transmission line is used between the antenna and receiver, it is possible to consider it as part of the receiver and connect the calibrating generator to it, thus avoiding the need for separately determining the transmission line loss and compensating for such loss.

The theoretical computation of the receiving antenna factor is facilitated by the use of simple types of antennas. For example, a thin, short, vertical-monopole antenna (shorter than 0.1 wavelength) located over an extensive ground plane may be assumed to have a linear current distribution, making its effective length equal to half its physical length. Its impedance can be simulated approximately by a series capacitor connected between the standard signal generator and the input of the measuring receiver. A second example is the thin half-wave dipole antenna with assumed sinusoidal current distribution. This antenna has a theoretical effective length of  $\lambda/\pi$  and a radiation resistance of approximately  $70 \Omega$  in free space.

A standard thin cylindrical dipole must have a length several percent shorter than a half wavelength in order to achieve self-resonance. Its radiation resistance is then less than that of an infinitesimally thin  $\lambda/2$  antenna. This is due to the effect of finite cylinder thickness on the current distribution. The directional pattern of a practical dipole, however, does not differ much from that of an infinitesimally thin dipole, and thus its gain and available power remain close to the theoretical values. A practical dipole may therefore be considered essentially equivalent to a theoretical thin dipole plus a transformer to account for the change in radiation resistance. A balun is an additional transformer that may introduce error unless its impedance-matching characteristics are optimized and its loss taken into account, see [B75] .

The gain of an antenna may be measured by comparison with a standard-gain antenna that has been calibrated previously or whose gain can be calculated theoretically. The gain also may be determined from measurements using two “identical” antennas, one for transmitting and the other for receiving, see [B11]. Similar to a two-antenna gain calibration, the more general three-antenna technique can be used. In this case, it is not necessary to have two identical antennas. The gains of all three antennas are obtained from three sets of measurements, each involving a pair of two antennas, see [B69]. Measurements of antenna gain should include the loss of any associated balun, isolator, or impedance-matching devices that are used.

Calibrated-signal generators, attenuators, power meters, and thermoelements for measurement of rf current in terms of direct-current standards are the principal reference standards used in the calibration of instruments for measuring field strength. Accurate scales are used to determine antenna separation distance and the dimensions of standard-field transmitting loops, aperture antennas, and standard receiving dipoles. Other calibrated standard-gain transmitting antennas also may be required.

## 2. Methods of Measuring Field Strength

### 2.1 General

Two general methods are applicable to the measurement of field strength, see [B39], [B85] and [B93]. One method consists of measuring the received power or open-circuit voltage induced in a standard receiving antenna by the EM field to be measured, and then computing the field strength in terms of the measured power or voltage and the dimensions and form of the standard receiving antenna. The other method consists of comparing voltages produced in an antenna by the field to be measured and by a standard field, the magnitude of which is computed from the type and dimensions of the transmitting antenna, the net power delivered to the transmitting antenna (or its current distribution), the antenna separation distance, and the effect of the ground. For the standard receiving antenna method, there are special requirements for the antenna and the power or voltage measuring equipment. Field-strength measurements are often made using commercially available meters that have been calibrated in a known field determined by either of the two above methods. A calibration service is maintained by the National Institute of Standards and Technology for field-strength meters in the frequency range of 30 Hz to 30 GHz. For information, write to the National Institute of Standards and Technology, Electromagnetic Fields Division, 325 Broadway, Boulder, CO 80303, USA.

#### 2.1.1 Standard Receiving Antenna Methods

In the standard receiving antenna method (see [B77] and [B85]), the antenna is of some standard form such as a thin dipole or a pyramidal horn (see [B49]) of known gain. The open-circuit rf voltage,  $V_{oc}$ , induced in a dipole by an incident field,  $E_{inc}$ , of given strength and polarization can be calculated theoretically, see [B77]. The ratio of this voltage to the component of electric field producing it is called the effective length,  $L_{eff}$ , of the antenna. If the direction of polarization of  $E_{inc}$  is parallel to the dipole,

$$L_{eff} = V_{oc}/E_{inc} \quad (4)$$

Effective length is usually defined as a vector such that  $\vec{V}_{oc} = \vec{L}_{eff} \cdot \vec{E}_{inc}$ . The scalar,  $L_{eff}$ , is equal to the magnitude of  $L_{eff}$  times the cosine of  $\theta$ , where  $\theta$  is the angle between  $L_{eff}$  and  $E_{inc}$ , the incident vector field.

The method of coupling a standard dipole to the voltage-measuring instrument is generally such that the voltage measured is not the total voltage induced by the field but is a fixed fraction of it. For example, the ratio of the detected dc voltage to the rf voltage induced in the antenna by the field is called the voltage transfer ratio. A separate determination of this ratio is usually required, see [B19].



A standard voltage source is often provided as part of the field-strength measuring apparatus for calibrating the voltmeter. The term “voltmeter” is used here for the radio receiving portion of the overall instrumentation. The calibrating voltage may be fixed or adjustable over a wide range. It may be inserted in series with the standard antenna or at some point in the coupling circuit, or it may be applied to the input terminals of the voltmeter in place of the coupling circuit. The voltmeter is calibrated at various levels corresponding to the voltages being measured. The method used to determine the voltage-transfer ratio generally depends on the particular form of antenna and circuit arrangement employed.

The field-strength value is calculated on the basis of the antenna gain or its effective length and voltage-transfer ratio. In some instances, it is difficult to calculate accurately the effective length of an antenna being used. In such cases, the product of the effective length and voltage-transfer ratio can be determined directly from a measurement of the output voltage of the antenna across a specified load impedance when it is placed in a standard field. The reciprocal of this product is called the antenna factor. This is the procedure usually followed in the calibration of commercial field-strength meters. The value of the antenna factor includes the gain of the receiver.

The form of standard receiving antenna that may be used is governed by the signal frequency, by the polarization of the field component to be measured, by the nature of the field (i.e., high or low strength, fading or steady, etc.), and by the ease of handling the antenna, see [B23] , [B24] , [B38] , [B68] , [B77] and [B78] . The types generally employed are the loop antenna, the vertical or horizontal dipole having a length of  $\lambda/2$  or less, a vertical monopole with output between its lower end and ground, a pyramidal horn, and an open-ended waveguide (OEG). Other types of antennas can also be used, in which case they are calibrated in terms of one of the simpler antennas just mentioned.

Measurements of field strength in radio communication services are often made with complex types of antennas and arrays. They generally offer the advantages of operating convenience, increased sensitivity, and interference discrimination. The complex antenna can be calibrated by comparison with a simple standard antenna. This calibration, however, may be in error because of differences in the directivity of the antenna under test and the simpler comparison antenna. For example, a standard dipole will not discriminate as effectively against a ground-reflected ray as will a high-gain log periodic antenna.

### 2.1.2 Standard Transmitting Antenna Methods

In this method a standard (known) field is produced by a local transmitter. The field at the receiving antenna location is computed from the gain or effective length of the transmitting antenna, its radiated power or current distribution, the distance from the transmitting antenna to the field point, and the effect of reflections from the ground, chamber surfaces, or other objects, see [B39] and [B85] . The measuring equipment associated with the receiving antenna generally consists of a power meter or a frequency-selective voltmeter, such as a spectrum analyzer or an interference meter (receiver). When the angles of arrival of the transmitted standard and measured fields differ, a correction must be applied for any resultant difference in the voltages caused by directivity of the antennas and effects of the ground.

### 2.1.3 Simplifications and Precautions

The problem of field-strength measurement is generally complicated by the complex nature of the field resulting from variability of the transmission medium (see [B11], [B19] , [B69] , and [B75] ), by the nature of the emission (see [B21] and [B24] ), and by the influence of the ground or disturbing structures, see [B24] , [B38] , and [B87] -[B89] . Simplifying approximations are introduced into measurement methods to keep the instrumentation from becoming unduly complex and to facilitate analysis of the data.

In the case of ionospheric or tropospheric wave fields, a complete measurement would involve determination of

- 1) The field strength of the vertical, longitudinal, and lateral electric and magnetic field components for each signal frequency at the receiving point
- 2) The phase angles between the different field components
- 3) The direction of arrival of each signal component

Fortunately, considerable practical information can be obtained by measuring only three mutually-orthogonal components of the field, see [B23] . Simple forms of receiving antennas are especially useful, employing (for example) three crossed dipoles or three crossed loops to measure the “total magnitude” of the electric or magnetic field strength.

A simplification in the measurement of complex emissions may often be achieved by removing the transmitter's modulation so that a single-frequency field (the carrier) is measured. In emissions without a carrier wave, such as a single-sideband transmission, it is sometimes desirable to restrict the modulation to a single frequency so that a single-frequency field is measured. Occasions arise, however, when it is necessary to retain the normal modulation, such as when considering the interfering effect of cross-modulation products or single-sideband emissions. When measuring the harmonic content of a given field, special precautions should be taken to avoid errors caused by harmonics and other spurious responses that may be produced within the measuring receiver.

A typical field-strength measuring system is a frequency-selective voltmeter designed to indicate the average, peak, and/or quasi-peak value over a chosen time interval of the voltage of a signal that has passed through the receiver's frequency-selective circuits. For single-frequency fields, the RMS voltage value is measured, usually with an average-indicating circuit calibrated by a single-frequency signal of known RMS voltage. For pulse-modulated signals, the value during the pulse on-time is usually measured by means of a peak-indicating circuit calibrated with an unmodulated single-frequency signal of known RMS value. In this case, the necessary receiver bandwidth and the optimum charge and discharge time constants of the indicating circuit depend on the pulse duration and repetition rate, see [B1] , [B2] , [B4] , [B5] , [B9] , [B71] , and [B72] . Calibrated oscilloscopes are also employed to determined pulse-peak values.

In the case of frequency-modulated waves, the effect of modulation on the measured field-strength value may be essentially eliminated if the receiver has an overall passband slightly greater than twice the maximum frequency deviation. When making field strength measurements of television video transmissions, the RMS value of the maximum levels that occur during synchronizing pulses is usually measured, see [B32] , [B36] , and [B48] . The time constant of the second detector and the overall receiver bandwidth should be such that the meter essentially will be peak reading. If it indicates at least 85% of the peak levels, the reading will be essentially independent of the modulation content. A measurement of the RMS value of the video carrier can be made but, because the result depends on the modulation percentage, this measurement is seldom used. Measurement of the sound carrier of a television station is more easily accomplished because of the narrower receiver bandwidth required and the lack of amplitude variation with modulation. The latter method often yields sufficient information for field survey purposes, provided that the radiation pattern of the transmitting antenna is essentially the same for the video-carrier and sound-carrier frequencies.

In analyzing or reporting measured data, it is important to consider the influence of the receiving site and surrounding topography, electrical properties of the ground (see [B38] , [B87] -[B89] ), the proximity of any disturbing structures (see [B24] ), and the orientation and height above ground of the receiving antenna. The true measurement of an incident field is often an indirect measurement involving a knowledge of the angle of arrival and electrical properties of the ground.

In the case of ionospheric waves at frequencies above 3 MHz, a useful expedient is to use a horizontal receiving dipole for measuring field strength (see [B23] , [B24] , and [B32] ) regardless of whether vertical or horizontal polarization is transmitted. Similar values of each polarization are often present in the ionospheric field incident upon the receiving antenna for any type of polarization at the transmitting antenna. The advantage of measuring only the horizontal component is that the measured field strength is practically independent of differences in the electrical properties of the ground at different receiving points. Hence, as long as the height of the receiving antenna above ground is stated, measurements at different sites (at a given frequency) may be directly compared.

In some cases, even though the vertical component of the electric field is greatly influenced by the nature of the ground, it should nevertheless be the field component measured since it provides a more accurate picture of the actual conditions encountered in service. This is particularly true for measurements in the AM broadcast band (540 to 1600 kHz) where the antenna generally responds mainly to the vertical component of the  $E$  field.

When a shielded loop antenna is employed for measuring field strength, it is the magnetic field component ( $H$ ) that is being measured. Unfortunately, the results are often erroneously expressed in V/m of an electric field ( $E$ ) by means of the following relation, which does not necessarily apply except for a linearly polarized plane wave in free space:

$$E = \eta_0 H \quad (5)$$

where

$$\begin{aligned} \eta_0 &= \text{impedance of free space, equal to approximately } 377 \, \Omega \\ H &= \text{magnetic field strength in A/m} \end{aligned}$$

It should be emphasized that the above relationship is not necessarily valid near a discontinuity such as the surface of the earth or near other reflecting surfaces. Under some conditions, standing waves may exist where the field is composed of two or more traveling waves for which Eq 5 also is not valid. Consequently, inconsistent results may be obtained if measurements of field strength are made using both a loop antenna and a vertical monopole or a horizontal (or vertical) dipole. In cases where the magnitude of the electric field is desired, and Eq 5 is not applicable, a calibrated  $E$  field antenna should be used with appropriate correction for proximity of the ground.

## 2.2 Field Strength Calibrations Using Loop Antennas, 30 Hz to 30 MHz

### 2.2.1 Standard Receiving Loop

A loop antenna may be used to measure magnetic field strength from about 30 Hz up to about 30 MHz. Below about 3 MHz, especially for low-arrival angles, the loop antenna has the advantage of convenience in handling, availability of a zero-response orientation, and a simple expression for the voltage produced by a given field strength. At higher frequencies, above about 100 MHz, a loop of practical size may no longer be electrically small. The loop then has a significant response to the electric field as well as the magnetic field, see [B91]. The characteristics expected for an electrically small loop are no longer present (e.g., a figure-eight field pattern with sharp nulls). Thus, a position of minimum response is ill-defined, and there is difficulty in defining the polarization for the field component being measured. Also, as the frequency is increased, there is difficulty in securing a balanced connection of the loop antenna to the voltage-measuring receiver.

The shielded-loop antenna has come into general use for direction finders and can be used for measuring field strength. It is desirable that the shield have a gap at the top so that the antenna is symmetrical with respect to ground. This also minimizes the response of the loop to a vertically-polarized electric field, see [B91].

Figs 2 and 3 illustrate methods of calibration for balanced and shielded unbalanced loops, respectively. These employ direct voltage substitution in which the rf voltage induced in the receiving loop is compared directly against a known rf voltage obtained from a standard signal generator. In this case the receiver is used merely as an uncalibrated comparison voltmeter. The loop antenna is oriented and the receiver is tuned for maximum response at the carrier frequency. The receiver gain is adjusted to produce a convenient indication and is left at this setting. The loop antenna is then turned until the receiver indication is a minimum. The standard signal generator is turned on and tuned to the frequency of the field to be measured. The signal generator voltage is adjusted to give the same indication as that obtained when receiving the unknown field. The signal-generator voltage then equals the induced voltage acting in series with the loop. If the signal-generator impedance is mismatched, or if resistors are added in series with the loop resistor to match the output, a special voltage calibration of the generator will be required.

When the self-resonant frequency of the loop antenna is several times the frequency of the field being measured and the electric and magnetic fields are related by Eq 5, the value of field strength  $E$  in V/m is

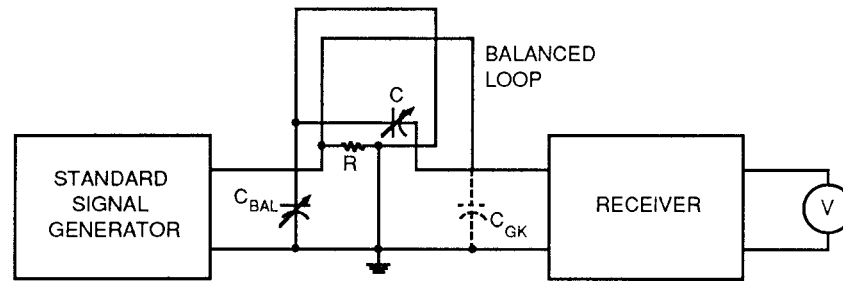


Figure 2—Direct-Substitution Method of Calibration for a Balanced Loop

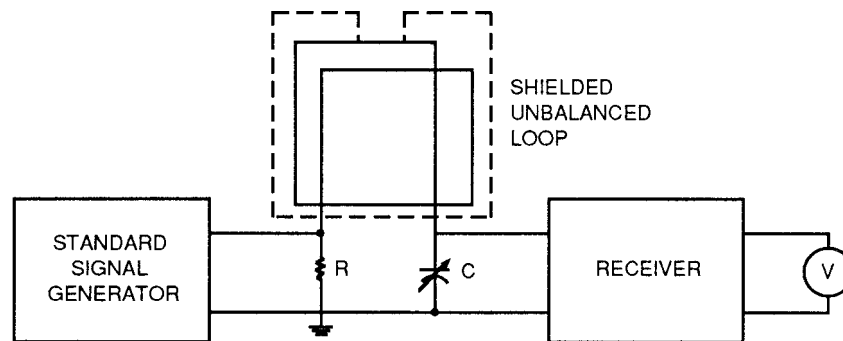


Figure 3—Direct-Substitution Method of Calibration for a Shielded Unbalanced Loop  
(The resistor  $R$  is of the order of  $1 \Omega$ )

$$E = \frac{V_{oc}}{L_{eff}} \quad (6)$$

where

$V_{oc}$  = output voltage of the standard signal generator across the loop resistance  $R$   
 $L_{eff}$  = effective length of the loop antenna, in m

$$= \left[ \frac{2\pi AN}{\lambda} \right] \left[ 1 - \left( \frac{\pi A}{2\lambda^2} \right) \right] \quad (7)$$

and  $A < 0.01\lambda^2$

where

$A$  = area of the loop in  $m^2$   
 $N$  = number of loop turns

The above method usually requires a signal generator having a low output impedance. The value of the resistor  $R_s$  in series with the loop antenna is limited to about  $1 \Omega$  so that it does not appreciably lower the  $Q$  of the loop circuit. This value of series resistance should remain the same when the receiver is tuned to the unknown field and when the standard voltage is injected for calibration purposes.

An error in calibration may occur due to the difference in voltage step-up or  $Q$  of the tuned loop antenna for a distributed induced voltage, compared with a lumped voltage inserted at one point. This error is caused by distributed capacitance in the loop antenna and is a function of the ratio of operating frequency  $f$  to the self-resonant frequency  $f_0$ .

of the antenna:  $f/f_0$ . The error may be greater than 15% for a tuned loop antenna operated near its resonant frequency. This and other sources of error in field-strength measuring equipment have been studied, see [B39]. For a balanced, unshielded loop antenna, the correction factor  $F$  that one must apply to the lumped voltage introduced at one end of the loop to get its true distributed effect is given by

$$F \cong 1 - \left(\frac{f}{f_0}\right)^2 \quad (8)$$

If the calibrating voltage is injected at the midpoint of the winding of a balanced loop, as is usually the case, the correction factor is given by

$$F \cong 1 + 0.27\left(\frac{f}{f_0}\right)^2 \quad (9)$$

It is possible to compensate partially for the distributed capacitance effect by operating the loop antenna unbalanced with respect to ground. In such a case, a shield must be employed to minimize the electric-dipole response of the loop to a vertically-polarized field, see [B91]. Also, the effect of distributed capacitance of the loop on the accuracy of field-strength measurements may in some instances be minimized by using an untuned loop, see [B27].

Another source of error is distortion of the field caused by the presence of the field-strength meter case (see [B24] and [B27]) or a vehicle, see [B47]. Such errors may be greater than 10%, but they can be reduced by mounting the antenna away from perturbing objects.

In another method of loop-antenna calibration, a calibrating voltage is applied to the input of the receiver and adjusted for equality with the voltage developed across the receiver input by the loop antenna immersed in an EM field. A measurement of the  $Q$  of the loop is required in order to calculate the voltage induced in the loop by the EM field. The field strength is then computed from induced voltage and effective length of the antenna.

The method of measuring the loop  $Q$  is designed to avoid errors arising from the distributed capacitance between turns of the loop. It utilizes a precision variable capacitor connected in parallel with the receiver tuning capacitor. The auxiliary capacitor is set at its mid-value for resonance of the loop antenna and is then adjusted on either side of resonance until the voltage across the capacitor drops to 0.707 of its value at resonance. If  $C_1$  and  $C_2$  are the two capacitance values of the auxiliary capacitor, the  $Q$  of the loop antenna is given by

$$Q \cong \frac{2C}{(C_2 - C_1)} \cong \frac{2C}{\Delta C} \quad (10)$$

where

$C$  represents the total capacitance at resonance.

Thus, if  $V$  is the total voltage across the tuning capacitor, the field strength is

$$E = \frac{V}{L_{\text{eff}}Q} = \frac{V\Delta C}{2L_{\text{eff}}C} \quad (11)$$

The error in determining the loop antenna  $Q$  is generally less than 5% if the self-resonant frequency of the loop is at least 2.5 times the operating frequency.

Many commercial field-strength meters employ a high-level calibrating voltage to standardize the overall gain of the receiver, see [B37]–[B39]. When so standardized, the factor  $k$  (in  $\text{m}^{-1}$ ), relating field strength to the meter indication of the set, is determined in terms of a standard field and the relation

$$k = \frac{E}{V_1} \quad (12)$$

where

$E$  = standard field strength, in V/m  
 $V_1$  = receiver indicated voltage

Recall from Eq 1 that  $K = 20 \log_{10} k$ , in dB ( $\text{m}^{-1}$ ). The meter indication  $V_1$  is roughly proportional to frequency for a given value of  $E$ . The calibration factor  $k$  is usually called the antenna factor or the antenna coefficient. It actually comprises the following terms:

$$k = \frac{1}{g' L_{\text{eff}} Q} \quad (13)$$

where

$g'$  is the standardized value of receiver gain.  
 $L_{\text{eff}}$  and  $Q$  are defined previously.

In this case, the product of the three terms  $g'$ ,  $L_{\text{eff}}$ , and  $Q$  can be determined from a single measurement of the receiver indication produced by a standard field.

Unknown values of field strength can then be measured with the calibrated receiver using the relation

$$E = k V_2 \quad (14)$$

where

$k$  is given in  $\text{m}^{-1}$  by Eq 12.  
 $V_2$  is the receiver indication produced by the unknown field  $E$ .

For commercial field-strength meters using direct voltage substitution in series with the loop antenna, the receiver is merely an uncalibrated comparison voltmeter. The receiver gain  $G$  and loop antenna  $Q$  do not enter directly into the measurement. The resulting calibration factor  $k$  then reduces to the reciprocal of the effective length

$$k = \frac{1}{L_{\text{eff}}} \quad (15)$$

where

$k$  can be evaluated in terms of a standard field using Eq 12 as before.

The factor  $V_1$  in Eq 12 is now the voltage of the standard signal generator and thus is the value of injected voltage in series with the loop antenna.

Unless the loop antenna is mounted above the field-strength meter case at a distance equal to its diameter, it may be necessary to specify the angular position of the loop relative to the case at the time of calibration. Otherwise additional errors up to  $\pm 10\%$  may occur in field-strength measurements using other relative loop orientations.

### 2.2.2 Standard Transmitting Loop

At frequencies up to about 100 MHz, a transmitting-loop antenna can establish a known free-space magnetic induction field at the location of a relatively close receiving loop. This technique minimizes the effect of reflections from the ground or nearby objects. While loop antennas are used for both transmitting and receiving, and the near-zone field is primarily magnetic, it is convenient to express the value of magnetic field strength in terms of the equivalent free-space electric component  $E$  that would exist in a far-zone radiation field. This relationship between  $E$  and  $H$  is given by Eq 5. The resulting value of the equivalent electric field for a single-turn transmitting loop averaged over the area of a receiving loop is given by Eq 16 (see [B37] and [B39]) for the case of a coaxial receiving loop, as shown in Fig 4.

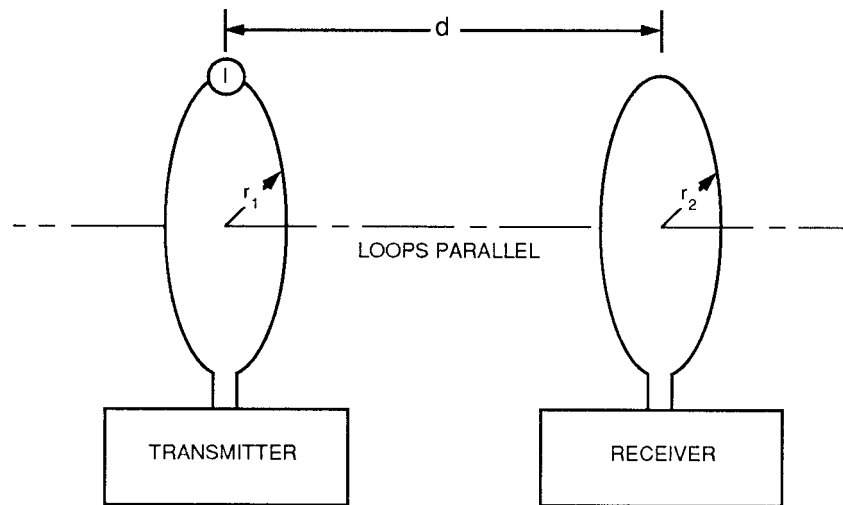


Figure 4—Induction-Field Method Using Coaxial Loops

$$E \cong \frac{60AI}{(R_o)^3} \left[ 1 + \left( \frac{2\pi R_o}{\lambda} \right)^2 \right]^{1/2} \quad (16)$$

where

- $R_o$  =  $[d^2 + r_1^2 + r_2^2]^{1/2}$ , in m
- $E$  = magnitude of the equivalent free-space RMS electric field strength, in V/m
- $A$  = area of the transmitting loop antenna, in  $m^2$
- $I$  = RMS current in the transmitting loop, assumed uniform around the loop, in A
- $\lambda$  = free-space wavelength, in m
- $d$  = distance between the centers of the transmitting and receiving loop antenna, in m
- $r_1$  = mean radius of the transmitting loop antenna, in m
- $r_2$  = mean radius of the receiving loop antenna, in m

In order to ensure an essentially uniform current in the transmitting-loop antenna, the circumference of the single-turn balanced circular loop should satisfy the inequality

$$2\pi r_1 < \frac{\lambda}{8} \quad (17)$$

In case a rectangular receiving loop antenna is used, the mean radius  $r_2$  is approximately that of the circle having the same area as the rectangle. An antenna separation distance  $d$  of 1–2 m is generally used. Since the induction field strength decreases essentially as the inverse cube of  $d$ , the distance to the nearest reflecting object need be only two or three times the antenna separation distance.

It can be seen that, except for the correction term  $(2\pi R_0 \lambda)^2$  under the radical, the induction field strength produced by a given loop is independent of frequency. For a separation distance of 1 m, this correction is negligible for frequencies below 5 MHz. For a radius  $r_1$  of 0.1 m, a spacing  $d$  of 1 m, and a current  $I$  of 0.1 A, the equivalent electric field  $E$  will be approximately 0.2 V/m at the receiving loop. The current  $I$  in the transmitting loop can be measured with a thermocouple ammeter. A VHF type of vacuum thermocouple may be built into the loop so that the lead lengths involved are short and do not appreciably alter the mean radius of the loop.

The dc output of the thermocouple may be measured with a suitable millivoltmeter and calibrated using direct current. This calibration will be valid at frequencies up to 100 MHz where the frequency error of such thermocouples in a balanced system is generally less than 1%. The standard induction-field method becomes impractical at frequencies above about 100 MHz because such a small loop is required to ensure a uniform current.

## 2.3 Field Strength Calibrations Using Dipole Antennas, 3 MHz to 1 GHz

The same general methods used for calibrating field-strength meters using loop antennas can be applied to sets with dipole-type antennas. These sets generally use a fixed high-level calibrating voltage to standardize the receiver gain. The calibration factor for the set can be determined by placing its dipole antenna at a specified height above ground in a standard horizontally-polarized radiation field, evaluated as described in 2.3.1 or 2.3.2. Eq 14 is used to measure an unknown field with the calibrated dipole. This calibration strictly applies only for the antenna height and ground constants prevailing at the time of calibration. Fortunately, the corrections involved are small for antenna heights exceeding about one wavelength, see [B38].

### 2.3.1 Standard Receiving Dipole

An electric dipole antenna is applicable at frequencies above about 3 MHz. Above 30 MHz, a self-resonant half-wave ( $\lambda/2$ ) dipole is easily portable. The methods of calibration described for a loop antenna may, in general, be applied to a thin  $\lambda/2$  dipole. A direct-substitution method of calibration (see [B34]) is illustrated schematically in Fig 5.

A calibrated signal generator is substituted for the dipole antenna at the end of the transmission line. The generator voltage is adjusted to produce the same receiver indication that is produced by the field being measured. The impedance at the signal generator should be equal to that of the dipole at the pertinent signal frequency. The voltage injected by the calibrated signal generator is equal to the voltage induced in the antenna by the field being measured. The measured value of field strength is given by

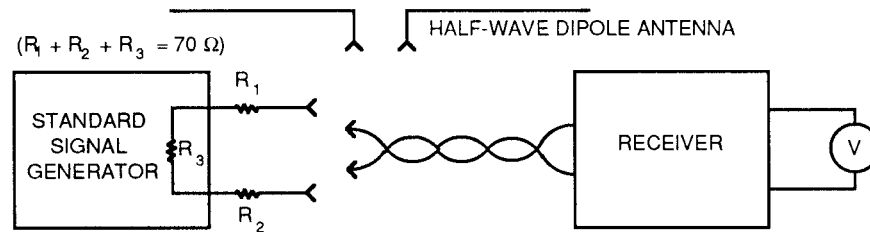
$$E = \frac{V_{oc}}{L_{eff}} \quad (18)$$

where

$$\begin{aligned} V_{oc} &= \text{open-circuit voltage at the output of the standard signal generator, in V} \\ L_{eff} &= \text{effective length of the dipole, in m} \end{aligned}$$



(The total resistance of  $R_1$ ,  $R_2$ , and  $R_3$  is of the order of  $70 \Omega$ .)



**Figure 5—Direct-Substitution Method of Calibration for a Thin Half-Wave Dipole.**  
(The total resistance of  $R_1$ ,  $R_2$ , and  $R_3$  is of the order of  $70 \Omega$ )

The maximum magnitude of the effective length of a thin half-wave dipole in free space is approximately  $\lambda/\pi$ . The physical length of a dipole at resonance is somewhat shorter than a half-wavelength, depending on its length-to-diameter ratio and on the amount of stray capacitance across its center terminals, see [B77]. The input impedance of a self-resonant dipole antenna in free space is resistive and will range from as low as  $50 \Omega$  for low values of length-to-diameter ratio, to as high as  $70 \Omega$  for very thin dipoles, see [B77]. Approximate values for both the antenna shortening and the radiation resistance as functions of length-to-diameter ratio are shown in Figs 6 and 7, respectively. The values should be checked experimentally for any dipole used to measure field strength. The effective length and gain of such an antenna will increase slightly as the length-to-diameter ratio is decreased.

The impedance of a dipole antenna is a function of its height above ground, the electrical properties of the ground, the conductivity of the wire, and its distance from any nearby reflecting surfaces or objects other than the ground. Consequently, the values given above will be in error for heights or distances less than about one wavelength, see [B38].

If the standard receiving dipole has a length that is shorter than that required for self resonance, the effective length is given approximately by

$$L_{\text{eff}} = \frac{\lambda}{\pi} \tan \frac{\pi L}{2\lambda} \quad (19)$$

where

$L$  is the overall physical length of the dipole antenna

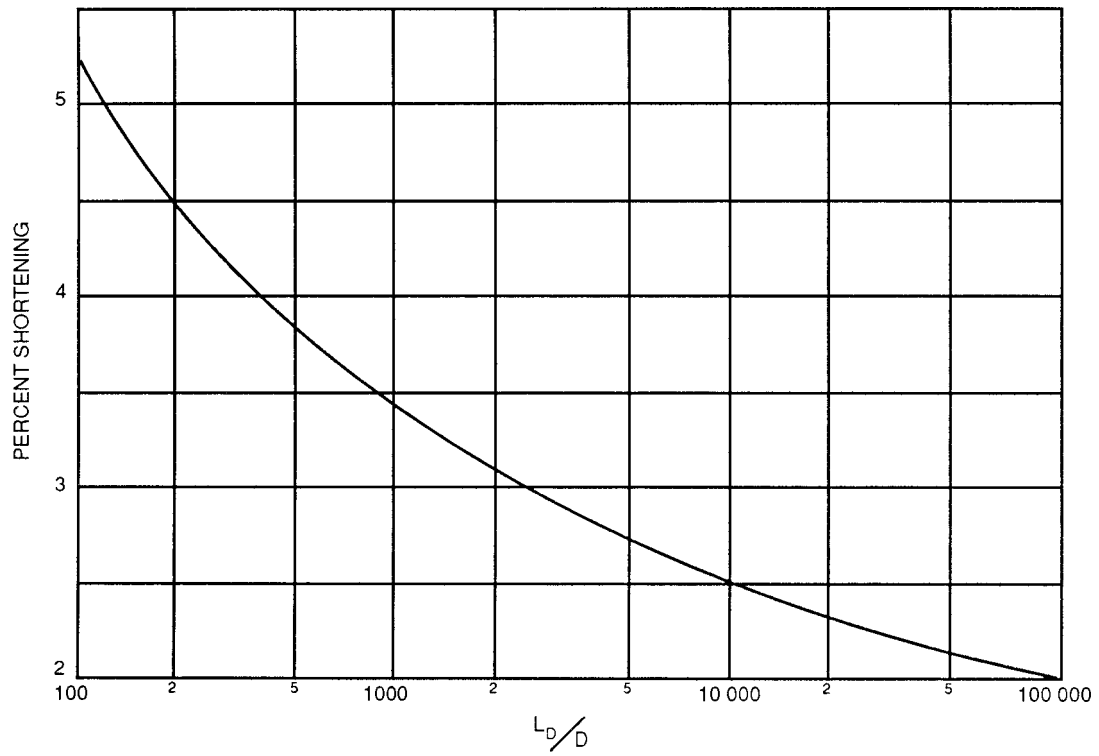


Figure 6—Percent of Shortening  $100(1-2L_D/\lambda)$  for a Thin Half-Wave Dipole at its First Resonance

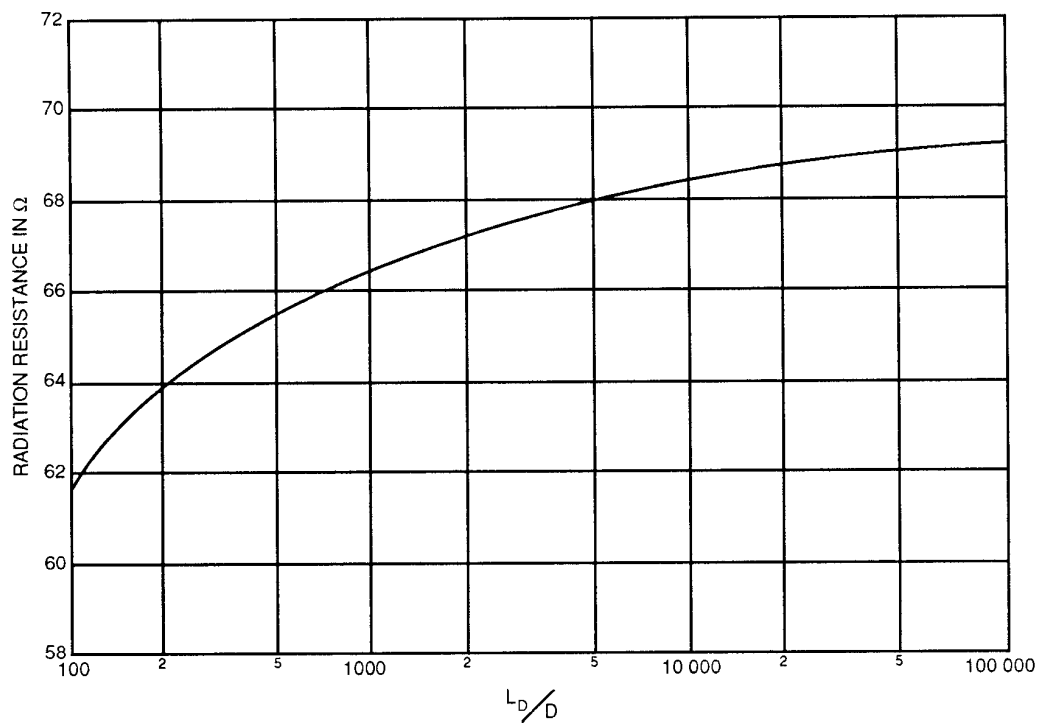
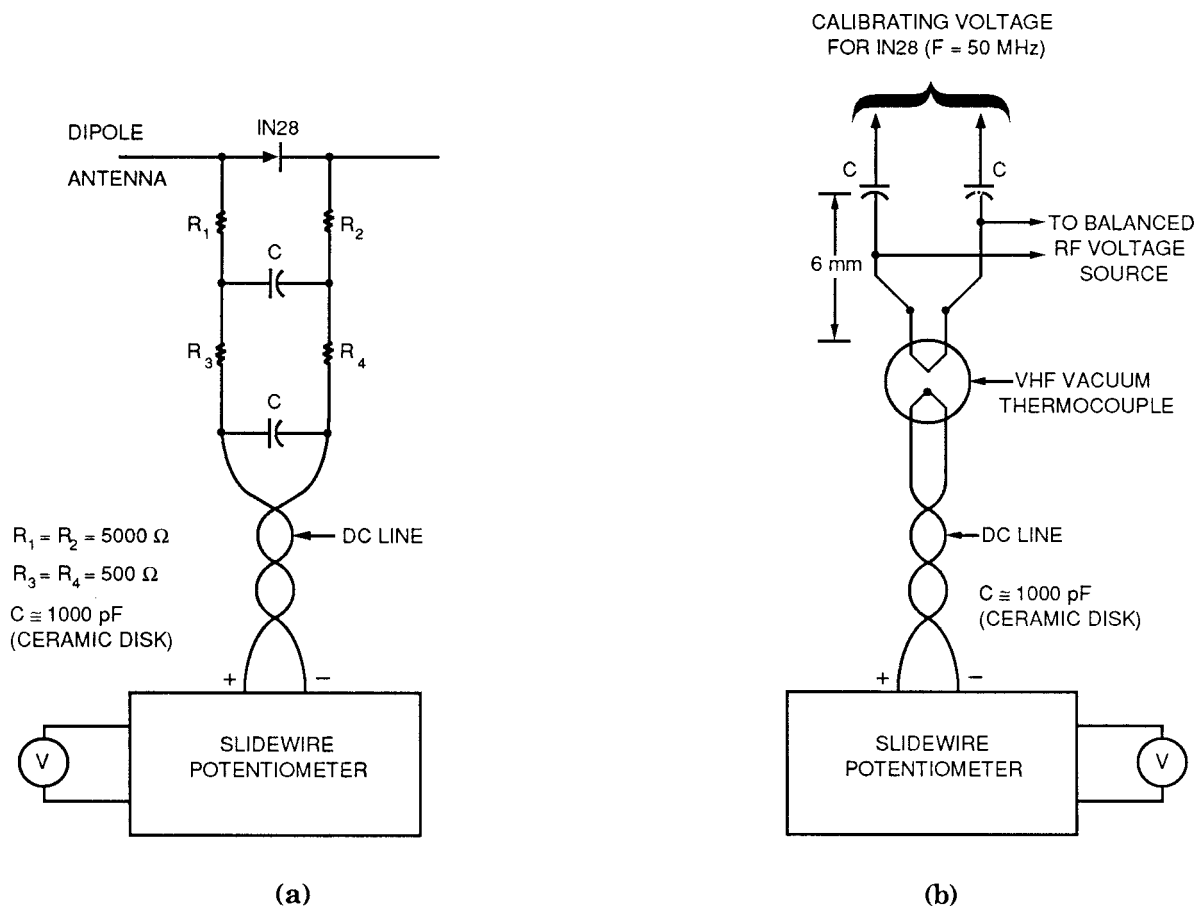


Figure 7—Radiation Resistance of a Thin Half-Wave Cylindrical Dipole at its First Resonance

When a receiving dipole is very short compared to a wavelength, the effective length is approximately equal to half the physical length.

Another method (see [B39]) that produces accurate field-strength measurements is shown in Fig 8. Here the open-circuit rf voltage induced in a standard  $\lambda/2$  dipole is measured directly by means of a high-impedance semiconductor diode detector circuit. This eliminates the necessity for knowing the impedance of the antenna and thus reduces the effect of the ground beneath the antenna on its voltage-transfer ratio. The rf-to-dc voltage-transfer ratio is also affected slightly by the ambient temperature of the diode, especially at low detected voltages corresponding to weak fields. The diode voltmeter thus has a good short-time stability but has the disadvantage of requiring daily recalibration for accurate results. Because of the lack of frequency selectivity in the receiving antenna circuit, it cannot be used in the presence of interfering signals whose fields are significant relative to the field being measured. This technique is best suited to measuring relatively strong locally-generated fields produced for calibrating electromagnetic interference (EMI) antennas and commercial field-strength meters. With proper precautions, it can be used to calibrate over a detected amplitude range of roughly 0.2-2 V dc with an accuracy of 1 dB or better. It can be used for either vertically or horizontally polarized fields, but the standard  $\lambda/2$  dipole is generally calibrated in a horizontal configuration.

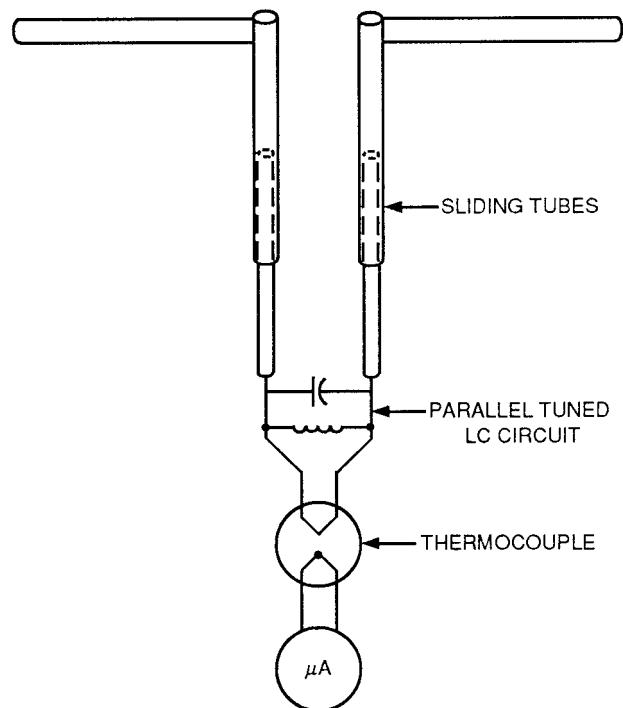


**Figure 8—Typical Calibration Circuit**  
(The parameter values shown are for illustrative purposes.)

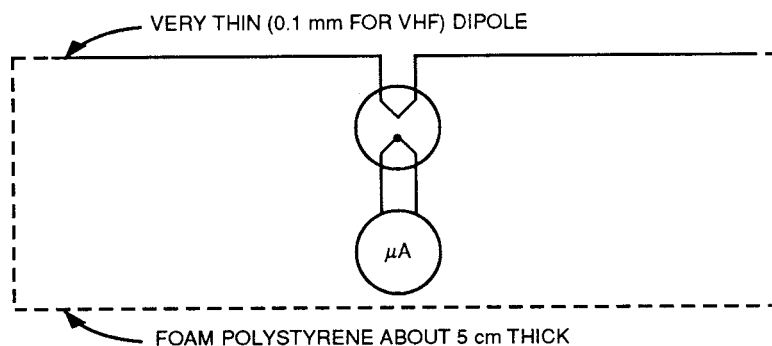
The rf diode voltmeter uses a low-capacitance (0.5 pF or less) silicon semiconductor diode. The diode is mounted in the gap at the center of the dipole. A balanced resistance-capacitance network is used, as shown in Fig 8, to filter the detected output and to prevent rf signals picked up on the dc transmission line from reaching the diode detector. The balanced transmission line can be made of conducting twinlead or of semiconducting plastic material, see [B39]. A dc

voltmeter with an impedance of about  $100\text{ M}\Omega$  is used to measure the output of the detector. It is preferably located at a distance of 3-10 m from the measuring dipole.

Details of the diode-calibration procedure are contained in [B39]. This calibration can be made at a frequency of 10 to 50 MHz where the error is small. Capacitors of the ceramic-disk type are used as shown in Figs 8(a) and (b). The rf-to-dc voltage calibration may then be used at signal frequencies up to 1 GHz where the diode frequency error has been found to be less than  $\pm 0.5\text{ dB}$ . Values of field strength are determined, as in the system described previously, by use of Eq 18.



**Figure 9—Matched-Load Field-Strength Meter**



**Figure 10—Field-Strength Meter With Thin Dipole Supported by Polystyrene Foam**

Two other forms of standard field-strength meters using dipoles are shown in Figs 9 and 10. Characteristics relative to these are given in [B60]. In Fig 10, the use of foamed polystyrene, which has a dielectric constant of about 1.02, makes it possible to support a very thin dipole and UHF-type thermocouple in such a manner that the thermocouple heater leads form a continuation of the antenna. The thermocouple lead is thus located at the center point of a thin straight

conductor, as required for reliable calculations. The shunt capacitance of the glass bead is less than 0.1 pF and the leads cause negligible inductance. However, care must be exercised to avoid rf pickup on the dc thermocouple leads.

It is possible to compute the gain and input impedance of a dipole using method-of-moment models for antennas over an imperfect earth (see [B16] -[B18] ) with known electrical properties, see [B40] .

### 2.3.2 Standard Transmitting Dipole

At frequencies above 3 MHz, it is practical to use a dipole for the transmitting antenna at a distance sufficiently large that the test antenna is in the radiation field, as shown in Fig 11. The procedure and calculations are simplified if conditions are such that only the direct wave and ground-reflected wave need be considered, and if the ground constants and angle of incidence are such that the reflection coefficient is very nearly equal to -1. This is generally the case for frequencies between 3 MHz and 1 GHz if horizontal polarization is used and if the antenna separation distance is large compared with the sum of the heights of the two antennas above ground (i.e., grazing incidence). A method for accurately measuring the transmitting dipole current using bead-type thermistors instead of thermocouples is given in [B39] . Procedures for determining the gain of half-wave transmitting dipoles at HF and VHF over imperfect ground have been developed, see [B30] and [B31] .

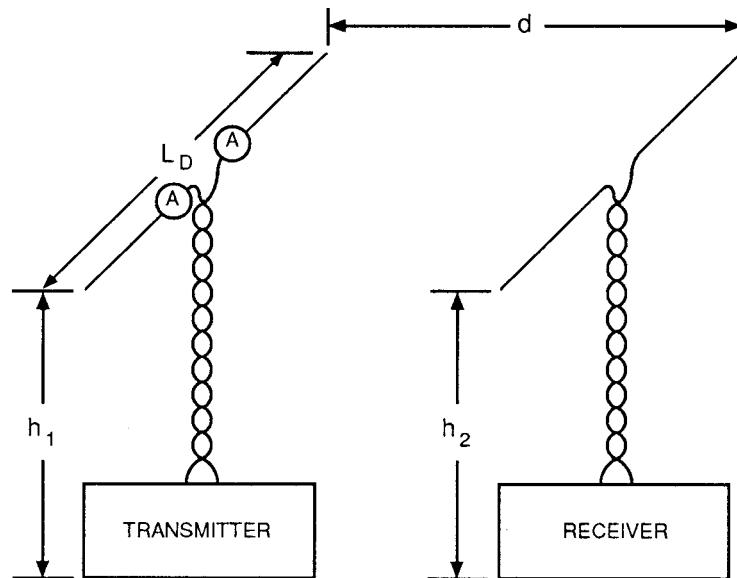
A flat clear site should be selected in which the distance to the nearest reflecting object such as trees or small buildings is at least three times the separation distance used. When  $(2\pi h_1 h_2 / \lambda d) \leq 1/4$  and  $d \geq 10(h_1 + h_2)$ , where  $h_1$  and  $h_2$  are as shown in Fig 11, the standard field can be calculated from the following equation, assuming horizontal polarization over plane earth:

$$E = \frac{\eta_0 L_{\text{eff}} I}{\lambda d} \sin \frac{2\pi h_1 h_2}{\lambda d} \quad (20)$$

$$\cong \frac{2\pi \eta_0 h_1 h_2 L_{\text{eff}} I}{\lambda^2 d^2} \quad (21)$$

where

$L_{\text{eff}}$  is the effective length of the transmitting antenna, in m.  
 $I$  is the transmitting dipole current, in A.



**Figure 11—Radiation-Field Method Using Parallel Dipoles**

The reflection coefficient of the ground is approximately -1.

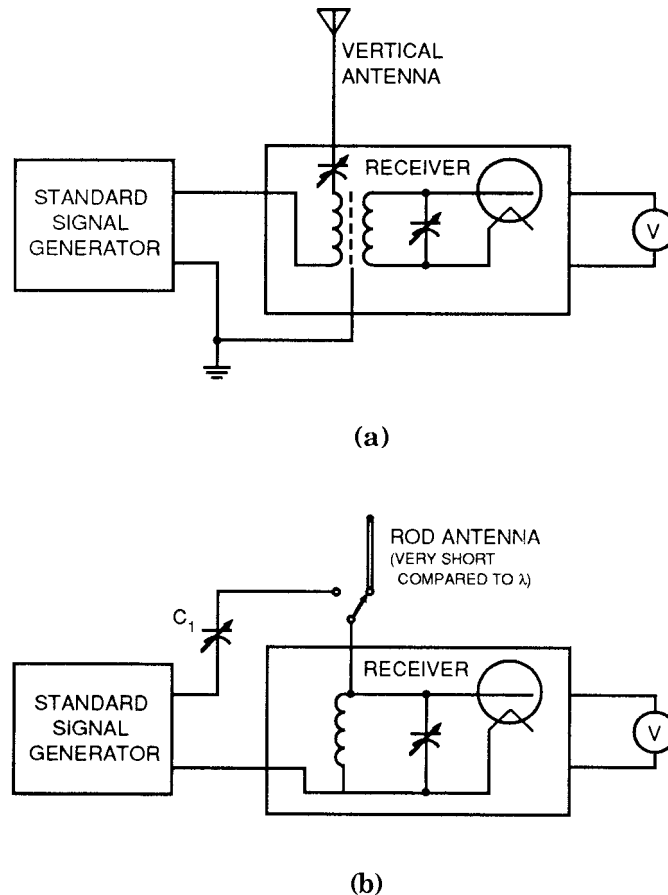
For a thin dipole with a length not greater than  $\lambda/2$ , the effective length is given by Eq 18. If it is desirable to use a single-turn transmitting loop antenna, the effective length of the loop is given by Eq 7. A uniform current is assumed in the transmitting loop (see Eq 17).

In case the field-strength measuring set uses another type of antenna, consideration must be given to such factors as path geometry and the size and directivity of the antenna so as to avoid errors due to differences in the effect of the ground-reflected component or of the wave-front curvature.

## 2.4 Field Strength Calibrations Using Monopole Antennas, 30 kHz to 300 MHz.

### 2.4.1 Standard Receiving Monopole

A vertical monopole antenna with rf output taken between its lower end and the ground is usable at frequencies up to about 300 MHz. Care must be taken to make sure that the rf output voltage is not due partly to voltages induced in the transmission line and ground system. A large symmetrical counterpoise or ground plane is usually desirable. Arrangements for measuring field strength with a vertical antenna are shown in Fig 12(a) and (b). The signal generator is used to calibrate the antenna as part of the measurement procedure.



**Figure 12—Field-Strength Meter Arrangements Using a Simple Antenna**

The effective length of a thin grounded vertical monopole antenna (in m) may be determined from the equation

$$L_{\text{eff}} = \frac{\lambda}{2\pi} \tan \frac{\pi h}{\lambda} \quad (22)$$

where

$h$  = overall physical height of the vertical monopole (assumed to be  $\lambda/4$  or less)

When a vertical monopole is very short compared to one wavelength and end effects are negligible, the effective length (sometimes called effective height) is approximately one-half of the overall physical height.

Fig 12(b) shows a calibration method that is applicable to monopole antennas that are very short compared to a wavelength. In this arrangement, the receiver is calibrated through a capacitance  $C_1$  that is equal to the source capacitance of the antenna. For best results, a metal ground plane should be used beneath the antenna ( $h \leq 0.01\lambda$ ) having dimensions at least  $\lambda/6$  in radius, see [B66]. Otherwise the effective length will be a function of the ground constants at the measuring site. This problem can be handled with method-of-moment techniques (e.g., the Numerical Electromagnetics Code [NEC-3], see [B16] -[B18]) that treat wires that penetrate the air-ground interface. Reasonable agreement between theory and measurement has been obtained for a short monopole with buried ground radials (see [B42]) where the ground constants were carefully measured, see [B40]. At low frequencies, the effect of the ground makes it difficult to calculate accurate values of the field strength at the receiving site.

Some short (e.g., 2.74 m) vertical monopoles have been installed on the roofs of vehicles to provide a mobile field-strength and radio-noise measurement capability at medium and high frequencies. It generally is necessary to calibrate each specific installation to determine the antenna factor (or effective length) for the antenna as installed, see [44].

## 2.5 Field Strength Calibrations Using Aperture Antennas, 200 MHz to 30 GHz.

### 2.5.1 General

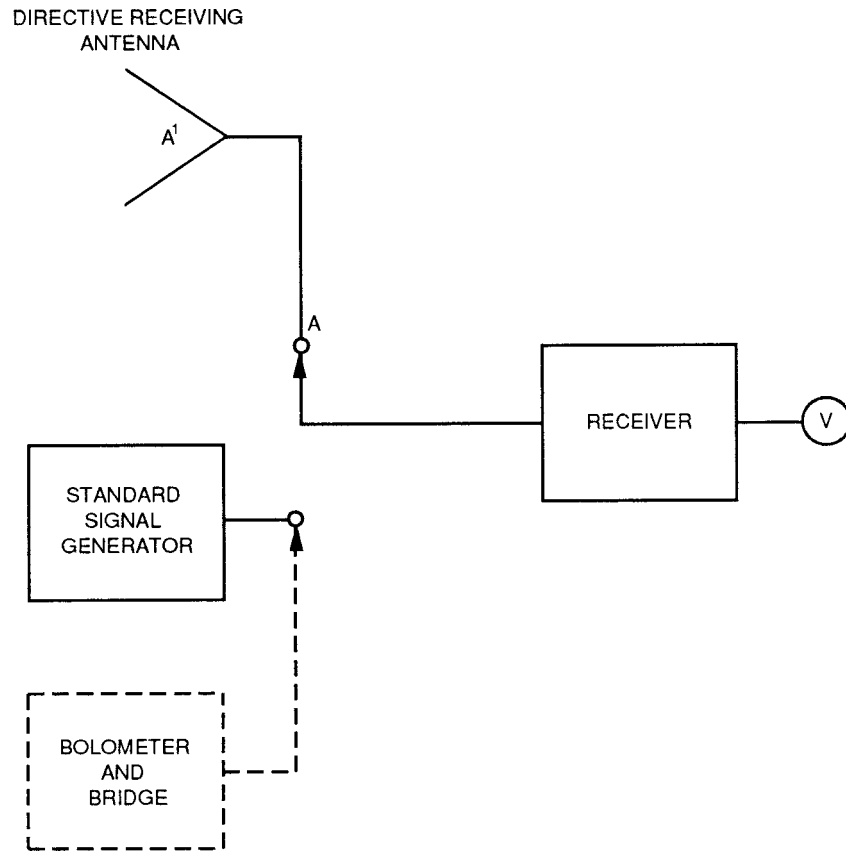
The techniques described in the previous sections may be extended to higher frequencies by using circuitry and methods appropriate to the frequency (see [B7] , [B21] , [B25] , [B33] , [B64] , [B67] , [B71] , [B72] , and [B76] ). Lumped resonant circuits used at the lower frequencies give way to tuned transmission lines or waveguide cavities at the higher frequencies (see [B45] and [B59] ). Silicon diode detectors are widely used in reception, see [B86] . Special triodes, klystrons, and magnetrons are used for generating high rf power, see [B33] . At these higher frequencies, attenuators may be constructed from a section of waveguide into which is inserted a conducting card to increase loss, see [B33] . At frequencies up to about 10 GHz, precision waveguide-below-cutoff attenuators may be employed, see [B48] .

For the transmitting and receiving antenna systems (see [B78] ), half-wave dipoles are often used with a parasitic reflector or director elements or a corner reflector to increase the antenna gain and to minimize the effect of reflecting objects to the side and rear of the antenna. A  $\lambda/2$  dipole at the focus of a parabolic reflector provides a considerable increase in gain. Biconical, sectoral, or pyramidal horns are often used. These antennas have relatively high gain; that is, they confine most of the radiated energy into a narrow beam. At frequencies above 200 MHz, it becomes more practical to measure power than to measure current or voltage. Bolometers using thin-film thermocouples, small thermistor beads, or diodes generally take the place of voltmeters (see [B48] and [B67] ). For many applications under far-zone free-space conditions, it is more convenient to measure an incident field in terms of power density ( $W/m^2$ ) rather than field strength ( $V/m$  or  $A/m$ ). In the near field, however, the  $E$  and  $H$  fields must be measured independently.

### 2.5.2 Standard Receiving Antenna Equations

A practical setup for measuring field strength or power density at frequencies above 200 MHz is shown in Fig 13. The antenna and receiver inputs should be matched to the transmission line impedance. The line used is generally either coaxial cable or hollow waveguide, depending on the frequency and length of line. The output of a receiving antenna can be measured with a power meter or by calibrating the receiver output response characteristic against a standard signal generator, which in turn is calibrated with a power meter.





**Figure 13—Field-Strength Measuring Set for Microwave Frequencies**

The setup shown in Fig 13 can be used to measure the received power  $P_r$  flowing through point A. If the gain of the receiving antenna system to point A is known, the power density of the field at the receiving antenna is given by

$$S_{\text{inc}} = \frac{4\pi P_r}{g\lambda^2} \quad (23)$$

where

- $S_{\text{inc}}$  = power density of incident field, in W/m<sup>2</sup>
- $P_r$  = received power into a matched load, in W
- $g$  = gain of the receiving antenna system relative to a lossless isotropic antenna, see [B77]

Since the power density of a free-space plane wave is equal to  $E^2/\eta_0$ , the field strength may be obtained from the equation

$$E = \frac{1}{\lambda} \sqrt{\frac{4\pi\eta_0 P_r}{g}} \cong \frac{68.8}{\lambda} \sqrt{\frac{P_r}{g}} \quad (24)$$

where

- $E$  = incident field strength, in V/m

### 2.5.2.1 Antenna Effective Area

When a plane-wave field is incident on a receiving antenna, the received power accepted by a matched load is equal to the effective area of the antenna (see [B83] ) times the power density of the incident wave. The effective area of a lossless isotropic antenna is

$$A_{\text{iso}} = \lambda^2 / 4\pi \quad (25)$$

With directive gain equal to  $g$  and no dissipation, the effective area is

$$A_{\text{eff}} = g(\lambda^2 / 4\pi) \quad (26)$$

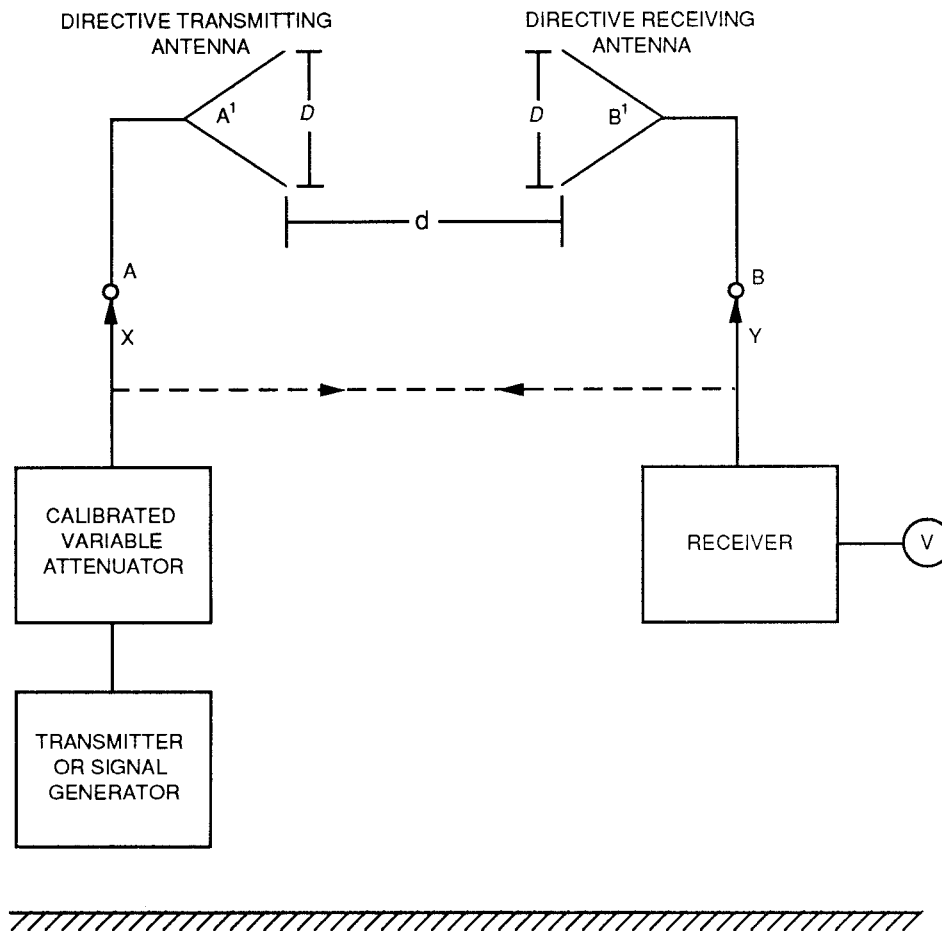
The effective area of a lossless half-wave dipole is

$$A_{(\lambda/2)} \cong 0.13\lambda^2 \quad (27)$$

### 2.5.2.2 Antenna Gain and Mismatch Factor

The power gain of an antenna system can be determined either by direct measurement or by comparison with an antenna having a known gain. In some cases, the gain can be calculated theoretically (see [B6] , [B16] –[B18] , [B63] , [B67] , [B71] , [B72] , and [B78] ). In this section, it is assumed that the transmitting and receiving antenna systems have the same polarization and that measurements are made at the point of maximum response on the pattern of each antenna (boresight angle).

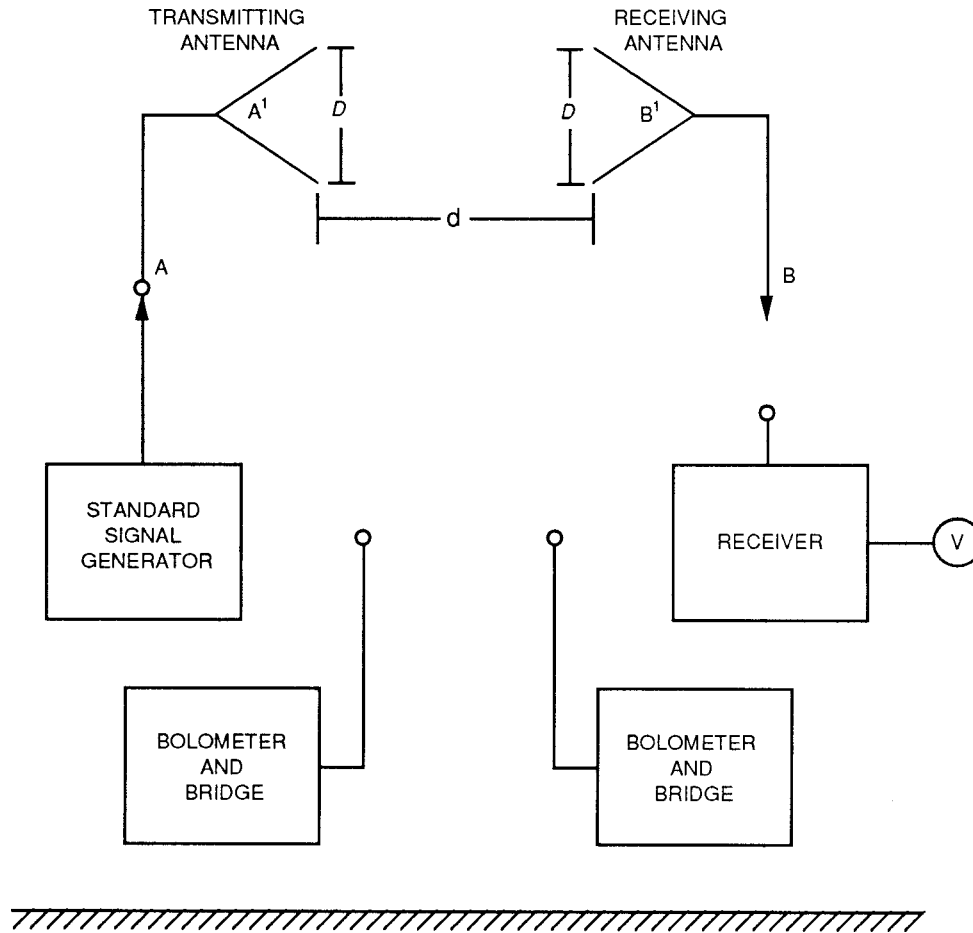
A convenient setup for measuring antenna gain is shown in Fig 14. The procedure involves measuring the ratio of power input  $P_1$  at point A to the power output  $P_2$  at point B. The two antenna systems, A and B, must be identical. This can be checked by using first one and then the other to receive an ambient signal or a locally generated signal from a third antenna. This third antenna should be at a sufficient distance that it produces a plane-wave field at antennas A and B.



**Figure 14—Arrangement for Measuring Antenna Gain**

To measure the ratio,  $P_1/P_2$ , terminal X is first connected to A and terminal Y is connected to B, and the calibrated variable attenuator and receiver output response are adjusted to give a convenient receiver indication. Then, without changing the receiver gain, terminals X and Y are connected together and the variable attenuator is adjusted to give the same receiver indication as noted previously. The required change of the attenuator setting is a measure of the ratio,  $P_1/P_2$ , if the standing-wave ratio is close to unity in each direction.

Another setup that may be used for measuring the ratio  $P_1/P_2$  is shown in Fig 15. The power in both cases (direct connection of the cables vs. insertion of the radiated path via the two identical antennas) is measured with a calibrated power meter. Care is necessary to achieve uniform tightness of the connectors used in performing the switching operations shown. If the power received via the radiated path is too small for accurate measurement, it can be measured by means of a calibrated receiver. By making use of an attenuator in the intermediate-frequency amplifier of the receiver, this range can be extended to much lower field strengths, down to about  $50 \mu\text{V/m}$ , see [B49].



**Figure 15—Alternative Arrangement for Measuring Antenna Gain**

In the calibration setups of Figs 14 and 15, it is preferable to use an open, fiat field site free of reflecting objects. The distance  $d$  should be greater than  $2D^2/\lambda$ , where  $D$  is the maximum dimension of the antenna aperture, see [B25]. If  $d = 2D^2/\lambda$ , there is a  $22.5^\circ$  phase difference in the incident wave from aperture center to aperture edge (a distance  $D/2$ ) of the receiving antenna. The height above ground and the antenna directivity preferably should be such that any ground-reflected component of energy is negligible. If this is not possible, the effect of the ground-reflected ray can be evaluated by plotting the ratio  $P_1/P_2$  versus antenna height or versus antenna separation distance, over a range sufficient to indicate maximum and minimum values in the  $P_1/P_2$  ratio, provided the ground surface is planar, see [B11]. Another expedient method is to reduce the effect of ground reflections by using one or more parallel diffraction edges, see [B25], which may be made of wire mesh, perpendicular to the transmission path between the antennas.

Having determined the ratio  $P_1/P_2$  by one of the methods shown in Figs 14 and 15, the gain  $g$  of each of the two identical antennas relative to an isotropic radiator may be calculated using the expression (see [B32]):

$$g = \frac{4\pi d}{\lambda} \frac{P_2}{P_1} \quad (28)$$

Expressed in dB,

$$G(\text{dBi}) = 10 \log_{10} g \quad (29)$$

where

$i$  indicates the gain is with respect to an isotropic radiator.

If the gains of the two antennas are slightly different, then the value of  $g$  obtained above satisfies the relation

$$g = \sqrt{g_1 g_2} \quad (30)$$

where

$g_1$  and  $g_2$  are the gains of the two antennas relative to an isotropic radiator.

In order to obtain each of the two gains individually, it is necessary to use a third antenna and employ a so-called three-antenna calibration method, see [B8] .

The intrinsic gain of a thin cylindrical dipole antenna with no dissipative loss, whose length  $L$  is short compared to one wavelength ( $L \ll \lambda$ ), is 1.5 (i.e., 1.76 dBi). For a thin self-resonant half-wave dipole, the gain is 1.64 (i.e., 2.15 dBi). In the latter case, the gain increases slightly as the length-to-diameter ratio of the cylindrical antenna is decreased, see [B60] . The gain of more complex antennas has been determined by several writers (see [B92] , [B77] , [B78] and [B86] ). When these antennas are used for reception, the given values of gain are exact only if the incident field is a uniform plane wave, i.e., having constant phase and amplitude over the aperture of the receiving antenna. These conditions will apply at distances from the transmitting antenna that are large compared with the antenna aperture in wavelengths (i.e.,  $> 2D^2/\lambda$ ).

In making antenna gain measurements, the impedance of antennas and loads should be matched to the transmission line impedance. The transmitter should be loosely coupled to the line, such as by use of a matched attenuator pad, so that removing the antenna does not affect the power output from the transmitter. This can be verified by using a voltmeter in the line between the transmitter and attenuator.

The impedance mismatch factor  $q$  can be written as (see [B83] ):

$$q = \frac{4RR_L}{(R + R_L)^2 + (X + X_L)^2} \quad (31)$$

where

$$\begin{aligned} Z &= R + jX = \text{antenna impedance} \\ Z_L &= R^L + jX_L = \text{load impedance} \end{aligned}$$

This expression is in agreement with that of Kraus [B61] for the case of a lossless antenna. When the antenna has both loss resistance  $R_1$  as well as radiation resistance  $R_r$ , the impedance mismatch and resistive loss can be combined in the following expression for  $q$ :

$$q = \frac{4RR_L}{(R_r + R_1 + R_L)^2 + (X + X_L)^2} \quad (32)$$

If both the antenna loss resistance  $R_1$  and the load reactance  $X_L$  are zero, then Eq 32 can be written

$$q = \frac{4RR_L}{(R + R_L)^2 + X^2} = 1 - |\Gamma|^2 \quad (33)$$

where

$\Gamma$  is the input reflection coefficient of the antenna when  $R = Z_0$  is the real characteristic impedance of the transmission line or waveguide feed, see [B44].

Since the magnitude of  $\Gamma$  ( $|\Gamma|$ ) can be measured,  $q$  is easily determined from Eq 33 and there is no need to measure the input resistance and reactance of the antenna. The main restrictions on Eq 33 are that the antenna resistive losses must be negligible and the transmission line or waveguide feed must be single moded at the antenna terminals.

The impedance mismatch factor relative to  $50 \Omega$  can be computed, in dB, from impedance data using the following equation:

$$q(\text{dB}) = 10 \log \left[ \frac{4(\text{VSWR})}{(\text{VSWR} + 1)^2} \right] \quad (34)$$

where

$$\begin{aligned} \text{VSWR} &= (1 + |\Gamma|)/(1 - |\Gamma|) = \text{voltage standing wave ratio} \\ |\Gamma| &= |(Z - Z_0)/(Z + Z_0)| = \text{magnitude of reflection coefficient} \\ Z &= \text{Measured or computed input impedance, in } \Omega \\ Z_0 &= \text{characteristic impedance of the transmission line, } 50 \Omega \end{aligned}$$

### 2.5.3 Standard Transmitting Antenna Equations

A transmitting antenna in free space that radiates  $P_1$  watts and has a gain  $g_1$  in a given direction will radiates  $P_1 g_1 / 4\pi$  watts per steradian at a distance that is large compared to the antenna aperture. The power density in  $\text{W/m}^2$  at a distance  $d$  from the antenna will be  $P_1 g_1 / 4\pi d^2$ . In terms of the field strength existing at the same distance, the power density is  $E^2 / \eta_0$ . By equating these two expressions, the free-space field may be determined from the relationship

$$E = \sqrt{\frac{\eta_0 P_1 g_1}{4\pi d^2}} \cong \frac{\sqrt{30 P_1 g_1}}{d} \quad (35)$$

where

$$\begin{aligned} E &= \text{free-space RMS electric field strength, in V/m} \\ P_1 &= \text{input power to the transmitting antenna, in W} \\ g_1 &= \text{gain of the transmitting antenna in the direction toward the receiving point relative to an isotropic radiator} \\ d &= \text{distance from the transmitting antenna to the receiving point, in m} \end{aligned}$$

For calibration purposes, an anechoic chamber or a flat-field site free of reflecting objects should be used. It is then possible to establish an essentially free-space field if the antenna separation distance is small compared with the heights of the antennas above ground, and the distance to the nearest reflecting object is large compared with both these distances.

At larger distances over plane earth, Eq 35 will not apply since it is then necessary to take into account the ground-reflected wave. For values of  $2\pi h_2 / \lambda d < 1/4$ , and  $d > 10(h_1 + h_2)$ , the field strength may be determined from the following relationship assuming that the reflection coefficient of the ground at the grazing angle of incidence is -1:

$$E = \frac{2h_1 h_2}{\lambda d^2} \sqrt{\pi \eta_0 P_1 g_1} = \frac{68.8 h_1 h_2 \sqrt{P_1 g_1}}{\lambda d^2} \quad (36)$$

where

$$E = \text{the generated field strength, in V/m,}$$

$h_1$  = height of the transmitting antenna, in m  
 $h_2$  = height of the receiving antenna, in m

Eq 36 also assumes that the angle of incidence of the ground-reflected wave and the beam width of the transmitting and receiving antennas are such that there is no appreciable discrimination between the direct and ground-reflected waves.

### 2.5.3.1 Measurement of Power Delivered to a Transmitting Antenna

Fig 16 shows a typical setup for measuring net power to a transmitting antenna. The four-port dual-directional coupler allows measurement of the power incident upon and reflected from the antenna. The net power to the antenna is then

$$P_{\text{net}} = P_{\text{inc}} - P_{\text{refl}} \quad (37)$$

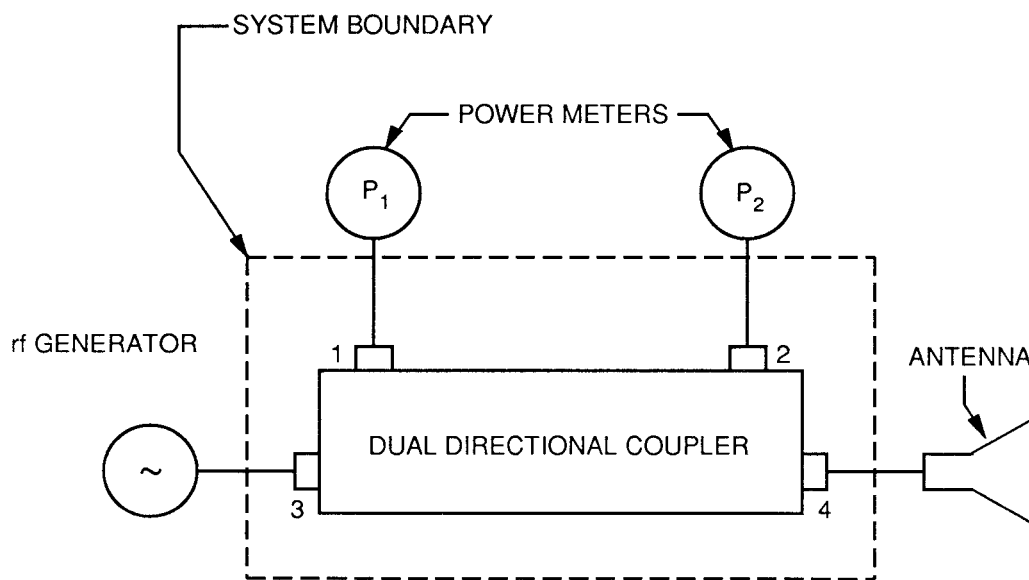


Figure 16—System for Measuring rf Power Delivered to an Antenna

Reprinted from [B55] .

The ratio of power emerging from port 4 (i.e.,  $P_{\text{inc}}$ ) to the power  $P_1$  read by the power meter on port 1 is the incident coupling factor  $C_{\text{inc}}$ . The reflected coupling factor  $C_{\text{refl}}$  is the ratio of power entering port 4 (i.e.,  $P_{\text{refl}}$ ) to the power  $P_2$  read by the power meter on port 2. (Measuring coupling factors with respect to port 4 eliminates the need for a separate measurement of coupler insertion loss.)

Manufacturers furnish nominal coupling factors over the frequency range of the coupler. For greatest accuracy in obtaining  $P_{\text{net}}$ , however, both coupling factors should be known at each measurement frequency. Then, at a given frequency,

$$P_{\text{net}} = C_{\text{inc}}P_1 - C_{\text{refl}}P_2 \quad (38)$$

Eq 38 is valid for use with directional couplers having high directivity (at least 25-30 dB), but may give inaccurate values for  $P_{\text{net}}$  if the coupler has low directivity and if the port-4 load is not well matched to the coupler or transmission line impedance. An analysis of the directional coupler as a four-port junction has been done in terms of the reflection

coefficients and scattering parameters of the system, see [B55] . The net power delivered to the transmitting antenna is given by

$$\begin{aligned}
 P_{\text{net}} &= P_{\text{inc}} - P_{\text{ref}} \\
 &= \frac{P_1}{(1 - |\Gamma_1|^2)} \left| \frac{S_{34}}{S_{13}} \right|^2 |g(S, \Gamma)|^2 \\
 &= \frac{P_2}{(1 - |\Gamma_2|^2)} \frac{1}{|S_{24}|^2} |h(S, \Gamma)|^2
 \end{aligned} \tag{39}$$

The symbols  $\Gamma_1$  and  $\Gamma_2$  represent the reflection coefficients observed looking into power meters 1 and 2.  $S_{ij}$  is the scattering parameter defined as the ratio of the complex wave amplitude emerging from port  $i$  to that incident upon port  $j$ , and  $g(S, \Gamma)$  and  $h(S, \Gamma)$  are functions of the system  $S$ -parameters and the reflection coefficients of ports 1, 2, and 4. For an ideal coupler, i.e., one having zero reflection coefficient for all input ports and infinite directivity ( $S_{11} = S_{22} = S_{33} = S_{44} = S_{14} = S_{41} = S_{23} = S_{12} = S_{21} = 0$ ), with matched power meters at ports 1 and 2 ( $\Gamma_1 = \Gamma_2 = 0$ ), the terms  $g(S, \Gamma) = h(S, \Gamma) = 1$ . Unless the magnitudes and phases of the system  $S$ -parameters and the reflection coefficients of a real system are well determined,  $g(S, \Gamma)$  and  $h(S, \Gamma)$  are not calculable. The deviation of  $g(S, \Gamma)$  and  $h(S, \Gamma)$  from unity is, therefore, taken to be part of the uncertainty in the determination of the net power delivered to the standard antenna. Although the degree of deviation from unity is a function of the system  $S$  and  $\Gamma$  parameters, it is found to be, in general, less than 1%, see [B55] .

To compute the net power using Eq 39, the terms  $S_{34}/S_{13}$  and  $1/S_{24}$  need to be determined. Although the magnitudes of  $S_{13}$ ,  $S_{24}$ , and  $S_{34}$  could be measured with a network analyzer, the system described here can be made flat-plate short and a matched termination. When a short ( $\Gamma_4 = -1$ ) is placed at port 4, the ratio of power measurements  $P_2$  and  $P_1$  gives

$$\frac{P_2}{P_1} = \left| \frac{S_{24} S_{34}}{S_{13}} \right|^2 \frac{1 - |\Gamma_2|^2}{1 - |\Gamma_4|^2} (1 + |\Delta_1|^2) \tag{40}$$

where

$\Delta_1(S, \Gamma)$  is a complex quantity much less than unity, see [B57] .

The second step in evaluating the  $S$ -parameter coefficients in Eq 39 is to move the power meter from port 2 to port 4 and terminate port 2 in  $50 \Omega$ . The ratio of the two power measurements  $P_1$  and  $P_4$  is

$$\frac{P_1}{P_4} = \left| \frac{S_{13}}{S_{34}} \right|^2 \frac{1 - |\Gamma_1|^2}{1 - |\Gamma_4|^2} (1 + |\Delta_2|^2) \tag{41}$$

where

$\Delta_2$  is another complex quantity much less than unity.

From the four power measurements and values for the power meter reflection coefficients  $\Gamma_1$ ,  $\Gamma_2$ , and  $\Gamma_4$ , the quantities  $|S_{34}/S_{13}|^2$  and  $|S_{24}|^2$  may be computed from Eqs 40 and 41. The terms  $\Delta_1$  and  $\Delta_2$  involve products of the system  $S$ -parameters and reflection coefficients  $\Gamma$  whose phases are not known. Therefore, the magnitudes of  $\Delta_1$  and  $\Delta_2$  cannot be determined, and the deviation of  $\Delta_1$  and  $\Delta_2$  from zero is unknown. Moreover, the uncertainty in the power measurements and in the values for the reflection coefficients also contribute to the uncertainty in the determination of  $|S_{34}/S_{13}|$  and  $|S_{24}|$ . The detailed discussion on this topic is given in [B57] .

### 2.5.3.2 Standard Field Equations for an Anechoic Chamber

Electromagnetic fields in an anechoic chamber are usually established with standard transmitting antennas. The antennas commonly used are open-end wave-guides (OEG) below about 450 MHz and rectangular pyramidal horns



above about 450 MHz. Eq 35 is used to calculate the electric field on the axis (boresight) of the antenna. The quantities to be determined are net power  $P$ , distance  $d$ , and antenna gain relative to an isotropic antenna  $g$ . The net power delivered to the antenna is measured as described in 2.5.3.1., and  $d$  is measured by any suitably accurate procedure. Evaluating the radiated electric field then reduces to determining the transmitting antenna gain at each measurement frequency and distance.

Early work to determine the field pattern and gain of large OEG radiators, both theoretically and experimentally, is described in [B14]. An equation for the gain of OEGs as a function of frequency and aperture dimensions has been determined by an analysis of data from a two-antenna calibration using two identical OEGs of 2:1 aspect ratio, see [B15]. This empirical equation for calculating the OEG gain  $g$ , is

$$g = 21.6fw \quad (42)$$

or, in dB

$$G = 10\log_{10}g = 10\log(fw) + 13.34, \text{ in dB} \quad (43)$$

where

$$\begin{aligned} f &= \text{frequency, in GHz} \\ w &= \text{width (large side) of the 2:1 OEG, in m} \end{aligned}$$

Eq 42 is accurate to approximately  $\pm 0.5$  dB ( $3\sigma$  value) when the distance from the OEG aperture to the field point is greater than  $2w$ . At on-axis points farther than about  $2w$  from the aperture, the OEG gain is independent of distance. Deviations from Eq 42 can be due to reflections in the anechoic chamber and higher-order modes excited in the waveguide interior.

In deriving the far-field gain of a pyramidal horn (Fig 17) by the Kirchhoff method, Schelkunoff accounted for the effect of the horn flare by introducing a quadratic phase error in the dominant-mode field along the aperture coordinates, see [B77]. Geometrical optics and single diffraction by the aperture edges yield essentially the Kirchhoff results. The proximity effect in the near field (i.e., an on-axis point is not equidistant from points in the aperture plane) can also be approximated by a quadratic phase error in the aperture field. Taking into account the preceding considerations, the near-field gain  $g$  of a pyramidal horn radiating at a wavelength  $\lambda$  m is given by [B51]

$$g = \frac{32ab}{\pi\lambda^2} R_E R_H = 113.3f^2 ab R_E R_H \quad (44)$$

where

$$f \text{ is in GHz}$$

The terms  $R_E$  and  $R_H$  are often written in terms of Fresnel integrals (see [B50]) and are functions of the horn dimensions and the on-axis distance from aperture plane to field point. The numeric gain may be more easily obtained from tabulated values for  $R_E$  and  $R_H$  in dB (see [B51]) to which polynomial expressions have been fitted, see [B62].

$$R_E(\text{dB}) = -(0.1\beta^2)(2.31 + 0.053\beta) \quad (45)$$

$$R_H(\text{dB}) = -(0.01\alpha)(1 + 10.19\alpha + 0.51\alpha^2 - 0.097\alpha^3) \quad (46)$$

where

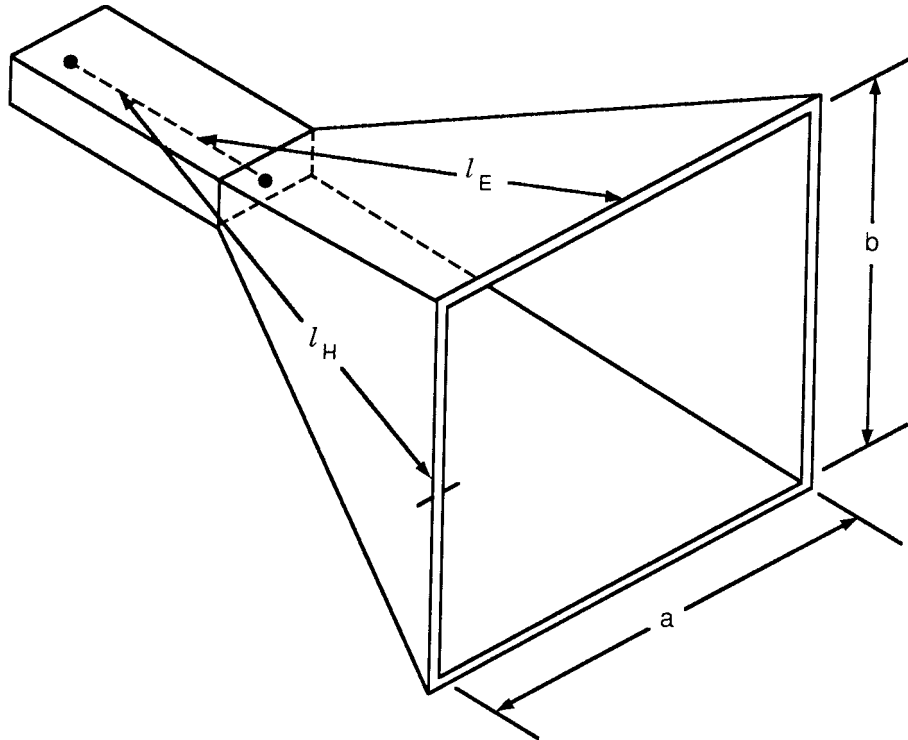
$$\alpha = \left(\frac{a^2 f}{0.3}\right)\left(\frac{1}{l_H} + \frac{1}{d}\right); \beta = \left(\frac{b^2 f}{0.3}\right)\left(\frac{1}{l_E} + \frac{1}{d}\right) \quad (47)$$

$a, b, l_E, l_H$  = horn dimensions, in m

$f$  = frequency, in GHz

$d$  = on-axis distance from horn aperture to field point, in m

The terms  $R_E$  and  $R_H$  in Eq 44 are gain reduction factors that are positive and less than unity. Though their values in decibels are negative, the tabulation in [B51] gives these decibel values as positive. Therefore, a minus sign (not present in [B62]) is included in Eqs 45 and 46 to make Eq 48 consistent with Eq 44.



**Figure 17—Sketch of a Pyramidal Horn Showing the Dimensions**

From Eq 44 for the gain  $g$ , the horn gain  $G$ , in dBi, is

$$G = 10 \log G = 20.54 + 10 \log(ab) + 20 \log(f) + R_E(\text{dB}) + R_H(\text{dB}) \quad (48)$$

The numeric gain  $g$  is then

$$g = 10^{G/10} \quad (49)$$

The standard field generated in an anechoic chamber by an OEG or a pyramidal horn antenna can be computed from Eq 35 using gain values obtained from Eqs 42 or 49. Anechoic chamber measurements of a 450 MHz electric field were made using both an OEG and a pyramidal horn as transmitting antennas, see [B57]. The measured fields from 1–4 m from the apertures were within  $\pm 1$  dB of the computed OEG field, and within  $\pm 0.6$  dB ( $3\sigma$  value) of the computed horn field. The discrepancies were due primarily to reflections within the chamber.

### 2.5.3.3 Calibration of Receiving Antenna Factor in an Anechoic Chamber

A test antenna is often calibrated in terms of its antenna factor measured over a range of frequencies and electric field strengths. This calibration is accomplished by placing the antenna in a known incident electric field  $E_{\text{inc}}$  (on the boresight of the transmitting antenna) and measuring the voltage  $V_{50\Omega}$  across the antenna terminals with a receiver of  $50\ \Omega$  input impedance. The test and transmitting antennas are polarization matched. The antenna factor is then obtained from Eq 50.

$$k = \frac{E_{\text{inc}}}{V_{50\Omega}} \quad (50)$$

During the calibration, a  $50\ \Omega$  cable of attenuation  $a$  may be required to connect the antenna to the receiver. The voltage measured at the receiver is then  $V_{\text{rec}}$  and the voltage at the antenna terminals is still denoted  $V_{50\ \Omega}$ . The cable attenuation at the measurement frequency is  $a = V_{\text{rec}}/V_{50\ \Omega}$  and the antenna factor with reference to the voltage at the antenna terminals is then

$$k = \frac{E_{\text{inc}}}{V_{50\Omega}} = \frac{E_{\text{inc}}}{V_{\text{rec}}/a} = \frac{E_{\text{inc}}a}{V_{\text{rec}}} \quad (51)$$

In dB,  $K$  is

$$K = 20 \log k = E_{\text{inc}}(\text{dBV/m}) - V_{\text{rec}}(\text{dBV}) + A(\text{dB}), \text{ in dB(m}^{-1}\text{)} \quad (52)$$

where

$E_{\text{inc}}(\text{dBV/m})$  is in dB with respect to 1 V/m.

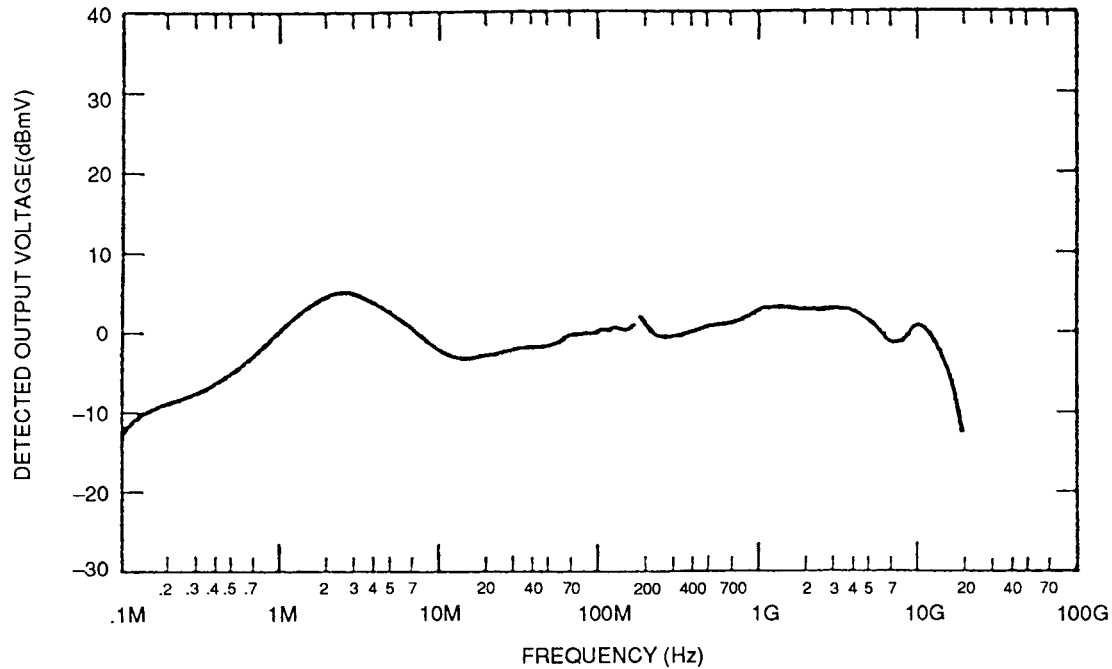
$V_{\text{rec}}(\text{dBV})$  is in dB with respect to 1 V.

The attenuation  $a$  is a positive number less than unity, and therefore  $A$  in decibels is a negative number.

## 2.6 Electrical and Magnetic Field Probes as Transfer Standards

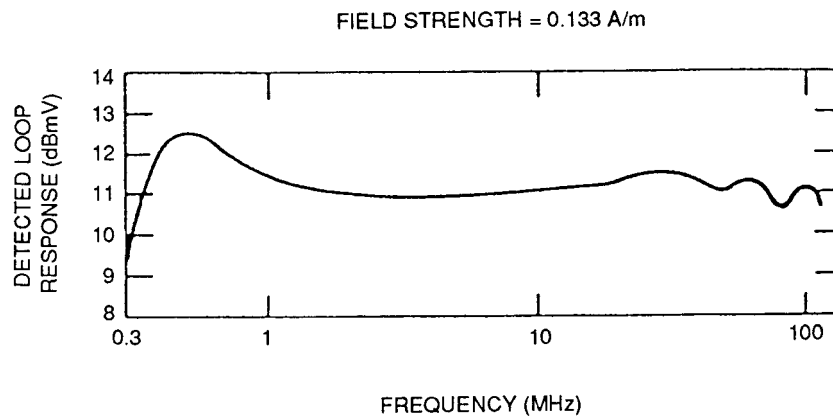
The term “transfer standard,” as used here, refers to an electrically small antenna or *rf* probe. This can be a short dipole for sensing  $E$  fields or a small loop for  $H$  fields, which has a known response over a given and relatively small range of frequency or amplitude. Such an antenna can be used to measure or verify the field strength as a function of frequency (e.g. in an anechoic chamber). This check is especially useful if the transfer standard has been calibrated by an independent approach over the same frequency range. A transfer probe can also be used to check the field level as a function of distance from the transmitting antenna, such as in an anechoic chamber. One method used is to adjust (and measure) the power delivered to the transmitting antenna until the probe output reaches a given (constant) indication. In this way, the field strength also remains constant, since the frequency and probe response are maintained constant, and the transmitting power required to obtain this (constant) response is recorded as a function of separation distance between the transmitting antenna and receiving probe. Then, from required transmitting power, the (near-zone) gain reduction of the antenna, or the “ripple” in field strength due to reflections inside the anechoic chamber, can be evaluated.

Most total electric field *rf* probes consist of three electrically-short, mutually-orthogonal dipole antennas with three diode detectors. The characteristics of these electric field *rf* probes have been analyzed by many researchers (see [B52], [B54], [B62], and [B79]). Fig 18 shows the dc output voltages of an 8 mm dipole with resistive loading as measured in a standard field of 10 V/m from 100 kHz to 18 GHz, see [B56].



**Figure 18—Measured Frequency Response of One Element of the Broadband E-Field Probe at 10 V/m**

Reprinted from [B56]



**Figure 19—Response of a Compensated Open-Circuit-Voltage Loop Antenna with Shielding (5-Turn, 10 cm Diameter)**

Reprinted from [B58]

For magnetic field measurements, three basic designs for broadband magnetic field probes have been analyzed, see [B58]. Three configurations are discussed here.

- 1) A short-circuit current loop
- 2) An open-circuit voltage loop
- 3) A compensated open-circuit voltage loop

To make the response of a loop antenna flat over the frequency range of interest, the  $Q$  of the loop antenna is reduced by means of a loading resistor. As an example, the response of a 5-turn compensated loop with a 10 cm diameter is shown in Fig 19.

The electric and magnetic field probes discussed above measure either the electric or magnetic field only, and therefore cannot measure complicated EM fields such as those with reactive near-field components, multipath reflections, etc. For this reason, a single sensor capable of performing simultaneous electric and magnetic field measurements was developed at NIST, see [B53]. In this case, a loop is loaded at diametrically opposite points with equal impedances. It can be shown that across one load the magnetic-loop response adds to the electric-dipole response, whereas across the other load the magnetic-loop response subtracts from the electric-dipole response. Thus, by taking the sum and difference of currents across loads at diametrically opposite points, the magnetic-loop response and electric-dipole response can be separated. That is, the sum current gives a measure of the magnetic field, whereas the difference current gives a measure of the electric field. This device is intended not only to measure the polarization ellipses of the electric-field and magnetic-field vector in the near-field region, but also to measure the time-dependent Poynting vector and describe the energy flow.

### 3. Using Field Strength Measurements to Determine the Power Radiated From an Antenna

#### 3.1 General

This section describes methods for measuring the radiated power from antennas under several conditions, see [B21], [B25], [B46], [B67], and [B78]. In general, the radiated power is calculated by comparing the measured field strength with the theoretical value predicted for an ideal antenna having the same height above ground.

#### 3.2 Effective Radiated Power for a Ground-Based Vertical Monopole Antenna

This method is useful at frequencies below 5 MHz for an antenna less than  $1 \lambda$  above ground. The equivalent power for ground-wave transmission is determined by measuring the field strength near the ground and at suitable distances from the antenna, see [B28] and [B46]. If the ground is a perfect conductor, and if the radiation pattern of the antenna is omni-directional in a horizontal plane, the radiated power can be obtained by measurement of field strength using the equation:

$$P = \left( \frac{E_1}{E_2} \right)^2 \quad (53)$$

where

$E_1$  = Field strength at  $i$  km radiated by the ground-based vertical monopole antenna when accepting  $P_1$  W  
 $E_2$  = Field strength at 1 km radiated by an ideal ground-based vertical antenna having the same effective length as the antenna used and a radiated power of  $P_1 = 1$  W

The field strength  $E_2$  produced by an ideal antenna is obtained from Fig 20. The relation between field strength and antenna length shown in Fig 20 (see [B12]) assumes an ideal antenna having a sinusoidal current distribution.

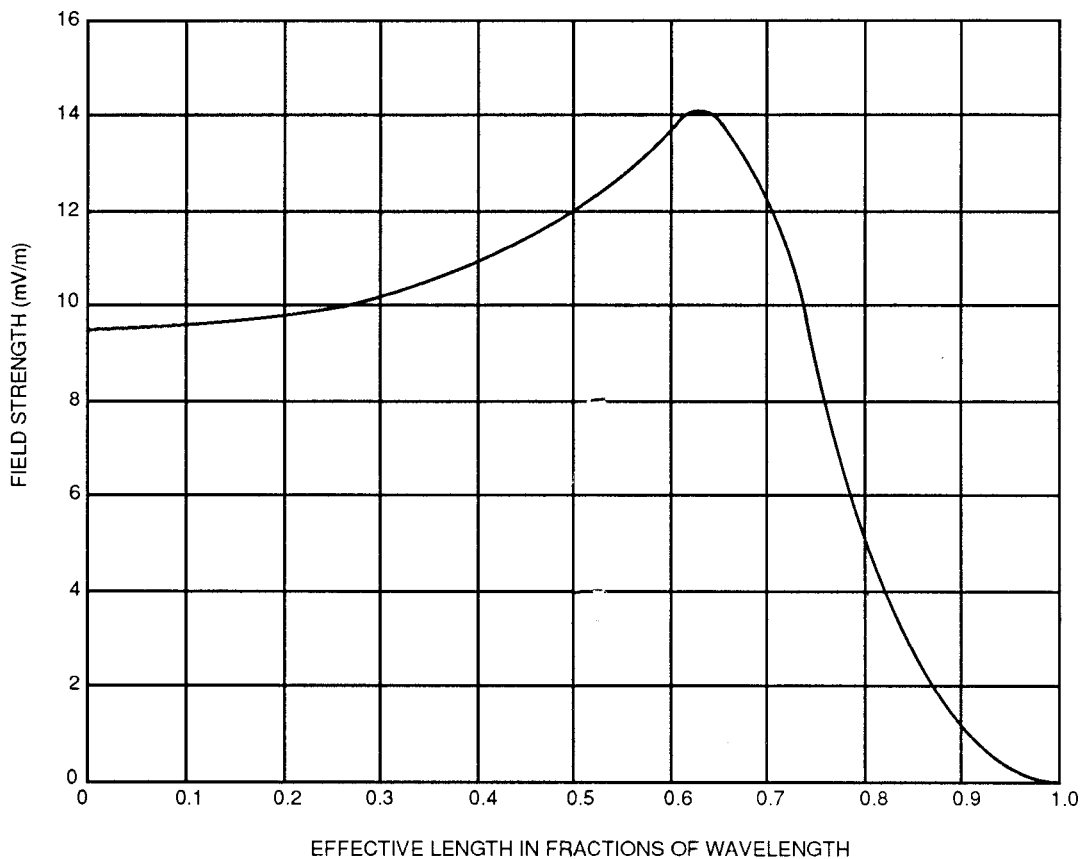
Actually, the conditions listed above cannot be expected in practice. Also, the accurate determination of field strength at 1 km distance is more involved than making a single measurement. Possible sources of error include

- 1) Determination of the ground conductivity

- 2) Possibility that the field strength at a given measuring point is not representative, but is influenced by directivity in the radiated pattern, uneven terrain, proximity of overhead wires, etc.
- 3) Determination of the distance between the measurement point and transmitting antenna
- 4) Inaccuracy of the field strength meter and other equipment used to measure  $E_1$

In practice, measurements of field strength are made on a number of equally-spaced radials. These measurements are made along each radial outward from the antenna in the general direction in which the field strength is desired. Measurements should be made at 10 to 20 evenly spaced intervals out to a distance (in km) from the antenna of  $8000/f^{1/3}$  where  $f$  is the operating frequency in Hz. Data should not be taken at distances less than one wavelength from the antenna or five times the vertical height of the antenna, whichever is greater.

In order to determine field strength at 1 km, the data should be plotted on log-log graph paper, with the same scales as shown in Fig 21, with field strength as the ordinate and distance as the abscissa. The field strength and distance may be in arbitrary units. The data sheet is then placed over the graph shown in Fig 21 and shifted horizontally and vertically (keeping the vertical lines on both sheets parallel) to determine which one of the family of curves best conforms to the plotted curve. When this best fit has been established, the line in Fig 21 marked  $1/p$  (inverse distance), should be traced on the data sheet and extended until it intersects the 1 km abscissa. The value of field strength that corresponds to that point is the desired value of unattenuated field strength at 1 km. If it is desired to reconstruct Fig 21 with a different scale of coordinates, the data given in Table 2 may be used.



**Figure 20—Unattenuated Field Strength at 1 km From an Ideal Ground-Based Vertical Antenna for 1 W Radiated**

The values of the ground conductivity and the dielectric constant of the terrain should be computed in order to determine that the best fit, as established above, represents a reasonable value. The numerical distance  $p$  and angle  $b$

are related to the actual distance  $d$  expressed in m, the free-space wavelength  $\lambda$  in m, the velocity of light  $c$  expressed in m/s, the frequency  $f$  in Hz, the ground conductivity  $\sigma$  in siemens per meter, and the relative permittivity (dielectric constant)  $\epsilon_r$  of the ground, as follows:

$$p = \frac{\pi d \cos^2 b''}{x \lambda \cos b'} \cong \frac{\pi d}{x \lambda} \cos b \quad (54)$$

$$b = 2b'' - b' \cong \tan^{-1} \frac{\epsilon_r + 1}{x} \quad (55)$$

$$x = \frac{2 \cdot 10^{-7} c^2 \sigma}{f} \cong \frac{1.80 \cdot 10^{10} \sigma}{f} \quad (56)$$

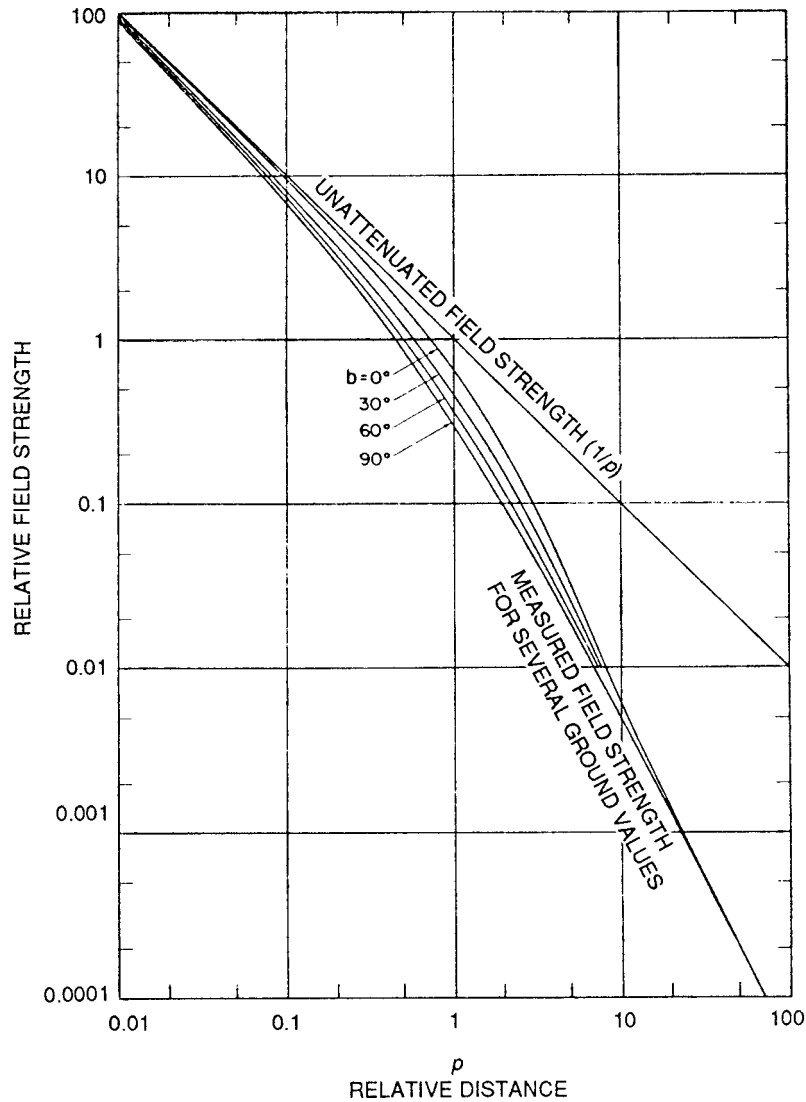
$$\tan b' = \frac{\epsilon_r - 1}{x} \quad (57)$$

**Table 2—Relative Field Strength Versus Relative Distance From a Vertical Monopole Over a Plane Earth (Fig 21, Eqs 54-58)**

$p$	$b=0^\circ$	$b=5^\circ$	$b=10^\circ$	$b=15^\circ$	$b=20^\circ$	$b=25^\circ$	$b=30^\circ$
0.01	99.572	98.808	98.054	97.312	96.583	95.868	95.167
0.02	49.573	40.037	48.511	47.995	47.491	46.999	46.520
0.05	19.576	19.244	18.922	18.609	18.306	18.013	17.730
0.1	9.581	9.354	9.136	8.927	8.726	8.534	8.350
0.25	3.595	3.464	3.342	3.226	3.118	3.016	2.920
0.5	1.617	1.538	1.466	1.399	1.337	1.280	1.228
1.0	0.6565	0.6154	0.5787	0.5456	0.5159	0.4891	0.4649
2.0	0.2199	0.2042	0.1906	0.1786	0.1680	0.1587	0.1504
3.0	0.1009	0.09408	0.08818	0.08304	0.07851	0.07452	0.07096
4.0	0.05384	0.05078	0.04809	0.04572	0.04360	0.04171	0.04000
5.0	0.03186	0.03047	0.02922	0.02807	0.02702	0.02606	0.02518
8.0	0.01020	0.01007	0.009925	0.009763	0.009591	0.009420	0.009245
10.0	0.006080	0.006044	0.005999	0.005934	0.005874	0.005819	0.005734
13.0	0.003395	0.003392	0.003380	0.003363	0.003342	0.003317	0.003289
18.0	0.001694	0.001694	0.001691	0.001686	0.001681	0.001673	0.001665
50.0	0.000263	0.0002063	0.0002062	0.0002061	0.0002059	0.0002056	0.0002054
$p$	$b=35^\circ$	$b=40^\circ$	$b=45^\circ$	$b=50^\circ$	$b=55^\circ$	$b=60^\circ$	$b=65^\circ$
0.01	94.483	93.816	93.166	92.535	91.923	91.330	90.759
0.02	46.054	45.601	45.162	44.737	44.327	43.932	43.552
0.05	17.457	17.194	16.941	16.699	16.467	16.245	16.033
0.1	8.174	8.007	7.847	7.696	7.552	7.415	7.285
0.25	2.830	2.746	2.667	2.593	2.524	2.459	2.398
0.5	1.179	1.135	1.093	1.055	1.020	0.9880	0.9581
1.0	0.4430	0.4232	0.4052	0.3888	0.3739	0.3603	0.3479
2.0	0.1430	0.1364	0.1305	0.1251	0.1203	0.1159	0.1119
3.0	0.0778	0.06493	0.06236	0.06003	0.05792	0.05601	0.05426
4.0	0.03846	0.03706	0.03578	0.03461	0.03354	0.03257	0.03167
5.0	0.02436	0.02361	0.02292	0.02227	0.02168	0.02112	0.02061
8.0	0.009074	0.008905	0.008740	0.006592	0.008430	0.008282	0.008141
10.0	0.005644	0.005593	0.005507	0.005455	0.005356	0.005283	0.005202
13.0	0.003260	0.003229	0.003196	0.003163	0.003130	0.003097	0.003065
18.0	0.001655	0.001645	0.001634	0.001623	0.001612	0.001600	0.001588
50.0	0.0002050	0.0002047	0.0002043	0.0002038	0.0002034	0.0002029	0.0002024
$p$	$b=70^\circ$	$b=75^\circ$	$b=80^\circ$	$b=85^\circ$	$b=90^\circ$		
0.01	90.208	89.679	89.172	88.687	88.226		
0.02	43.187	42.838	42.504	42.187	41.886		
0.05	15.831	15.639	15.457	15.284	15.122		
0.1	7.163	7.047	6.938	6.836	6.740		
0.25	2.342	2.289	2.240	2.195	2.152		
0.5	0.9305	0.9050	0.8816	0.8600	0.8401		
1.0	0.3366	0.3263	0.3168	0.3082	0.3003		
2.0	0.1083	0.1050	0.1020	0.09932	0.09684		
3.0	0.05267	0.05122	0.04989	0.04867	0.04756		
4.0	0.03084	0.03008	0.02938	0.02874	0.02815		
5.0	0.02013	0.01969	0.01929	0.01891	0.01856		
8.0	0.008009	0.007885	0.007764	0.007652	0.007549		
10.0	0.005126	0.005072	0.004998	0.004949	0.004883		
13.0	0.003034	0.003003	0.002973	0.002944	0.002916		
18.0	0.001576	0.001565	0.001553	0.001542	0.001531		
50.0	0.0002019	0.0002014	0.0002008	0.0002003	0.0001998		

For  $p > 100$  ( $A/p = 1/2 p^2$ ).





**Figure 21—Decay of Surface-Wave Component of the Ground Wave for a Plane Earth**

$$\tan b'' = \frac{\epsilon_r}{x} \quad (58)$$

Using the approximations on the right of Eqs 54 and 55,  $\sigma$  and  $\epsilon_r$  may be estimated from the graphically determined values of numerical distance  $p_1$  at 1 km and angle  $b$  as follows:

$$\sigma \cong \frac{(5.83 \cdot 10^{-16})(f^2 \cos b)}{p_1} \quad (59)$$

$$\epsilon_r \cong x \tan(b) - 1 \quad (60)$$

These equations and Fig 21 assume a plane homogeneous earth that limits the usefulness of this procedure to distances not greater than  $8000/f^{1/3}$  km. Due to irregularities of data caused by terrain effects and variable conductivity, it may be difficult to determine which of the curves represents a given condition most accurately.

The above measurement process is repeated in as many directions as are required to establish the complete broadcast groundwave radiation pattern. Six to ten radials are usually sufficient. The radius of a circle, the area of which is equal to the area bounded by this pattern, leads to the effective unabsorbed field strength at 1 km. This value corresponds to  $E_1$  in Eq 53.  $E_2$  is determined from Fig 20 for an ideal antenna having the same effective length. The effective radiated power  $P$ , in W, may then be computed from Eq 53.

The radiation efficiency of an antenna is defined as the percentage of input power that is actually radiated. That is,

$$\eta = 100 \frac{P_e}{P_i} \quad (61)$$

where

$P_e$  = effective isotropically radiated power (EIRP)

$P_i$  = input power to the transmitting antenna terminals

### 3.3 Ground-Wave Effective Radiated Power Measurements Above 5 MHz

If the field strength of the direct wave can be measured at a known distance  $d$  from the transmitting antenna, the effective isotropically radiated power (EIRP) in that direction can be determined (see [B2] , [B4] and [B21] ) from Eq 35 in the form

$$\text{EIRP} \cong P_T g_T = \frac{E_o^2 d^2}{30} \quad (62)$$

where

$\text{EIRP}$  = effective isotropically radiated power in a given direction, in W  
 $P_T$  = transmitter power, in W  
 $g_T$  = gain of transmitting antenna in a given direction  
 $E_o$  = free space direct-wave field strength, in V/m  
 $d$  = distance, in m

This “direct-wave” approach can also be used below 5 MHz when the numerical distance  $p$  is small enough that  $E \propto 1/d$  for  $E$  measured at the surface of the earth.

In some cases, especially at frequencies above 200 MHz, the height, distance, and directivity can be made such that the ground-reflected ray is negligible and the magnitude of the direct ray can be determined by measurement of the field strength. In case the ground-reflected ray is not negligible, it is sometimes possible to determine the direct ray by measuring the field strength at a series of heights or a series of distances, and averaging the measured field-strength values by either a graphical or analytical process, see [B11].

In other cases where it is not possible to avoid the ground-reflected ray, the effective radiated power in a given direction can be measured by comparing the received field strength from the antenna under test (AUT) with the field strength received when a known amount of power is radiated by a standard antenna of known gain from the same transmitting antenna location. This method is valid if the vertical directivity of the AUT and standard antenna are equal, and thus the effect of the ground-reflected ray is the same for both cases.

For frequencies up to about 50 MHz, a loop antenna may be used as a standard transmitting antenna. For frequencies between 25 MHz and 1 GHz, a thin half-wave dipole is a convenient form of standard transmitting antenna. For frequencies above 200 MHz, it is generally desirable to use a more directive aperture antenna or array as a standard antenna, which in turn requires a determination of its gain. (See 2.5.3 and [B5] .)

The field radiated by an N-turn loop in free space, at a distance d greater than several wavelengths (see [B25] ), is given by

$$E_o = \frac{\pi \eta_o N A I}{\lambda^2 d} \quad (63)$$

Combining Eqs 62 and 63, an expression for the effective power radiated by a loop in directions of maximum radiation in free space is

$$\text{EIRP} = \frac{4\pi^3 \eta_o N^2 A^2 I^2}{\lambda^4} \quad (64)$$

For a thin half-wave dipole, the power gain relative to an isotropic radiator in directions of maximum radiation in free space is 1.64 (see 2.5.2.2). The maximum effective power radiated is therefore 1.64 times the power input to the dipole. In general, the EIRP in the direction of maximum radiation in free space is the power gain of the standard antenna multiplied by the power input to the antenna.

If the path geometry and vertical directivity of a standard antenna and the AUT are such that the ground-reflected ray has the same effect on the resultant field, the EIRP in a given direction is determined by measuring the ratio between the field strengths produced by the AUT and by the standard antenna. The effective power of the AUT is given by

$$(\text{EIRP})_x = \left( \frac{E_x}{E_s} \right)^2 (\text{EIRP})_s \quad (65)$$

where

$(\text{EIRP})_x$  = effective radiated power of the AUT in the direction of measurement. field strength produced by the transmitter and AUT in the direction of measurement.

$E_x$  = field strength produced by the transmitter and AUT in the direction of measurement.

$E_s$  = field strength at the same point produced by a known effective power radiated by the standard antenna over the same propagation path.

$(\text{EIRP})_s$  = effective power radiated by the standard antenna oriented for maximum radiation in the direction of measurement. The value of  $(\text{EIRP})_s$  is given by the product of power input and gain for the standard antenna.

Since only the ratio  $(E_x/E_s)$  is required, it is sufficient to measure only relative values of  $E_x$  and  $E_s$ . This comparison method, while simple in theory, requires considerable care in taking the data. One difficulty is preventing distortion of the pattern of the standard antenna by undesired coupling with the AUT or with the structure on which it is mounted.

In determining the ratio of field strengths produced in a given direction when power is fed alternately to a standard transmitting antenna and the AUT, it is desirable to make measurements at several locations along the radial in order to obtain an average value for the ratio.

When a half-wave dipole is used as the standard transmitting antenna, it may be fed through a coaxial cable if a balun is used. The balun should match the dipole impedance to the cable and system impedance, and its insertion loss should be measured. This can be conveniently accomplished in practice by measuring the insertion loss of two identical baluns back-to-back and dividing the result by 2 (or subtracting 3 dB). The dipole should be located at a height above ground equal to the height of the electric center of the AUT. The distance to the receiving point should be great enough so that a small movement of the dipole with respect to the AUT produces no appreciable change of field strength at the receiving point. If the standard antenna is mounted on a conducting tower, field perturbation may be caused by currents induced on the tower. Metal tower guy wires can cause similar effects. It has been found that a horizontal separation between the two transmitting antennas of at least twice the greatest dimension of either antenna is required.

Another method for calculating the effective radiated power of a transmitting antenna is to multiply the measured power input to the antenna by the antenna gain. The power input can be determined by measurements in the transmission line to the antenna (preferably near the antenna), and the gain can be determined by one of the methods described in 2.5.3.2 and [B8] . In the special case of stacked arrays used for TV and FM broadcasting, the common practice is to measure the vertical pattern by mounting the antenna array horizontally on a turntable located on a large area of flat ground free of obstructions that might produce spurious reflections. Interactions due to reflection from the ground immediately under the array may be minimized by the use of “space cloth” supported about 1/4 wavelength above ground. (Space cloth is material with the characteristic impedance of free space:  $v_0$ .) After the array is mounted on the tower where it is to be used, the horizontal pattern can be measured with the aid of a blimp, airplane, or helicopter, see [B13] , [B41] , and [B82] .

### 3.4 Effective Isotropically-Radiated Power for Ionospheric Wave Transmission

In the frequency range of 2-30 MHz, the ionospheric wave (also called sky wave) is often predominant over the ground wave at distances sufficiently far from the transmitter. In this case, the received radiation is usually greater at some angle to the horizon than it is along the horizon. Thus, antenna measurements made near the ground are not necessarily indicative of the primary path used for the transmission, see [B23] , [B28] , and [B80] . A method used extensively under these circumstances involves determining the amount of power that must be transmitted by a standard antenna, such as a thin half-wave dipole, to produce the same field strength at the distant receiving point as established by the AUT. The comparison dipole should be oriented for the same polarization as that of the AUT. The height and orientation (horizontal or vertical) of the comparison dipole should be stated in reporting the data, together with other information such as the ground constants at the transmitting site. The location of the comparison dipole should be such that its radiation is not affected by the presence of the AUT or other nearby antennas or by obstacles such as buildings, towers, transmission lines, etc.

The general procedure is to transmit alternately at short intervals (from a second to several minutes) on the standard dipole and the AUT with known power levels into each, and record the received field strength at the distant measurement point so as to average out the effects of fading. This technique can also be used to measure the relative gain of receiving antennas by switching receiving antennas at 10 s intervals and using a sampling algorithm, see [B41] . Since the status of the ionosphere is variable with time of day, the measurements should be made over the various hours of interest.

## 4. Transmitting Antenna Effectiveness in Ground-Wave Transmission and Equivalent Gain

### 4.1 Relative Communication Efficiency (RCE)

The effectiveness of a test antenna in launching a ground wave can be determined by first calculating a ratio called the RCE (see [B29] ) that compares the test antenna with a short dipole as each radiates from the same point of transmission. A procedure for computing the RCE at a distance  $d$  from the transmission point is given in [B26] using the expression

$$\text{RCE} = \frac{|E_t|^2 P_d}{|E_d|^2 P_t}$$

where

- $E_t$  = electric field strength at distance  $d$  from the test antenna, in V/m
- $P_t$  = power delivered to the test antenna to establish  $E_t$ , in W
- $E_d$  = electric field strength at distance  $d$  from the short dipole, in V/m
- $P_d$  = power delivered to the short dipole to establish  $E_d$ , in W

In [B50], the quantities  $E_t$ ,  $P_t$ ,  $E_d$ , and  $P_d$  are computed by an antenna algorithm based on values for these quantities previously obtained from the computer program Numerical Electromagnetics Code, version 3 (NEC-3), see [B16] - [B18]. Parameters required for computing  $E_t$ ,  $P_t$ ,  $E_d$ , and  $P_d$  are the ground conductivity, ground relative permittivity (dielectric constant), frequency, azimuth, and the geometries of the test antenna and short dipole, including their height above ground.

## 4.2 Equivalent Gain

An equivalent gain is defined in [B26] as a measure of the effectiveness of the test antenna in transmitting ground waves. This equivalent gain is obtained by multiplying the RCE by the gain of a short dipole antenna at the *surface* of a lossy ground  $g_d$  given in [B84] as

$$g_d = \frac{E^2 d^2}{(173)^2 P A^2} \quad (67)$$

where

$E$ , in mV/m, is established by the dipole at a distance  $d$  in km, with a dipole input power  $P$  in kW.

The term  $A$  is the Norton approximation (see [B70]) to the Sommerfeld attenuation function (see [B81]), and takes into account the effect of ground loss.

The antenna algorithm computes  $g_d$  using the ground and antenna parameters required in evaluating the RCE. The product of Eqs 66 and 67 is the equivalent gain  $g_e$  of the test antenna over a lossy ground and referenced to an isotropic radiator in free space. In dB relative to an isotropic antenna,  $G_e = 10 \log_{10} g_e$ :

$$G_e(\text{dBi}) = G_d(\text{dBi}) + \text{RCE}(\text{dB}) \quad (68)$$

The term  $A$  (see pages 804-805 in [B84]) is a function of the parameters given in Eqs 54-58 and is the same quantity that is plotted in Fig 19 as "relative field strength." The common source of these relationships is Norton's analysis of propagation over a lossy ground, see [B70].

## 5. Miscellaneous

### 5.1 Presentation of Field Strength Measurement Data

In view of the several factors that influence measured values of field strength, information about the following should be provided (see [B20] - [B22]):

- 1) The field component measured
- 2) The measurement location and antenna setup, including topography, electrical properties of the ground, proximity of disturbing structures, orientation of the receiving antenna, its height above ground, etc.
- 3) The measuring equipment and procedure
- 4) In case of fading fields, the time distribution of field strengths throughout the time occupied by the measurements
- 5) For measurements made near the ground, a statement of which field was measured, i.e., incident or resultant
- 6) Estimated accuracy of the measurement
- 7) For modulated waves, the bandwidth and detector characteristics of the receiver
- 8) The background noise level and signal-to-noise ratio during the test

## 5.2 Automatic Recording of Data

When measuring fields of variable strength, such as ionospheric fields and electromagnetic noise fields, the acquisition of data is facilitated by using automatic recording methods, see [B90] , [B35] , [B41] , [B80] , and [B84] . The rectified output of the receiver intermediate-frequency (IF) amplifier is often used for automatic gain control (AGC) and also for operating a recorder. A logarithmic (dB) relation between the recorder deflection and field strength is convenient in view of the great variation of signal strength often encountered. The optimum time constant for the recorded data depends on the use to be made of the data. For example, one common use is to show the effects of variation in the transmission medium from hour to hour, season to season, etc.

One useful presentation of measured data for fading fields is a graph of field strength values exceeded for various percentages of the time. The distribution of field strengths over a short time will usually correspond to the Rayleigh distribution curve (see [B74] ) and will remain the same for time periods during which absorption, turbulence, and path geometry remain constant (i.e., for periods of a few minutes up to one hour). The variations caused by changes in the propagation medium are usually much larger than the short-time variations due to phase interference. An integrating circuit of about 1 min charge and discharge time constant is useful for many purposes. It is often desirable to plot field strength versus percent-of-time distribution curves by means of automatic analyzing equipment. The curves indicate the total time that the field strength has been equal to or greater than each of several calibrated field-strength levels, see [B35] .

For propagation studies, it is desirable to have a site with a low ambient noise level that is free of disturbing reflections. Continuous recording for determining wave propagation in connection with a broadcast service in the frequency range 3 MHz to 1 GHz is generally made with receiving antennas 30 ft above ground. Automatic field-strength recording equipment may also be operated in a moving vehicle (see [B32] , [B43] , and [B73] ) to correlate field strength data with distance, azimuth, or geographical location. If a peak-reading detector is used for pulse-modulated transmissions, the overall time constant of detector and recorder should be such as to indicate the peaks of the pulses but not the peaks of any standing-wave pattern through which the vehicle passes. The standing-wave pattern is a function of both the signal frequency and speed of the vehicle. Modern data-acquisition systems involve analog-to-digital (A/D) conversion and magnetic or optical storage of the data. The data from such systems can be digitally processed to simulate various detectors and time constants.

## 5.3 Accuracy of Field Strength Measurements

The accuracy attainable in practice depends not only on the technique and instrumentation used but also on the conditions under which the measurements are made, see [B10] , [B21] , [B39] . Assuming ideal facilities (weather, personnel, etc.) an accuracy of  $\pm 1/4$  dB can be attained for H field calibrations at frequencies below 5 MHz. The error increases to  $\pm 1$  dB for E field measurements at frequencies above 30 MHz. Under field operating conditions, the errors are often greater. It is desirable that any reported data be accompanied by a statement of the estimated accuracy of measurement.

Table 3 gives a summary of typical echelon designations of accuracy for field strength measurements and calibrations. Tables 46 list the estimated accuracies available in each of the three echelons over the designated ranges of magnitude and frequency. Experience in actual comparisons of accuracy between the echelons is very limited, both in the number of comparisons and in the range of magnitudes and frequencies that have been compared. The values given in the tables are based on the combined information and judgment of the members of the Subcommittee on Field Strength Measurement of the Wave Propagation Standards Committee of the IEEE Antennas and Propagation Society. In some cases, there is partial overlapping of the ranges of values given for two echelons. At a given magnitude and frequency, a higher echelon is always expected to have the higher accuracy.

**Table 3—Echelon Designations for Levels of Field Strength Calibration Accuracy**

Accuracy of Designation	Calibration Description	Typical Agencies or Facilities Operating at This Level	Reference Standards Used or Maintained at This Level
Echelon I	Highest available within a country	Specialized laboratories, including government agencies. In the USA, the National Institute of Standards and Technology for electrical quantities.	Direct reference standards may be established, or calibration may be performed with other parameters for which standards exist, such as frequency, length, and time.
Echelon II	Intermediate level, which may be divided into sublevels	Standards laboratories or universities, military services, industrial research and development laboratories, and standard laboratories of instrument manufacturers.	Standards calibrated by agency operating at the Echelon I level of accuracy. Other standards of reduced accuracy may be obtained by calibrations within the Echelon II level of accuracy. The number of permissible sublevels is limited by the accuracy required in Echelon III operations.
Echelon III	Levels at which measuring instruments are calibrated prior to use by the ultimate customer	1) Production line test departments and service departments of instrument manufacturers 2) Instrument repair and calibration facilities of instrument users	Standards calibrated with an accuracy level obtained at the Echelon II level. The reference standard used should be cited in a calibration certificate

**Table 4—Accuracies Attainable for Calibration of Continuous-Wave Field Strength Echelon I Agency**

(This agency calibrates in terms of the national standards)

Frequency Range V/m	30 Hz to 5 MHz		5 MHz to 30 MHz		30 MHz to 1000 MHz	1000 MHz and above
	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	± Percent	
10 <sup>-6</sup> to 10 <sup>-4</sup>	6 to 8	Calibration	8 to 10	Calibration	15 to 7	Calibration
10 <sup>-4</sup> to 10 <sup>-2</sup>	4 to 6	service	6 to 8	service	13 to 15	service
10 <sup>-2</sup> to 1	3 to 4	not available	5 to 6	not available	12 to 13	not available
1 to 10	4 to 5	available	6 to 7	available	13 to 14	available

**Table 5—Accuracies Attainable for Calibration of Continuous-Wave Field Strength Echelon II Agencies—Standards Laboratories of Manufacturers of Field-Strength Meters, Government Agencies, etc.**

(These agencies use standards calibrated by Echelon I agencies to calibrate other standards.)

Frequency Range	10 kHz to 5 MHz		5 MHz to 30 MHz		30 MHz to 1000 MHz	1000 MHz to 10 000 MHz
V/m	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	± Percent	± Percent
$10^{-6}$ to $10^{-4}$	7 to 10	10 to 25	9 to 20	15 to 40	16 to 25	16 to 25
$10^{-4}$ to $10^{-2}$	5 to 7	10 to 25	7 to 10	15 to 30	14 to 20	14 to 20
$10^{-2}$ to 1	4 to 5	10 to 25	6 to 8	15 to 30	13 to 15	13 to 15
1 to 10	5 to 6	10 to 25	8 to 11	15 to 30	15 to 17	15 to 17

**Table 6—Accuracies Attainable for Calibration of Continuous-Wave Field Strength Echelon In Agencies — Manufacturers of Field-Strength Meters, and Maintenance Areas of Users**

(These agencies calibrate equipment for ultimate users in terms of standards obtained from Echelon II agencies. The calibration accuracies listed above should be attained in production-line field strength meters.)

Frequency Range	10 kHz to 5 MHz		5 MHz to 30 MHz		30 MHz to 1000 MHz*	1000 MHz to 10 000 MHz*		
V/m	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	Loop Antenna, ± Percent	Short Monopole Antenna, ± Percent	± Percent	± dB	± Percent	+ dB
$10^{-6}$ to $10^{-4}$	10 to 20	15 to 40	15 to 40	20 to 40	20 to 50	1.5 to 3.5	30 to 50	2.5 to 3.5
$10^{-4}$ to $10^{-2}$	7 to 15	10 to 30	10 to 20	20 to 40	20 to 40	1.5 to 3	30 to 50	2.5 to 3.5
$10^{-2}$ to 1	5 to 10	10 to 30	8 to 15	20 to 40	20 to 40	1.5 to 3	30 to 50	2.5 to 3.5
1 to 10	8 to 12	15 to 40	12 to 18	20 to 40	20 to 40	1.5 to 3	30 to 50	2.5 to 3.5

\*Both percentage and dB values are included because the percentage figures are rather large



## 6. Bibliography

- [B1] ANSI C63.2-1987, American National Standard for Electromagnetic Noise and Field Strength, 10 kHz to 40 GHz Specifications.
- [B2] ANSI C63.4-1991, American National Standard for Methods of Measurement of Radio-Noise Emission from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz.
- [B3] ANSI C63.5-1988, American National Standard for Calibration of Antennas Used for Radiated Emission Measurement in Electromagnetic Interference (EMI) Control.
- [B4] ANSI/CISPR 16-1987, Specification for Radio Interference Measuring Apparatus and Measurement Methods.
- [B5] IEC 106 (1974), Recommended Methods of Measurement of Radiated and Conducted Interference From Receivers for Amplitude-Modulation, Frequency-Modulation and Television Broadcast Transmissions.
- [B6] IEC 597, — Aerials for the Reception of Sound and Television Broadcasting in the Frequency Range 30 MHz to 1 GHz (Parts 1–4).
- [B7] IEEE Std 139-1988, IEEE Recommended Practice for Measurement of Radio Frequency Emission from Industrial, Scientific, and Medical (ISM) Equipment Installed on User's Premises (ANSI).
- [B8] IEEE Std 149-1979 (Reaf 1990), IEEE Standard Test Procedures for Antennas (ANSI).
- [B9] IEEE Std 187-1990, IEEE Standard on Radio Receivers: Open Field Method of Measurement of Spurious Radiation from FM and Television Broadcast Receivers (ANSI).
- [B10] IEEE Std 189-1955 (W), IEEE Standard Method of Testing Receivers Employing Ferrite Core Loop Antennas.
- [B11] MIL-STD-461A, Military Standard Requirements for Equipment Electromagnetic Interference Characteristics, Aug. 1985.
- [B12] Ballantine, S. "High Quality Radio Broadcasting and Transmission," *Proceedings of the IRE*, vol. 22, pp. 564-629, May 1934.
- [B13] Barker, G. E. "Measurement of the Radiation Patterns of Full-Scale HF and VHF Antennas," *IEEE Transactions on Antennas and Propagation*, vol. AP-21, no. 4, pp. 538-544, Jul. 1973.
- [B14] Barrow, W. L. and Greene, F. M. "Rectangular Hollow-Pipe Radiators," *Proceedings of the IRE*, vol. 26, no. 12, pp. 1498-1519, Dec. 1938.
- [B15] Bowman, R. R. "Calibration Techniques for Electromagnetic Hazard Meters: 500 MHz to 20 GHz," National Bureau Standards (US) NBSIR 75-805, 33 p., Apr. 1976.
- [B16] Breakall, J. K., Burke, G. J. and Miller, E. K. "The Numerical Electromagnetics Code (NEC)," Proceedings of the 6th Symposium and Technical Exhibition on Electromagnetic Compatibility, pp. 301-308, Zurich, Mar. 1985.
- [B17] Burke, G. J. "Numerical Electromagnetics Code — User's Guide Supplement for NEC-3 for Modelling Buried Wires," Lawrence Livermore Laboratory, Livermore, CA, Oct. 1983.
- [B18] Burke, G. J. and Poggio, A. J. "Numerical Electromagnetics Code (NEC)-Method of Moments," NOSC Technical Document 116, vol. 1, Contract MIPR-N0095376MP, Lawrence Livermore Laboratory, Livermore, CA, Jan. 1981.

- [B19] Camell, D. G., Larsen, E. B., and Cruz, J. E. "NBS Calibration Procedures for Horizontal Dipole Antennas (25 to 1000 MHz)," NBS Technical Note 1309, U.S. National Bureau of Standards, 43 pp., Apr. 1987.
- [B20] CCIR, "Characteristics and Applications of Atmospheric Radio Noise Data," Report 322-3, Recommendations and Reports of the CCIR, 1986, 16th Plenary Assembly, vol. VI, International Communication Union, Geneva, Switzerland, 1986.
- [B21] CCIR, "General Methods of Measuring the Field Strength and Related Parameters," Report 227-3, vol. V, Documents of the 10th Plenary Assembly, International Intercommunication Union, Geneva, Switzerland, pp. 27–35, 1986.
- [B22] CCIR, "Measurement of Atmospheric Radio Noise From Lightning," Report 254-6, vol. VI, Documents of the 16th Plenary Assembly, International Telecommunication Union, Geneva, Switzerland, pp. 228–235, 1986.
- [B23] CCIR, "Measurement of Sky-Wave Signal Intensities at Frequencies Above 1.6 MHz," Report 253-4, vol. VI, Documents of the 16th Plenary Assembly, International Telecommunication Union, Geneva, Switzerland, pp. 245–256, 1986.
- [B24] Chapin, E. W. "Field Strength Measurements," Federal Communications Commission Laboratory Division, Report 6.3.1, Oct. 1, 1958.
- [B25] Cutler, C. C., King, A. P., and Kock, W. E. "Microwave Antenna Measurements," *Proceedings of the IRE*, vol. 35, pp. 1462-1471, Dec. 1947.
- [B26] DeMinco, N. "Automated Performance Analysis Model for Ground-Wave Communication Systems," National Telecommunications and Information Administration (US) NTIA Report 86–209, 111 p., Dec. 1986.
- [B27] Diamond, H., Norton, K. A., and Lapham, E.G. "On the Accuracy of Radio Field-Intensity Measurements at Broadcast Frequencies," *Journal of Research NBS*, vol. 21, pp. 795–818, Dec. 1938, RP1156.
- [B28] FCC *Rules and Regulations*, vol. 3, pt. 73, "Radio Broadcast Services," Jan. 1964.
- [B29] Fenwick, R. C. and Weeks, W. L. "Submerged Antenna Characteristics," *IEEE Transactions on Antennas and Propagation*, vol. AP-11, pp. 296–305, May 1963.
- [B30] FitzGerrell, R. G. "Gain Measurements of Vertically Polarized Antennas Over Imperfect Ground," *IEEE Transactions on Antennas and Propagation*, AP-15, no. 2, Mar. 1967a.
- [B31] FitzGerrell, R. G. "The Gain of a Horizontal Half-Wave Dipole Over Ground," *IEEE Transactions on Antennas and Propagation*, AP-15, pp. 569-571, Jul. 1967b.
- [B32] Fricke, H. "Field Strength Measurements in High Frequency Fields," *Arch. Tech. Messen.*, pp. 15–16, Jan. 1959; pp. 89–92, May 1959; pp. 133–136, Jul. 1959.
- [B33] Gaffney, F. J. "Microwave Measurements and Test Equipments," *Proceedings of the IRE (Waves and Electrons)*, vol. 34, pp. 775–793, Oct. 1946.
- [B34] George, R. W. "Field-Strength Measuring Equipment at 500 Megacycles," *RCA Review*, vol. 5, p. 69, Jul. 1940.
- [B35] George, R. W. "Signal Strength Analyzer," *Electronics*, vol. 24, pp. 75–77, Jan. 1951.
- [B36] Grantham, R. E. and Freeman, J. J. "A Standard of Attenuation for Microwave Measurements," *AIEE Transactions*, vol. 67, pt. 1, pp. 329335, Jun. 1948.

- [B37] Greene, F. M. "Calibration of Commercial Radio Field-Strength Meters at the National Bureau of Standards," *NBS Circular*, vol. 517, Dec. 1951.
- [B38] Greene, F. M. "The Effect of the Ground on the Calibration and Use of VHF Field Intensity Meters," *Journal of Research NBS*, vol. 44, Feb. 1950, RP2062.
- [B39] Greene, F. M. and Solow, M. "Development of VHF Field-Intensity Standards," *Journal of Research NBS*, vol. 44, May 1950, RP2100.6262.
- [B40] Hagn, G. H. "Ground Constants at High Frequencies (HF)," Proceedings of Third Annual Review of Progress in Applied Computational Electromagnetics, Monterey, CA, March 1987.
- [B41] Hagn, G. H. and Harnish, L. O. "Measurement Techniques for HF Tactical Antennas," Proceedings of the DARPA-AFCEA-IEEE Technical Conference on Tactical Communications, The Next Generation, Fort Wayne, Indiana, 22–24 Apr. 1986.
- [B42] Harnish, L. O. Lee, M. and Hagn, G. H. "Comparison of Measured and NEC-Calculated Characteristics of a Vertical Monopole with Buried Ground Radials," Proceedings of Third Annual Review of Progress in Applied Computational Electromagnetics, Monterey, CA, 24–26 Mar. 1987.
- [B43] Head, H. T. and Prestholdt, O. L. "The Measurement of Television Field Strengths in the VHF and UHF Bands," *Proceedings of the IRE*, vol. 48, pp. 1000–1008, Jun. 1960.
- [B44] Hill, D. A. and Francis, M. H. "Out-of-Band Response of Antenna Arrays," *IEEE Transactions on Electromagnetic Compatibility*, vol. EMC-29, no. 4, pp. 282–288, Nov. 1987.
- [B45] Isely, F. C. "A New Approach to Tunable Resonant Circuits for the 30- to 3000-Mc Frequency Range," *Proceedings of the IRE (Waves and Electrons)*, vol. 36, pp. 1017–1022, Aug. 1948.
- [B46] Jakes, W. C., Jr. "Gain of Electromagnetic Horns," *Proceedings of the IRE*, vol. 39, pp. 160–162, Feb. 1951.
- [B47] Jesch, R. L. "Measured Vehicular Antenna Performance," *IEEE Transactions on Vehicular Technology*, VT-34, pp. 97–107, 1985.
- [B48] Jones, T. L. and Colebrook, F. M. "An Appraisal of Laboratory Radio Measuring Techniques," *Journal of the IEE*, vol. 95, pt. 3, p. 315, 1948.
- [B49] Jull, E. V. *Aperture Antenna and Diffraction Theory*, Stevenage: Peter Peregrinus Ltd., 1981.
- [B50] Jull, E. V. "Errors in the Predicted Gain of Pyramidal Horns," *IEEE Transactions on Antennas Propagation*, vol. AP-21, no. 1, pp. 25–31, Jan. 1973.
- [B51] Jull, E. V. "Finite-Range Gain of Sectoral and Pyramidal Horns," *Electronic Letters*, vol. 6, pp. 680–681, Oct. 15, 1970.
- [B52] Kanda, M. "Analytical and Numerical Techniques for Analyzing an Electrically Dipole with a Nonlinear Load," *IEEE Transactions on Antennas and Propagation*, AP-28, pp. 71–78, Jan. 1980.
- [B53] Kanda, M. "An Electromagnetic Near-Field Sensor for Simultaneous Electric and Magnetic Field Measurements," *IEEE Transactions on Electromagnetic Compatibility*, EMC-26, pp. 102–110, Aug. 1984.
- [B54] Kanda, M. "A Relatively Short Cylindrical Broadband Antenna With Tapered Resistive Loading for Picosecond Pulse Measurements," *IEEE Transactions on Antennas and Propagation*, AP-26, pp. 439–447, May 1978.

- [B55] Kanda, M. and Orr, R. D. "A Radio-Frequency Power Delivery System: Procedures for Error Analysis and Self-Calibration," National Bureau of Standards (US) Technical Note 1083, August 1985.
- [B56] Kanda, M. and Driver, L. D. "An Isotropic Electric-Field Probe with Tapered Resistive Dipoles for BroadBand Use, 100 kHz to 18 GHz," *IEEE Transactions on Microwave Theory and Techniques*, MTT-35, pp. 124–130, Feb. 1987.
- [B57] Kanda, M. and Orr, R. D. "Near-Field Gain of a Horn and an Open-Ended Waveguide: Comparison Between Theory and Experiment," *IEEE Transactions on Antennas Propagation*, vol. AP-35, no. 1, pp. 33–40, Jan. 1987.
- [B58] Kanda, M., Ries, F. X., Driver, L. D., and Orr, R. D. "Design Considerations for Broadband Magnetic-Field Sensors," *IEEE Digest; Conf. on Precision Electromagnetic Measurements*, Boulder, Colorado, Jun. 1982.
- [B59] Karplus, E. "Wide-range Tuned Circuits and Oscillators for High Frequencies," *Proceedings of the IRE*, vol. 33, pp. 426–440, Jul. 1945.
- [B60] King, D. D. "Two Standard Field-Strength Meters for Very-High Frequencies," *Proceedings of the IRE*, vol. 38, pp. 1048–1051, Sep. 1950. Correction, *Proceedings of the IRE*, vol. 39, p. 174, Feb. 1951.
- [B61] Kraus, J. D. *Antennas*, New York: McGraw-Hill, 1950.
- [B62] Larsen, E. B. and Ries, F. X. "Design and Calibration of the NBS Isotropic Electric-Field Monitor [EFM-5], 0.2 to 1000 MHz," National Bureau Standards (US) Technical Note 1033, Mar. 1981.
- [B63] Lesk, I. A. and Suran, J. J. "A Circularly Polarized Corner Reflector Antenna," *Electrical Engineering*, vol. 79, pp. 270–277, Apr. 1960.
- [B64] Levin, E., Muchmore, R. B., and Wheelon, A.D. "Aperture-to-Medium Coupling on Line-of-Sight Paths: Fresnel Scattering," *IRE Transactions on Antennas and Propagation*, vol. AP-7, pp. 142–146, Apr. 1959.
- [B65] Magrab, E. B. and Blomquist, D. S. *The Measurement of Time-Varying Phenomena*, New York: Wiley Interscience, 1971.
- [B66] Maley, S. W. and King, R. J. "Impedance of a Monopole Antenna with a Radial-Wire Ground System on an Imperfectly Conducting Half-Space, part II," *Journal of Research NBS*, vol. 68D (Radio Science), no. 2, pp. 157–163, Feb. 1964.
- [B67] Montgomery, C. G., Ed., *Technique of Microwave Measurements*, MIT Rad. Lab. Ser. vol. 11. New York: McGraw-Hill, 1949.
- [B68] Nakano, H. *Helical and Spiral Antennas—A Numerical Approach*, New York: John Wiley & Sons, 1987.
- [B69] Newell, A. C., Baird R. C., and Wacker, P. F. "Accurate Measurements of Antenna Gain and Polarization at Reduced Distances by an Extrapolation Technique," *IEEE Transactions on Antennas and Propagation*, vol. AP-21, pp. 418–431, Jun. 1973.
- [B70] Norton, K. A. "The Calculation of Ground-Wave Field Intensity Over a Finitely Conducting Spherical Earth," *Proceedings of the IRE*, vol. 29, no. 12, pp. 623–639, 1941.
- [B71] Norton, K. A. "System Loss in Radio Wave Propagation," *Journal of Research NBS*, Jul. 1959.
- [B72] Norton, K. A. "Transmission Loss in Radio Propagation," pt. 1, *Proceedings of the IRE*, vol. 41, pp. 146–152, Jan. 1953; pt. 2, NBS Technical Note 12, Jun. 1959.
- [B73] Preston, J. G. "Measuring TV Field Intensities by Helicopter," *TeleTech*, vol. 11, p. 64, Jun. 1962.

- [B74] Lord Rayleigh, "On the Resultant of a Large Number of Vibrations of the Same Pitch and of Arbitrary Phase," *Phil. Magazine*, vol. 10, pp. 73–78, 1880. Also, *Theory of Sound*, 2nd ed. London: Macmillan, 1894, paragraph 42a. This problem is also treated in the "Collected Works" of Lord Rayleigh.
- [B75] Reed, H. R. and Russell, C. M. *UHF Propagation*, 2nd ed. Boston: Boston Technical Publishers, 1964.
- [B76] Saxton, J. A. and Harden, B. N. "The Performance of Directive Arrays in Complex UHF Fields," *Proceedings of the IEE*, vol. 106, pt. B, p. 315, May 1959.
- [B77] Schelkunoff, S. A. and Friis, H. T. *Antennas Theory and Practice*, New York: John Wiley & Sons, 1952.
- [B78] Silver, S., Ed. *Microwave Antenna Theory and Design*, MIT Rad. Lab. Ser., New York: McGraw-Hill, 1949.
- [B79] Smith, G. S. "Limitations on the Size of Miniature Electric-Field Probes," *IEEE Transactions on Microwave Theory and Techniques*, MTT-32, pp. 594–600, Jun. 1984.
- [B80] Smith, W. B. "Recording Sky-Wave Signals from Broadcast Stations," *Electronics*, vol. 21, p. 112, Nov. 1945.
- [B81] Sommerfeld, A. "Propagation of Waves in Wireless Telegraphy," *Ann D. Phys.*, vol. 28, pp. 665–736, March 1909.
- [B82] Stehle, R. H., Des Brisay, G. A., Whitson, A. L., and Hagn, G. H. "RELEDOP: A Full-Scale Antenna Pattern Measurement System for High-Powered HF Transmitting Antennas," *IEEE Transactions on Broadcasting*, vol. 34, no. 2, pp. 210–220, Jun. 1988.
- [B83] Tai, C. T. "On the Definition of the Effective Aperture of Antennas," *IEEE Transactions on Antennae and Propagation*, vol. AP-9, pp. 224–225, 1961.
- [B84] Terman, F. E. *Electronic and Radio Engineering*, New York: McGraw-Hill, 1955, pp. 803–804.
- [B85] Terman, F. E. and Pettit, J. M. *Electronic Measurements*, New York: McGraw-Hill, 1951, ch. 11.
- [B86] Van Voorhis, S. N., Ed., *Microwave Receivers*, M.I.T. Rad. Lab. Ser., vol.23. New York: McGraw-Hill, 1948.
- [B87] Vogler, L. E. and Noble, J. L. "Curves of Ground Proximity Loss for Dipole Antennas," NBS Technical Note 175, 1963.
- [B88] Wait, J. R. and Surtees, W. J. "Impedance of a Top Loaded Antenna of Arbitrary Length Over a Circular Grounded Screen," *Journal of Applied Physics*, vol.25, p. 553, 1954.
- [B89] Wait, J. R. "Effect of the Ground Screen on the Field Radiated from a Monopole," *IEEE Transactions on Antennas and Propagation (Communications)*, vol. AP-4, pp. 179–181, Apr. 1956.
- [B90] Watt, A.D., et al. "Performance of Some Radio Systems in the Presence of Thermal and Atmospheric Noise," *Journal of Research NBS*, pt. B, Jul. 1959.
- [B91] Whiteside H., and King, R. W. P. "The Loop Antenna as a Probe," *IEEE Transactions on Antennas and Propagation*, vol. AP-12, pp. 291–297, May 1964.
- [B92] Wilson, A. C. and Cottony, H. V. "Radiation Patterns of Finite-Size Corner-Reflector Antennas," *IRE Transactions on Antennas and Propagation*, vol. AP-8, pp. 144–157, Mar. 1960.
- [B93] Wind, M. *Handbook of Electronic Measurements*, vol. 2, Polytechnic Institute of Brooklyn, New York: Interscience, 1956, ch. 8.