# DRAFT Recommended Practice for Measurements and Computations with Respect to Human Exposure to Radiofrequency Electromagnetic Fields, 3 kHz to 300 GHz

Sponsor

#### IEEE Standards Coordinating Committee 28 on Non-Ionizing Radiation Hazards

**Abstract:** IEEE C95.3-199X specifies techniques and instrumentation for the measurement of potentially hazardous electromagnetic (EM) fields both in the near field and far field of the electromagnetic source. In doing so, this document combines and extends the specifications previously set forth in IEEE C95.3-1991 [IEEE, 1991]. This Draft *Recommended Practice* includes leakage and near-field measurements. This also document contains a description of the concepts, techniques, and instruments that can be applied to the measurement of specific absorption rate (SAR) or the electric field strength in organisms (including humans) exposed to electromagnetic fields. Below 100 MHz, the current flowing through the body to ground is measurable and can be used to determine the SAR and, therefore, a brief treatment of the low-frequency body current measurement is also included.

This DRAFT *Recommended Practice* is intended primarily for use by engineers, biophysicists, and other specialists who are familiar with basic electromagnetic (EM) field theory and practice, and the potential hazards associated with EM fields. It will be most useful to bioeffects researchers, instrument developers and manufacturers, those developing calibration systems and standards, and individuals involved in critical hazard assessments or surveys.

**Keywords:** Electromagnetic field measurements, electromagnetic field computation, exposure assessment, RF/microwave survey instruments, RF/microwave surveys, RF/microwave hazard assessment, non-ionizing radiation.

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# Foreword

(This Foreword is not a part of IEEE C95.3-199X, DRAFT Recommended Practice for Measurements and Computations with Respect to Human Exposure to Radiofrequency Electromagnetic Fields, 3 kHz to 300 GHz.)

In 1960, the American Standards Association approved the initiation of the Radiation Hazards Standards project under the co-sponsorship of the Department of the Navy and the Institute of Electrical and Electronics Engineers.

Prior to 1988, C95 standards were developed by an accredited standards committee C95, and submitted to ANSI for approval and issuance as ANSI C95 standards. Between 1988 and 1990, the committee was converted to Standards Coordinating Committee 28 under the sponsorship of the IEEE Standards Board. In accordance with policies of the IEEE, C95 standards will be issued and developed as IEEE standards, as well as being submitted to ANSI for recognition.

The present scope of IEEE SCC28 is:

"Development of standards for the safe use of electromagnetic energy in the range of 0 Hz to 300 GHz relative to the potential hazards of exposure of man, volatile materials, and explosive devices to such energy. It is not intended to include infrared, visible, ultraviolet, or ionizing radiation. The committee will coordinate with other committees whose scopes are contiguous with SCC28."

The IEEE Standards Coordinating Committee 28 is responsible for this DRAFT *Recommended Practice*. There are five subcommittees concerned with:

- I Techniques, Procedures, and Instrumentation
- II Terminology, Units of Measurements and Hazard Communication
- III Safety Levels with Respect to Human Exposure, 0-3 kHz
- IV Safety Levels with Respect to Human Exposure, 3 kHz-300 GHz
- V Safety Levels with Respect to Electro-Explosive Devices

Three standards, two guides and two recommended practices have been issued. Current versions are:

IEEE C95.1-1991, IEEE Standard for Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 3 kHz to 300 GHz. (Reaffirmed in 1997 – Replaces ANSI C95.1-1982)

ANSI C95.2-1981, American National Standard Radio Frequency Radiation Hazard Warning Symbol. Reaffirmed in 1989.

ANSI C95.3-1973, IEEE Standard Techniques and Instrumentation for the Measurement of Potentially Hazardous Electromagnetic Radiation at Microwave Frequencies. Reaffirmed in 1979.

IEEE C95.3-1991, IEEE Standard Recommended Practice for the Measurement of Potentially Hazardous Electromagnetic Fields - RF and Microwave. (Replaces ANSI C95.3-1973 and ANSI C95.5-1981.)

IEEE P1460-1996, IEEE Guide for the Measurement of Quasi-Static Magnetic and Electric Fields.

ANSI C95.4-1978, American National Standard Safety Guide for the Prevention of Radio-Frequency Radiation Hazards in the Use of Electric Blasting Caps.

ANSI C95.5-1981, American National Standard Recommended Practice for the Measurement of Hazardous Electromagnetic Fields - RF and Microwave.

Changes in the latest revision of C95.3 include an expanded frequency range, limits on induced body current to prevent radio-frequency (RF) shock or burn, a relaxation of limits on exposure to magnetic fields at low frequencies, and exposure limits and averaging time at high frequencies that

are compatible at 300 GHz with existing infrared maximum-permissible exposure (MPE) limits. Important improvements in rules for valid measurement of electromagnetic fields have been introduced, and expanded rules for relaxing the exposure limits for the case of partial body exposure have been developed. Also, a distinction is made between controlled and uncontrolled environments relative to safe exposure limits.

IEEE C95.1-1991 prescribes MPEs to prevent biological injury from exposure to electromagnetic radiation. Revisions of the original version of this standard (ANSI C95.1-1966) were made in 1974 and major revisions of ANSI C95.1-1974 were made in 1982 to take into account the significant expansion of the data base, improvements in dosimetry, and the increasing number of people in the general population exposed to RF fields. The changes in the standard included a wider frequency coverage, frequency dependence resulting from the recognition of whole-body resonance and incorporation of dosimetry. In addition to those changes, the present standard also includes a distinction between controlled and uncontrolled environments and guidelines for partial-body and near-field exposures. Exposure limits in the uncontrolled environment are lower than in the controlled environment under certain conditions, such as resonance, or when exposure is complicated by associated hazards like RF shock or burn.

IEEE C95.1-1991 contains a detailed discussion of both the rationale and the limitations of the recommended guidelines based on the present database.

IEEE C95.1-1991 was reaffirmed in 1997 and is now undergoing revision. Several issues approved by SCC-28 for the next revision were deemed sufficiently important by the committee to warrant a *Supplement* before the entire revision has been completed. This DRAFT *Supplement* modifies the induced and grasping-contact current limits specified in IEEE C95.1-1991, specifies electric field-strength below which induced and contact currents do not have to be determined, modifies measurement distance requirements and provides more precise definitions for spatial averaging, averaging volume, and radiated power as applied to hand-held wireless devices.

This DRAFT *Recommended Practice* was prepared by Subcommittee I on Techniques, Procedures and Instrumentation of IEEE Standards Coordinating Committee 28. The following persons participated on the Subcommittee I balloting committee that approved this DRAFT *Recommended Practice*:

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# DRAFT Recommended Practice for Measurements and Computations with Respect to Human Exposure to Radiofrequency Electromagnetic Fields, 3 kHz to 300 GHz

# 1. Scope

# 1.1 Purpose.

The purpose of this recommended practice is to specify techniques and instrumentation for the measurement of potentially hazardous electromagnetic (EM) fields<sup>1</sup> both in the near field and far field of the electromagnetic source. In doing so, this document combines and extends the specifications previously set forth in IEEE C95.3-1991 [IEEE, 1991]. This recommended practice includes leakage and near-field measurements. Further, IEEE Std C95.1-1991 [1]<sup>2</sup> personnel exposure standard includes specific absorption rate (SAR) in biological tissues as a parameter to be measured. Therefore, this document contains a description of the concepts, techniques, and instruments that can be applied to the measurement of SAR or the electric field strength in organisms (including humans) exposed to electromagnetic fields. Below 100 MHz, the current flowing through the body to ground is measurable and can be used to determine the SAR and, therefore, a brief treatment of the low-frequency body current measurement is also included.

This document is intended primarily for use by engineers, biophysicists, and other specialists who are familiar with basic electromagnetic (EM) field theory and practice, and the potential hazards associated with EM fields. It will probably be most useful to bioeffects researchers, instrument developers and manufacturers, those developing calibration systems and standards, and individuals involved in critical hazard assessments or surveys. Interested parties looking for a less technical document describing how to measure potentially hazardous fields may refer to NCRP Report 119 [NCRP, 1993]. However, the material in Sections 3, 4, and 5 of this document that treats measurement problems, desirable instrument characteristics, and procedures for measuring external fields should be of value to anyone concerned with EM hazards.

<sup>&</sup>lt;sup>1</sup> It is beyond the scope of this recommended practice to define what constitutes a hazardous nonionizing electromagnetic field. For the present U.S. Maximum Permissible Exposure Guide, refer to IEEE Std C95.1-1991. It should be kept in mind that the limits given in these documents are under review and are subject to change as new information becomes available.

<sup>&</sup>lt;sup>2</sup> The numbers in brackets correspond to those of the references listed in 1.5 or 1.6.

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Subcommittee 1 on Evaluation Techniques and Instrumentation was originally organized on April 7, 1960, to establish specifications for techniques and instrumentation used in evaluating hazardous radio frequency (RF) radiation. In June 1985 the scope was clarified and the purpose was extended to establish specifications for techniques and instrumentation to be used in evaluating RF hazards to mankind from exposure to manmade sources of EM radiation or from the exposure of volatile materials and explosive devices to such radiation. It is not intended to include infrared, visible, ultraviolet, or ionizing radiation.

Although the subcommittee's scope includes hazards resulting from exposure of flammable volatile materials and explosive devices to EM radiation, this document is devoted exclusively to personnel hazards. This is because the necessary specifications and definitions pertaining to fuels and explosive devices are still being developed, and there are implications that the instrumentation requirements may be substantially different from those addressed here. However, the measurement techniques and instruments described here are applicable to the measurement of fields in the vicinity of flammable materials and explosive devices, even though exposure standards for these situations have not been established.

# 1.2 Frequency Range.

The techniques and instrumentation specified herein are useful for field measurements over the frequency range of approximately 300 kHz to 100 GHz.<sup>3</sup> No single measurement technique or instrumentation arrangement is valid over the wide frequency range covered by this recommended practice. In general, measurement techniques and instrumentation developed for use in the frequency range above about 300 MHz only measure the electric field strength. Magnetic field strength measurements are, however, required to evaluate near field hazardous situations at frequencies below 30 MHz. Here, measurement of both the electric and the magnetic field strength is required within one wavelength of the source. A series of commercially available instruments has been developed for this purpose. For body currents, measurements can be made with simple, portable laboratory instruments over the frequency range of 0 to about 100 MHz. SAR can be measured using RF-transparent temperature sensors over the frequency range of 0 to about 10 GHz. Above this frequency, energy absorption is confined to the surface of a biological system. Thermographic cameras can be used to measure the surface SAR up to about 300 GHz.

# 1.3 Quantities and Parameters to be Measured.

For frequencies above 300 MHz, and for measurements performed at distances from the source greater than  $a^2/\lambda$  with a probe whose dimensions are much less than a, power density (W) is usually a meaningful quantity that has been widely adopted as a hazard indicator (a is the largest dimension of the effective aperture of the source, and  $\lambda$  is the wavelength). However, power density is difficult to determine, except in cases of stationary, single-component, plane-wave fields of known polarization. In fact, no existing instrument actually measures power density directly; all measure one or more

<sup>&</sup>lt;sup>3</sup> These frequency limits should not be considered rigid, but are given as a general guide. Normally, a given instrument does not abruptly become invalid at a specific frequency; rather its accuracy, or sensitivity, or both, deteriorate over some frequency range until finally it is no longer usable. For example, an instrument designed for the lower frequencies may cover the entire broadcast band down to 0.5 MHz. One should be aware of the frequency limitations of a given instrument when using it.

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components of the electric field strength (E), or the magnetic field strength (H), or both, and then infer an equivalent power density from the far-field, plane-wave relationship W =  $|E|^2/377$  or 377  $|H|^2$  where E is the RMS electric field strength in volts per meter, H is the RMS magnetic field strength in amperes per meter, and S is the power density in watts per meter squared. In the near field, and where scattering, reflections or multiple sources are present, the field configuration can be extremely complex and one must measure both E and H separately.<sup>4</sup> In many instances, measurements will be required in the reactive or radiating near-field regions<sup>5</sup> where reactive fields or standing waves exist. Under these more general conditions, the time-average power density is not a reliable hazard indicator, even if practical probes could be developed for measuring power density. Insofar as hazards are due to energy absorption, the parameters  $|E|^2$  and  $|H|^2$ , or the corresponding energy densities  $U_E = (1/2) (\epsilon' |E|^2)$  and  $U_H = (1/2) (\mu' |H|^2)$  are more reliable indicators of potential hazards [B124]. Here  $\varepsilon$  ' and  $\mu$ ' are, respectively, the real parts of the permittivity  $\varepsilon$  and permeability  $\mu$ , of the medium, and |E| and |H| are the RMS magnitudes of the electric and magnetic field vectors. The units of energy density are J/m<sup>3</sup>. The advantage of energy density is that the intensities of the electric fields, magnetic fields, and electromagnetic fields can all be expressed in the same units, and energy absorption is conceptually easier to relate to units of energy per unit volume than to  $|\mathbf{E}|^2$  or  $|\mathbf{H}|^2$ .

Since for plane waves all of the preceding parameters are simply related to power density, no loss of generality would occur with the use of these parameters for far-field measurements. As stated previously, some commercially-available meters actually measure  $|E|^2$  or  $|H|^2$ , while indicating equivalent plane-wave power density. Therefore, they do provide a valid indication of the exposure level and, hence, the potential hazard, even though they may not indicate the actual power density at the point of measurement. Because RF-hazard measurement instrumentation is currently not available to measure absolute peak values, except for pulses of RF of duration greater than a few milliseconds, RMS values will be considered unless otherwise stated. As the state-of-the-art advances, recommendations will be extended to include absolute peak quantities. For internal dosimetry (within body tissues) the quantities used are SAR (W/kg), internal RMS and peak electric field strength (V/m), internal current (A) and current density (A/m<sup>2</sup>).

# 1.4 Types of Situations Covered.

This recommended practice is intended for use in situations typified by the following:

# **1.4.1 External Fields**

- (1) Leakage fields
- (2) Radiation fields
- (3) Reactive fields

Leakage situations are generally interpreted as those involving the unintentional emission of RF energy, whereas radiation situations are generally interpreted as those involving

<sup>5</sup> See definitions in Section 2

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<sup>&</sup>lt;sup>4</sup> For well defined sources where analysis shows that only one field component is sufficient to show compliance with a particular maximum permissible exposure (MPE) level or radio frequency protection guide (MPE), only that component need be measured

equipment designed to intentionally radiate EM energy. Reactive fields are present in the immediate vicinity of the sources of both leakage and radiation fields and tend to be stronger near radiators of small dimensions with respect to wavelength. Multipath interference due to the scattering and reflection of energy from objects or conducting surfaces is an additional complicating factor and is nearly always present to some degree in all of the above situations. Accurate measurements should include procedures for estimating or avoiding multipath effects and near-field measurement errors. Some useful techniques for measurements in multipath environments are given in 4.5.1, 5.3.2, and 5.3.3.

# 1.4.2 Specific Absorption Rate (SAR).

The need for SAR measurements within models of exposed personnel is twofold. First, SAR is the underlying basis for many recent RF maximum permissible exposures (MPEs), including those of IEEE Std C95.1-1991 [1]. It is reasoned that the true initiator of biological effects is the internal electric field strength that leads to the concept of SAR. IEEE Std C95.1-1991 [1] external field strength MPEs are based on a whole-bodyaveraged SAR of 0.4 and 0.08 W/kg for the controlled and uncontrolled environments, respectively. In addition, a local (or spatial-peak) SAR of less than 8 and 1.6 W/kg in any gram of tissue in the shape of a cube, is the criterion for exclusion from the corresponding field strength MPEs.. Generally, accepted methods of measurement of SAR [B97] include the measurement of the rate of temperature rise within the exposed object or the measurement of the internal electric field strength. The temperature rise may be characterized by a whole-body-averaged (calorimetric) measurement, a point measurement (via a thermometer implanted in the body being exposed), or thermographic camera analyses of bisected phantom models that have been exposed to large RF fields. The internal electric field strength may be measured with an implantable E-field probe.

The second need for measuring SAR is that even under far-field plane-wave exposure conditions, the SAR varies widely with frequency and spatial location within an object. Therefore, the measurement of the SAR provides a much-needed insight into the spatial distribution of absorbed energy, and hence, the true nature of the potential hazard with respect to different organs of the exposed subject.

# 1.4.3 Measurement Problems for Internal Body Currents.

Internal body currents are induced in personnel when partial or whole-body exposures to RF fields occur. The issue of induced body currents generally becomes a consideration at lower frequencies, typically below 100 MHz, and especially below 30 MHz. Most often, the evaluation of excessive induced body current occurs in the near-field region of the RF source. For greatest accuracy in predicting what these induced currents will be in personnel, use of a full size model is required. The model is placed in the near-field region of the source in question, simulating the position, posture and size of the potentially exposed individual. Then, special measurement techniques are used to evaluate the RF-induced currents. A complication associated with evaluating the magnitude of induced RF current relates to pathways through which these currents flow in the body. For example, with electric field exposure, the induced currents flow through the body, or parts of the lowest potential surface in contact with the body). In this case, use of instrumentation, which is in effect placed in series with the body and ground, can provide a measure of these electric field induced currents. In the case of magnetic fields,

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however, the currents that are induced in the body, called eddy currents, most commonly circulate about the cross sections of the anatomy, with the greater magnitudes being at the outer periphery of the body, near the surface. These circulating currents tend not to exit the body in the same fashion as electric field induced currents and, consequently, represent a major measurement challenge. Assessments of induced currents should, however, give consideration to both electric and magnetic field contributions.

# 1.5 References.

This recommended practice should be used in conjunction with the following documents:

[1] IEEE Std C95.1-1991, IEEE Standard for Safety Levels With Respect to Human Exposure to Radiofrequency Electromagnetic Fields, 3 kHz to 300 GHz.<sup>6</sup>

[2] IEEE Std 100-1988, IEEE Standard Dictionary of Electrical and Electronics Terms (ANSI).

[3] ANSI/IME 20-1978, Safety Guide for the Prevention of Radio Frequency Radiation Hazards on the Use of Blasting Caps.<sup>7</sup>

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# 2. Definitions

**anechoic chamber.** A room or enclosure in which reflections of EM waves from the boundary surfaces have relatively small values.

**antenna.** A device designed for radiating (or receiving) electromagnetic energy. **antenna array.** A system of antennas coupled together for the purpose of enhancing radiation in one or more directions and reducing radiation in other directions. **antenna directivity.** The ratio of the transmitted RF radiation intensity in a specified direction to the radiation intensity averaged over all directions.

NOTE: The average radiation intensity is equal to the power radiated by the antenna divided by  $4\pi$ . **antenna effective aperture or effective area.** In a given direction, the ratio of the power available at the terminals of a receiving antenna to the power density of a plane wave incident on the antenna from that direction; the wave being polarization-matched to the antenna.

- NOTES: (1) A plane wave of given power density and from a given direction is polarization-matched to an antenna if its polarization is that which results in maximum power at the antenna terminals.
  - (2 The (total) effective aperture of an antenna may be decomposed into the sum of two partial effective apertures referred to two specified, orthogonal polarizations, respectively.
  - (3) For a lossless and perfectly matched antenna, effective area in a given direction is equal to the square of the operating wavelength, times the gain in that direction, divided by  $4\pi$ .

**antenna field regions.** Classification of the important spatial subdivisions of an antenna's electromagnetic field. The subdivisions, at non-uniquely defined distances from the antenna, include the reactive near-field region adjacent to the antenna, the radiating near-field region (for large antennas commonly referred to as the Fresnel region), a

transition zone and furthermost, the far-field region, also known as the Fraunhofer region. See also: near-field region and far-field region.

antenna gain (absolute gain in a given direction). The ratio of the radiation intensity, in a given direction, to the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically.

NOTES: (1) Gain does not include losses arising from impedance and polarization mismatches.

- (2) The radiation intensity corresponding to the isotropically radiated power is equal to the power accepted by the antenna divided by 4[P].
- (3) If an antenna is without dissipative loss, then in any given direction, its gain is equal to its directivity.
- (4) If the direction is not specified, the direction of maximum radiation intensity is implied.
- (5) The term "absolute gain" is used in those instances where added emphasis is required to distinguish gain from relative gain; for example, absolute gain measurements.

**antenna gain, partial (for a given polarization).** In a given direction, that part of the radiation intensity corresponding to a given polarization, divided by the radiation intensity that would be obtained if the power accepted by the antenna were radiated isotropically. NOTES: (1) The (total) gain of an antenna, in a specified direction, is the sum of the partial gains

for any two orthogonal polarizations.

(2) The gain of aperture antennas, e.g., parabolic reflectors, is usually referenced to an isotropic radiator; The gain of linear antennas, dipoles, co-linear arrays, is usually references to a half-wave dipole.

**antenna, parabolic.** An antenna consisting of a parabolic reflector and a source at or near its focus.

**antenna pattern.** The spatial distribution of a quantity that characterizes the electromagnetic fields radiated by an antenna.

NOTES: (1) The distribution can be expressed as a mathematical function or as a graphic representation.

- (2) The quantities that are most often used to characterize the radiation from an antenna are proportional to or equal to, power density, radiation intensity, directivity, phase, polarization and field strength.
- (3) The spatial distribution of the EM field characteristics over any surface is also an antenna pattern.
- (4) When the amplitude or relative amplitude of a specified component of the electric field vector is plotted graphically, it is called the amplitude pattern, field pattern or voltage pattern. When the square of the amplitude or relative amplitude is plotted, it is called a power pattern.

antenna terminal impedance or driving-point impedance. The ratio of complex voltage to complex current at the terminals of a transmitting antenna, or the ratio of the open-circuit voltage to the short-circuit current at the terminals of a receiving antenna. aperture (antenna). A surface, near or on an antenna, on which it is convenient to make assumptions regarding field values for the purpose of computing fields at external points.

NOTE: The aperture is often taken as that portion of a plane surface near the antenna, perpendicular to the direction of maximum radiation, through which the major part of the radiation passes.

**athermal effect (nonthermal effect).** Any effect of electromagnetic energy absorption not associated with or dependent upon the production of heat or a measurable rise in temperature.

**attenuation.** A general term, expressed as a ratio, used to denote a decrease in magnitude of a field quantity in the transmission from one point to another.

NOTE: The attenuation may be expressed as a ratio, or, by extension of the term, in decibels. **average power (P).** The time-averaged rate of energy transfer;

$$\overline{P} = \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} P(t) dt$$

**average power density.** The instantaneous power density integrated over a source repetition period.

**average power output, amplitude-modulated transmitter.** The radio frequency power delivered to the transmitter output terminals averaged over a modulation cycle. **backward scatter.** Energy directed (scattered) into the rear hemisphere with respect to the direction of incident energy.

**beamwidth, half-power.** In a plane containing the direction of the maximum lobe of the antenna pattern, the angle between the two directions in which the radiated power is one-half the maximum value of the lobe.

**bolometer.** A device capable of absorbing radiant energy and of using the heat so developed to change its electrical resistance, thus indicating the amount of energy absorbed.

**conductivity.** The ratio of the electric field strength in a medium to the conductioncurrent density.

dielectric constant. See: permittivity.

**diode.** A semiconductor device having two terminals and exhibiting a non-linear voltagecurrent characteristic; in more restricted usage, a semiconductor device that has the asymmetrical voltage-current characteristic exemplified by a single p-n junction.

**dipole.** (A) Any one of a class of antennas producing the radiation pattern approximating that of an elementary electric dipole, (B) A linear radiator, usually fed in the center, producing a maximum of radiation in the plane normal to its axis. The length specified is the overall length.

NOTE: Common usage considers a dipole antenna to be a metal radiating structure that supports a line-current distribution similar to that of a thin straight wire, a half-wavelength long, so energized that the current has a node only at each end.

#### driving-point impedance. See: antenna terminal impedance. duty cycle. Deprecated. See: duty factor.

**duty factor.** The ratio of the pulse duration to the pulse period of a periodic pulse train. **duty ratio.** The ratio of average to peak pulse power.

effective aperture. See: antenna effective aperture.

effective area. See: antenna effective aperture.

**electric dipole.** A pair of equal and opposite charges separated by an infinitesimal distance. When the charges are oscillating, the dipole becomes an elementary radiating electric dipole.

**electric field (general).** A vector field of electric field strength or of electric flux density. NOTE: The term is also used to denote a region in which such vector fields have a significant magnitude. **electric field strength (E).** (A) (General.) At a given point, the magnitude (modulus) of the vector limit of the quotient of the force that a small stationary charge at that point will experience, by virtue of its charge, to the charge as the charge approaches zero in a macroscopic sense.

(B) (signal-transmission system) The magnitude of the potential gradient in an electric field expressed in units of potential difference per unit length in the direction of the gradient.

(C) (radio wave propagation) The magnitude of the electric field vector. The electric field strength is expressed in units of volts per meter (V/m).

electric field vector. The force on a stationary positive charge per unit charge.

NOTE: This may be measured either in newtons/coulomb or in volts/meter. This term is sometimes called the electric field intensity, but such use of the word intensity is deprecated, since intensity connotes power in optics and radiation.

**electric flux density (displacement).** A vector equal to the product of the electric field strength and the permittivity of the medium. In an anisotropic medium, the permittivity is a function of direction; hence, the electric flux density is not necessarily in the same direction as the electric field strength. Electric flux density can be considered as a surface charge density and is expressed in units of coulombs per square meter (C/m<sup>2</sup>). **electrical length.** The length of any electrical conductor, such as an antenna or transmission line, expressed in wavelengths, radians, or degrees.

NOTE: When expressed in angular units, it is distance in wavelengths multiplied by  $2\pi$  to give radians, or by 360 to give degrees.

**electromagnetic field.** Electromagnetic phenomena expressed in scalar or vector functions of space and time.

**electromagnetic radiation.** The propagation of energy in the form of EM waves through space. (Not intended to describe propagation along waveguides and other transmission lines.)

**energy density (electromagnetic field).** The EM field energy contained in a given infinitesimal volume divided by that volume. Energy density is expressed in units of joules per cubic meter (J/m<sup>3</sup>).

NOTES: (1) The term energy density may be used with modifiers such as instantaneous, maximum, and peak.

(2) In speaking of average energy density in general, it is necessary to distinguish between the space average (at a given instant) and the time average (at a given point).

equivalent plane-wave power density. The normalized value of the square of the electric or the magnetic field strength at a point in the near-field of a radiating source. The value is expressed in  $W/m^2$  and is computed as follows:

 $W = |E|^2/377 = |H|^2 \times 377 \text{ W/m}^2$ 

**far-field region.** That region of the field of an antenna where the angular field distribution is essentially independent of the distance from the antenna. In this region, (also called the free space region) the field has a predominantly plane wave character, i.e., locally, very uniform distributions of electric field strength and magnetic field strength in planes transverse to the direction of propagation. For larger antennas especially, the far-field region is also referred to as the Fraunhofer region.

**horn antenna.** An antenna consisting of a waveguide section whose cross-sectional area increases toward the open end that is the aperture through which electromagnetic energy is radiated or received.

horn radiator. See: horn antenna.

**incident wave.** A wave, traveling through a medium, in a specified direction, which impinges on a discontinuity or a medium of different propagation characteristics. **insertion loss.** The loss resulting from the insertion of a component in a transmission system. It is the ratio of the power delivered to the load when connected to the generator, to the power delivered to the load when the component is inserted. Insertion loss is usually expressed in decibels.

**internal body current.** The current that is induced in a biological subject that is exposed to low-frequency RF fields.

**intrinsic impedance (of free space).** The ratio of the electric field strength to the magnetic field strength of a propagating electromagnetic wave. The intrinsic impedance of a plane wave in free space is approximately 377 ohms.

**ionizing radiation.** Any electromagnetic or particulate radiation capable of producing ions directly or indirectly in its passage through matter. Examples are X-rays and gamma rays.

isotropic. Having the same properties in all directions.

**isotropic antenna.** An antenna capable of radiating or receiving equally well in all directions, and equally responsive to all polarizations of electric and/or magnetic fields. NOTE: In the case of transmitting coherent electromagnetic waves, an isotropic antenna does not

exist physically, but represents a convenient reference antenna for expressing directional properties of an actual transmitting antenna.

**lobe, antenna.** A part of the antenna radiation pattern between adjacent minima. **loss tangent.** The ratio of the imaginary component of the complex permittivity of a material to the real component of the complex permittivity.

**magnetic dipole.** A magnetic field moment caused by current flowing in an infinitesimally small loop. When the current is oscillating, the dipole becomes an elementary radiating magnetic dipole.

**magnetic field strength (H).** The magnitude of the magnetic field vector, expressed in units of amperes per meter (A/m).

**magnetic field vector.** A field vector that is equal to the ratio of the magnetic flux density to the permeability, expressed in units of amperes per meter (A/m).

**magnetic flux density (magnetic induction).** The vector quantity  $\overline{B}$  producing a torque  $\overline{T}$  (in Joules) on a plane current loop in accordance with the relation  $\overline{T} = IA\overline{n} \times \overline{B}$ , where  $\overline{n}$  is the positive unit vector normal to the loop, A is its area in m<sup>2</sup> and I is its current in amperes. Magnetic flux density is expressed in units of tesla (T), formerly

current in amperes. Magnetic flux density is expressed in units of tesla (T webers per square meter.

**maximum normalized field strength.** The value of the maximum spatial electric field strength E or the maximum spatial magnetic field strength H under a set of specific near-field exposure conditions (in a prescribed volume of space) that is equal to the corresponding value of E or H associated with a plane wave of given power density.

NOTE: The Maximum Normalized Field Strength concept is used to aid in the assessment of the relationship between exposure of an object to external near-fields and the resulting internal SAR.

**maximum permissible exposure (MPE).** The RMS and peak electric and magnetic field strengths, their squares or the plane-wave equivalent power densities associated

with these fields and the induced and contact currents to which a person may be exposed without harmful effect and with an acceptable safety factor.

**microwaves.** A term used rather loosely to signify radio waves in the frequency range from about 300 MHz upward.

**modulation.** The process, or result of the process, whereby some characteristic of one wave is varied in accordance with another wave or signal.

NOTE: For the purpose of this recommended practice, continuous wave (CW) operation is considered to be a special form of modulation, that is, zero modulation.

**multipath error.** The error caused by the reception of a composite signal that arrives via two or more different paths.

**multipath transmission (radio propagation).** The propagation phenomenon that results in signals reaching the receiving antenna by two or more paths.

**near-field region.** A region in the field of an antenna, located near the antenna, in which the electric and magnetic fields do not have a substantially plane-wave character, but vary considerably from point to point. The term "near-field region" is only vaguely defined and has different meanings for large and small antennas. The near-field region is further subdivided into the reactive near-field region, which is closest to the antenna and contains most or nearly all of the stored energy associated with the field of the antenna, and the radiating near-field region. If the antenna has a maximum overall dimension that is not large compared with the wavelength, the radiating near-field region may not exist. For antennas large in terms of wavelength, the radiating near-field region is sometimes referred to as the Fresnel region on the basis of analogy to optical terminology. NOTE: For most antennas, the outer boundary of the reactive near-field region is commonly taken

to exist at a distance of one-half wavelength from the antenna surface.

nonthermal effect. See: athermal effect.

**nonionizing radiation.** Any electromagnetic radiation incapable of producing ions directly or indirectly. Microwaves and RF energy are forms of nonionizing radiation. **peak power density.** The maximum instantaneous power density occurring during the interval when power is transmitted.

**peak power output.** In a modulated carrier system, the output power averaged over a carrier cycle, at the maximum amplitude that can occur with any combination of signals to be transmitted.

**peak pulse amplitude.** The maximum absolute peak value of a pulse excluding those parts considered to be unwanted, such as spikes, provided that the energy is small (5% or less) compared with the total energy at the pulse.

NOTE: Where such excursions are made, it is desirable that the amplitude chosen be illustrated pictorially.

**penetration depth.** For a plane electromagnetic wave incident on the boundary of a medium, the distance from the boundary into the medium along the direction of propagation in the medium, at which the field strengths of the wave have been reduced to 1/e of their boundary values. Penetration depth is expressed in meters (m).

**permittivity (complex).** The properties of a dielectric material or biological tissues, which are frequency dependent.

**polarization (radiated wave).** That property of a radiated electromagnetic wave describing the time-varying direction and amplitude of the electric field vector; specifically, the figure traced as a function of time by the extremity of the E-field vector at a fixed location in space, as observed along the direction of propagation.

NOTE: In general, the figure is elliptical and it is traced in a clockwise or counterclockwise sense. The commonly referenced circular and linear polarizations are obtained when the ellipse

becomes a circle or a straight line, respectively. Clockwise sense rotation of the electric vector is designated right-hand polarization and counterclockwise sense rotation is designated left-hand polarization.

**power.** A physical quantity describing the rate of delivery or transmission of energy. In this document, power will refer to radio frequency power with units of watts (W). **power density.** Power per unit area normal to the direction of propagation, usually expressed in watts per meter squared (W/m<sup>2</sup>) or, for hazard assessment, mW/cm<sup>2</sup>. See *also:* **poynting vector.** 

**power gain.** Of an amplifying device, the ratio of the RF power delivered to a specified load impedance to the RF power absorbed by its input. Of an antenna, *see:* **antenna gain.** 

NOTE: Power gain is usually expressed in decibels.

**power level.** At any point in a transmission system, the ratio of the power at that point to some arbitrary amount of power chosen as a reference. This ratio is usually expressed as decibels referred to 1 mW (dBm) or decibels referred to 1 W (dBW).

**poynting vector.** A vector  $\overline{P}$ , defined as the vector product of the electric and magnetic field vector at the point in question, that is,  $\overline{P} = \overline{E} \times \overline{H}$ .

NOTE:  $\overline{P}$  is expressed in units of watts per square meter and, for a uniform plane wave, represents the electromagnetic energy flow per unit area per unit time across a given surface element.

**probe-length (probe antenna-length).** The maximum physical dimension of the antenna of a near-field probe, or the dimension of the largest antenna in a multiple array. **pulse repetition frequency (PRF).** In a pulse-modulated RF system using recurrent pulses, the number of pulses per unit of time.

pulse repetition rate. See: pulse repetition frequency.

**pulsed RF.** A continuous-wave RF carrier signal that is amplitude-modulated at a known PRF with a controlled duty factor.

**radar.** A system that radiates pulsed or frequency-modulated electromagnetic waves and utilizes the reflection of such waves from distant objects to determine their existence or position.

NOTE: The name is derived from initial letters of the expression radio detection and ranging. **radiation intensity.** In a given direction, the power radiated from an antenna per unit solid angle in that direction. The units are watts per steradian (W/sr).

radio frequency protection guide (RFPG). Deprecated. See: maximum permissible exposure.

**radio frequency radiation hazard meter (monitor).** An instrument that is capable of measuring spatially localized electric and/or magnetic field strengths under near and far-field conditions. The instrument consists of a sensor with an antenna suitable for the wavelength under study, plus a means for transmitting information from the sensor to a suitable field strength indicator.

**radio beam.** Radiowaves whose energy is essentially confined within a relatively small angle in at least one plane.

radio frequency (RF). A frequency that is useful for radio transmission.

NOTE: The present practicable limits of radio frequency are roughly 10 kHz to 300 GHz.

**reactive field.** Electric and magnetic fields surrounding an antenna or other electromagnetic devices that result in storage rather than propagation of electromagnetic energy.

**reflected wave.** A wave in a medium produced by reflections from objects or discontinuities in the medium or from a boundary of a different medium.

**reradiated field.** An electromagnetic field resulting from currents induced in a secondary, predominantly conducting object by electromagnetic waves incident on that object from one or more primary radiating structures or antennas. Reradiated fields are sometimes called "reflected" or, more correctly "scattered fields." The scattering object is sometimes called a "reradiator" or "secondary radiator." *See:* **reradiator. reradiator, passive or parasitic.** Electrically conducting structures that, when illuminated by a primary RF source or ambient electromagnetic fields, act as a secondary radiating source because of currents induced in the structure. In some cases, reradiators can produce localized EM fields significantly more intense than the fields at the location of the reradiator that are associated with the prime source. **response time.** The time required for a field-measuring instrument to reach some specified percentage of the final value after being placed in the field to be measured. In this document, 90% of the final value is assumed.

**scattering.** The process that causes waves incident on discontinuities or boundaries of media to be changed in direction, frequency, phase or polarization.

**specific absorption rate (SAR).** The time derivative (rate) of the incremental energy (dU) absorbed by (dissipated in) an incremental mass (dm) contained in a volume element (dV) of a given density (r).

$$SAR = \frac{d}{dt} \left( \frac{dU}{dm} \right) = \frac{d}{dt} \left( \frac{d}{\rho dV} \right)$$

SAR is expressed in units of watts per kilogram (W/kg).

NOTES: (1) SAR can be related to the electric field at a point by

$$SAR = \frac{\sigma/E^2}{\rho}$$

where:

s = conductivity of the tissue (S/m) r = mass density of the tissue (kg/m<sup>3</sup>) E = Total RMS electric field strength (V/m)(2) SAR can be related to the increase in temperature at a point by  $SAR = \frac{c\Delta T}{\Delta t}$ 

where:

 $\Delta T$  = change in temperature ( °C)

 $\Delta t$  = duration of exposure (seconds)

c = specific heat capacity (J kg<sup>1</sup> degree C<sup>1</sup>)

This assumes that measurements are made under "ideal" non-thermodynamic circumstances, i.e. no heat loss by thermal diffusion, heat radiation, or thermoregulation (blood flow, sweating, etc.).

**standing wave.** A spatially periodic or repeating field pattern of amplitude maxima and minima that is generated by two equal-wavelength propagating waves traveling in different directions. For any component of the field, the ratio of the amplitude at one point to that at any other point does not vary with time.

**standing wave ratio.** The ratio of maximum field strength to minimum field strength along the direction of propagation of two waves traveling in opposite directions on a transmission line.

**thermocouple.** A pair of dissimilar conductors so joined at two points that an electromotive force is developed by the thermoelectric effect when the junctions are at different temperatures.

**time averaged power density.** The average power density integrated over the longest source period. This period can be as a result of rotational or directional characteristics of the source. Units of this quantity are  $W/m^2$ .

**waveguide.** An enclosed system capable of guiding electromagnetic waves from one place to another. Usually it consists of a hollow metallic tube or a solid dielectric material. **waveguide component.** A device designed to be connected at specified ports in a waveguide system.

# 3. The Measurement Problems of Radio Frequency (RF) Hazard Determination

# 3.1 Characteristics of RF Electromagnetic (EM) Radiation

# 3.1.1 Parameters Determined by the Source.

Sources of EM radiation have widely different characteristics that impose a requirement for versatility upon the monitoring equipment. Pertinent characteristics are the following:

- (1) *modulation*. The specific characteristics of the signal in both the time and frequency domains.
- (2) *radiation pattern*. In the near field, patterns change with distance from the source, whereas in the far field no significant change with distance occurs. At any given point in space, the field strength continually changes if either mechanical or electronic scanning techniques are used.
- (3) *frequency*. Energy may be present over several decades of frequency, and may be predominately associated with either the E or H fields.
- (4) polarization. In the far field of a single radiating source, only one polarization exists over a broad physical area (vertical linear, horizontal linear, elliptical or circular). However, in the near field, any one of these polarizations may exist at any given point, and the polarization changes with small variations in location from the RF source.

# **3.1.2 Interference Patterns.**

In any environment where RF measurements are to be made, field strength will commonly be variable with position. Such variability is caused by interference patterns produced by the combining of energy received directly from the source(s) and reflections (or reradiation) from natural or man-made objects (multi-path radiation). Since the phase of the reflected signal can be at any angle with respect to that of the direct signal, the effect of the reflection can be to enhance or diminish the signal strength that would be found at the measuring location in the absence of a reflected signal. The distances between maxima and minima are a function of wavelength so they may vary from a fraction of a centimeter to many meters. At locations with emitters operating at various frequencies, the field strength pattern is likely to be particularly chaotic.

Scanning sources, such as radars and other sources that operate intermittently, introduce time variations with their own interference patterns. Planning of measurement programs should take into account both spatial and time variabilities. That need is important from both the viewpoints of collecting all data pertinent to the objective and ensuring that the personnel are protected from excessive exposure.

# 3.1.3 Radiation Leakage.

Radiation leakage from electronic equipment presents special problems because the source of energy may not be clearly defined a priori. It could emanate from a crack in the shielding cabinet, or from poorly joined or non-shielded, connecting cables or sections of

waveguide. The polarization of the EM field and the location of the leak are not generally known. This is a special case of the general near-field situation and the same problems can exist for all near-field measurements, whether the emitted fields are intentional or accidental. Whereas RF antennas (dipoles, horns) include structures intentionally designed to radiate or receive EM energy efficiently, devices designed to process materials with RF energy (RF-dielectric and induction heaters, electrosurgical devices, and arc welders) are not intentionally designed to radiate, but may contain structures that can function as antennas. Workers operating such equipment can receive very intense exposures because of their proximity to these radiating elements.

A completely general theoretical treatment of the leakage problem is very difficult and is beyond the scope of this recommended practice. The survey techniques differ from those associated with the radiation fields from antennas. In the leakage case, the location of the source is generally found by trial and error. A nondirectional, nonpolarized "isotropic" detector is generally desirable in order to probe in the immediate vicinity of the equipment where directive pickup antennas would give inaccurate readings because of their inability to respond to multipath signals and because of inaccurately known gain-reduction factors in their near field. However, at microwave frequencies, a somewhat directive system consisting of a small horn or waveguide probe, thermistor, attenuator, and power meter is easy to assemble, and may be useful for locating a source of leakage when accurate knowledge of the level is not required. However, both the vertical and horizontal polarization components should be measured separately by rotating the probe 90° about its axis of propagation.

In the case of leakage sources that result in highly non-uniform exposure of personnel, measurements of induced currents may be a more accurate indicator of exposure than the magnitude of the electric and magnetic field strengths. Induced currents will provide a measure of the capacitive coupling between the source and a nearby individual that field strength alone cannot. For example, in a case where the hands are exposed to particularly strong EM-fields, the induced current flowing in the fingers, hands and wrists may become the limiting factor, when interpreted in terms of the resulting SAR.

# 3.1.4 Reactive Near Field.

The reactive near field is the region close to the radiation source where energy storage fields are important. In this region of space immediately surrounding the leakage source or antenna, reactive components of the field predominate over the radiating near-field and radiating far-field components. Reactive fields can be inductive (low E/H ratio) or capacitive (high E/H ratio) in nature, and either type of field can predominate. The characteristics of reactive fields relate to the inductance and capacitance of the radiating structure. Part of the associated electromagnetic energy is stored, i.e., it is not propagated beyond the near-field region. This stored energy is transferred periodically between the radiating structure and the near field. Although the extent of the reactive region varies for different equipment, the practical outer limit is of the order of a few wavelengths. For example, at a distance of between two and three wavelengths from an ideal dipole, all reactive components are less than 10% of the radiating components. Although the reactive components do not contribute to the net flow of radiated energy, they can couple into biological or other material, and thus affect energy absorption. Consequently, for many hazard assessments it is important that the reactive fields be measured using the appropriate instrumentation and measurement techniques (see 5.3).

Furthermore, both the electric and the magnetic fields should be measured to fully evaluate the hazard, since both contribute to the induced SAR in biological objects [B88].

# 3.1.5 Passive, Parasitic or Secondary Reradiators.

Conducting objects that are illuminated by RF energy, either in the near or the far field of a primary source, are termed passive or parasitic reradiators. High E- and/or H-fields (with respect to ambient RF fields) can usually be measured in close proximity to a reradiator. These are generally reactive fields that decay rapidly with distance, as discussed in 3.1.4. A passive reradiator can theoretically and physically be replaced by an equivalent transmitting antenna (reciprocity) because an antenna has the same field pattern whether it is transmitting or receiving. The magnitude of the near and far fields produced by a reradiator depends on the RF currents induced in it by the primary source. The size and orientation (with respect to the primary source) also determine the magnitude of the induced currents and, hence, the local fields.

Examples of passive reradiators are electrically-large, linear, metal objects, such as flag poles, metal sign posts and electrical house wiring. Closed loops, such as metallic automobile steering wheels and metal window frames are examples of a second type of reradiator. Large metal surfaces, such as metal fences, metal signs, and the walls of metallic buildings are examples of a third type of reradiator. Each individual reradiator has a unique scattering cross section that determines the amount of energy intercepted from an ambient RF field. The scattering cross section of any single object is dependent on the frequency of the incident radiation and the physical configuration of the object. Resonances and focusing at specific frequencies may occur in a given object that results in significantly enhanced reradiated fields.

There are virtually no data in the literature that relate to the SAR induced in a person exposed to the near fields of a passive reradiator. In spite of this fact, near-field theory, antenna theory, and published dosimetric data associated with active radiators can be utilized to draw some general conclusions. Personnel hazards due to passive reradiators are "reduced" or "enhanced" with respect to exposure to plane waves of equal maximum normalized field strength. Caution should be exercised when interpreting so-called, localized "hot spots" caused by reradiating objects since the amount of energy that may be coupled to an individual contacting the reradiator may be very small, despite the existence of relatively high-surface field strengths. Further discussion of this subject appears in 5.6.

# 3.1.6 Other Considerations.

In the near field, three orthogonal components of the electric field with arbitrary relative phases and amplitudes exist. Similarly, there are three orthogonal components of the magnetic field with arbitrary phases and amplitudes.

The electric field is elliptically polarized in an arbitrary plane and the magnetic field, in general, is elliptically polarized in another plane. Consequently, in the near field, measurements of the phase and amplitude of each of the three components of the electric (magnetic) field generally provide no information about the magnetic (electric) field at the same point. Thus, use of instruments capable of measuring either the electric or the magnetic field, and that respond simultaneously to all polarizations is indicated. Field measurement devices utilizing three orthogonal dipoles or loops that detect the amplitude, but not the phase of the electric or magnetic field, cannot provide complete

information about the elliptically polarized field. Specifically, the maximum instantaneous field vector is not measured with these types of devices. Only an averaged total field strength is measured, with the averaging occurring over one cycle of the oscillation of the field (the carrier frequency). Currently available hazard instrumentation does not provide both phase and amplitude measurement capability. This means that power density is not actually measured under these circumstances even when power density is the measure displayed, but is only inferred from measurements of  $|E|^2$  and  $|H|^2$ .

When characterizing hazardous EM fields, a distinction should be recognized between emission levels and exposure levels. An emission standard specifies the maximum field strength or power density at a specified (usually small) distance from an emitting source; an exposure standard generally specifies the maximum field strength or power density to which personnel should be exposed as a function of exposure duration. In most cases where an emission standard applies, the sources are small apertures, e.g., localized leakage around the periphery of a microwave oven door. In these situations the radiated fields follow approximately an inverse- square law reduction of power density with distance or an inverse dependence of field strength with distance. Such inverse-distance dependence has been verified for leakage emission from microwave ovens for distances of 5 cm to several feet. The rapid decay from permissible emission levels to acceptable exposure levels at a distance of several feet from the typical microwave oven may not apply if the leakage source consists of a large radiating aperture (for example, a leaking viewing window).

An RF radiation monitoring instrument can respond to pulsed fields in an entirely different manner than it responds to CW fields. For low duty factor pulsed fields, the instrument may become a peak detector and produce meter indications that exceed the actual, time-averaged field values by factors of 10 to 20 dB.

In general, the maximum level of exposure to personnel will not be equivalent to the measured emission level of a source of RF energy. Furthermore, the exposure area will generally decrease as one approaches the source. Thus, as the source is approached, a plane should be scanned by personnel surveying a leakage field to determine the location of the localized leakage radiation beam.

# 3.2 Summary of the Measurement Problem

# 3.2.1 External Field Measurement Problems.

The EM environment is determined by many factors including the following:

- (1) the direction of energy propagation from the sources
- (2) the directions, distances, and relative orientations of the sources and prominent features of the physical environment, with respect to the field point, and

(3) the polarization, frequency, type of modulation, and power of the sources The variable nature of these factors and their effects on the resulting EM field should be understood to successfully design and operate instruments that measure the EM environment, and to obtain sufficient data to ensure personnel safety.

In general, the character of the near-field of an RF source is composed of both reactive and radiation components that exhibit both spatial and temporal variations. These

variations will be functions of the physical environment, as well as the properties of the RF source. As a result of the extremely wide variety of possible situations, each of which could be essentially unique, the calculation of near-field intensities for each situation is generally not practical due to the complex nature of near fields. Hence, one should usually rely on measurements. The following material is applicable to both near-field and far-field external fields and SAR measurement applications.

# 3.2.2 Problems of Time and Spatial Averaging.

Many recommendations and standards including IEEE Std C95.1-1991 [1], specify the maximum permissible values of the RF field strength or power density as averaged over a specified averaging time, e.g., any continuous 6-min interval. If, for example, RF exposure in the VHF band is considered, a maximum time-averaged power density of 1 mW/cm<sup>2</sup> is permitted. The time-averaging provision of the MPE permits higher exposure values if the exposure time is reduced to less than 6 min. Another way of expressing this is:

W (mW/cm<sup>2</sup>) = t (min) = 6 mW min/cm<sup>2</sup>

Thus, for example, if the exposure duration is only 3 min, a maximum power density of 2 mW/cm<sup>2</sup> is permitted. Figure 3.1 illustrates the application of the time-averaging provision of the MPEs. In Fig 3.1a the time-averaged exposure value is 6 mW-min/cm<sup>2</sup> in both periods 1 and 2. During the remainder of the illustrated 6-min period, no exposure is allowed in order to keep the time-averaged value from exceeding 6 mW-min/cm<sup>2</sup>. In reality, RF exposures usually vary continuously with time due to source characteristics or movement of the individual within the RF exposure field. This is represented by Fig 3.1b, where the area under the curve within any 6-min window of time does not exceed 6 mWmin/cm<sup>2</sup>. The time-averaging feature of the MPE may introduce substantial complications in determination of compliance, depending upon the particular exposure situation. Assessing time-averaged RF exposures in some complex environments may be accomplished accurately only through the use of instrumentation designed to acquire and average the real-time variations in the measured field strengths [B10]. Such measurements may be performed with portable datalogging devices adapted to the averaging time of the MPE [B120]. In less complicated exposure environments, such as when the RF exposure is intermittent, but not otherwise varying in level, the use of stripchart recorders connected to the recorder output of a broadband RF field-strength meter may suffice for determining the time- averaged exposure values.

The assessment of RF fields that vary substantially with location also presents a problem when attempting to specify whole-body average exposure levels. In this case, the same approach employed for the measurement of time-averaged exposure levels may be used successfully. For example, [B120] describes a measurement technique using a datalogging device wherein a uniform velocity scan is performed for a vertical planar scan of space and the resulting value of RF exposure, averaged over the total planar scanning time is equivalent to the spatial average of the RF field. Commercially available RF survey meters now provide for integrated data logging devices connected to a basic field meter. The use of spatial averaging of fields provides a more meaningful description of the exposure, particularly in areas where extremely localized and high intensity fields may exist, but where only limited exposure of the body actually occurs. RF exposure fields may be averaged as a straight line linear spatial average or averaged

over the projected area of the body. The difference in using the projected area of the

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body is that non-uniformity of the fields is weighted according to the projected area of the body and, due to the variation of the projected area with height, the RF fields are not linearly averaged. For example, the projected area of the head represents a relatively small area relative to its vertical extension compared with the torso over an equal vertical dimension. Thus, in cases where the fields may be maximum near the location of the head, and relatively small over the rest of the body, the projected area average will result in a smaller value than would occur via simple linear averaging. However, for fields that are highly variable with spatial maxima in the region of the torso and lower body, the use of projected area averaging may produce higher spatial averages than simple linear averaging. The outcome of the spatial averaging process will be dependent on the spatial characteristics of the RF fields in relation to the posture of the exposed subject.

Tell (1996) has evaluated the differences between these two methods of spatial averaging for vertical co-linear type antennas, common to the type used for paging and other wireless communications. In an analysis of an 800 MHz band antenna with alternative mounting heights of 0, 4 and 6 feet, he found that with the simulated head height configuration, the projected area averaging resulted in an averaged power density almost 29% less than linear averaging. But, for field distributions producing larger fields only slightly higher in elevation, the reduction was determined to be nominally less, about 8% and 1%, respectively. However, for the field distributions of the co-linear antennas with 0 and 4 foot mounting heights, the projected area approach actually resulted in slightly higher values for the spatially averaged power density, being 7% and 15% greater, respectively.

This finding is not surprising based on the considerably complex situation of reflected fields encountered at telecommunications antenna sites. If the power density of the local field is relatively high in the regions of the trunk of the body, the increased projected area of the body throughout the trunk can result in weighted power densities that are actually greater than if the fields were simply linearly averaged. On the other hand, it is clear that basing spatial averaging on the body's projected area can result, in some cases, especially with highly localized fields, in notably lower values of exposure. While the IEEE C95.1 standard specifies MPEs in terms of spatial averages based on projected areas, the practical complication that this presents to compliance studies is recognized in Section 6 of the standard, yielding to the acceptability of performing linear averaging. Individuals involved in compliance studies of telecommunications sites should be aware of the possible differences that either technique can have on the results.

# 3.2.3 In- and Out-of-Band Interference in RF Hazard Meters, Including Cable Pickup.

In- band interference associated with RF pickup in the cable connecting the probe with the readout of an RF hazard meter often occurs. Errors of as great as 10 dB can occur at frequencies below a few MHz in improperly designed instruments. For more detailed information, see 5.3.5.

Out-of-band performance of RF radiation hazard monitors becomes increasingly important in areas where multiple signals are present. When illuminated only by frequencies within its designated band, a monitor may provide accurate measurements. Signals outside this band may produce uncalibrated meter responses giving rise to false characterization of the fields actually present. In the case of frequencies higher than the upper usable frequency, erroneously high responses associated with the probe structure generally produce this form of error. For frequencies below the operating band, the

instrument response to the reactive field, or a scalar potential field, may produce false indications [B91]. This type of error can be minimized by avoiding measurements close to energized low-frequency elements, where capacitive coupling to the survey instrument may produce this type of false response. Techniques developed for 60 Hz field strength measurements, where the potential effects are canceled [B94], can, in some situations, be utilized to correct for this error. The effect of low-frequency interference can be determined at the measurement site using the technique described in 5.3.5.

# Fig 3.1 Application of 6-Minute Time Averaging

# 3.2.4 Effects of Sensor Size and Measurement Distance.

When an isotropic, near-field probe is used to make RF measurements close to an RF radiator, or near a reflecting or reradiating object, several types of errors arise. The errors can readily exceed many dB if the following effects are not avoided.

- (1) Field Gradients. Measurement data can be distorted when using an isotropic "near-field probe" to map steep spatial gradients close to the radiating elements of an RF emitter (an antenna or an unintentional radiator). These gradients may cause the amplitude of the field being measured to vary significantly over the volume of space occupied by the probe's antennas. This introduces measurement errors due to spatial averaging. This problem imposes limits on the maximum size of the array antennas or sensors in the probe. Also, a minimum distance exists where accurate measurements can be made of the near field, for a given probe size.
- (2) Interaction of an Active Source with the Probe. Coupling of reactive near fields to the measurement probe can result in erroneously high-measured values when using a near- field probe in proximity to an active radiator or a passive reradiator.

The degree of probe interaction (or coupling) is a function of the probe antennas' or sensors' size and the separation distance between the RF source and the probe.

(3) Probe-Antenna Loading Effects from Nearby Objects. When the probe is close to a reflecting object or reradiator, a probe-loading error is produced. This effect alters the "source impedance" of the probe's antennas and changes the equivalent electrical circuit of each antenna and its respective detector. For a given type of detector, this loading error is dependent on the size of the antenna, the distance to the reflecting object and the frequency of the field being measured.

# 3.2.5 SAR and Internal Body Current Measurement Problems.

The measurement of SAR in exposed biological subjects at radio frequencies is a challenging task, both under near- and far-field exposure conditions. In the far-field case, internal fields are highly dependent on the size, orientation (with respect to polarization), and composition (complex permittivity) of the object. In a sphere (such as the human head) or cylinder (such as the arm or leg) resonances may occur, causing large gradients in the internal field-strength distribution [B61] with focal points, or hot spots, occurring near the center of the sphere with standing waves throughout the volume of the exposed object. However, the SAR on the "front" surface of the sphere will typically be high, and is usually the maximum value of local SAR. Under localized (partial body) nearfield exposure conditions, the internal fields decay exponentially with distance from the exposed external surface. The rate of decay depends on the conductivity of the tissue. Thus, the determination of SAR is easier for this class of near-field exposure, compared with far-field exposure, since the internal fields are confined primarily to the volume directly adjacent to the exposure aperture of the apparatus [B61]. In both the near- and far-field cases, internal regions with differing permittivities create reflections and standing waves and, thus, complicate the measurement problem [B70]. Also, when the whole body of a person or animal is exposed to plane-wave or near-field RF energy, localized regions deep within can be heated selectively. The resulting intense, local "hot spots" are due to resonant conditions existing in these localized regions. The local SAR ( $E^2$ ) may exceed the average whole-body SAR by factors exceeding 100 times [B47, B117].

Measurements of induced body currents are at present practically restricted to determining the induced currents caused by electric fields (see 1.4.3). The measurement of induced currents has been discussed [B40, B49, B54, B60, B122]. In practice, a current measuring device is placed in series with the feet or hands and the current that then flows to earth or some other grounded surface is measured. A method for determining the current induced in the arms of towers climbers has been described [B45] in which a broadband, thermocouple type RF milli-ammeter is used. A narrowband measurement technique in which the RF voltage (that is proportional to the current flowing in the body) developed across a low-impedance resistor is measured with a high sensitivity tunable receiver is discussed by [B122]. By using this method, the currents induced at different exposure frequencies can be distinguished. The use of clamp-on type RF current transformers is an alternative method for measuring induced or contact currents.

# 3.2.6 Limitations on the Measurement Accuracy of the RF Energy Absorption Rate (SAR).

The local SAR values and the SAR distribution in biological objects cannot be measured without producing relatively large measurement uncertainties, regardless of the
instrumentation used. Under ideal plane-wave exposure conditions, the maximum local (point) SAR can be 20 to 100 times greater than the whole-body-averaged SAR [B115]. Thermodynamic factors and large gradients in the internal E-fields increase the magnitude of the SAR measurement error whether it is measured with thermal instrumentation or with implantable E-field probes. A measurement uncertainty of about ±1 to 2 dB is usually the best that can be achieved when attempting to determine the maximum and minimum internal EM fields or the SAR that exists anywhere within an irradiated biological object. Calorimetric measurements of the whole-body- averaged SAR can be performed with absolute accuracy and precision that is better than 10% [B48]. However, the whole-body-averaged SAR, as well as the local SAR at various points in an exposed object vary significantly as the position of the object is changed with respect to the exposure-field vectors. Therefore, measured SAR data should be expressed with realistic precision (no more than 2 significant figures) and the limits of uncertainty of the SAR measurement should be stated explicitly.

#### 3.2.7 Limitations in the Use of Near-Field Instrumentation.

An undesirable situation often arises when a hazard survey is performed using isotropic survey instruments. Attempts are made to assess the degree of hazard using only field strength data taken in the near field of a radiating RF source or a passive reradiating object. The surveyor measures a high field strength that decays rapidly as the probe is moved away from the source. At distance of a few centimeters, the measured field strength may exceed the applicable MPE expressed in units of far-field electric or magnetic field strength or equivalent plane wave power density. However, the coupling of these localized RF fields to any absorbing object (such as a person) may not exceed the SAR value upon which the MPE was based. (Section 5.5.3 discusses the possibilities for estimating RF absorption using only external field measurements. However, often the estimate will be so inexact that it is useless to attempt to determine the degree of potential risk using external field measurements.)

One such situation is the high E-field strength near the tip of the monopole of a hand held transmitter operating at a wavelength that has similar dimensions as the human head. Very high field strengths (relative to the prevailing MPE) are measured but only within a few centimeters of the antenna tip. It can be shown that RF absorption may indeed be high if a person's head is placed a few centimeters from the antenna tip, but it is not possible to predict the local SAR at the surface of the head using only the measured external field strength. This and many other situations should be assessed through the use of internal field SAR measurements (dosimetry) or induced currents in realistic models of the part of the anatomy that is immersed in the local high fields. The use of dosimetric analysis via physical or mathematical models is a well accepted technique in ionizing radiation health physics. The tools and models to perform RF dosimetry are commercially available. These dosimetric techniques should be used as the principal means for hazard assessment under certain circumstances where the risk or the economic scale of an RF exposure problem make the relatively high cost and complexity of dosimetry worthwhile. Therefore, dosimetric techniques are discussed in detail in this document, to enable persons responsible for RF hazard assessment to utilize the available dosimetric technology.

## 4. Instrumentation

## 4.1 Instrumentation for External Field Measurements.

#### 4.1.1 RF Survey Instruments.

RF radiation hazard meters (monitors, survey instruments) are usually the preferred means for measurement and assessment of potential RF hazards. As shown in Fig 4.1, an RF monitor can be divided into three basic parts: probe (sensor), leads, and metering instrumentation. The probe consists of an antenna in combination with a sensor or detector. The design and characteristics of the probe determine to the greatest extent the performance and application of the unit. The detected output from a probe with a flat frequency response is a direct measure of the EM-field strength. One exception is the specially-designed hazard probe which has a frequency-shaped response that conforms to a particular MPE (see 4.4.5), i.e., the detected output is appropriately weighted at each frequency. The "leads" refer to that part of the instrument that carries the detected signal to the metering instrumentation. To accomplish this without causing perturbation of the field, the leads may take the form of high-resistance wires, or if they are metallic they

should be carefully oriented so as to minimize coupling with the field. They may also take the form of an optical fiber [B18, B19, B127]. Metering instrumentation includes signal-conditioning circuitry and display devices.

## Fig 4.1 Basic Components of an RF Survey Instrument

To make meaningful near-field measurements, the following conditions should be satisfied:

- The probe should respond to a particular electromagnetic field parameter and not have spurious responses (for example, it should respond to the E-field with no spurious H-field response).
- (2) The dimensions of the probe sensor in its surrounding medium should be less than a wavelength at the highest operating frequency.
- (3) The probe should not produce significant scattering of the incident electromagnetic fields.
- (4) The probe response should be isotropic (independent of orientation), nondirectional and nonpolarized. A probe with a nonisotropic response is useful if the polarization of the measured quantity (E or H) is known, or if some provision is made to rotate the probe to find the orientation for maximum output.
- (5) The leads from the sensor to the meter should not interact significantly with the field or conduct RF current from the field to the sensor.

#### 4.2 Desirable Electrical Performance Characteristics

#### 4.2.1 Power Supply.

The instrument should employ a self-contained power supply that is isolated from external fields by appropriate shielding and filter decoupling. If batteries are used, provision should be made for indicating their condition. The instrument should be capable of at least 8 hr of operation within its rated accuracy before replacement or recharging of the batteries becomes necessary.

### 4.2.2 Polarization.

The combination of probe antennas should be responsive to all polarization components of the EM field. This performance may be accomplished either by inherent design using multiple dipoles or loops or by physical rotation of a single antenna about its axis.

## 4.2.3 Quantities and Units.

For personnel hazards, the instrument should indicate one or more of the following parameters [B124]:

- (1) average "equivalent plane-wave" power density in milliwatts per square centimeter (mW/cm<sup>2</sup>)
- (2) mean-squared electric field strength in volts squared per meter squared ( $V^2/m^2$ )
- (3) mean-squared magnetic field strength in amperes squared per meter squared  $(A^2/m^2)$
- (4) field strength in amperes per meter (A/m) or volts per meter (V/m)

(5) energy density in picojoules per cubic meter ( $pJ/m^3$ ).

Average power density (mW/cm<sup>2</sup>) is displayed by some survey meters intended for use primarily above 300 MHz; such meters are often used in near-field and reactive-field regions where true power density is neither measurable nor the most meaningful quantity ( $E^2$  or  $H^2$  are preferred). Some instruments indicate "equivalent plane-wave" power density as derived from the field quantity being measured. Instruments having a probe with a shaped frequency response should read in terms of "percent of exposure limit" based on the MPE of interest, e.g., IEEE Std C95.1-1991 [1] (or current revision thereof).

## 4.2.4 Range.

An adequate dynamic range for instruments with a shaped frequency-response is -10 dB to + 5 dB (10% to 300%) relative to 100% of the exposure limit, as defined by the MPE. For flat frequency response probes, the minimum recommended dynamic range of the instrument is 10 dB below the lowest value and 5 dB above the highest value of the MPE. Either a single logarithmic range or a number of linear ranges can be used to obtain the desired dynamic range.

## 4.2.5 Recorder Output.

The instrument should be equipped with a recorder output or other means that will enable the measurement of hazardous fields without endangering the operator, and will facilitate spatial and time averaging. Alternate provisions to avoid endangering the operator may be extension cables between the probe and the meter or a maximum-hold mode of operation where the maximum measured value is maintained until the instrument is rezeroed by the operator.

## 4.2.6 Shielding.

The instrument housing and antenna cables should provide adequate shielding to ensure that the measurement uncertainty remains within stated limits when the instrumentation portion or accessory cables (that may act as unintended receiving elements) are exposed to the same field strength as the probe. This shielding should be effective under

conditions in which the maximum coupling or "pickup" occurs for the unintentional receiving elements.

#### 4.2.7 Modulation.

The instrument should indicate RMS parameters, independent of any modulation. However, it is permissible to have a detector or indicator time-constant switch for the CW and amplitude-modulated continuous-wave (AM-CW) modes. Also, the instrument should be capable of averaging the narrowest pulse-modulated envelope of a non-continuous- wave field that is expected to be encountered by the surveyor.

#### 4.2.8 Static Electricity.

Equipment should not indicate false levels due to a response to static electric charges. These static charges are often induced on the probe of the survey instrument, or on the system being surveyed, e.g., cathode-ray tube displays. As an example, when making measurements during windy conditions and/or low humidity, static charge on the operator can influence the readings of the survey instrument.

#### 4.2.9 Response to Other Radiation.

The specified accuracy of the instrument should include effects of exposure to ionizing radiation, artificial light, sunlight, or corona discharge.

#### 4.2.10 Response Time.

The response time is generally defined as the time required for the instrument to reach 90% of its final value when exposed to a step function of CW RF energy. This parameter should be known to the user. Multiple operator-selectable response times are desirable with a fast response time no greater than 1 s.

#### 4.2.11 Special Functions.

It is desirable that the instrument should be provided with one or more of the following functions:

- (1) A maximum-hold function that indicates the maximum reading during the measurement period.
- (2) An audio signal function that is proportional to the measured field strength and/or an audible indication that a preset level has been exceeded.
- (3) A data-logging function that can provide an average, maximum and minimum value of the field quantity being measured. These data are stored for later utilization. This feature could also provide a real-time average of the measured fields with an averaging time specified by the user, e.g., 6 min. This average value should be updated every few seconds providing the user with an indication of the behavior of spatial or time varying fields.
- (4) A time-averaging function with relatively long time constants (of the order of minutes) that, for a flat-frequency response probe, averages the measured quantity over a known time period. The output of the shaped-frequency response probe averaged over a period of 6 min is a direct measure of that portion of the current MPE for the controlled environment where 6 min is the appropriate averaging time.

#### 4.2.12 Stability.

The instrument should exhibit stability sufficient to permit accurate measurements of the RF exposure fields over periods of time that are consistent with the times normally required for the particular measurement. In practice, the instrument should be capable of operating for a minimum of 10 to 30 min without the need for rezeroing the meter (in the absence of RF radiation) on the range necessary for the measurement. Automatic electronic zero-circuitry can be used to avoid the requirement of shielding the sensitive probe from ambient RF fields during the zeroing process. This is desirable, particularly when performing RF surveys in difficult environments such as on communications/broadcast towers, where an RF-free location may not be available or where a surveyor, climbing a tower, may not be able to move freely or have the use of both hands to reset the zero level of the instrument. The instrument should be insensitive to thermal variations within the range of normally encountered temperature extremes. The instrument specifications should state the maximum zero drift for each range.

#### 4.2.13 Precision and Accuracy Considerations.

The instrument should be provided with calibration data that permits the user to assess the maximum uncertainty in determining RF field strength or power density when using the instrument in various types of fields of different frequencies. Calibration data should also include the sensitivity of the instrument to frequencies beyond the intended useful range (out-of-band response). A meter sensitive to out-of-band fields should not be used in an environment where such fields may be present at other than negligible levels. Absolute field-strength calibration uncertainties (accuracy) of no greater than  $\pm 1$  dB are desirable but difficult to achieve. Uncertainties of  $\pm 2$  dB or even greater may be acceptable if the levels are well below the limits of the MPE, but as the MPE limit is approached, measurement certainty becomes of greater importance. In any event, the uncertainty factor should be known and included in the measurement report. The instrument specifications should address the instrument's ability to respond to amplitude modulated (AM) fields, such as pulsed radar signals, as well as a multiplicity of signals that might simultaneously illuminate the sensing probe (see 4.4). The instrument readout should permit resolution (precision) of the measured field strength to within 5-percent of the full scale value or less.

### **4.3 Desirable Physical Characteristics**

#### 4.3.1 Portability.

The instrument should be portable to permit convenient operation under restrictive conditions (e.g., climbing a tower).

#### 4.3.2 Weight.

The weight should be kept as low as is practical in keeping with good engineering practice.

#### 4.3.3 Volume.

The volume should be as small as is practical and convenient for hand-held operation (see 4.3.1).

## 4.3.4 Dependence on Temperature, Humidity, and Pressure.

Specified accuracy of the instrument should include the effects of temperature, humidity, and pressure variations, and the operating ranges for these parameters should be indicated.

## 4.3.5 Durability.

The indicating meter and other system components should be rugged enough to withstand vibration and shock resulting from transport. A carrying case is desirable.

## 4.3.6 Readability.

The meter dial markings should be large enough to be easily read at arm's length. The reading corresponding to the applicable protection guide should appear within the central one-third of the full scale reading of the dial if the readout is of the analog type. If more than one range of sensitivity is provided, the full scale value of the selected range should be indicated, and the units of interest should be readily interpretable. In any case, the analog or digital readout should provide a clear indication of the units being displayed.

## 4.3.7 Ease of Adjustment.

The instrument should have a minimum number of controls. They should be clearly labeled as to their functions. There should be no requirements for moving two controls at the same time. For mechanical meter movements, the electrical zero point should be coincident or upscale from the mechanical zero of the indicating meter.

## 4.3.8 Ease of Use.

Complicated operating procedures should be avoided. The average technician should be able to make accurate measurements with only the information supplied in the instruction manual.

## 4.4 Instrument Types for External Field Measurements.

The instruments described in the following sections are limited to those types that can provide reasonable accuracy in both near-field and far-field situations. Instruments that have large scattering areas (in terms of wavelength) are not included here.

## 4.4.1 Diode Rectifier-Based Instruments.

In these instruments, single or multiple diodes terminate small antennas. Multiple diodes and antenna elements arranged in suitable configurations can be used to sum all field components and enable measurements to be made independent of polarization and direction of incidence. A minimum of three elements in an orthogonal arrangement is required for an isotropic instrument that can be used in any orientation with respect to the field.

Some units now in use employ a single diode in combination with an electrically short dipole or small loop antenna. These instruments respond to only one field component and, consequently, should be oriented to read the maximum value, a process that can be tedious and time consuming. Such instruments are, however, useful for measuring individual field components.

A crossed-dipole or multiple dipoles arranged in a single plane will respond correctly to signals arbitrarily polarized in the plane of the sensor, but not to the total field. Such units

should be oriented so that the resultant (maximum) field vector is in the plane of the sensor.

Two dimensional H-field probes, which consist of two concentric orthogonal loops with diameters of 5.4 and 5.5 mm directly loaded with diodes [Schmid, ?], are commercially available. The resistive lines have been realized using thick-film technology enabling excellent decoupling of the lines from the loop with a filter. The loops are not resistively loaded, i.e., the detected signal is directly proportional to the square of the frequency (when the diode operates in the square-law region). Although this makes the probe unsuitable for measuring broad-band fields, it provides a higher sensitivity and better accuracy for narrow-band signals of known frequency. The sensitivity is of the order of 1 A/m at 1 GHz decreasing to 0.1 A/m at 100 MHz. Such probes can be used to measure surface currents on conductors if held orthogonal to the surface. One dimensional probes with a loop diameter of 3 mm and frequency range of 100 MHz to 2 GHz has also been developed and are commercially available. Fiberoptic H-field sensors are currently being studied but probes based on this approach are not known to be available commercially.

Diode instruments are basically non-linear with respect to power density or field strength. At low levels, the rectified voltage is proportional to power density, or the square of E (or H). At higher levels, the rectified voltage becomes directly proportional to E (or H) [B75]. This change in characteristic requires that the range of operation of the diode be restricted to low levels to provide a true indication of total power density. When the diodes are operated at higher levels, it is required that the output voltages of the individual elements be modified (generally squared) prior to their summation. When diode instruments are used in pulsed fields they usually change from an average to a peak detecting device and, hence, measurement errors may be large in fields of high peak to average ratio. The exception is when saturation of the diode takes place at very high peak field strengths or when circuitry is added to minimize the effects. Such linearization of the square-law response can be very effective for the measurement of single frequency fields, but may introduce errors when subjected to multiple frequency fields

Two errors are associated with measurements made with any diode-based probe. The first error arises with the use of any electrically short dipole antenna with a diode load and an RF filter transmission line. This is the multiple source, multiple-frequency error [B103] that is typically 1- 3 dB, but in some circumstances can exceed 10 dB. The second error arises from an effect discussed in 4.4.2. This is due to the fact that typical high-resistance signal-carrying leads act as a more efficient antenna at low frequencies, e.g., 540 - 1700 kHz used for AM broadcasting, than the short dipoles in the probes. Alignment of the leads with the E-field can result in erroneous measurements.

Schottky diodes, in general, exhibit some photovoltaic effect. Beamlead hybrid types exhibit this effect to a much greater extent and may produce erroneous readings when illuminated by sunlight or strong incandescent light. Therefore, optically opaque encapsulation is required to eliminate this effect.

When adapted to broadband operation, the upper frequency range of a diode based Efield instrument is presently above 26 GHz [B63, B78]. The low-frequency limit is below 400 kHz. The burn-out characteristics can be in the hundreds of mW/cm<sup>2</sup> range.

Diode detectors, depending on design, may exhibit a marked dependence upon ambient temperature. Variations in output with ambient temperature will typically be less than a 0.05 dB per degree C. Diode units also may be modulation sensitive if the square-law region is exceeded, resulting in errors dependent upon the form of modulation.

#### 4.4.2 Active Antenna.

It is difficult to make accurate, broadband E- and H-field probes that cover the longwavelength (100 meter - 3 MHz) region, using the conventional means cited above. This is due to the fact that the source-impedance of an electrically-small (10 cm) dipole antenna is extremely high and, also, the sensitivity of a small (5 -10 cm diameter) loop antenna is very low. In order to provide a flat frequency response and adequate sensitivity in a dipole probe, the load impedance of the detector and the high-impedance lead in combination should be greater than the antenna (source) impedance. One solution is to provide a high-impedance RF buffer amplifier that is connected directly to a monopole or loop antenna and that acts as the load. This is accomplished by placing the amplifier and battery in a metal enclosure that serves as a cubical 'ground plane' or asymmetric dipole element. In commercially available units, the size of this cube is small enough (10 -15 cm on a side) so as to not unduly perturb the field under test or the receiving-pattern of the device [B98]. This is practical for frequencies between 10 kHz and several hundred MHz. Commercially-available magnetic and electric field probes, using active electronics, operate at frequencies as low as 60 Hz [B33].

A second problem associated with probes without active electronics is that of isolating the signal-carrying leads from the antenna/detector combination. This problem may become severe below about 100 MHz, and particularly below 10 MHz. This is due to the fact that the typical high-resistance signal-carrying leads serve as a low-pass filter, and their ability to separate the low frequency detected signal from the RF field being measured becomes more difficult as the two frequencies approach each other. This results in excessive sensitivity and poor antenna patterns in passive probes. Finally, at frequencies above about 300 MHz, when "free-space" or uniform irradiation conditions exist, both the sensor and the metal enclosure of the survey instrument can be exposed to similar levels of RF [B75] and scattering from the enclosure to the sensor (probe) can cause significant measurement errors.

Active electronic probes eliminate the use of such leads entirely by including the visual display (readout) with the metal box containing the active electronics. Instruments are available commercially that use a fiber-optic data link to connect the box containing the active electronics (located at the base of the probe) to a remote readout.

#### 4.4.3 Displacement Current Sensors.

In addition to short dipoles and monopoles, a form of parallel plate capacitor, called a displacement-current sensor [B53], can be used to measure electric fields normal to its surface or normal to any large conducting surface. An instrument designed primarily for measuring fields associated with video display terminals, is based on the displacement-current sensor concept. The sensor is a double sided circuit board 30 cm in diameter. On one side, the front, an annular ring electrically isolates a smaller circular area. The smaller area, approximately 10 cm in diameter, is the active area; the larger annular ring serves as a guard ring to eliminate errors associated with fringing fields. The smaller area is connected through an integrator (an operational amplifier with capacitive

feedback) to the backplane of the sensor that serves as a reference ground. When a time varying- electric field is normally incident on the front surface of the device, the induced surface charge produces a current that flows to the backplane through the integrator circuit. This current, called a displacement current because it is proportional to the time derivative of the electric displacement, is also proportional to the time derivative of the incident electric field strength. Integration of the displacement current results in a voltage at the output of the operational amplifier that is directly proportional to the incident electric field strength (and the cross sectional area of the smaller circular area). The output of the operational amplifier drives a true RMS-voltmeter located within a shielded electronics package. The electronics package, which contains a microprocessor, digital readout and an optical fiber telemetry system for remote monitoring, is mounted to the edge of the sensor. Displacement-current sensors are typically used at frequencies in the LF and VLF regions, e.g., from DC to a few hundred KHz, but may effectively be used at frequencies as high as a few hundred MHz.

#### 4.4.4 Electro-Optical (Photonic) Sensors.

This type of electromagnetic field sensor utilizes a nonmetallic, passive sensing element (electro-optic modulator) with a very broadband response (DC to 20 GHz) that converts electromagnetic field strength information to instantaneous modulation of a laser beam. The laser energy is transmitted via fiber optics to a modulator. The modulator impresses amplitude modulation on the laser beam, in proportion to the instantaneous amplitude of the RF electromagnetic field to which the modulator is exposed. The amplitude-modulated laser beam is then carried from the modulator to a photodetector that converts the modulated optical beam to an electrical signal that represents the instantaneous amplitude of the RF field strength.

The above system has been used with electrically small dipoles [B16, B92, B93] as an electric field sensor, as well as with no antenna (where the electro-optic modulator itself serves as the E-field sensor) [B127]. In addition, conventional antennas can be connected to a commercially- available electro-optic modulator via a short lead, to provide a non-metallic, passive RF link to the antenna.

#### 4.4.5 Thermocouple.

Thermocouple-type radiation monitors generally utilize thin-film type thermocouples as the detection elements. Parts of the film perform the function of the antenna element. These units exhibit extremely good adherence to square-law characteristics in that the DC output from the thermocouple is proportional to the square of the electric field strength. These units, while relatively independent of ambient temperature, are generally less stable than diode type detectors. Hot and cold junctions of the couple are in close proximity to reduce the drift due to variations in ambient temperature. Variation in sensitivity is of the order of 0.1% per degree C. The major limitation of the element is the burnout characteristic which, in terms of average values, is typically approximately 3 times full scale. The full-scale ratings depend on the basic probe sensitivity. The utilization of thin resistive films provides very broad bandwidth.

Some commercial probes arrange alternate hot and cold thermocouple junctions in a continuous series in a linear arrangement to form a dipole or in a circular arrangement to form a loop. If any single thermocouple junction fails because the probe is subjected to fields well in excess of the maximum rating of the probe, the probe will be nonusable.

Some thermocouple devices utilize dual modes of operation for extreme bandwidth. They may function as a linear resistive dipole at frequencies typically below 18 GHz [B5, B64], and as a traveling wave or phase effect antenna above 18 GHz. Such devices have an upper frequency limit that exceeds 95 GHz [B8].

#### 4.4.6 Shaped Frequency Response.

Current MPEs are frequency dependent, and recommend that the fraction of safe level incurred within each frequency interval be determined and the sum of such fractions not exceed unity. Probes containing dipole-diode elements separately or in conjunction with thermocouple elements have been designed to have a sensitivity versus frequency characteristic that is the inverse of a particular MPE [B7]. This allows the summation and weighting of multiple-frequency signals in conformance with a frequency dependent protection guide. The resultant readout for this equipment is in "% of the Standard." The probes are tailored to specific MPEs such as that specified in IEEE Std C95.1-1991 [1].

Shaped frequency response probes may cover only a portion of the frequency range of typical MPEs. Additional probes may be used to complement each other and provide a wider measurement range. When complementary probes are used, it is necessary that their out-of-band performance be well defined; preferably, the instrument should be insensitive to out-of-band signals. Overlapping of frequency bands will also add to the uncertainty of the measurement.

#### 4.4.7 Combined Electric and Magnetic Field Probes.

Probes described in 4.4.1 to 4.4.6 have utilized one probe to measure the electric field and a second, separate probe to measure the magnetic field. In the near field of an RF source, the relative values of E and H vary considerably, with respect to one another, as a function of distance from the source. Also, under typical usage situations, potentially hazardous fields may change rapidly with time. Both of these factors present practical measurement problems. For example, in order to measure both electric and magnetic field strengths that vary over time and space, one should place an E-field probe and then an H-field probe at exactly the same point, sequentially in time. However, since the fields under study may change during the finite time that elapses between the performance of successive measurements, a measurement uncertainty results.

Methods for the simultaneous and mutually-independent measurement of RF electric and magnetic field strengths have been developed [B11, B12, B41, B76]. A broadband isotropic probe system for measuring the E- and H-fields simultaneously can be produced with a set of three mutually orthogonal dipole elements and a set of three mutually orthogonal loops that are physically located within the same very small (compared with wavelength) volume (Fig 4.2). Each antenna has associated circuitry comprised of a detector and an optional frequency- response-shaping filter to provide a frequency response that is flat over the desired frequency range. The lengths of the dipoles are kept electrically short so that the fields are not perturbed and the diameters of the loops are kept electrically small so that the E-field pickup of the loops will be negligible. Mutual coupling between any of the probe elements is also minimized by the use of electrically small antennas. Detectors based on the use of square law operated diodes or thermocouples are used to provide a signal to the electronic circuits which include an arrangement for analog or digital summing and data processing. An analog or digital display, data- logger, and/or recorder can be incorporated as the final stages of

this system. If three loop antennas are arranged with a common center, an isotropic response is achieved [B9]. This type of probe is commercially available.

#### Fig 4.2 Schematic Representation of Three-Dimensional Isotropic E/H Probe

#### 4.4.8 Personnel monitors.

Personal monitors are typically small, portable broadband detectors, suitable for attachment to workers' clothing, equipped with an alarm feature for alerting the subject to the presence of high-level RF fields that may approach the MPE limits of various exposure standards. Most personal monitors provide an alarm with a preset threshold of RF exposure is exceeded [Tell, 1995]. In the region between 1 and 100 GHz, they use resistive thermoelectric dipoles as sensors with a background of lossy material to reduce the effect of scattering from the body. In the frequency range between 30 and 1000 MHz, the sensor generally monitors the magnetic field to reduce the effects of scattering from the body. Between 0.1 and 30 MHz, the monitors use surface charge detection as sensors thus monitoring the radial fields which predominate near re-radiators in this frequency range. These monitors may incorporate sensors for both electric and magnetic fields and some contain frequency dependent sensors that automatically weight the detected RF fields in accordance with frequency-dependent RF exposure limits. This feature makes them especially useful in multiple frequency environments such as broadcast and wireless telecommunications antenna sites.

Because personal monitors can only respond to RF fields at the position where the monitor is located on the body, some care should be used in its placement and interpreting exposure during an alarm condition. For example, wearing a belt mounted personal monitor when the predominant exposure is at the eye may not provide sufficient warning of excessive exposure. Conversely, when used in non-uniform RF field environments such as VHF and UHF broadcast facilities, an alarm condition may be a very conservative indication of potential excessive exposure since the MPE limits of several exposure standards are based on spatial averages over the body dimensions.

Encountering fields that trigger the alarm at one point in space may not ultimately lead to whole-body-average exposures exceeding the MPE limit.

Use of personal monitors can from a valuable component of an RF safety program but a description of their use should be an integral part of any RF safety training program.

#### 4.4.9 Human equivalent antenna.

This device simulates a standard person permitting the measurement of induced currents without requiring the person to be subjected to exposure to fields that induce the current. This device also reduces the variability of the measurements with people of different stature. The frequency range of one commercially available device is 50 Hz to 100 MHz. These devices also have lower sensitivity to radial and magnetic fields and thus require the use of a correction factor where such fields predominate.

#### 4.5 Calibration of External Field Measuring Instruments.

Reliable calibration of the various instruments used for measuring EM fields is essential to ensure safety of personnel, to assure compliance with regulations, and to provide a basis for comparing the results of RF hazard or RF field-calibration research that has been performed by two or more independent groups and laboratories. Existing calibration methods are based on the premise that a known field strength can be established through measurement, calculation, or a combination of both. The device to be calibrated is placed in this standard field and the meter indication is compared with the known field value. There are three basic approaches for producing a standard calibrating field:

(1) the free-space standard-field method;

(2) guided wave methods, and;

(3) the standard-probe or transfer-standard method.

The choice of technique will depend on the type and size of probe, frequency range, available facilities and equipment, and the accuracy requirements [B77, B86, B95]. The most widely used techniques are described in the following sections.

NOTE: The RF output of the generator used in all of the following methods should be free of harmonics, i.e., the harmonics should be at least 20 dB below the fundamental frequency. A low-pass filter can be placed on the output port of the generator to accomplish this.

## 4.5.1 Free-Space Standard-Field Method.

There are several variations of this method, but the objective is to establish a known calibration field in free space. The most common experimental arrangement for use at microwave frequencies is shown in Fig 4.3. The power density W at a point on the axis at a distance d from a transmitting antenna is given by:

$$W = \frac{P_T G}{4\pi d^2} \tag{Eq 4.1}$$

where  $P_T$  is the net power delivered to the antenna and *G* is the effective antenna gain with respect to an isotropic antenna. The gain is normally determined in advance, and  $P_T$ and *d* are measured as part of the regular calibration procedure.

The most convenient method for determining  $P_T$  is by means of a dual directional coupler, as indicated in Fig 4.3. The incident power  $P_i$  and the reflected power  $P_r$  are monitored at the coupler sidearms and  $P_T$  is obtained from the relationship  $P_T = P_i - P_r$ .

High-quality, broadband couplers are available, along with methods for calibrating and using them to determine PT [B23, B43, B44, B65, B66]. The methods cited are for calibrating power meters, but the same techniques can be applied to antennas if corrections are made for impedance mismatch effects. The method of Bramall allows the use of a low-power, calibrated bolometer to provide accurate determination of high transmitted power levels.  $P_T$  can be determined to within 1 or 2% if mismatch corrections are included (as they should be for accurate measurements)<sup>8</sup> One way of accounting for mismatch effects utilizes the power-equation technique [B43, B44], which also permits use of directional couplers that are not restricted in terms of directivity and do not require precision connectors. The power-equation technique and cascaded-coupler method were combined [B17] for the precise determination of transmitted absolute power from an antenna used to calibrate near-field instruments.

#### Fig 4.3 Free-Space Standard-Field Calibration Method

Alternatively, modern automatic network analyzers with built-in microprocessors for realtime correction of system errors may be used to determine  $P_r$  as well as  $P_T/P_r$ . All components used in the system being calibrated should use precision connectors and coaxial-to-waveguide adapters. This includes the high-power coaxial and waveguide directional couplers and adapters that are used during actual generation of far-field calibrated power densities.

<sup>&</sup>lt;sup>8</sup> The uncertainties stated here are intended to include all known, significant sources of error and correspond roughly to 95% confidence limits. However, uncertainties quoted from the published literature should be interpreted as indicated in the original articles.

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The foregoing assumes that the device being calibrated is sufficiently small and far enough away from the transmitting antenna that the amount of energy reflected back into the transmitting system is insignificant. If this condition cannot be met, it is possible to obtain a correction for the effect of the reflected energy by varying *d*, observing the (approximately) sinusoidal variations in  $P_r$ , and then averaging  $P_r$  over at least one full cycle.

**4.5.1.1 Sources of Error.** The principal sources of error in the free-space method are multipath interference, reflections from metallic or dielectric objects used for measuring the fields in the calibrating system and uncertainties in the antenna gain determination. Multipath effects are often overlooked, but every calibrating facility will have some scattering associated with the walls, equipment, and even dielectric probesupport structures. They may cause the power-density in the calibration region to be significantly different from that predicted by Eq 4.1. Calibration errors due to multipath effects can be reduced by observing the probe response as a function of position and averaging the results. Useful discussions of methods for the reduction of multipath errors are given by [B121] and [B118]. Additional errors can be caused by backscatter from cables, metering components, etc. that are a fixed distance behind the probe-undertest. The effects of this type of scattering can be reduced by multiple-position averaging if the probe can be moved with respect to the source of scattered energy (see 5.3.1 and [B118]). Alternatively, absorbing material should be placed in front of all such items that could reflect energy in the direction of the probe.

**4.5.1.2** Antenna Gain Determination. Problems associated with the determination of antenna gain (G) add to the measurement difficulty. It is relatively easy to obtain accurate gain values at large distances, but large distances require greater transmitter power and the multipath situation is often worse. On the other hand, there are some fundamental difficulties associated with accurate gain determinations at short distances.

The effective gain *G* of an antenna is a function of distance and approaches a constant  $G_{\infty}$  as *d* approaches infinity. This is illustrated in Fig 4.4 that shows the estimated gain reduction for a representative example [B73] plotted as a function of the parameter  $n = Id/a^2$ , where *a* is the largest aperture dimension and *I* is the free-space wavelength. In establishing a calibrating field, one should use the correct value of *G* for the specific distance involved; otherwise, significant errors can result. The far-field gain  $G_{\infty}$  of pyramidal horns can be calculated with sufficient accuracy ( $\approx 0.3 \text{ dB}$ ) for many purposes [B24, B68, B73, B110, B112] and may be measured to within about 0.1 dB if necessary [B30, B72, B81, B99, B114, B123]. The measured values of  $G_{\infty}$  obtained by these methods hold for distances greater than about  $8a^2/1$  (n > 8) (as indicated in Fig 4.4). It may also be possible to calculate *G* to within about 0.3 dB for pyramidal horns [B74, B86] at distances down to two  $a^2/1$ . However, the accuracy of these near-zone calculations has not been definitely established.

#### Fig 4.4 Estimated Gain Reduction for a Representative Antenna

There are also problems in experimentally determining the near-field gain. The usual farfield gain measurement approach involves measuring the power transmitted between a pair of antennas and applying the equation

$$G_T G_R = \frac{P_R}{P_T} \left(\frac{4\pi d}{\lambda}\right)^2$$
(Eq 4.2)

where  $P_R$  is the received power,  $G_T$  and  $G_R$  are the gains of the transmitting and receiving antennas, respectively, and *d* is the distance between the antennas. Equation 4.2 holds rigorously only in the far field. At shorter distances,  $G_T$  and  $G_R$  cannot be separated into the individual factors [B114]. Nevertheless, since one can measure  $P_R/P_T$ , it is tempting to apply Eq 4.2 to the case of two identical antennas in the near field and obtain  $G_a^2 = (P_R/P_T) (4\pi d/1)^2$ , where  $G_a$  is the measured apparent near-field gain of the two antennas. However,  $G_a$  obtained in this manner is not the correct nearfield gain. In other words,  $G_a$  will not yield the correct on-axis power density when used in Eq 4.1. This fact can perhaps be seen intuitively.  $P_R$  is the result of an integration (or averaging) of the incident field distribution over the receiving aperture and unless the incident field is a plane wave, there is no simple relationship between  $G_a$  and the desired on-axis power density. The error decreases as d becomes smaller. The near-field gain error can be approximated empirically by plotting measured data (smoothed to eliminate standing wave oscillations) and comparing it with a theoretical curve that falls off as  $1/d^2$ . By determining the deviation from  $1/d^2$ , the smoothed, experimental data can be evaluated.

For rectangular apertures,

$$W = \frac{\eta P_T}{n^2 A} \left(\frac{b}{a}\right)^2 \tag{Eq 4.3}$$

where *W* is the power density at the receiving aperture, A = ab is the physical area of the aperture, *a* and *b* are the aperture dimensions (a being the larger), and *h* is the aperture efficiency defined as  $A_e$  /*A*, with  $A_e$  the effective aperture area. Equation 4.3 is simply a modified form of Eq 4.1 obtained by use of the relations  $G = 4\pi A_e / l^2$  and

 $n = Id/a^2$ . For horns of a given geometric and electrical design (i.e. a family of "standard gain horns" from a particular manufacturer for use at the various waveguide operational frequency bands), the ratio b/a and h are approximately constant and, according to Eq 4.3, the power density for a particular value of n is inversely proportional to the aperture area. It is desirable to have n as large as possible to reduce the gain uncertainty; therefore, if  $P_T$  is limited, it is necessary to use smaller apertures in order to achieve the required calibration field strength.

Hence, if one desires to calibrate antennas at short distances, because a long-distance range is not available, or to avoid the expense of high-power systems and to avoid the complications caused by standing waves due to multipath reflections from an imperfect anechoic chamber, the near-zone gain should be known. Two possible techniques for determining the near-zone gain follow. If one antenna is small (an open-ended waveguide, for example) and its far-field gain is known, it can be used to determine the effective on-axis gain of a larger antenna at relatively short distances by means of Eq 4.2. The measurements should be reasonably accurate ( $\approx 0.5 \text{ dB}$ ) so long as d is greater than four times  $a^2/l$  for the small antenna. With respect to distance from the larger antenna. the primary considerations are that the field gradients be small in the calibration region and the wavefront should approximate a plane-wave. These conditions will be satisfied reasonably well at  $a^2/I$  for the large antenna. The dimensions of the receiving aperture or the sensing elements of the probe being calibrated should also be less than the aperture dimensions of the small antenna. This procedure was followed by [B126], which claims an overall uncertainty in the calibrating field of ±0.5 dB from 1 GHz to 18 GHz and  $\pm 1$  dB up to 35 GHz.

**4.5.1.3 Small Apertures.** In view of the discussion in the two preceding paragraphs, there is no advantage in using a large antenna as a source. In fact, one can operate at closer distances with less transmitter power if the source antenna is kept relatively small. Open-ended waveguides are perhaps the smallest practical source antennas. They are readily available, do not have serious mismatch problems, and yet have sufficient directive gain to concentrate the energy in the calibration region and facilitate the suppression of scattered energy in the test chamber. Further, one can easily operate at distances greater than four  $a^2/1$ . However, an open-ended waveguide antenna should consist of a section of waveguide whose aperture end extends several wavelengths from any flanges or bends. Also, the aperture (radiating) end should be very cleanly cut in the plane perpendicular to the axis of propagation of the guide. For common open-ended waveguide apertures with a two-to-one aspect ratio, i.e., a/b = 2, the far-field gain is approximated by the equation [B79].

(Eq 4.4)

(Eq

where *f* is the frequency in GHz and *a* is the width (larger dimension) of the waveguide aperture in meters.

When it is necessary to calibrate a large number of nominally identical hazard meters, the extrapolation method described in [B99] is useful when applied as follows Let  $B_d$  be the meter indication with the probe at an arbitrary near-field distance d, and  $B_0$  the indication with the probe at a large distance  $d_0$  where far-field conditions hold. One can write the relations

$$B_{\rm o} = KW_{\rm o}$$

4.5)

$$B_d = KW_d$$

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where  $W_o$  is the far-field power density,  $W_d$  is the equivalent plane-wave power density in the near field, and *K* is a proportionality factor that relates the meter indication to the incident power density. In the extrapolation technique,  $B_d$  is measured over a range of *d* distance, and a power series is fitted to the product  $B_d d^2$  over the measurement interval. This series is then used to determine  $B_0 d_0^2$  by extrapolation. One can then obtain the ratio

$$\frac{B_d d^2}{B_o d_o^2} = F_d \tag{Eq 4.6}$$

 $F_d$  (the near-field correction factor) can also be determined without recourse to the extrapolation method if a long enough range is available to measure  $B_0 d_0^2$  directly. Combining Eq 4.5 and Eq 4.6 yields

$$W_d = F_d W_o \left(\frac{d_o}{d}\right)^2 = \frac{F_d P_T G_T}{4\pi d^2}$$
(Eq 4.7)

since

$$W_{\rm o} = P_T G_T / (4\pi d_{\rm o}^2).$$

 $P_T$  can be measured, and  $G_T$  (the far-field gain of the transmitting antenna) can be obtained by the extrapolation method or other methods previously referenced. The near-field correction factor  $F_d$  is a function of *d*, and should be determined for every combination of radiator and probe type. However, once  $F_d$  has been obtained for a given probe, it can be used to calibrate other probes of the same type with little additional error.

Practical near-zone calibration facilities can be established for use at frequencies above about 300 MHz. Standard-gain horns are normally used as radiators above 1.1 GHz (that is, WR650 and smaller waveguides). The transmitter power required to produce a 10 mW/cm<sup>2</sup> calibration field is only 10 to 20 W. Between approximately 300 MHz and 2.6 GHz open ended waveguides (WR2100-WR430), are more suitable for the radiating elements because, as shown in Eq 4.3, less power is required to produce a calibrating field of sufficient power density and uniformity with known on-axis gain. In this case, transmitter powers of 50 W or less will produce a 10 mW/cm<sup>2</sup> field. Note that the near-field gain calculations for horns do not yield accurate results for waveguides, so the near-field gain should be measured. A good discussion of free-space methods of calibrating hazard meters from 500 MHz to 20 GHz is given by [B22], which indicates that an overall uncertainty of  $\pm$ 1.0 dB or less can be achieved if sufficient care is taken.

#### 4.5.2 Calibrations Using Rectangular Waveguides.

The fields inside a rectangular waveguide can be calculated and, in some cases, are sufficiently uniform to be considered for calibration purposes. The main advantage of such a system is that considerably less power and space are required. One disadvantage is that the maximum transverse dimension of a rectangular waveguide should be less than the free-space wavelength at the highest calibration frequency in order to avoid higher-order modes that result in complicated field distributions. Hence, the method is generally used only for frequencies below 2.6 GHz (WR430), since the device being calibrated should be small compared with the guide dimensions. The distribution of the known field in the waveguide is an approximation to leakage fields propagated from a leak in a microwave oven door. The field in the latter case would decay rapidly with distance from the radiator, and the probe sensor would experience the major effect of the field, i.e., the handle and cable are illuminated by a greatly decreased field. Therefore, waveguide calibrations may provide a lower uncertainty in the calibration

of leakage probes than calibration in a plane-wave field where the entire probe is more uniformly illuminated. A careful error analysis of this problem has not been completed, but it appears that if the maximum probe dimension is less than one-third the smallest waveguide dimension, the total uncertainty will not exceed  $\pm 1$  dB [B6, B126].

#### Fig 4.5 Rectangular Waveguide Calibration System

Figure 4.5 shows how a section of rectangular waveguide can be used for calibrations. A reflectionless load is connected to the output end to prevent standing waves that would cause serious errors in the calibration. The probe to be calibrated is usually inserted into the waveguide through a hole in the side wall (as in Fig 4.5) and positioned in the center of the guide where the field is most nearly uniform. (Entry through the top wall is not recommended because spurious pickup by the leads that are then aligned with the E-field is greater.) The access hole should be as small as possible to minimize perturbation of the field distribution. Equations for calculating the field distribution from  $P_n$  (the net power delivered to the section) and the guide dimensions can be found in [B59, B86]. The "equivalent power density" can be determined in terms of  $E^2$  (not  $E \ H$ ) at the center of a rectangular guide in which the width, *a*, is twice the height, *b*, from

$$W = \left(\frac{4P_n}{a^2}\right) \left[1 - \left(\frac{1}{2a}\right)^2\right]^{-1/2}$$
(Eq 4.8)

A slide and slot in the side wall of the guide may be used to evaluate and reduce the uncertainty produced by any standing wave within the guide.  $P_n$  is determined in the same way as  $P_T$  in Eq 4.1. It is difficult to estimate the total uncertainty of this method because the field strength at the probe sensor being calibrated will be modified by the size and nature of the probe. The influence of the conductive walls on probe calibration in a waveguide is analyzed and evaluated in [B69]. The error assessment of the probe voltage for dipoles of lengths of 20 and 30 mm having terminations of 100 ohms in a WR430 waveguide is 1% and 2.5%, respectively. (The 100 ohm termination impedance is typical for thermocouple, but not for diode based probes used for microwave oven leakage measurements). [B126] describes a system that operates from 400 MHz to 600

MHz with an estimated uncertainty in power density of  $\pm 12\%$  (0.5 dB). Later results at 2450 MHz have been reported by [B6], which claims an accuracy of  $\pm 5\%$  (0.2 dB). In this case, the probe diameter was 1.6 cm compared with the narrowest guide dimension of 5.46 cm.

## 4.5.3 Calibrations Using TEM Cells.

Another guided wave method suitable for calibrating EM-field probes below frequencies of about 500 MHz is the use of a transverse electromagnetic or TEM Cell. Like the rectangular waveguide, this device is fully shielded, and does not emit energy that may be hazardous or cause interference with nearby electronic equipment. Other advantages are excellent long-term stability of the calibration system and moderate cost (compared with an anechoic chamber). The basic TEM cell is a section of two-conductor transmission line operating in the transverse electromagnetic (TEM) mode, hence the name. As shown in Fig 4.5, the main body of the cell consists of a rectangular outer conductor and a flat center conductor located midway between the top and bottom walls. The dimensions of the main section and the tapered ends of the cell are chosen to provide a 50  $\Omega$  characteristic impedance along the entire length of the cell [B34]. When the cell is properly designed and terminated in a reflectionless load, the input voltagestanding-wave ratio (VSWR) is usually less than 1.05 for frequencies below the cutoff limit. In the center of the calibration zone, halfway between the center conductor and the top (or bottom) wall, the E-field will be vertically polarized and quite uniform. Also, the wave impedance (E/H) will be close to the free space value of 377 ohms. Introduction of the probe into this region will alter the field distribution in the vicinity of the probe, but the total uncertainty in the field strength is less than 1 dB [B34, B35, B90] if the maximum probe dimension is less than b/3 where b is the distance from the top wall to the center plate. Cells can be made in various sizes to suit particular needs and to cover specific frequency ranges. However, since the width (surface parallel to the surface of the center plate) should be less than a half-wavelength to avoid higher order modes in the cell, the upper useful frequency of a TEM cell is approximately 500 MHz unless the cell walls are lined with RF absorbing material (see 4.5.3.6).

#### Fig 4.6 Typical Transverse Electromagnetic (TEM) Cell

For proper use of TEM cells, several factors should be considered, including the following:

- (1) the electrical characteristics of the cell
- (2) higher order modes
- (3) relative size of the probe being calibrated with respect to the plate separation, and
- (4) stability and calibration of the voltmeter, directional couplers and power meters used in conjunction with the cell to produce field strengths with an absolute, known value.

Several of these issues have been considered in more detail by [B58] and [B90]. 4.5.3.1 Electrical Characteristics. The rectangular TEM cells available

commercially are designed to have a characteristic impedance of approximately 50 ohms. This value can be calculated from the equation:

$$Z_{o} = \frac{94.2}{\frac{w}{d} + \frac{2}{\pi} \ln\left(1 + Coth\frac{\pi g}{2b}\right)}$$
(Eq 4.9)

where the dimensions *w*, *b* and *g* are given in Fig 4.6 [B89]. The characteristic impedance can be measured with a time domain reflectometer (TDR). The TDR can also be used to check for and correct impedance mismatches, particularly at the transitions. The fields at the test point, i.e., the geometrical center of the center plate (septum) and midway between the center plate and the upper (or lower) wall of the cell, can be calculated from:

$$E = V / b = \frac{\sqrt{P_n Z_o}}{b}$$
 (V/m) (Eq 4.10)

H = E/377 (A/m)

where *V* is the voltage at the input or output port of the cell,  $Z_0$  is the real part of the characteristic impedance of the cell and *b* is the distance from the upper wall to the center plate.  $P_n$  (the net power delivered to the cell) is determined in the same way as  $P_T$ , and the discussion in 4.5.1 applies. The equivalent plane wave power density *W* can be calculated from:

$$W = E^2/377 (W/m^2)$$
  
or (Eq 4.11)

 $W = 377 H^2 (W/m^2)$ 

These field values apply only at the test point for a well-matched cell and significant variation will be seen closer to or farther from the septum.

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**4.5.3.2 Higher Order Modes and Standing Waves.** The maximum operating frequency of a cell is determined by calculating the cutoff frequencies for the higher order modes. The TE10 cutoff frequency  $f_c$  is given by the relationship

$$f_c = (75 / a)\sqrt{1 + (4a / \pi b) \ln(8a / \pi g)}$$
(Eq 4.12)

where  $f_c$  is expressed in MHz and a, b and g are the cell dimensions in meters. The TEM cell should be used well below this frequency to assure proper operation for probe calibration. Actually, some difficulties may be presented by the TE01 mode that can be determined by methods published in [B51] or [B125]. However, if the frequency of operation is limited to half the TE10 cutoff frequency problems should not occur, and uniform fields should be obtained if the cell is properly designed. (Additional analyses on higher order modes have been performed by Hill [B58].) However, slight errors in design or construction may sometimes lead to impedance discontinuities in the cell, particularly in the taper regions. These mismatches can produce standing waves, which generate errors in the values of the fields at the calibration point. To assure proper cell operation, field maps at each desired frequency of operation should be performed in a plane above the septum, half way between the septum and outer wall of the cell. Maps can be made by using E-field and H-field probes with small sensors. When using the cell, forward and reflected power measurements should be made at the input port. Constant values assure consistent cell calibration. The error due to standing waves can be estimated by calculating the ratio of the field at the test point to the average field above the septum, the average being taken from one end of the cell to the other along the centerline.

**4.5.3.3 Probe Sensor Size with Respect to Plate Separation.** If the sensor being calibrated occupies 1/3 or less of the distance *b* (from septum to outer wall of the cell), field perturbation error is less than 10% for the E-field and can be corrected to within 1% using methods described in [B36]. However, as the probe length is increased, the probe response is increased over that expected from the calculations. This field enhancement is due to loading of the chamber, and, since there is no accurate way to correct for this error, one should limit the space used to less than *b*/3, effectively eliminating the problem. This limits the useful range of a TEM cell to frequencies below 500 MHz for probes with sensors having a diameter of 5 cm.

**4.5.3.4 Power Measurement Stability.** The accuracy of probe calibrations using a TEM cell is directly related to the accurate determination of the cell voltage or the power flow through the cell. There are basically two ways to measure power flow through the cell. The first uses directional couplers to measure input, reflected, and output power thereby determining the net power delivered to the cell. This method ensures that the accuracy of the field strength is associated with the power meter and coupler calibrations that usually have less than 1% uncertainty. The other method involves use of a high-power attenuator that is attached to the load end of the chamber, and a power meter that is attached to the attenuator to measure the power flow through the chamber. The uncertainty of this measurement (approximately 1%) is associated with the attenuator and power meter calibrations. As long as all components remain constant, this is an accurate method for cell calibration. However, since changes in the cell power or power measuring instruments cannot be detected with a single power measurement at the cell output, this is not the preferred method.

**4.5.3.5 TEM Cell Operated with a Termination Impedance Different from 50 W.** If a transmission line (other than a waveguide) is terminated in its characteristic impedance, only the TEM mode is present and the intrinsic impedance (*E*/*H* ratio) has a free-space value of 377  $\Omega$ . However, a TEM cell can be operated without a resistive

(Eq 4.13)

termination at frequencies up to about 30 MHz. When the cell is open-circuited, it acts as a parallel plate capacitor rather than a transmission line, and produces a field with a high intrinsic impedance (high E/H ratio). Large E-fields can then be produced by connecting a broadband step-up transformer at the cell input terminal. By contrast, when the cell is short-circuited it acts as a single turn loop and produces a field having a very low intrinsic impedance. This configuration is useful for producing large H-fields.

**4.5.3.6 Wideband TEM Cells.** If a TEM cell is "loaded" in strategic locations with microwave absorbing material (such as the carbon-filled foam that is used in anechoic chambers), higher-order modes can be suppressed. The resulting cell can therefore generate a calculable, relatively pure TEM mode, at frequencies that are much higher than would be possible using standard TEM cells [B35]. This condition degenerates whenever reflecting (metallic or dielectric) objects are introduced in the cell. For calibrating probes with very small sensors and nonperturbing leads (resistive or fiber optics), specially designed absorber-loaded TEM cells may be useful over a very broad range of frequencies (DC-18 GHz) [B83]. Careful design, testing, and error-analysis should be performed for each specific combination of TEM cell and sensor under test, to ensure reasonable performance at any frequency in the range where undesirable higher-order modes may exist.

#### 4.5.4 Magnetic Field Generators Using Coils.

At low frequencies, the axial magnetic field (in A/m) at the center of a circular loop of wire is simply the current (in amperes) divided by the loop diameter (in meters). For a single-turn coil in free space, a loop becomes self-resonant when the circumference approaches the free space wavelength. For multiturn coils, the resonant frequency is lower because of capacitance between the turns. Using a coil with a total wire length less than I/10, the input impedance is very low, but the field strength value is easily calculable and given by

$$H = \frac{NI}{D}$$

where:

H = Magnetic field strength at the center of a flat coil (A/m),

N = Number of turns in the coil

I = RMS Current (A)

D = Diameter of the coil (m).

The magnetic field at points on the axis of a flat coil is

$$H = \frac{NIr^2}{2(r^2 + d^2)^{3/2}}$$
 (Eq 4.14)

where:

r = Radius of the coil (m)

d = Distance along the axis from the coil center to the field point where the H-field is to be computed (m)

This type of coil is useful for probe calibration purposes up to about 30 MHz. Figure 4.7 is a sketch of a Helmholtz coil arrangement that consists of two flat coils on the same axis, both carrying current in the same direction. This type of coil system generates a more uniform H-field over a larger volume than the single coil. For the purpose of generating a uniform H-field in which a typical hazard probe can be calibrated, Helmholtz coils are useful up to about 10 MHz. This frequency limit is dictated by the dimensions of the coil that should be small compared with a wavelength. The wave impedance (E/H

ratio) is low for frequencies below self-resonance. The spacing between the two coils is generally made equal to the radius of each coil. In this case,

$$H = 1.43 \frac{NI}{D}$$
 ......(Eq

4.15)where:

H = Magnetic field strength at point P midway between the two coils (A/m)

N = Number of turns on each of the coils

D = Diameter of each coil (m).

Since  $B = \mu H$ , it is possible to calculate the magnetic flux density *B* in microtesla from the equation

$$B = 1.80 \frac{NT}{D} \tag{Eq 4.16}$$

where B = Magnetic flux density at the midpoint (in  $\mu$ T).

# Fig 4.7 Helmholtz Coils for Generating an H-field to Calibrate Hazard Probes at Frequencies Below about 10 MHz

#### 4.5.5 Standard-Probe Method.

This method is the simplest, and may be the best method of calibrating hazard meters for general field use. The principle of this method is to have a stable and reliable probe that has been calibrated accurately (by one of the previously discussed techniques) for use as a "transfer standard." The standard probe is used to measure the field strength produced by an arbitrary RF field-generating device, e.g., antenna or TEM cell, over a particular region in space (or in a waveguide system). Then an uncalibrated probe is placed at the same location in the field that the standard probe occupied, and the uncalibrated probe's meter reading is compared with the known, measured value of the field, based on data obtained with the standard probe. The transmitter and field-generating device used during this process should generate a field that has the desired magnitude and that is constant with time, and the field should be uniform over the region where the unknown probe is placed. Accuracy of about  $\pm 2$  to 3 dB are readily attainable with this method and improved accuracy is possible if special care is taken. The

advantages of this approach are convenience, reliability, and simplicity. A potential source of error when using the transfer standard to calibrate another probe is the possible difference in the receiving patterns of the two probes. Also, in the near field of a radiator, the size of the probe's sensor is important. Ideally, the standard and unknown probes should be nominally identical and the calibration should be conducted in a field relatively free of spatial variations due to multipath interactions between the probe, the radiator, the anechoic chamber and other field generating components, and near-field gradients. In TEM cells or parallel plate transmission systems, capacitive coupling between the probe and the center plate and walls of the cell can create calibration errors.

The transfer standard probe should be stable, rugged, and not easily burned out; it should have a large dynamic range, cover a broad frequency range, and possess an isotropic response. Organizations that cannot justify the construction and maintenance of a calibration facility could have a transfer standard (probe) calibrated by a reputable laboratory. This secondary reference standard can then be used to calibrate a field-generating system at the user's facility. The system, in turn, can be used to calibrate other probes. The transfer standard should be recalibrated at intervals appropriate to the particular standard, based on experience with the stability of the calibration factor with time. See Appendix A for additional calibration techniques.

#### 4.5.6 Evaluation of Field Survey Instruments.

In order to determine electric and magnetic fields accurately, the characteristics of the survey instruments used should be defined in some detail. A series of tests, when properly performed, define worst-case uncertainties that can occur when that instrument is used to make field strength measurements. The tests also indicate what operational procedures can be used to minimize measurement errors. Details of tests for microwave instruments (for frequencies greater than 0.9 GHz) can be found in [B57] and details of electric and magnetic field probe performance evaluations at 27 MHz can be found in [B98]. The minimum test parameters that should be observed are:

- (1) Absolute Calibration -- Should be performed at field strengths that produce indications that equal or exceed the instrument's mid-scale readout display.
- (2) *Instrument Linearity* -- Measurements should be made at 25, 50, 75, and 100 percent of full scale, on each range of the instrument's readout device.
- (3) Amplitude-modulation Response -- This test should be tailored to the desired instrument usage conditions. For example, if the modulation of concern is from unfiltered power supplies, such as in industrial microwave heaters, 60 or 120 Hz modulation with a 50% duty factor would be appropriate. If the instrument is to be used around radar equipment, duty factors as low as 0.001 should be used during the evaluation process.
- (4) Frequency Response -- The response of the instrument over the frequency band of interest should be determined. The response should be relatively flat over the design frequency range, ([R]1 to 3 dB).
- (5) Out-of-Band Response -- The sensitivity of the instrument, probe, cable and readout should be evaluated for exposure to fields at frequencies far from the specified usable range of the instrument. This is important when instruments are used in mixed-frequency fields.

- (6) Near-Field Response -- The response of an instrument in very high- and lowimpedance fields (*E*/*H*) should be investigated to determine the instrument's response to extraneous fields. For example, the H-field response of an E-field instrument, or E-field response of an H-field instrument, should be evaluated with appropriate field-generating devices.
- (7) *Polarization Ellipticity* -- The variation in response as the probe is rotated about the axis of its handle should be defined.
- (8) *Receiving Pattern Isotropy* -- Variations in response can occur as the handle of a probe is rotated through the E-plane so that the handle is parallel to the E-field at one position during the test.
- (9) *Lead Pickup* -- Tests to detect and quantify extraneous pickup by the probe handle are important (see 4.4.2), [B91].
- (10) *Temperature Response* -- Changes in the instrument's response to a given field strength over the temperature range of interest, should be determined.
- (11) Supply-Voltage Response -- Several commercially available RF hazard survey instruments utilize one or more batteries including single battery-packs with several separate voltage terminals. It has been found that deviations from the nominal voltage rating of one or more of the batteries can cause errors in the overall accuracy of the instrument [B57]. Therefore, adequate battery-voltage testing is important for battery operated RF survey meters.
- (12) Radio Frequency Interference (RFI) -- The response of the instrument readout device and cable should be determined with the probe, cable, and readout device exposed to E- and H-fields whose magnitudes lie in the range of use for the probe during typical hazard surveys.
- (13) *Drift and Noise* -- Short-term and long-term stability of the instrument should be determined with respect to full scale on each measurement range of the instrument, in the absence of EM-fields.

Based on all of the above tests, a worst-case error analysis can be performed. As in any experiment, care should be taken to design each test to measure only the parameter of interest, while all other parameters are held constant.

#### 4.5.7 Practical Measurement Accuracy.

Several methods for calibrating hazard meters have been discussed and the uncertainties associated with each method were estimated. It is important to understand that one cannot expect to achieve the same accuracy when using the meters for practical measurement applications. Some of the reasons are as follows:

- (1) Hazard meters are usually calibrated in nominally plane-wave or uniform fields. Such fields are seldom encountered in practice, and the sensor may not respond in the same way to nonplanar fields (fields with large spatial gradients).
- (2) In most calibration methods, only the sensor (probe) is exposed to the field while, in practice, the complete system, including the indicating unit and connecting cable, is

immersed in the field. Significant errors can result from spurious responses from other parts of the instrument including readout meter (case) and cable.

The overall uncertainty added by the above factors is difficult to assess and will vary with the type of meter and usage situation. However, if good measurement procedures are followed, accuracy of  $\pm 1$  to 3 dB can be expected in practice, with greater uncertainties in near-field situations and at higher frequencies (shorter wavelengths), or in areas where large reflecting objects are present.

#### 4.6 Instruments for Measuring Internal Fields, SAR and Induced Curent

#### 4.6.1 Implantable E-Field Probes for SAR Measurements.

Implantable E-field probes provide the ability to measure the local electric field strength in tissue at a specific point, or they can be used to obtain a continuous real-time or linear scan in tissue. A typical isotropic device consists of three orthogonally arranged dipoles which are directly loaded with Schottky (metal-barrier) diodes. The RF signal is directly converted to a DC voltage by the diodes and the DC signal is transmitted to a data conversion unit by high resistance (RF transparent) lines. The high resistance lines have been realized either by carbon impregnated Teflon strips or by thin- or thick-film technology on a ceramic or quartz substrate. A typical probe consists of a thin-film dipole, 0.6 to 3 mm in length, and a pair of high-resistance leads placed on a thin plastic or glass substrate and encapsulated in a low-dielectric constant insulating material. A beam-lead diode is placed across the gap of the dipole to provide RF detection capability of the order of 1 mV per mW/cm<sup>2</sup> in free space [B19]. Typically, a 1 to 5 mm spherical or cubical volume is required to house the 3 orthogonal dipoles of an isotropic internal E-field probe. This means that field strength and, hence, SAR data may be obtained with a spatial resolution of better than several millimeters. This is about one-half of a wavelength in high-water-content tissue such as muscle, brain or internal organs at a frequency of 3 GHz. A low-frequency limit of about 300 MHz is achievable. The lower limit is due to the fact that the high-resistance leads cannot reject RF pickup adequately at low frequencies, (as discussed in 4.4.2). Implantable E-field probes with 1.5 to 2.5 mm long dipoles have been produced in small quantities by several commercial firms. Reviews of the theory of this type of probe are discussed in [B19; and Schmidt, et al., 1996].

Recent improvements in decoupling the diodes from the high resistance line have led to the realization of true RMS sensors. The improvement is achieved by employing thick-film techniques, which allows the use of multiple sheet resistances on the same substrate, to provide high-resistance lines of several k $\Omega$ /square. Measurements have shown that the dielectric material around classically designed probes can significantly disturb the reception pattern, leading to deviations from an isotropic pattern that are larger than ±2 dB. Possibilities for optimization have been studied using numerical modeling. Solutions have been found that will reduce the deviation from an isotropic response in all planes and polarizations to better than ±0.35 dB [Pokovic, et al., 1996]. Additional methods for determining SAR are given in Appendix C.

#### 4.6.2 Calibrating induced current meters.

Induced current meters can be used for determining SAR in the ankles or wrists. Calibration is accomplished using current injection and power measurement techniques in a terminated RF circuit. To calibrate a stand-on meter an RF source is connected,

through a suitable series resistance, to electrodes that contact the upper and lower plates of the induced current meter. An RF current transformer is used to measure the current applied to the induced current meter and the output of the transformer is monitored using an RF voltmeter (Tell, 1995). (See Figure 4.8)

## Fig. 4.8 Current Injection Method for Calibrating Stand-on Induced Current Meters.

The clamp-on induced current meter is calibrated in a manner similar to that used for common RF current transformers. A special fixture is used that connects to a 50 ohm coaxial line and expands the outer shield to allow access to the inner center conductor (similar in principle to a TEM cell). The size of the fixture must accommodate the larger dimension of the clamp-on induced current meter when it is connected around the center conductor of the test fixture. (See Figure 4.9.) An insulating spacer is often used to center the conductor within the aperture of the induced current meter. RF power from a suitable source is transmitted through the fixture to a termination with a means for measuring total transmitted power, e.g., a feed-through termination. The current flowing in the coaxial line and, hence, through the aperture of the induced current meter is calculated using Ohm's law.

## Fig. 4.9 Equipment used for Calibrating Clamp-on Induced Current Meters

## 4.6.3 Techniques for Calibrating Temperature Probes

For the calibration of thermal probes used in SAR measurements, a capability to resolve temperatures down to tenths of degrees C is required. For clinical settings with human subjects, even higher precision is needed [CETAS]. For comparisons to reference thermometers, the ability to produce a stable thermal environment over the range of approximately 10ø C to 50ø C is also required.

In earlier years, it was impractical for a small laboratory to possess a standard platinum resistance thermometer (SPRT) along with the bridge circuitry developed by the National Bureau of Standards. Therefore, many laboratories maintained ultra-stable standard reference thermistors from which highly accurate comparisons were conducted using four-lead resistor-measurement circuits with current-reversing switches and thermal "calibration blocks" to ensure near-zero temperature gradients inside a controlled water bath [Hochuli]. During the past 10 years, however, self-contained and self-calibrating platinum resistance thermometers have become commercially available such that tedious resistance measurements that were fed into a computer-based curve-fitting program of the thermistor equation [Steinhart & Hart ] became unnecessary.

Still required is an insulated, stirred water-bath to provide a homogeneous environment for probe-to-probe comparisons. Sufficient measurement time is also a significant factor in the accuracy of thermal probe calibrations. To avoid the negative effects of stray thermal gradients between the compared probes, several minutes, at least, should elapse before recording new thermal levels.

#### 4.7 Calibration of SAR Instruments

#### 4.7.1 Implantable E-field Probe Calibration Techniques.

The response of E-field probes implanted in biological tissue or tissue simulating material is enhanced by a factor that depends on the dielectric properties of the material. Therefore the implantable E-field probe should be calibrated in lossy dielectric media at points where the absolute value of E is known. Calibrations have been performed in

spheres [B15, B116] and in waveguides filled with lossy dielectric liquids, e.g. salt water, [B59]. These lossy media are exposed to a known external E-field and then placed in the dielectric object in which the internal E-field distribution has been calculated using electromagnetic field theory. The internal E-field calibration factor for the probe is determined from its response compared with the calculated internal field. If this is done at several frequencies for a specific probe design, the calibration uncertainty will be typically 1 to 2 dB when implanted in any lossy, high dielectric-constant object, such as biological tissue that contains a large percentage of water (muscle, brain, and internal organs, but not bone or fat). (See also 5.5.2.)

Another technique for calibrating probes in tissue equivalent material is based on the boundary condition that the tangential component of the electric field is continuous across any interface [B59]. This technique utilizes a waveguide section with a thin plastic septum separating two segments, one containing air and the other segment containing the tissue simulant. (See Figure 4.10.) Various recipes for tissue-equivalent material can be found in the literature, e.g., [Hartsgrove, 1987] and [Chou, 1984].

The probe output is measured in various positions on each side of the septum and the resulting curves are extrapolated to the septum interface. The field in the tissue-equivalent material decays exponentially whilst the field in air varies sinusoidally due to reflections at the air-tissue simulant interface. To minimize interaction with the field, the probe is passed through the narrow wall of the waveguide to place the probe axis and lead wires normal to the E-field. Measurements should be made as close as possible to the interface. The resulting curves should plot as straight lines on semi-log paper.

When distances between the measurement points and the distance between the septum interface and the closest measurement point are equal, the following equation may be used to measure the probe-tissue enhancement factor  $F_{TE}$ :

# Fig 4.10 Boundary Condition Method for Calibrating Implantable E-field Probes in Tissue-Equivalent Material

$$F_{TE} = \frac{V_0}{V_0'} = \left(\frac{V_1}{V_1'}\right)^2 \left(\frac{V_2'}{V_2}\right)$$
(Eq 4.17)

where

 $F_{TE}$  = the tissue enhancement factor

 $V_0$  = voltage measured in air,

 $V_0' =$  voltage measured in tissue,  $V_1$ ,  $V_1'$ ,  $V_2'$  and  $V_2'$  are the voltages measured at the locations indicated in Figure 4.8.

Care should be taken when calibrating the probe in a region of the lossy dielectric object where SAR spatial gradients are large. This is often the case for probes calibrated at the higher microwave frequencies in high-dielectric constant media (e.g., saline-based phantom materials). An alternate method for calibrating implantable E-field probes is to measure the SAR in an irradiated object with a precisely calibrated temperature probe, and then place the E-field probe at the exact location where the SAR was measured. Techniques for calibrating temperature probes

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## 5. Measurement of Potentially Hazardous Exposure Fields

#### 5.1 **Preliminary Considerations**

#### 5.1.1 Source and Propagation Characteristics.

Before carrying out a survey of potentially hazardous EM fields, it is important to determine as many of the known characteristics of the sources of these fields as possible and estimate their likely propagation characteristics. This knowledge will permit a better estimate of the expected field strength and, consequently a more appropriate selection of test instruments and test procedures. A check list of source characteristics should include the following:

- (1) Type of RF generator and the output power;
- (2) Carrier frequency(ies), signal duty factor, pulse width, pulse-repetition frequency, etc.;
- (3) Modulation characteristics, e.g., peak and average values, waveform, etc.;
- (4) Intermittency, e.g., scanning beams, operational duty factors;
- (5) Number of sources–If more than one source is present, are some or all of the signals coherent? Are intensities likely to add linearly or will they create interference patterns (standing waves etc.)?
- (6) Spurious frequencies including radiated harmonics.

A check list of propagation characteristics may include the following:

- (1) Distance of source to test site;
- (2) Type of antenna and properties including gain, beamwidth, orientation, scanning program, physical size with respect to the distance to the area being surveyed (i.e., near field, etc.);
- (3) Polarization of the E- and H-fields (linear, elliptical);
- (4) Existence of absorbing or scattering objects likely to influence the field distribution at the test site.

A review of such a check list, is a necessity if the surveyor is to avoid some simple, but often surprising, situations. For example, it is necessary to know the location of the source and RF propagation path during surveys with hand held probes. Only then can an appropriate assessment of the effect of the presence of the surveyor's body be made, and measurement errors avoided. Another example common in leakage situations is the possibility that the levels of the EM fields may be hazardous to the surveyor and may produce RFI in the instrument electronics if it was not designed for operation in the presence of such fields.

#### 5.1.2 Estimate of Expected Field Strength.

If the fields are far fields or radiating near fields of an antenna, then the material on theoretical calculations of exposure fields (Appendix B) can be used to obtain field strength estimates. General references on antennas and hazard surveys are useful [B69, B71, B84, B111, Tell B, C].

If more conservative values for the on-axis power density maxima  $W_m$  are desired for tapered aperture distributions, the curves in Appendix F can be used. However, one should always keep in mind that equations in Appendix B and the curves in Appendix F are only suitable for obtaining approximate power densities for estimation purposes. More definitive values will require careful measurements.

Field enhancement due to ground reflections could increase *W* by as much as a factor of four times and even more if focusing effects are present. It should be recognized that such fields measured in the absence of a person may be misleading relative to hazards. For example, a person exposed in front of a reflecting plane reduces the magnitude of the standing wave by his or her absorption.

In the case of low frequencies or small-aperture antennas, the existence of potentially hazardous reactive near fields becomes relevant. These fields can be accurately calculated only for well-defined antennas of simple geometry. Since radiator geometries are rarely simple or well defined for low-frequency (1 MHz - 300 MHz) sources in commercial use, measurements of *E* and *H* are usually required. However, one can always utilize the general property [B82] that reactive fields predominate at distances *d* close to sources where  $2pd/l \ll 1$ . Reactive near-field amplitudes diminish as  $1/d^2$  or faster, whereas radiation far-field amplitudes diminish as 1/d. General texts [B82] can sometimes be used to estimate E- and H-field values at these lower frequencies, and specific literature [B26, B32, B37, B39, B56, B121] on the propagation characteristics of various broadcasting and communication antennas can be used to estimate either near or far-fields from these sources.

#### 5.1.3 Determination of Type of Instrument Required.

Although many instruments designed for potentially hazardous fields are broadband in nature, there are none that cover the entire frequency range of interest and all parameters of potential interest. Some general considerations in the selection of an instrument include the following:

- (1) Frequency-Frequencies should be determined in advance so that proper instruments and measurement methods can be selected. The presence of several frequencies dictates the use of a broadband device with true RMS response. (If the operating frequency cannot be readily determined, e.g., from the equipment label or user information, the use of a spectrum analyzer or frequency counter should be used to determine the frequency.)
- (2) Response Time-It is usually desirable to begin a survey using a hazard instrument with a response time (integrator time-constant) of one second or less (the "fast" setting on some commercial instruments). This enables a coarse measurement or the detection of pulse-modulated or intermittent fields, e.g., those created by a scanning radar beam. A "peak hold" feature on some survey instruments can provide an accurate indication of moderately fast bursts of RF energy (duration greater than several milliseconds). Once a high- field-strength zone is located, a slower time constant (3 s or more) should be used to obtain the time averaged value of the field strength. If the hazard meter still indicates that an intermittent field exists, other means of recording and averaging should be used. Data-logging systems are available specifically for use with RF hazard meters.
- (3) *Peak Power Limitations*–A knowledge of the peak power limitations of the instrument is necessary to protect probes from damage in some low-duty-factor pulsed fields, such as those associated with radars.
- (4) *Polarization*–A knowledge of the polarization of the fields enables a surveyor to use a non-isotropic probe for hazard surveys. In the absence of such knowledge, an isotropic probe is highly desirable both for ensuring accuracy and ease of performance of the survey in a reasonable period of time.
- (5) Dynamic Range–The maximum anticipated field strengths should be estimated before measuring emissions from an RF source. A survey instrument capable of withstanding continuous exposure to field strengths ( $E^2$  or  $H^2$ ) of at least ten times the predetermined value should be chosen in order to avoid destruction of the probe sensing elements or the high-resistance leads connected to those elements. In addition, adequate sensitivity is required to ensure a reasonable signal-to-noise ratio when the minimum expected field strengths are being measured.
- (6) *Near Field Measurement Capabilities*—If a leakage situation exists, or if the fields in close proximity to a source are to be measured, care should be taken to select a suitable instrument (see 5.3.6).

#### 5.2 Safety Precautions

Personnel should take appropriate safety precautions while conducting surveys, and the degree of care exercised should increase in proportion to the power levels associated

with the systems being surveyed. The nature of the precautions will also differ for leakage surveys compared with measurements of deliberate radiating systems (antennas).

#### 5.2.1 Hazards Not Directly Associated with the Survey.

Before discussing precautions directly related to the surveying process, it is worthwhile to consider potential hazards, other than RF exposure, which might be associated with the electronic equipment or system being surveyed.

- (1) High Voltage. Electrical and electronic equipment can present potential and lethal shock hazards. Ordinary precautions such as not defeating interlock protection systems, exercising care around necessarily exposed high voltage leads and terminals and avoiding working alone near high-voltage systems, should be remembered. It is to be noted that, in many high-power systems, a prime RFleakage source may be the high-voltage electrodes of the transmitting tubes. Additional caution is advised when performing measurements in the vicinity of conductive structures, such as tall cranes or long vertically suspended cables, that are located near high-power, low-frequency RF sources. In such circumstances, large open-circuited voltages can exist on the structures that are exposed to ambient RF fields; these voltages may reach levels of several kilovolts and have the potential for arcing to a grounded body, leading to strong startle responses, and in some cases, severe RF burns. Appropriate precautions should be exercised before contacting improperly grounded objects in strong RF fields.
- (2) X-ray Hazards. In high-power systems utilizing high-voltage transmitting or other high- power tubes (greater than about 20 kV), there is generally the potential for Xray emission. It, therefore, may be desirable to first conduct a survey of X-ray emissions before an RF survey is conducted in close proximity to such transmitters. One should take care that the X-ray survey instrument is not susceptible to RFI.
- (3) *DC Magnetic Fields*. Very high-power systems may include sources of strongstatic and low-frequency magnetic fields. Survey personnel should avoid sustained close proximity to such sources even though definitive exposure standards do not exist for static fields Also, tools have been known to fly out of the surveyor's pockets, etc., causing personnel injury.
- (4) *Indirect RF Hazards*. It is important to remember that the presence of RF fields can produce hazards, or at least undesirable effects, besides those arising from exposure of body tissue. Since surveys may be conducted not only in controlled laboratory conditions, but also near mobile transmitters, in industrial situations, and even in homes, one should at least be aware of the following possibilities:
  - (a) Serious hazards are associated with the potential exposure of electroexplosive devices (EEDs), combustible gas or flammable materials to EM fields. ANSI/IME 20-1978 [3], (or current revision thereof), on safe exposure distances for EED's relative to various transmitters should be followed.

- (b) It is generally important to realize that potential EM interference to electronic devices or systems often occurs at levels far below those that cause bodily harm. This interference may be only an annoyance, for example the effects of low-elevation radar operation on TV or consumer electronic devices in homes, or it may be more serious, such as causing the possible reprogramming or "upsets" of microprocessor-controlled medical devices, e.g., pacemakers, or causing errors in digital computers controlling industrial processes. In any case, such interference is undesirable, and care should be exercised when operating systems, particularly in abnormal modes of operation that may be useful for survey purposes. A clear assessment of the impact of potential interference problems should always be made before beginning a survey. It is also understood that surveys are to be made with responsible adherence to the rules of the Federal Communications Commission and other bodies that regulate against RF interference.
- (5) *Burns*. (Associated with high-power fields.) One should take care to prevent RF burns resulting from handling conducting objects exposed to these fields or RF cables with exposed connectors [B105]. In addition, one should follow ordinary precautions in the operation of RF heating systems and plastic sealers during surveys, e.g., avoid handling test loads, sealing bars and superheated liquids.
- (6) Abnormal Modes of Operation. One should be aware that electronic systems have the potential for abnormal modes of operation in which spurious frequencies and unintended leakage radiation are generated at significant power levels. The operation of a system should not be attempted by the surveyor without qualified personnel present to assess the normality of operation, in order to avoid such situations.
- (7) *Fall Hazards.* Personnel performing RF field surveys must always be aware of the potential for falling from roof-top or tower antenna sites, elevated antenna pedestals, etc. The use of safety lines and climbing harnesses is recommended.

## 5.2.2 Precautions During the Survey Process.

Serious precautions are clearly required when surveying a radiating system, such as a high-power radar. Such precautions include the following:

- (1) Based on known parameters of the system, the survey process should be planned so as to limit exposure of all personnel to levels below the MPE specified in the latest version of IEEE C95.1-1991 [1], or other applicable exposure guides or standards. This limitation relates not only to power density, but also exposure duration. If survey personnel are exposed to field strengths in excess of those specified in the guides for continuous exposure, they should be accompanied by other personnel who can ensure that the exposure duration does not exceed the time recommended in the MPE for exposure to higher level fields (see 3.2.2). In such situations, it may be desirable to conduct the survey with radiation-emitting equipment operating at a reduced power level and use power scaling to compute the corresponding field levels that would exist during full-power operation.
- (2) The operation of movable or scanning antennas should be done with full allowance for safety precautions. These precautions range from avoiding injury from bodily collision with rotating or moving structures, to avoiding start-up
operation of RF generators with antennas pointed in the direction of personnel. Before a survey begins, antennas should be adjusted far from the most potentially hazardous position with the surveyors approaching from out of the beam path toward the beam. Furthermore, if measurements are to be conducted while the antenna is scanning, one should first determine whether or not the response time of the instrument is fast enough to respond to the scanning beam. In addition, one must be aware of the burnout characteristics of thermocouple type probe elements which can reach their burnout limit in a pulsed field without reaching the RMS limit of the probe.

- (3) A theoretical examination of the radiation patterns should be made before beginning actual system operation or conducting the survey.
- (4) Antennas should not be pointed toward metal structures, and metal objects should not be inadvertently located close to antennas. These not only create scattering and multipath situations, but are also a potential source of RF burns. However, if the normal area of transmission includes such metal objects, measurements should be conducted in those areas with the objects in place. The presence of secondary structures such as towers, guy wires, fences, reflecting surfaces, etc., can enhance the fields and produce RF hot spots. Allowance for such effects should be made when undertaking a survey. During the survey, the surveyor should be in continual communication with the operator of the RF source so that the source may be controlled in accordance with survey requirements.

When performing leakage surveys, the following precautions should be taken:

- (1) The possibility of leakage exists at the site of the RF generator, along any transmission line or waveguide conveying power from the generator (particularly at waveguide joints), and at all access doors and panels of enclosures housing the generator. Normally, leakage energy drops off as the inverse square of the distance. Therefore, in conducting a survey, one should begin by approaching the generator, antenna, or any unintended radiating/leaking structures from a "safe" distance. The survey instrument should be set to a "high" range to alert the surveyor of the possibility of over-exposure (for example, greater than 10 mW/cm<sup>2</sup>).
- (2) The possibility of RF burns exists, so contact should be avoided with any metallic structure on or near a point where high field strengths could exist.
- (3) When opening access doors or panels to insert or remove a test load (for example, in a microwave oven or RF-exposure test chamber), the equipment should be shut down first and the interlock systems left operative.
- (4) In checking for possible inoperative interlocks at an RF enclosure access port, one should ascertain leakage levels while the source is on and the port is closed. Then the surveyor may slowly open the port to observe any increase in leakage and possible interlock failure.

- (5) Foreign objects (especially metallic objects) should not be inserted into any opening or port of the RF enclosure. This applies particularly in the case of high-power industrial systems that use conveyor belts carrying materials through ports of the RF enclosures.
- (6) With the source switched off, the surveyor should visually inspect all flexible waveguides that carry high power. This inspection should determine signs of fatigue, aging, damage at joints, and lack of adequate support, etc.

# 5.3 Measurement Procedures for External Fields

### 5.3.1 General Considerations.

Prior to making measurements, one should estimate the expected field strength and determine the type of instrument required, as discussed in 5.1. Additional approaches and equations for calculating field strength in various situations are given in Appendix B. The measurement procedures to be used may differ, depending on the source and propagation information available.

If the information is adequate, then the surveyor, after making estimates of expected field strengths and selecting an instrument, may proceed with the survey. The surveyor should use a high-power probe with the range switch set on the most sensitive scale The high-intensity field areas, e.g., the main beam of a directional antenna, should be approached from a distance to avoid probe burnout. The surveyor then gradually proceeds to move progressively closer to the regions of higher field strength. Extreme care should be exercised to avoid overexposure of the surveyor and survey instrument. For low frequencies the E-field should be measured first since it represents a greater hazard.

On the other hand, if the information is not well defined (for example, reports of strong, intermittent interference), then it may be difficult to make a hazard survey without first conducting an empirical, hazard assessment. A survey for potentially hazardous fields of unknown frequency, modulation, distribution within an area, etc. may require use of several instruments. Examples of such instruments are spectrum analyzers or field-strength meters that display frequency-domain information with a means to analyze amplitude modulation characteristics, and which have a wide dynamic range, e.g., 60 dB in power. After this preliminary procedure is performed, it may be possible to continue a more meaningful survey with isotropic hazard survey instruments.

### 5.3.2 Far-Field, Single-Source Conditions.

The measurement of a linearly polarized plane- wave field whose source location, frequency, and polarization are known, may be performed with a tunable field-strength meter of acceptable accuracy, which covers the frequency range of interest. This instrument is used with a calibrated conventional antenna such as a standard-gain horn or dipole. Alternatively, an isotropic hazard probe may be used.

Multipath reflections may create highly nonuniform field distributions, particularly at frequencies in excess of 300 MHz. To judge the level of exposure at any specific location, a series of measurements should be made over a square area whose sides are

approximately one or two meters in extent. The spatial average of the field within that area should be considered as the appropriate level for comparison with whatever protection guide is being employed as a criterion. Measurements near metallic objects should be made with the edge of the probe at least 3 "probe lengths," e.g., 20 cm, from the object (see 5.3.6.1).

In an analysis of RF field measurement data obtained during a filed survey in a metropolitan area adjacent to a large antenna farm, Tell [D] reported on the degree of variability in the measured field magnitudes obtained at the same measurement location, using the same instrumentation with measurements performed by the same individual but with different orientations of the observer relative to the measurement point. A broadband electric field probe equipped with a datalogger was used to acquire the minimum, maximum and average square of the RF field strengths along a vertical path two meters in height with the observer facing the measurement point from four different directions. Measurements were performed at 171 different locations along neighborhood streets near a broadcast site with six FM radio antennas. Observer induced field perturbations accounted for a mean value of 71 percent (2.3 dB) variation in the spatial average measurement of the RF fields (ratio of maximum spatial average to minimum spatial average obtained from four successive vertical scans at each measurement point). RF fields were found to be non-uniform over the body dimension with a mean value of the maximum to minimum field readings of 10.3 (10.1 dB). These variables in measured fields should be carefully considered when interpreting measurements used for showing compliance with exposure limits.

While mounting or holding the measuring antenna or probe, care should be taken to avoid reflections or perturbations of the field by support structures or by the operator's body. Where required, to avoid field perturbation, metallic portions of the measuring device, or support structure, should be covered with absorbing material of appropriate quality. Where possible, probe interconnect cables should be oriented normal to the electric field. When that is not practical, or where severe multipath effects produce fields originating from multiple directions, metallic cables should be covered with absorber unless tests demonstrate that the cable position does not affect the measurement. Dielectric fixtures should be as small as possible (minimum reflection cross section) and should be of low dielectric-constant material, or be less than one-quarter wavelength in effective thickness  $T_E$ . The effective thickness is given by:

### $T_E = T(e_r)^{1/2}$

(Eq 5.1)

where *T* is the physical thickness, and  $e_r$  is the relative permittivity. Even dielectric slabs  $(e_r > 2)$  can significantly alter plane wave fields if the effective thickness is greater than 0.1 wavelength.

For highest accuracy, sources of error can be accounted for so that the true field strengths may be ascertained with less than  $\pm 2$  dB of uncertainty. To obtain this level of accuracy at frequencies above approximately 300 MHz, a scanned measurement or many fixed-point measurements per wavelength should be performed in order to obtain information on the variations in field strength in that area due to multipath and other reflections.

# 5.3.3 Far-Field, Complex Sources.

When measuring the fields from multiple, distant sources of unknown frequency, polarization, or direction of propagation, a broadband isotropic probe is required. Since

standing-wave effects and multiple-source field interactions should be accounted for, it is necessary to scan a volume of space in the zone of interest. The area should be divided into a grid of one meter squares, and measurements should be taken at each grid intersection. Scans should also be made in the vertical plane at grid intersection points.

In the case of multiple sources of unknown polarizations, a single-axis probe (linear dipole) cannot be used to provide accurate data in a reasonable length of time, since measurements with three orthogonal orientations of the probe should be performed to ensure that all components of the field are accounted for. If a single-axis probe or linearly polarized antenna must be used, one should be sure that the field being measured is time invariant. Even if an isotropic probe is used, it should be relatively free of sources of measurement errors caused by reflections from the probe, cables, readout case, and the surveyor. The use of long (many meters) high- resistance or fiber-optic probe interconnect cables will minimize the reflection problems mentioned above.

# 5.3.4 Near Fields.

Since large field gradients exist in the near field of an active radiator or passive reradiator, their measurement requires the use of a probe with an electrically small array of three orthogonal dipoles and, for frequencies below approximately 300 MHz, an array of three electrically small orthogonal loops, in order to provide satisfactory performance for the resolution of these spatial gradients. Otherwise, a spatially-averaged value will be measured by a large probe (one with an effective area greater than one-quarter wavelength in cross section). In addition, a small antenna array produces minimal perturbation of the field and the radiation characteristics of the source are not modified (alteration of reactive near fields). Since the polarization of the fields in near-field situations is usually unknown, under most circumstances an isotropic probe should be used. If the frequency and polarization are known, a broadband instrument is not required. Instead, a narrowband probe with uniform response in a single plane (similar to some commercial, microwave-oven survey instruments with two-orthogonal dipoles) may be used (see also 5.3.6).

### 5.3.5 External Field-Measurement Instrumentation—Usage Considerations.

The scattering effects of the operator, cable, support structure, and readout device have been mentioned. These problems are most significant when a spatially uniform field (plane wave) illuminates all of these objects and the probe sensor with the same approximate field strength. This produces reflections that are directly related to the scattering cross section of the various objects and their distance from the probe. Scattering from the operator's body may introduce errors in the equivalent plane-wave power density of more than 2 dB [B27]. Reflections from readout cables aligned with the incident plane-wave electric field and placed approximately 30 cm behind the sensor of an isotropic probe can cause measurement variations of ±1.5 dB at 915 MHz and ±0.75 dB at 2450 MHz [B14, B17]. These effects become more significant at lower frequencies where the cable length and RF wavelength are comparable. The magnitude of the reflections from a cable, the operator, or other objects increase as frequency decreases for a fixed geometrical relationship between the probe and cable. Therefore, greater care should be taken at frequencies below 1000 MHz to avoid large errors when measurements are performed in spatially uniform fields. Orientation of the cable perpendicular to the incident electric field, or covering it with absorber, will reduce this problem. One should bear in mind, however, that most absorbers are not effective at low

frequencies. Therefore, fiber optically coupled instruments are preferable at frequencies below 300 MHz.

RFI also becomes a significant problem at lower frequencies, since it is more difficult to shield readout electronics and cables at frequencies below about 500 MHz. An operational test of the system can be made by totally shielding the sensor tip of an electric-field probe with metal foil. This procedure allows the determination of the existence of RFI or capacitive coupling between the cable and readout and nearby radiating objects during the measurement. Similarly, reorienting the probe leads in the field, with the probe sensing element fixed, should not alter the reading of the meter appreciably (±6 dB) if RFI is not problematic. It is not uncommon to find large errors in readings from older designs of isotropic instruments when they are used to measure electric and magnetic fields below about 1000 MHz, especially when the readout device and probe cable are exposed to a field of the same magnitude as that which illuminates the probe [B16, B109]. This often is due to faulty RFI gaskets in the readout device or poor system RFI design. Fiber-optic cables, high-resistance cables, or double-shielded coaxial and signal cables can be used by designers of hazard survey instruments to connect the probe to the readout device, in order to minimize the effects of RFI. Also, a self-contained (active antenna) probe with integrated antenna, sensor, and readout in the same housing can minimize the problem of RFI otherwise induced in cables or readout [B19].

Instrument linearity as a function of the peak-to-average power ratio of amplitudemodulated fields should be considered when performing measurements in an unknown situation. Pulse duty factor or amplitude-modulation waveform can be quantified through the use of a conventional antenna with a coaxial output, e.g., log periodic, dipole, and a coaxial attenuator and diode detector. The detector output can be fed via coaxial cable into an oscilloscope of appropriate bandwidth. The duty factor of pulsed sources and the waveform of the amplitude modulation can then be observed, and the ratio of the peak to average power or energy density can be ascertained. It is not necessary to maintain square-law operation of the detector for the evaluation of 100% rectangular amplitudemodulated waveforms. For other pulse modulated waveforms, one can only be assured that both the peak and average field strengths are correct if the detector is operated in the square law region [B118]. Also, with square law detection, the ratio of the peak to average equivalent power density, ( $E^2$  or  $H^2$ ) can be quantified.

Measuring the effects of a slowly time-varying field generated by a rotating radar antenna, or a microwave oven with a mode stirrer, can be accommodated by use of the appropriate time constant of the readout instrument. Some instruments provide "fast" and "slow" time constants. Any observable periodic variation of the field strength, as indicated by the instrument in the fast mode of operation, should be considered as low-frequency field modulation and the slow time constant should be used. Care should be taken to avoid mistaking slow variations in reflected signals (due to objects moving in the field) for a true time-varying source modulation. This implies that any movement of the operator, probe, or surrounding objects, should be eliminated during the measurement procedure where time-varying fields are involved.

An ellipticity (non-isotropic reception condition) in the receiving pattern of multi-axis or isotropic probes is always present to a certain degree, (typically 2% to 20%). Only if the ellipticity of an instrument is accounted for, can highly accurate data be taken. In fields where components do not exist along the direction of the probe handle, a rotation of the probe about the handle axis should be performed during a survey (if the surveyor desires

to eliminate this error) at each fixed point of interest within the spatial volume being surveyed. The minimum and maximum values should be recorded and an average value computed. This can be done manually in near-field situations, e.g., microwave oven surveys, without introducing error, if the surveyor's hand is in an area of minimal field strength with respect to the field at the sensor (probe tip). In situations where uniform illumination exists over the entire probe body (tip and handle), larger errors are usually introduced by the presence of the operator's hand, while performing this rotation, than are introduced by probe ellipticity. In this case, if precise data are required, the probe should be supported by a dielectric support and error limits associated with the probe ellipticity should be assigned to the measurement, rather than attempting to manually rotate the probe.

# 5.3.6 Interaction of RF Hazard Probes with Nearby Passive Scattering Objects (Reradiators) and Active Radiators.

When measurements are made with a hazard probe placed close to conducting or highdielectric-constant objects, (scatterers or "passive reradiators") large errors may result. Two situations are addressed in this section. One situation occurs when a hazard probe with an "electrically large" antenna (larger than about 0.25 wavelength) is placed close to field-perturbing objects such as a person's body, or large conducting objects, e.g., a metal pole or metal shed. A second error-producing situation occurs when measurements are made with the probe antennas less than a few probe "antennalengths" or "probe lengths" from an active RF radiator such as the monopole antenna of a mobile radio transmitter or a leaking microwave oven. (The term probe/antenna-length is discussed in 5.3.6.4.)

Inaccurate performance of a hazard probe located near either passive reradiators or active radiators is due to several factors including the following:

- (1) Reflections from a reradiating object that produces standing-waves (or interference patterns in the EM-fields) that extend a distance of several wavelengths from the scatterer. When a probe is not tightly coupled (via the reactive near-fields of the reradiator), the techniques of 5.3.2 can be used to minimize measurement errors. These techniques remove the effects of standing-waves through the use of spatial averaging. When the probe is tightly coupled to the reradiator, the data in 5.3.6.1 can be used;
- (2) A perturbing object (whether it is an active radiator or a passive reradiator) "loads" or distorts the measurement characteristics of the probe's antenna/detector combination [B106]. This occurs when the antenna is large compared with the wavelength of the RF energy being measured (5.3.6.1 addresses this situation);
- (3) An electrically large probe in the reactive near-field of an active radiator alters the fields being radiated by the source and spatially averages the nonuniform near fields being measured. This averaging occurs over the effective aperture area of the probe antenna, i.e., dipole length or loop diameter.

**5.3.6.1 The Effects on Measurement Accuracy of the Separation Distance Between Survey Probes and Nearby Passive Radiators.** An analysis can be performed to determine the degree of probe interaction (or coupling) with nearby objects, such as passive reradiators or scatterers including exposed personnel. When a probe is

tightly coupled to a reradiator, a probe-loading error has the effect of altering the impedance match between the probe's antenna and its detector. Although, this loading error is dependent on two electrical parameters of the antenna and detector equivalent circuits, the ratio of the antenna's "source impedance" to its "load (detector) impedance" is the critical one. The "source impedance" refers to the complex output impedance (at the frequency of interest) of the antenna: the "load impedance" is the complex impedance of the RF detector that is placed at the output terminals of the antenna. For any type of detector, e.g., diode, thermocouple, if the magnitude of the complex impedance of the load (detector) is "high" (about ten times larger than the source antenna impedance), the probe will be less susceptible to performance degradation due to a small separation distance between it and a large field perturbing object.

Analyses and specific calculations have been performed to quantify the probe-to-passivereradiator effects discussed above. Estimations have been made to quantify the errors caused when measurements of field strength are made with a hazard probe that is in close proximity to a large, passive reradiator. A thorough analysis has been carried out for the situation where a dipole-based hazard probe is used to measure a plane-wave field close to an infinite, conducting plane [B115]. The effect of proximity of the conducting plane to the electric field probe was studied analytically, and confirmed experimentally. The degradation of the measurement is quantified in terms of the true field-strength distribution (including the standing-wave) along a path normal to, and close to the large reflector. Two cases were studied for dipole-diode probes of total lengths equal to 0.2 and 0.4 wavelengths. (It should be noted that in [B115], the results are expressed in terms of the half-length of the dipole antenna, while the following will relate the results to the full "probe antenna length".

The performance of the antenna/detector combination was analyzed as a function of distance between the antenna and the reflecting plane and the results quantified the probe error with respect to the true electric field (including the large standing wave associated with the reflector). The errors associated with the measurement of the E-field were examined for very high- impedance (open-circuit) antenna-terminal loads (detectors), and for very low-impedance (short-circuit) loads. The measurement error of the dipole/detector combination was determined by analyzing the behavior of the equivalent circuit of the probe antenna and detector. The analysis performed by Smith using the assumptions above, yield a fractional E-field measurement error  $\Delta V$  that is defined as follows:

$$\Delta V = \frac{\left|V\right| - \left|V_0\right|}{\left|V_0\right|}$$

(Eq. 5.2)

where:

 $V_0$  = the voltage from the E-field detector in the absence of the scatterer

V = the measured voltage in the presence of the scatterer

While probes with a complex load (detector) impedance of at least ten times greater than the source (antenna) impedance are less susceptible to errors induced by a nearby fieldperturbing object, a low-capacitance, high-impedance diode detector across the antenna terminals will reduce the worst-case errors by a factor of about two times.

Results and conclusions of the analysis of measurement errors due to the interaction between a probe and a nearby passive scatterer are presented below. Selected data from [B115] are included in Table 5.1. Note that errors in equivalent power density or field-strength-squared in Table 5.1 have been calculated using the relationships shown in Eq 5.3. For the equivalent power density error, it was assumed that the RF hazard probe is square law. Here, the fractional error encountered when measuring the equivalent power density with and without the reflecting plane present can be calculated from Eq 5.2 and expressed as indicated below:

$$\Delta W = \frac{V^2 - V_0^2}{V_0^2}$$
(Eq 5.3)

where:

 $\Delta W$  = field-strength-squared error (or equivalent power density error), and V and V<sub>0</sub> are defined above.

The following trends are revealed in [B115]. At frequencies between 300 MHz and 3 GHz, and separation distances between 0.2 and 0.5 wavelengths, field-strength measurement errors were found to be less than 13% (28% for field-strength-squared or equivalent power density) for antennas that are less than 20 cm in length. Data are lacking for dipole antennas used below 30 MHz that are shorter than 200 cm. Therefore, practical hazard probe error analyses cannot be obtained from [B115] for frequencies at or below 30 MHz.

|           |               |                        | Measurem |        |                                      |
|-----------|---------------|------------------------|----------|--------|--------------------------------------|
| Frequency | Probe Length  | Separation<br>Distance | (FS)*    | (FS²)† | Separation<br>"Probe-Antenna Length" |
| (MHz)     | cm/Wavelength | cm/Wavelength          | %        | %      | cm                                   |
| 300       | 20/0.2        | 20/0.2                 | 10       | 21     | 1                                    |
| 3000      | 2/0.2         | 2/0.2                  | 10       | 21     | 1                                    |
| 3000      | 4/0.4         | 5/0.5                  | 13       | 28     | 1.25                                 |

# Table 5.1 Measurement Errors for Field Strength Measurements Made in Close Proximity to an Electrically Large, Passive Reradiator (from [B113])

Notes:

\* (FS) = field Strength

† (FS<sup>2</sup>) = Field Strength Squared or Equivalent Power Density

Overall findings from the analysis presented above indicate that worst-case measurement errors, over the frequency range of 300 to 3000 MHz, are no greater than 10% (21% for field strength squared), under the following worst-case conditions:

(1) The detector (load) impedance is low, with respect to the antenna's source impedance (i.e., the detector draws a relatively high RF current from the receiving antenna). In contrast, a low-capacitance, high-impedance diode detector will reduce the probe loading errors by about a factor of two times.

- (2) The perturbing object (passive reradiator) may have any scattering cross section, i.e., its size can be much greater than several wavelengths at the frequency being measured. Small scatterers will introduce lower measurement errors.
- (3) The dipole electrical length is less than or equal to 0.4 wavelengths tip-to-tip.
- (4) The distance between the probe and the perturbing object is greater than 20 cm at 300 MHz (0.2 wavelengths), and greater than 2 cm (0.2 wavelengths) at 3000 MHz.

Tell [E] described a test designed to evaluate the possibility that a measurement probe might, when placed very close to a strong localized RF source, capacitively couple to the source, thereby altering the source characteristics and, thus, the fields being measured. In these tests, the electric field near one end of a re-radiating resonant rod was measured with one instrument while other probes were brought near the opposite end of the resonant rod. Any interaction of the probe with the re-radiating rod could conceivably be manifested as changed readings of the field at the other end of the resonant rod. Of the probes used in these tests, only the magnetic field probe and one electric field probe produced any observable change in the electric field being monitored at the other end, and even in this case the effect was minimal. Data related to the magnetic field probe that produced the greatest interaction are summarized in the table.

| Distance from Probe Surface to Rod (cm) |    | Change in Square of Electric Field Strength |  |  |
|---|----|---|--|--|
|   |    | (%)   |  |  |
|   | 20 | 0   |  |  |
|   | 10 | 0   |  |  |
|   | 5  | 0   |  |  |
|   | 4  | 1.2   |  |  |
|   | 3  | 1.2   |  |  |
|   | 2  | 3.5   |  |  |
|   | 1  | 7.1   |  |  |
|   | 0  | 17.6  |  |  |

Table 5.2 Effect of Probe Interaction with Resonant Dipole Antenna at 144 MHz Caused by the Presence of a 10 cm dia. Isotropic Magnetic-Field Probe (from Tell [E])

These data show that only when the probe surface was placed in contact with the resonant rod was there sufficient coupling between the probe and the rod to significantly change the electric field strength at the other end of the rod. In the case of the magnetic field probe tested, it was observed that the probe has loop windings on its spherical surface. Hence, the windings were essentially placed in direct contact with the source and, consequentially, there was no distance isolation afforded as there was with the other probes tested evaluated.

In the evaluation of a broadband isotropic electric field with a 40 GHz response, direct contact between the probe housing and one end of the re-radiating rod resulted in a maximum change of 2.2 percent in the field observed at the opposite end of the rod.

Other evaluations related to deformation of the pattern of the probe response due to near field coupling phenomena were reported [Tell, E]. These tests revealed only minor interactions with the re-radiating dipole rod. These test data, when taken collectively, appear to support the remarkable conclusion that measurement probe interaction with RF hot spots does not appear to be a significant issue, at least for the conditions evaluated in this project. For applications in the VHF broadcast band, the test results appear to suggest that no significant interactions occur which would lead to erroneous field strength readings for most commonly used isotropic broadband probes. This finding seems to hold even for measurements taken with the spherical shell of the probe in direct contact with the source. However, in the case of one magnetic field probe, a 2 cm spacing between the probe and the source appeared necessary to minimize any interaction with the source.

A set of measurement errors with similar values can be expected for an electrically-small loop antenna (H-field probe) due to the impedance "loading" effect from a nearby conducting plane. Here the loop diameter would represent the probe antenna length.

5.3.6.2 Establishment of the Minimum Separation Distances Between a Survey Probe and an Active Radiator. The accuracy of measured data can be affected when using a near-field probe to map large spatial gradients very close to radiating elements of an RF emitter (an antenna or a leakage source). These gradients may cause the amplitude of the field to vary significantly over the volume of space that is occupied by the probe antennas, thereby introducing measurement error due to spatial averaging. As the separation distance between the probe and the radiator increases, the field throughout the entire volume occupied by the probe's antennas becomes more uniform. Based on the fact that the largest field gradients are associated with the reactive or quasi-static Efield near a radiating device, it is possible to predict the minimum distance between a near-field probe and an active radiator that will avoid significant measurement errors. For example, the fields near an electrically small dipole decrease inversely with the cube of the separation distance d between the radiator and the point of measurement (see Eq D9). An analysis of the magnitude of the E-field components indicates that the reactive field component  $(1/a^3)$  dominates at distances of less than 0.15 wavelengths from the source. Further examination of the data at distances less than 0.15 wavelength provides data that defines the range of separation distance over which  $E^2$  varies less than a factor of  $\pm 3 \text{ dB}$ . Using this as a criterion and Eq D9, a worst-case simplified analysis using the single  $1/wa^{\beta}$  term in the expression for  $E_q$  yields a minimum separation distance of 5 receiving "probe-antenna lengths" (see Table 5.1), e.g., approximately 20 cm or more for typical commercial near-field instruments where the size of the sensor elements is small compared with 20 cm. Here, an antenna length is equal to the tip-to-tip dimension of a simple dipole or the diameter of a loop. A worst-case estimate of the "probe-antenna" length" can be readily obtained following the procedure described in 5.3.6.4.

It should be noted that the above worst-case analysis yields errors that are significantly larger than the actual errors encountered for many types of radiators. Specifically, many radiation sources, such as a crack in a microwave oven door, do not produce "radial" components with  $1/d^3$  decay with separation distance, even in their reactive near-field region.

**5.3.6.3 Conclusions on the Minimum Spacing Between an RF Probe and a Radiating Source or a Passive Reradiator.** The analyses presented in 5.3.6.1 and 5.3.6.2, and the information contained in 4.5.3 may be applied to a near-field

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measurement problem. This problem involves the worst-case uncertainties that occur when measurements are made at a minimum distance from an active radiator or passive reradiator. Table 5.3 provides information on this subject, in terms of ranges for which the probe sizes and frequencies are valid for the particular analysis technique. Also, the minimum separation distance is expressed with respect to wavelength and in terms of the probe-antenna length (the maximum dimension of a single antenna or the array of orthogonal antennas of the hazard probe).

From the data in Table 5.3, estimates can be made regarding the errors that will result if the minimum probe to source or scattering object separation distances are not exceeded. The largest error is due to the gradient of the radial field over the probe antenna aligned with it. This error is an extreme worst case, and many situations will not produce this large of an error, even at smaller separation distances. In general, for a typical hazard probe with an array of 5 or 10 cm dipoles or loops, a separation of 3 to 5 probe lengths will ensure a maximum error of 3 dB at frequencies lower than 500 MHz, while the maximum error at higher frequencies may be lower due to the fact that the gradients in the radial components are less steep at distances greater than 0.15 wavelengths. Therefore, for most situations, when appropriate near-field survey instruments are used, a minimum separation distance of 20 cm is reasonable.

| Analysis Method           | Antenna Length | Wavelength | Frequency Range | Error |
|---------------------------|----------------|------------|-----------------|-------|
|                           |                |            | MHz             | dB    |
| TEM Cell                  | 1              | <0.25      | <500            | 1     |
| Passive Radiator          | 1 - 1.25       | 0.2 - 0.5  | 300 - 5000      | 1     |
| Active Radiator $(1/a^3)$ | 5              | <0.15      | <<470           | 3     |

# Table 5.3 Worst-Case Errors Associated with the Minimum Probe to RF Source Separation Distance

From the data in Table 5.3, estimates can be made regarding the errors that will result if the minimum probe to source or scattering object separation distances are not exceeded. The largest error is due to the gradient of the radial field over the probe antenna aligned with it. This error is an extreme worst case, and many situations will not produce this large of an error, even at smaller separation distances. In general, for a typical hazard probe with an array of 5 or 10 cm dipoles or loops, a separation of 3 to 5 probe lengths will ensure a maximum error of 3 dB at frequencies lower than 500 MHz, while the maximum error at higher frequencies may be lower due to the fact that the gradients in the radial components are less steep at distances greater than 0.15 wavelengths. Therefore, for most situations, when appropriate near-field survey instruments are used, a minimum separation distance of 20 cm is reasonable.

**5.3.6.4 Estimation of the Physical Size of Hazard Probe Antennas.** Although the length of the dipoles or diameter of the loops of most survey instruments are not obvious, the maximum size of a probe's sensing antennas can, nevertheless, be estimated. Since a survey probe has antennas that are physically smaller than the dielectric radome or other physical object surrounding them, the size of the radome can be used to approximate the maximum size of the antennas within. Thus, the diameter of the spherical radome can be used as a worst-case estimate of the size of the enclosed

antennas. This, in turn, can aid in determining the minimum separation distance that should be used between the survey probe and any surrounding object.

Several popular broadband RF hazard probes with thermocouple sensors have antenna elements that are almost as large as the diameter of the radome that encloses them. Some active antenna-type hazard probes that have been designed to work below 200 MHz, utilize active electronics to match the impedances of their electrically short-monopole antennas to the RF detection circuitry. This type of instrument is equivalent to an asymmetric dipole, with one element (monopole) emerging from the center of a metallic box. The box houses the active electronic impedance-matching circuitry, as well as serving as the "ground plane" or second element of the dipole. Although the overall dipole length is greater than the sum of the length of the monopole receiving antenna and the height of the electronics-containing box, the combined length should be used to define the equivalent dipole length.

# 5.4 Induced Body Current and Contact Current Measurements

### 5.4.1 Induced Body Current.

Body currents are generally taken to be the induced current associated with exposure of the body to RF fields, but without any direct contact with objects other than the ground upon which the subject may be standing. Several common techniques are used for measuring body currents including clamp-on "loop" type current transformers for measuring current through the ankle or calf, and parallel plate "stand-on meters" for measuring currents that flow to ground through the feet.

Commercially available lightweight clamp-on current transformer instruments have been developed that may be worn (with care) about the subject's lower leg [Pasour, 1995]. A readout module, either mounted directly on the transformer or connected through an optical link for remote reading, provides a display of the current flowing through the aperture (primary circuit) of the transformer. Current sensing in these units may be accomplished using either narrowband techniques, e.g., spectrum analyzers or tuned receivers (which offer the advantage of being able to determine the frequency distribution of the induced current in multi-source environments [B122]), or broadband techniques, e.g., diode detection or thermal conversion. If diode sensing is used, care must be taken to ensure that the diodes are operated in their square-law response region to achieve a true RMS current indication. Sometimes additional circuitry is provided to extract the square root for linear indications of current. This mode of operation produces true RMS indications in the presence of multiple frequencies and/or amplitude-modulated waveforms.

True RMS current detection is usually achieved with thermal sensors which respond accurately to the simultaneous flow of currents at different frequencies and to low duty cycle pulsed currents. In most current transformers there is a tradeoff between the size of the opening (aperture) and the reliable high frequency response. Generally, as the aperture size increases, the high frequency response decreases. Therefore, current transformer-type instruments must be used with care since operation at frequencies above their specified high frequency limit can result in erroneous measurements. Aircore transformers, as opposed to the typical ferrite-core transformer, have been used to help extend the upper frequency response of these instruments. The lower weight of the

air-core sensors makes them useful for longer term measurements. Air-core instruments, however, tend to be significantly less sensitive than ferrite-core devices.

An alternative to the clamp-on device is the parallel plate stand-on meter. In this instrument, the body current flows through the foot (feet) to a conductive top plate, through some form of current sensor to the bottom plate and then to ground. The current flowing between the top and bottom plates may be determined by measuring the RF voltage drop across an low impedance resistor and using Ohm's law to relate the measured voltage to the corresponding current. Alternately, a small aperture RF current transformer may be placed around a conductor placed between the two plates and, with appropriate circuitry (which includes narrowband instruments), the current can be determined. Chen and Gandhi [B29] have described a parallel plate type of system in the form of bi-layered printed circuit board, in the shape of human feet. Commercial instruments with a flat frequency response between 3 kHz and 100 MHz are available as are shoe-insertable sensors. Another alternative is a direct reading RF thermocouple ammeter placed in series between the plates. This method is entirely passive since power supplies or other associated circuitry are not required. Two factors that may reduce the effectiveness of the thermocouple ammeter are its physical size and sensitivity to burnout.

There are several issues which should be considered when selecting an instrument for measuring induced current. First, stand-on meters are subject to the influence of electric-field induced displacement currents from fields terminating on the top plate. This means that such meters may produce some indication of current when subjected to strong electric fields, even without a subject standing on the meter. When a subject stands on the meter, however, the electric fields are generally shielded from significant interaction with the top plate, as they preferentially terminate on the surface of the subject. Therefore, when using parallel plate type meters, the displacement current indication, without the subject in place, should be ignored. That is, the induced current reading with the subject in place should be taken as the most accurate indication of the induced body current (the initial background displacement current should not be subtracted).

Another observation is that the sum of both ankle currents measured with clamp-on type meters tends to be slightly greater than the corresponding value indicated with stand-on type meters. This observation can be explained by the fact that fringing electric fields about the periphery of the top plate of the stand-on meter induce charge on the top plate and, consequently, displacement current that does not flow through the current sensing elements of the meter. This phenomenon, which tends to reduce the indicated currents for this type of meter, is a function the RF frequency and the geometry of the meter. While the current passing through the ankle, just above the foot, may be slightly greater than that passing through the bottom surface of the foot due to displacement current leakage off of the foot, this is not likely to be significant [USAF, 1996]. The currents measured with a clamp-on current meter should be a more accurate measure of the current actually flowing through the ankle. While the induced-current limits specified in the IEEE C95.1-1991 [1] standard are based on limiting the current flow through this region, i.e., the smallest cross-section of the leg, to limit the local SAR, the actual limits are specified as foot current, not ankle current.

An additional issue pertains to the problem of accurately relating the induced current indicated by a stand-on meter to the actual current passing through the foot when the subject is standing on a variety of ground surfaces. For example, differing ground

conductivity conditions and ground surface textures such as grass, gravel, concrete, steel decking, wood floors, etc., can result in different indicated body currents, for the same electric field strength, when measured with a stand-on meter [Lubinas and Joyner, 1992]. This is because of the differing degree of electrical contact between the bottom metal plate and the actual ground surface, i.e., the flat surface of the bottom plate does not necessarily make uniform contact with many surfaces on which it is placed. Also, the degree of contact can vary according to the weight of the subject. The inherent variability introduced by the stand-on meter suggests that a direct measurement of ankle current, using a clamp-on current meter, will be subject to less variability due to contact conditions and will yield a more meaningful measure of the current flowing in the ankle under realistic conditions of shoe contact with differing ground surfaces.

When discussing the variability in measurement results, the use of a Human Equivalent Antenna (see 4.4.9) must be considered. These devices eliminate the variability due to the difference in human physique, stance and footwear. They also allow current measurements to be made without requiring a person to be exposed to potentially hazardous currents and fields.

# 5.4.2 Contact Current.

Measurement of contact currents may be accomplished with the clamp-on type current transformer discussed in 5.4.1. Inserting a current measuring device between the hand and the object to be tested is an alternate technique for measuring contact currents. This may take the form of a metallic probe, one end of which is held by the individual and the other end of which is touched to the object under test. The contact surface area (and resulting surface impedance) between the subject and the object under test is an unknown variable when using this technique.

As a means of protecting the subject from excessive current flow, an impedance network may be used to simulate the body impedance during initial measurements. [B80] and [B84] discuss electrical models that are intended to represent equivalent body impedances in connection with common objects. The use of body impedance equivalent circuits to prevent excessive contact current from flowing between the subject and object under test must be approached with caution since the electric fields present in some objects can couple directly to the subject's hand, thereby producing RF currents in the hand, arm, and body of the subject while only a portion of the contact current flows through the probe being held by its insulated handle.

# 5.5 Internal Field Measurement Procedures (SAR)

# 5.5.1 Temperature Probes for SAR Measurement.

Certain temperature probes can be used successfully to make SAR measurements. The minimum requirements are that the temperature sensor and associated leads should be nonperturbing to the EM-fields, and the SAR should be large enough to produce a measurable temperature rise during a period of less than about 20 s. The first requirement is usually satisfied by using highly resistive materials or fiber optics, instead of metal components, for the temperature-sensing element leads [B25, B62]. The second requirement entails the measurement of SARs no lower than a few W/kg. This lower limit exists because the resolution of most temperature probes is typically 0.01 to 0.1° C, and the longest practical duration of irradiation that allows reasonably accurate

SAR measurement is typically 5 to 30 seconds. Irradiation of an inert, lossy dielectric object for longer durations causes local "hot spots" to lose their thermal energy to the surrounding area via conduction, diffusion and convection. In living biological systems, active thermo-regulation also degrades accuracy. It is acceptable to use an ordinary metal-wire temperature probe for RF and microwave dosimetry when the probe is not in place during irradiation, but is put in place immediately before and after. Such a method has limited application, but has been used successfully in small laboratory exposure systems, e.g., waveguide exposure systems. The use of metallic probes during irradiation is not acceptable. Even when the probe's metallic leads appear to be oriented orthogonal to the incident electric field vector, the depolarization of fields within finite sized dielectric objects induces errors. Limitations of many nonperturbing temperature probes include fragility and high cost. Without the strength of wire leads and rugged insulation, nonperturbing devices are much less durable than ordinary thermistor and many thermocouple probes. Also, high-resistance leads in temperature probes can be easily destroyed by local heating when the leads are in air and are exposed to high Efields and high E-field gradients along the length of the lead. The high cost of commercially available devices is due to the relatively small market for SAR-measuring apparatus and the fact that the parameter SAR itself has only been in widespread use for less than sixteen years.

The use of a nonperturbing temperature probe in this application is very simple in principle, but many pitfalls exist to complicate the practical situation if precise data are required. The objective is to measure the time rate of irradiation-induced temperature rise  $(\Delta T/\Delta t)$  at a specific location in tissue or phantom material. The SAR, which is proportional to  $(\Delta T/\Delta t)$ , can then be determined. When the temperature does not rise linearly during constant irradiation of the tissue or tissue-equivalent material under test, other factors such as heat loss by thermal conduction, convection etc., should be considered. Therefore, the simplest procedure is to produce a brief and relatively small  $\Delta T$  (no more than 1 - 2° C in 20 s) at the location of the temperature probe.

To obtain SAR, the probe indication or analog output is usually plotted or automatically recorded before and during irradiation, and the irradiation-induced rate of temperature rise is either graphically determined or found through the use of a slope-determining algorithm. A graphical record of  $\Delta T/\Delta t$ , as well as a simultaneously recorded plot of the RF power (to precisely determine when the power is turned off and on with respect to the temperature rise) is helpful in verifying the linearity of the slope. SAR is calculated from the initial linear slope of  $\Delta T/\Delta t$  from

$$SAR = \frac{\Delta T \cdot c}{\Delta t} \tag{Eq 5.4}$$

where *c* is the tissue (or phantom material) specific heat capacity, in  $J/kg^{\circ}$  C. Typical values for heat capacity are shown in Table 5.3

|   | Specific Heat<br>(kJ/kg/°C) | Mass density<br>(kg/m <sup>3</sup> x10 <sup>3</sup> ) |
|---|-----------------------------|---|
| Muscle Simulant*                            | 3.7                         | 1.0   |
| Muscle Simulant <sup>†</sup>                | 3.6                         | 1.1   |
| Brain Simulant*                             | 3.4                         | 0.98  |
| Adipose Tissue (fat) Simulant <sup>†#</sup> | 1.1                         | 1.4   |
| Muscle in-vitro                             | 3.5                         | 1.1   |
| Brain <i>in vitro</i>                       | 3.5                         | 1.1   |
| Adipose Tissue in vitro                     | 1.2-1.6                     | 1.05  |
| Bone  | 1.25-3.0                    | 1.25-1.8  |

| - | Table 5.4 | <b>Specific Heat Capacity and Mass</b> | Density              | of Tissue-Equivalent |
|---|-----------|--|----------------------|----------------------|
| ( | (Phantom) | Materials and Actual Biological T      | lissues <sup>§</sup> |                      |

Notes:

\* Data for tissue formulated for use at 2450 MHz

† Data for tissue formulated for use at 27 MHz

# Simulated fat material has dielectric properties that are almost identical to living bone

§ Data for this table was taken from [B87]

There are several sources of error when using temperature probes for SAR measurements. First, it is difficult to obtain repeatable results at a location of a large spatial SAR gradient. Small changes in probe location can cause large SAR changes in such situations, and the linear portion of the slope of  $\Delta T / \Delta t$  is rather brief (with respect to the initiation of irradiation). This is due to the presence of high-thermal gradients and the resulting thermodynamic effects that cause measurement degradation. For these reasons, it is desirable to determine locations in the subject where maximum values of SAR exist. Data should be taken at and on either side of the maxima. Second, regions of relatively high SAR (> 20 W/kg) should be scrutinized closely to ensure that the proper value is actually obtained, because such regions often exhibit high-thermal conduction losses, as well as high SAR. In high SAR regions, it is appropriate to halve the irradiation time and ensure that  $\Delta T$  is also reduced by one-half; if it is not, thermodynamic effects, such as those associated with thermal conduction, are present. It is also important to note that dielectric constants will generally change with temperature, which of course will alter the measured SAR. This would be relevant, say, if the measurements were being made inside a dead animal that was allowed to equilibrate to room temperature to obtain  $\Delta T / \Delta t = 0$  in order to null out the spatial thermal gradients.

Significant errors can occur when SAR is measured, using temperature probes, at a single point in an object with one or more "hot spots" near, but not coincident with the probe tip. The temperature, as measured by the probe, will behave as follows: The temperature may not rise immediately after the RF irradiation of the test object. After several seconds, the temperature, as monitored by the probe, begins to rise more quickly as heat is conducted from a nearby hot spot. When the irradiation ceases, the temperature continues to rise as heat is conducted away from the hot spots to cooler regions where the probe is located. Figure 5.1 illustrates this phenomenon. The rate of temperature rise seen by the probe (apparent slope) could be falsely interpreted as the

local SAR at the location of the probe tip. In fact, a very low value of SAR (as determined by the initial slope) may exist at that point.

# Fig 5.1 Typical Thermal Dosimetry Data: Temperature vs. Time—Before, After and During Irradiation

A third source of error arises from the dielectric heating of certain high-resistance leads of the temperature probe as they exit the object being irradiated. Some irradiation configurations cause significant heating of the leads at this point, resulting in an artificially exaggerated surface SAR. It is, therefore, better to obtain the SAR that exists at the surface of an object with the sensor tip of an embedded probe located at the measurement point and the leads exiting the object at a location removed from the sensor. There are obviously other sources of error in measurements of SAR using these techniques, but they can be minimized by a thorough knowledge of the capabilities and limitations of the temperature probe being used and by careful application of scientific methods [B25, B62].

An additional consideration that warrants attention is the multiple measurements of SAR in a given preparation or model. In large models, for example, SAR is needed in many locations, but often there are not enough nonperturbing probes available to obtain all the data simultaneously during a single irradiation. If one or only a few probes are used to map SAR in a large volume, ideally the pre-irradiation value of  $\Delta T/\Delta t$  would be zero (no "ambient" drift of the phantom's temperature) for subsequent irradiation after the first. However, after the first exposure, it is common to observe an exponential post-irradiation cooling curve that lasts for many minutes or hours if high values of  $\Delta T$  had occurred during prior RF irradiation (more than several degrees Celsius). For reasons of practicality and economy, it is important to obtain as much dosimetric data as possible

(Eq. 5.5)

during each day of laboratory experimentation. Therefore, a compromise should be made that provides good SAR data and also conserves time. A useful rule of thumb in deciding when to start another irradiation is to wait until the slope of the cooling curve is relatively constant, (about 5% of the prior RF-induced rate of temperature rise over the period to be used for the next irradiation), and the decrease in temperature before irradiation is relatively small compared with the expected irradiation-induced  $\Delta T/\Delta t$ . Repeated experiments using consistent techniques are essential for obtaining accurate results in SAR studies utilizing temperature probes. Finally, after several RF irradiations of the same object, its temperature may have increased above acceptable limits, and the phantom or biological material may degrade.

The majority of SAR measurements are made using temperature probes. However, many researchers are not fully aware of the many factors that degrade accuracy of these measurements. For example, thermodynamic factors will always limit the accuracy and precision of SAR measurements as will any uncertainty in the value of the specific heat capacity of the actual or simulated tissue being evaluated. The specific heat capacity is often mistakenly cited as that of water (about 15% higher than that of most high-water content tissue), which, even under optimal usage conditions, leads to uncertainties of at least  $\pm 1$  to 2 dB in the local SAR distribution in an object when measured by sampling the tissue volume with a temperature probe.

### 5.5.2. SAR Measurement with Miniature Electric-Field Probes.

Miniature isotropic, implantable E-field probes with high impedance feed lines, which have been commercially available for a number of years [Cheung, et al., 1975], have been used to measure SAR distributions in phantom models and in living, anesthetized animals [C2, C12]. [See also 4.6.1.] These probes have much higher sensitivity than thermal probes and are especially suitable for measuring E-fields within simulated or actual biological tissues of moderate to high water content, i.e., brain and muscle. While it is possible to measure SARs of the order of 1 W/kg using sensitive and precise thermal measurements ( $\Delta T/\Delta t \approx 0.1^{\circ}$  C/30 s), it is well within the domain of E-field probes to measure SARs as low as 10 mW/kg [Balzano, 1995]. SAR can be calculated using Eq. 5.5 and the data in Tables 5.4 and 5.5 that show typical values for simulated and actual tissues.

$$SAR = \frac{1}{2\rho} \omega \varepsilon_0 \varepsilon'' E_{int}^2 \quad W/kg$$
$$= \frac{\sigma}{2\rho} E_{int}^2$$

where:

 $r = mass density (kg/m^3)$ 

 $e_0$  = permittivity of free space (F/m)

*e*" = imaginary part of the complex relative permittivity

 $w = radian frequency (= 2\pi f)$ 

s =conductivity (S/m)

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## Table 5.5 Electrical Conductivity (S/m) of Simulated Tissues at RF Frequencies

|              | Frequency (MHz) |                    |                         |       |  |
|--------------|-----------------|--------------------|-------------------------|-------|--|
|              | 10              | 100 1000 2450      |                         |       |  |
| Muscle       | 0.7*            | 0.9*               | 1.3*                    | 2.2*  |  |
| Brain        | —               | $0.47^{\dagger}$   | 0.75*, 1.2 <sup>†</sup> | 1.2*  |  |
| Fat and Bone | _               | 0.008 <sup>†</sup> | 0.07*, 0.12*            | 0.18* |  |

#### Formulations:

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\* Polyethylene powder and TX 150 gelling agent used for muscle and brain simulants [C7].

+ HEC gelling agent and no polyethylene powder used for muscle and brain simulants [C9].

Note: Fat and bone simulants are solid versions for both formulations.

|                      | Frequency (MHz) |       |       |      |        |
|----------------------|-----------------|-------|-------|------|--------|
| _                    | 10              | 100   | 1000  | 3000 | 10,000 |
| Muscle               | 0.67            | 0.8   | 1.3   | 2.7  | 12     |
| Brain (Grey Matter)  | 0.29            | 0.66  | 1.1   | 2.6  | 9.3    |
| Brain (White Matter) | 0.17            | 0.36  | 0.76  | 1.6  | 5.7    |
| Fat                  | 0.027           | 0.033 | 0.054 | 0.1  | 0.58   |
| Bone (Cancellous)    | 0.071           | 0.15  | 0.34  | 0.74 | 3.7    |
| Bone(Cortical)       | 0.43            | 0.07  | 0.18  | 0.5  | 22     |

### Table 5.6 Electrical Conductivity (S/m) of Biological Tissues at RF frequencies

Notes:

- (1) Values are from Gabriel [1996]
- (2) Materials other than muscle have a large variation in their value depending on the author and the actual type of tissue.

Measurements may be made at individual points within a simulated or actual biological subject. Since there are large spatial gradients and standing waves in most biological objects that are exposed in the near or far field, many data points should be taken to accurately characterize the SAR distribution. At each site, the sum of the outputs of three mutually orthogonal dipoles and the location of the probe must be recorded. The accurate positioning of the sensing probe is essential to perform repeatable measurements; repeatability can usually only be achieved by means of a robot – not by hand. To reduce the magnitude of the effort of taking data throughout a volume of tissue, data can be taken while the probe is scanned through the volume. Since the E-field

probe has a response time of the order of a few milliseconds, continuous line scans of the internal E-fields may be dynamically recorded by means of a robot that generates position data as the probe is moved along a path. Extensive data in an object can, thus, be plotted in a relatively short period of time, and the possibility of missing a local peak is reduced [Bassen, et al., 1977].

There are several inherent sources of error associated with the use of implantable E-field probes for SAR measurements. Regardless of the quality of the specific probe used, calibration (in terms of absolute field strength in high water-content biological tissue or tissue simulant) is difficult (see below). Large gradients in the internal E-fields and imprecise knowledge of the conductivity and mass density of the biological tissue or tissue simulant add additional degrees of uncertainty. Although commercially available implantable probes are available, custom-designed probes are frequently developed, evaluated and calibrated by the developer and/or user. Therefore, instrument performance limitations, as well as errors introduced during the measurement procedure, should be understood by the user and steps taken to minimize errors introduced by these factors. Detailed discussions on the evaluation, calibration and use of implantable probes have been published [Bassen, et al., 1977, Stuchly, 1987]. In summary, SAR measurement uncertainties of at least  $\pm 2$  dB are to be expected, even under optimum measurement conditions.

The high sensitivity of the E-field probes make them ideally suited for the measurement of SARs associated with low-power (of the order of one watt or less) localized sources such as hand-held radio transceivers, e.g., cellular and personal communications equipment. The low power output and small size of the antenna make thermal measurements extremely difficult [Balzano, et al., 1978a; Balzano, et al., 1979; Balzano, et al., 1978]. Increasing the power of such sources by a factor of ten times so that thermal techniques could be used would entail a substantial modification of the device to the point of being non-representative of an actual transceiver. Since the exposure of concern from such low power sources localized sources is within about 5 cm of the antenna, accurate positioning of the sensing probe is crucial for performing repeatable measurements. The probe positioning should be performed by machine, e.g., a robot, rather than by hand.Simulated human tissue used for such measurements may vary from case to case depending on the specific device being evaluated. For cellular phones, for example, the head and upper part of the torso would usually be sufficient [Balzano, 1995], while a

150 MHz two-way radio strapped to the belt with the antenna operated by remote switching would require a full-size human phantom [Fujimoto, 1994].

The conductivity of the tissue simulant must be correct for the frequency being tested. The mixing of such materials and the measurement of their corresponding electrical properties present substantial challenges to achieve accuracy and repeatability. It is not possible to use a single formulation over a wide frequency range, e.g., more than an octave, without running into relatively large departures ( $\pm 5$  %) from published conductivities for biological tissues. To obtain repeatable results ( $\pm 3$  %) it is advisable to restrict the frequency band and to purchase the primary materials from the same supplier. Well documented mixing procedures should also be adhered to, e.g., accurate weights of the components, temperature of the liquids during mixing, length of the mixing time, RPM of the stirring device. Measurement of the dielectric properties of tissue simulants is also difficult to perform accurately. Acceptable results can be obtained

using open coaxial line methods [Stuchly, et al., 1987; Athey, 1982], but the slotted coaxial line method, if not more accurate, provides more repeatable results for liquid simulants. Moreover, the slotted coaxial line method provides a means to examine the attenuation of the RF wave as it progresses along the line so that the overall precision of the measurement can be assessed more accurately than that for single surface point measurement of the open coaxial line technique.

Accurate SAR measurements can only be made with probes that have been carefully calibrated in the simulant being used to represent biological tissue (see 4.7.1.). The calibration process, which is tedious and prone to error, requires simultaneous or subsequent measurement of the magnitude of the E-field and the temperature rise at the same location in a canonical model such as a flat or a spherical phantom of appropriate tissue-equivalent material. The calibration in a flat model is usually is performed using a relatively high power source coupled to a resonant dipole placed at a specified distance from the phantom [Kuster and Balzano, 1992]; calibration in a spherical model is usually performed under plane wave irradiation conditions.

As pointed out above, the experimental error associated with SAR measurements can be substantial (±2 dB) because of the multi-step nature of the process. The following factors contribute to the overall experimental accuracy that can be expected to be realized: accuracy of the electrical characteristics of the tissue simulant – ±3 % (if the measurements are limited to a narrow band of frequencies); accuracy of the temperature rise measurements during probe and errors associated with calibration – ±3 %; accuracy of the RF power measurements – ±5 %; positioning errors the non-isotropic response of the probe – ±6 %. This leads to a total relative error of approximately ±17 % or about ±0.6 dB.

Even for a narrow frequency band, the achievement of a total relative error of only  $\pm 17$  % requires specialized equipment for measuring the dielectric properties of the tissue simulant, accurately calibrated RF power meters, temperature probes, and well-trained personnel to measure rises in temperature of the order of  $0.10^{\circ}$ C with a  $0.03^{\circ}$ C measurement error. The procedure is time-consuming and the calibration of a single E-filed probe at a single frequency in two different media, e.g., brain and muscle tissue simulant, can take as long as two working days.

**5.5.2.1 Automated SAR Scanners.** The measurement of 3-dimensional SAR distributions within a phantom involves measurements at hundreds of points. At higher frequencies, especially with near field exposures from small localized sources which produce rapid spatial variations in the SAR distribution, the locations of the measurement points with respect to the phantom must be determined precisely. High precision is also necessary to accurately measure the spatial peak SAR. Automated scanning systems allow such measurements to be performed routinely. In order to move a small probe along unrestricted continuous paths, such systems are generally restricted to shell phantoms filled with liquids that simulate human tissue.

Although automated scanners based on temperature probes are possible, the maximum measurement speed would be unacceptably slow. Because of this and the low sensitivity of temperature probes, scanning systems that have been implemented are based on miniature E-field probes. Such systems range from one-dimensional

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positioners [Cleveland and Athey, 1989] to 3-axis scanners [Stuchly, et al., 1983] and, most recently, 6-axis robots [Balzano, et al., 1995].

A system designed for testing compliance of hand-held radio transceivers, e.g., cellular phones, with the spatial-peak SAR safety criteria, is described in [Schmid, et al., 1996]. It incorporates a high precision robot (working range greater than 0.9 m and a position repeatability better than  $\pm$  0.02 mm), isotropic E-field probes with diode-loaded dipole sensors, an optical proximity sensor for automated positioning of the probe with respect to the phantom surface (within  $\pm$  0.2 mm) and sophisticated software for data processing and measurement control. The useable frequency range extends from 10 MHz to at least 3 GHz, the sensitivity is reported to be better than 1 mW/kg and the dynamic range extends to 100 W/kg. Complex measurements, such as the spatial peak SAR value when starting with an unknown field distribution in the body can be completed within 15 minutes. Three dimensional magnetic field probes are also available for this commercially available system.

# 5.5.3 Calorimetric Determination of the Whole-Body Average SAR.

Average SAR may be measured using calorimetric methods. In the past, such methods have been used predominantly with small animals or animal models [B38, B101; Allan and Hurt, 1979; Blackman and Black, 1977]; recently, however, calorimetric twin-well methods have been successfully used to measure SAR in a full-size human model [B102]. The heart of the measurement system is the calorimeter device itself, and gradient-layer devices are commonly used. Gradient-layer calorimeters have a convenient voltage output signal that is proportional to the rate of heat energy flowing out of the device (positive voltage) or the rate of heat energy flowing inward (negative voltage). The signal generally has very low noise, and the sensitivity of a typical device is about 1.3 J/(mV-s).

In a laboratory setting, calorimetric SAR measurement begins with the thermal equilibration of the test object, usually a realistic animal model or a scaled human model. It is assumed that the laboratory temperature is ostensibly constant and is the same as that of the thermally stabilized test object and calorimeter. The test object is then irradiated for a measured period of time after which it is immediately placed inside the calorimeter. The calorimeter output voltage is then periodically monitored until all of the irradiation-induced heat energy has flowed from the object, and it is again at room temperature. This process may take hours or days depending on the size and mass of the object. By this time, the calorimeter voltage is zero, and the area under the curve described by the time-course of the calorimeter voltage is proportional to the energy deposition by the irradiation time in seconds yields the rate of energy deposition (power) in watts; average SAR is obtained by dividing the resulting power by the mass (kg) of the test object.

If two matched calorimeters are used in conjunction with two identical test objects, these procedures can be used in the absence of strict temperature control, such as in an outdoor environment. However, twice the physical effort is required for outdoor SAR measurements, and all of the apparatus should be given some sort of protection from the effects of direct sunlight, rain, etc.

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Although the calorimetric determination of the energy stored in the model once inside the calorimeter is quite accurate per se, the overall system accuracy in terms of SAR is limited by how closely the test object models the actual object and by the amount of irradiation-induced heat that escapes from the object without being measured. This method, moreover, requires sufficient time for the thermal equilibration processes and requires sufficient energy deposition in the test object to produce a calorimeter output signal that is enough above baseline to be measurable. However, the Dewar-flask method of calorimetry is a relatively simple, straight-forward way of determining the whole-body average SAR of small-bodied animals [Padilla and Bixby, 1986]. The calorimetric technique of determining a whole-body average temperature requires that the cadaver be immersed in a Dewar-flask containing a medium, such as water, at a known temperature; then the temperature of the cadaver/medium mixture.

# 5.6 Hazard Assessment; Estimation of Internal SAR from External-Exposure Field Measurement Data

# 5.6.1 General.

Contemporary MPEs are based on whole-body-averaged SAR thresholds for biological effects in RF-irradiated animals. The extrapolation of plane-wave exposures to wholebody-averaged SAR is supported by the many dosimetric studies, both experimental and mathematical, reported in the literature [B42]. For human exposure to plane-wave fields, the MPEs are often the maximum permissible plane wave exposure limit that ensures that a given whole body SAR will not be exceeded. The significance of exposure in close proximity to a near-field source, however, is difficult to estimate, especially when only external field strength data (obtained from radiation hazard survey meters) are available. The relationship between the spatial maximum field strength and the SAR is very complex varying considerably as the orientation and spatial distribution of the fields change with respect to the exposed object. Consequently, induced SAR in an object or a person near a radiator or passive reradiator is extremely difficult to estimate from measured data in the form of external field strengths. These measured data can only provide the basis for crude estimates of the maximum spatial SAR useful for establishing temporary personnel protection guides corresponding to a measured exposure situation (see also 5.6.2.4).

# 5.6.2 Enhanced or Reduced SAR in Personnel in Close Proximity to Passive Reradiators or Active Radiators.

Absorption in personnel exposed to reactive near fields may be either enhanced or reduced when compared with exposure in plane-wave fields. The reactive near-field region, for both active and passive radiators, i.e., where the reactive fields dominate over radiating fields, is limited to distances much smaller than a wavelength. Most MPEs are expressed in terms of those values of  $E^2$ ,  $H^2$  or *W* for *plane-wave* exposures that ensure that the *whole-body averaged* SAR is below some nominal value. In many cases, where or how these parameters are to be measured, e.g., spatial averaging, is not specified. In some cases, a maximum spatially localized SAR may be specified. This local SAR defines the maximum SAR that is allowed to exist in any small volume of tissue, e.g., one gram or one cm<sup>3</sup>. During RF hazard surveys, the exposure field strengths are measured, not the internal SAR. When a surveyor measures the field strength in a region that may be occupied by a person, and the field strength exceeds the applicable

MPE at one or more points in that region, a violation of the MPE is implied. However, exposure to nonuniform fields produces a nonuniform SAR distribution throughout the object occupying the region and the maximum induced SAR may or may not exceed the peak or the whole-body averaged SAR limits of the MPE depending on whether an "enhanced" or "reduced" absorption situation exists.

Although the degree of induced SAR in a person exposed to a near-field source is difficult to relate to the maximum value of the local field strength (E or H) it can be normalized with respect to a standard plane-wave-exposure situation. The term normalized plane-wave absorption is defined as the SAR induced by a plane wave whose field strength is equal to the maximum field strength (E or H) measured in the near-field exposure situation in question at the site that a person would occupy. The maximum spatial SAR induced by this normalized plane wave can then serve as a reference level for evaluation of the near-field exposure. If the corresponding maximum spatial SAR induced in a person's body by the near-field source exceeds the normalized plane-wave SAR, an enhanced absorption case exists; otherwise, a reduced absorption case exists.

The interpretation of field strength measurements near a passive reradiator or active radiator should be performed carefully to determine if an enhanced absorption case actually exists. When a typical, modern isotropic hazard survey meter is placed near a passive reradiator or an active radiator, the meter should accurately measure the field strengths if the guidelines in 5.3.6 are followed during the measurement procedure and, in many cases, these values may exceed the plane-wave limits of the MPE. However, the maximum RF energy that is actually absorbed in any part of a person's body and, hence, the maximum peak SAR, might not exceed the local SAR provisions of the same MPE. Conversely, some situations exist where the local SAR in a part of the body may indeed exceed the limits of the MPE [B46]. These two extreme cases (enhanced or reduced absorption) and their physical basis are examined in 5.6.2.1 and 5.6.2.2.

5.6.2.1 The Reduced Absorption Case. In any situations involving the exposure of a person to near fields from a passive reradiator or active radiator, the spatial maximum (local) SAR is relatively "low" with respect to the maximum local SAR that would be induced by a normalized plane-wave exposure, as discussed above. In virtually all cases, the whole-body-averaged SAR induced in a person under near-field exposure conditions is much less than the whole-body- averaged SAR associated with plane-wave exposure [B115]. Very high field strengths can be measured (with respect to the MPE) when a survey meter is placed close to a resonant (at the frequency of irradiation) passive reradiator. Also, high field strengths may occur in the near field of active radiators. The field strengths that are indicated by a survey meter, in close proximity to a passive reradiator or active radiator, may be much greater than the ambient field strengths existing only a fraction of a meter away (i.e., large field strength gradients exist). Thorough experimental and theoretical studies of near-field exposure from active radiators have been performed (see 5.6.2.3). Many cases result in reduced absorption. An example of the reduced hazard situation for a passive reradiator easily can be envisioned, but little experimental or theoretical SAR data are available to confirm this situation.

For example, high field strengths (relative to the ambient field strength) are generally measured in the immediate vicinity of small, resonant, passive reradiators. These passive reradiators include automobile steering wheels [B4, B85], and/or short linear metal objects. If a person is near, but not touching the resonant object, little energy

coupling exists between the reradiator and the person. It would be expected, but has not been confirmed, that the SAR induced in the person's body, is small relative to the local SAR limits of most MPEs.

**5.6.2.2 Enhanced Absorption.** Enhanced absorption is defined as the case where the spatial maximum SAR that is induced in a person's body by a reactive near-field source is not significantly less than the spatial-maximum SAR induced by a normalized plane-wave field. An example of the enhanced absorption involves personnel close to, or in contact with, an active radiator, or a large, passive reradiator, e.g., a large linear object equal in length to a multiple number of half wavelengths at the frequency of irradiation. In this case, high SARs, relative to normalized plane-wave SAR, can be induced in certain areas of the person's body. This has been demonstrated for the case of a full-size model of a person in very close proximity to, or in contact with a reradiator [B100].

**5.6.2.3 Studies of Induced SAR by Passive Near-Field Radiators.** Very little research has been performed to quantify the induced SAR associated with exposure to the near field of a passive reradiator. A preliminary series of experiments has, however, indicated that both reduced and enhanced absorption situations are encountered, depending on the spatial relationship of the reradiator to the model [B100].

**5.6.2.4 Studies of Induced SAR from Nearby Active Radiators.** Exposure of personnel to fields from nearby active radiators produces complex internal SAR distributions [B52]. Section 5.6.3 briefly describes several experimental and mathematical studies of a few common exposure situations and the resulting induced SAR distribution in nearby personnel (see also Appendix D6.2).

# 5.6.3 Using Near-Field Survey Data to Assess Potentially Excessive SAR in Exposed Personnel.

Current MPEs are expressed in terms of  $E^2$ ,  $H^2$  and W are based on whole-bodyaveraged SARs, below which adverse effects are not expected to occur. For most exposure situations, however, it is necessary to estimate the potential RF hazard that may exist by measuring the incident fields, i.e., the induced SAR in exposed personnel cannot be measured directly; only the external exposure-field parameters can. For uniform plane-wave, whole-body exposure, however, the whole-body-averaged SAR can be determined with reasonable accuracy using exposure field data since most MPEs are based on mathematical models and calculations of the whole-body SAR associated with exposure to plane-wave fields.

For specific non-plane-wave RF exposure environments, coarse, order-of-magnitude estimates can be made of the regional or local SAR in various areas of an exposed person's body for the far-field case and for certain near-field exposure situations. In certain instances, it may be possible to obtain these estimates, without actually performing invasive SAR measurements, e.g., thorough measurements of the exposure field-strengths in the particular RF environment in question could be compared with these values with published SAR data. References to the results of several comprehensive research studies of the local SAR for various exposure situations are found later in this section. The above measurements and comparisons will allow rough estimates to be made of the expected local SAR distribution in personnel exposed to fields similar to those of the RF environment surveyed. Under certain near-field exposure conditions, field strength data do not provide an adequate means for assessing potential RF hazards to personnel. When an RF radiator or reradiator illuminates a small portion of a person's body, and the spatial distribution of the fields is highly nonuniform over the volume in question, measurement of the local SAR distribution may be the most appropriate means

for hazard assessment [B13, B107, B108]. Therefore, under these circumstances, measurements of E and/or H may not be sufficient. This is especially true when the distance from the RF source to the exposed object is less than approximately 3 probe lengths (see 5.3.6.4).

Noninvasive techniques have been developed to estimate SAR from the measurement of induced body current. These measurements are made with devices that are located very close to or in contact with the body. For example, in exposure situations involving RF fields at frequencies below a few hundred MHz, the measurement of the total RF current flowing through the body, to ground, can be used to estimate the local SAR due to near-field coupling in various anatomical regions. This noninvasive technique is described in Appendix C3. Several techniques for the measurement of RF currents flowing in human extremities have been proposed. One technique, applicable over a wide range of frequencies, consists of a nonmetallic equivalent of a clamp-on ammeter. This technique was designed and studied via mathematical modeling and its feasibility has been demonstrated experimentally [B50, B55; Hagmann and Babij, 1992].

Information has been published that describes the measurement relationships between external exposure fields around an active, near-field radiator, and the resulting measured internal SAR in nearby simulated biological objects [B53, B115]. Other studies report experimental data on the local internal SAR distribution relative to the spatial location of active radiators [B27, B31, B115]. Mathematical models useful for predicting internal SAR distributions in humans exposed to nonplane-wave fields from various RF emitters have also been published [B27]. Similar data on the SAR induced by passive reradiators is not generally available.

In practice, the fundamental problem confronted in assessing human exposure to nonuniform fields is selecting an appropriate protocol that will yield meaningful measures of RF fields that can be related to whole-body averaged and local SAR. Determining the appropriate measurement distance from a near-field source is, at best, a compromise. However, when considering the potential constraints on measurement accuracy caused by inadequate probe-source separation distance and the importance of assessing whole-body exposure levels, a minimum measurement distance of 20 cm is recommended, in accordance with guidance in 5.3.6.3.

# Appendixes

These appendixes are not part of IEEE Std C95.3-1991, *IEEE Recommended Practice for Measurements and Computations with Respect to Human Exposure to Radiofrequency Electromagnetic Fields, 3 kHz to 300 GHz*, but are included for information only.)

# **Appendix A**

# Additional Calibration Techniques for External Field Measuring Instruments

#### A1. General.

It is sometimes advantageous to use open parallel plates or transmission lines having a characteristic impedance other than 50  $\Omega$ , instead of a TEM cell. The major limitations of TEM cells and parallel plate lines are the inverse proportionality between the size and upper frequency limit. For proper operation, the spacing between the cell conductors, i.e., the width of the septum, and the length of the uniform line (nontapered section), should all be less than I/2. Another limitation is that a relatively large amount of power is required to produce desirable field intensities, for the usual case, where the characteristic impedance  $Z_0$  and load resistance are 50  $\Omega$ . Some alternative methods are mentioned below.

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# Fig A1 Open Parallel Plate Calibration System

NOTES:

- (1) Top and bottom are metal clad plywood
- (2) Vertical supports are wood 2 X 4s
- (P) Location P is the optimum calibrating point

#### A2. Open Parallel Plates

Figure A1 is a sketch of a system using open parallel plates to generate a known electric field at frequencies up to about 30 MHz. The expression for calculating the E-field existing between two flat conducting plates is the same as that for a TEM cell, namely

$$E = V/b (V/m)$$

(Eq A1)

where *E* is the electric field strength, *V* is the voltage difference between plates and *b* is the plate spacing. The uncertainty of the field level is the same as that for a TEM cell. The wave impedance is greater than 377  $\Omega$ , similar to that of an unterminated TEM cell. Therefore, a larger E-field can be produced for a given RF input. If the line is made resonant by means of an inductor, larger field strengths can be produced, but more care must be taken to ensure the accuracy of calibration. In the example of Fig A1, an E-field of 1000 V/m can generally be achieved with a 1 W power source.

# Fig A2 Parallel-Plate Transmission Line for Generating E- and H-fields, f = 0.3 - 30 MHz

NOTES:

- (1) Top and bottom are metal clad plywood
- (2) Vertical supports are wood 2 X 4s

- (3) Input connector
- (4) Copper straps
- (5) Load resistors
- (P) Optimum calibration point

# A3. Parallel-Plate Transmission Line

Figure A2 is a sketch of a parallel-plate line that is useful for generating known E- and Hfields at frequencies up to 30 MHz. This type of TEM calibrating line is specified in MIL-STD-462  $[A4]^9$  and SAE Standard ARP-958 [A5]. The required value of load resistance for terminating the line with a minimum VSWR depends mainly on the plate spacing and width. The characteristic impedance of the line  $Z_0$  is also affected by its surroundings, especially if the line is located in a shielded enclosure, such as a screen room. Therefore, the optimum value of terminating resistor should be verified experimentally from measurements of the E-field uniformity. The value of  $Z_0$  for the parallel-plate line of Fig A2 is 80 to 100  $\Omega$ . The E-and H-fields are again given by Eq 4.10 in 4.5.3.

At frequencies approaching self-resonance of the system (similar to open-parallel plates) it is difficult to calculate the field level accurately and, it is then necessary to measure the field strength at the "calibrating point" with a small transfer probe, such as a 5 cm dipole that has been previously calibrated in another type of reference system.

# A4. Parallel-Wire Transmission Line

A two-conductor TEM line using two balanced thin wires can also be used to generate calculable values of E and H. This transmission line has a relatively high characteristic impedance ( $Z_0 = 300$  to  $1000 \Omega$  so the E-field midway between the two wires is relatively high, but the field uniformity is not as good as that produced by flat plates. For example, a field strength of 200 V/m can be produced by a 10 W source when using a two-wire line with a spacing of 30 cm. Since the *E*/*H* ratio is 377  $\Omega$ , the H-field level is calculable and also is relatively high.

Figure A3 shows the E-and H-field distribution in the vicinity of a two-wire line. The direction of the electric field vector at point P is also shown in the figure. H is perpendicular to E. If the length of the transmission line is greater than five times the distance between the wires, the equation for calculating the magnitude of E and H midway between the two wires is:

$$E = \frac{2V}{Z_0 \left(d + \sqrt{d^2 - r^2}\right)}$$

(Eq A2)

and

where:

E = Electric field strength midway between the 2 wires (V/m) H = Magnetic field strength midway between the 2 wires (A/m)

<sup>&</sup>lt;sup>9</sup> Numbers in brackets referred to throughout the Appendixes of this document refer to the Bibliography section at the end of the Appendix in which it is referenced.

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- V = RMS voltage on each line with respect to the neutral plane midway between the wires (V)
- $Z_0$  = Characteristic impedance of the transmission line ( $\Omega$ )
- =  $120 \cosh^{-1} (d/r)$ d = Distance from the field point P (midpoint of line) to the center of each wire (m)
- r = Radius of each wire (m)

Fig A3 Parallel Wire Transmission Line for Generating E- and H-Fields

#### Fig A4 Long-wire Antenna Chamber

#### A5. Long-wire Antenna Chamber

A standard screen room can be converted into a chamber with calculable E- and Hfields. This provides a large volume with a moderately uniform field distribution for the exposure of objects or for antenna calibrations at frequencies up to about 30 MHz. As shown in Fig A4, a single wire is stretched horizontally between two insulators that are attached at opposite ends of the screen room. For a typical room with a height of 8 ft, the wire height is usually about 6 ft above the floor. This TEM transmission line is usually fed via a coaxial transmission line at one end of the room, using a shunt resistor  $R_1$  to help terminate the coaxial line in its characteristic impedance. Resistor  $R_2$  is used to terminate the single wire line in its characteristic impedance (usually 300 - 600 ohms).

Theoretical investigations of the fields in a screen room, when energized by a single-wire transmission line, indicate that the field uniformity is quite poor. The distribution of the E and H fields below the line depends on optimization of the line-terminating impedance. Equations to determine the values of *E* and *H*, and the approximate values of  $R_1$  and  $R_2$  are given in the literature [A1, A2, A3, A4]. In the derivation of these equations, it is assumed that a pure resistance is an adequate termination for the wire transmission line. If the line height is greater than 2/3 the room height, the presence of the side walls and floor of the screen room is generally ignored when calculating the characteristic impedance of the line. The field strength at point P, near the center of the screen room, is given approximately by

$$E \approx \frac{60V}{\left(Z_0 + Z_g\right)} \left(\frac{1}{d} - \frac{1}{d+2a} - \frac{1}{d-2b} + \frac{1}{d-2a-2b} + \frac{1}{d+2a+2b} - \frac{1}{d-2a+4b} - \frac{1}{d+4a+2b} + \frac{1}{d-4a-4b}\right)$$
(Eq A3)

and

 $H \approx E/377$ 

where:

- E = Electric field strength at point P midway between the ends of the screen room (V/m)
- H = Magnetic field strength at the same point (A/m)
- V = Voltage on the single wire transmission line (V)
- $Z_0$  = Characteristic impedance of the line ([W])  $\approx 60 \ln (2a/r)$
- $Z_{\rm g}$  = Impedance of the generator ( $\Omega$ )
- r =Radius of the wire (m)
- a = Distance from the wire to the ceiling (m)
- b = Distance from the wire to the floor (m)

d = Distance from the wire to field point P (m)

## A6. Bibliography

[A1] ANSI C63.2-1987, American National Standard for Specifications for

Electromagnetic Noise and Field-Strength Instrumentation, 10 kHz to 40 GHz.

[A2] ANSI C63.3-1964 (R1969), American National Standard for Specifications for Radio-Noise and Field-Strength Meters, 20 [E] 1000 Megacycles/Second.

[A3] ANSI C63.4-1991, American National Standard for Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz.

[A4] MIL STD-462, Measurement of Electromagnetic Interference Levels, (1986).[A5] SAE Standard ARP-958, Electromagnetic Measurement Antennas; Standard

Calibration Requirements and Methods, (1991).

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# Appendix B

# **Theoretical Calculations of Exposure Fields**

## **B1. Technical Considerations of RF Source Characteristics**

Although the prediction of power-density levels in the vicinity of RF sources is complicated by many factors, useful estimates can be made. The quality of such calculations will depend on the analytical approach used, as well as on the accuracy of the values of the peak power, pulse duration, pulse repetition rate, antenna radiation patterns, antenna placement, and scanning rates that are used in theoretical computations. Corrections for near-field effects may also be appropriate. The operating parameters listed below must be specified adequately so that the true average-radiated power from the antenna, and the resulting power density at a distant point can be calculated.

For all sources (pulsed or CW), the antenna type and size, gain, antenna pattern including E and H plane beam widths and sidelobe distribution, antenna height above ground, operating frequency, antenna beam orientation (all possible cases) and the attenuation of the transmission line that connects the RF generator to the antenna should be known or estimated. For the calculation of the expected power density levels of pulse modulated sources, the maximum possible values of peak power, pulse duration, and pulse repetition rate that closely approximate, but do not exceed the maximum rated duty factor of a transmitter should be used. In the case of multiple sources, the contribution of each source should be considered when estimating the combined effect.

### B2. Antennas -- On Axis

The field in front of an antenna can be characterized by the following three regions.

- (1) Reactive Near-Field Region. This is the region of space immediately surrounding the antenna or leakage source where the reactive (nonradiating) components predominate and energy is stored in the field. The reactive near-field region extends to a distance of approximately one wavelength from the antenna, except for the case of electrically large antennas (whose physical size is greater, in any dimension, than several wavelengths).
- (2) Radiating Near-Field Region (Fresnel Region). In this region, which starts at a distance from the antenna where the reactive field region has diminished to an insignificant amount, the antenna gain and the angular distribution of the radiated field vary proportionally with distance from the antenna. This is because the phase and amplitude relationships of the various waves arriving at the observation point from different areas of the antenna change with distance. For reflector-type antennas, such as parabolic dishes, the radiation is somewhat more complex in its distribution pattern.
- (3) *Far-field Region (Fraunhofer Region).* This region is sufficiently far from the source that the phase and amplitude relationships of the waves arriving from different areas of the antenna do not change appreciably with distance. The antenna gain and angular pattern are essentially independent of distance, and the power density is inversely proportional to the square of the distance from the source. Although

the transition from the non- radiating near field is a gradual one, the far-field region is commonly assumed to begin at a distance of about  $2a^2/I$  for antennas with equiphase excitation and extends to infinity (*a* being the largest linear aperture dimension and *I* is the wavelength at the frequency of interest). This criterion is not adequate for all types of antennas and should not be applied indiscriminately.

To compute an approximate value for the maximum power density W in the Fresnel and far-field regions of an antenna, use the equation:

$$W = \frac{GP_T}{4\pi d^2} = \frac{A_e P_T}{\lambda^2 d^2}$$
(Eq B1)

where *G* is the far-field antenna gain (power ratio),  $P_T$  is the net power delivered to the antenna, *d* is the distance to the antenna, *l* is the wavelength, and  $A_e$  is the effective area of the antenna. If *G* is not known, a useful approximation for *W* can be obtained by substituting *A*, the physical aperture area, for  $A_e$  in Eq B1. Since *A* is generally larger than  $A_e$ , the estimated value of *W* will be somewhat larger than the actual value.

Equation B1 can be used to estimate W at distances greater than about  $0.5a^2/I$  where a is the largest aperture dimension. At closer distances, the values predicted by Eq B1 are too large and near-field estimates must be used. For commonly used horn and reflector antennas, the maximum power density  $W_m$  expected in the radiating near-field can be estimated by [B1, B2]:

$$W_m = \frac{4P_T}{A} \tag{Eq B2}$$

The values predicted by Eq B2 will be within  $\pm 3$  dB of the correct value (in the absence of reflections) for square apertures with uniform, cosine, and cosine<sup>2</sup> amplitude tapers, and for circular apertures with tapers ranging from uniform up to  $(1-q^2)^3$  [B2].

If a computation indicates that the approximate power density is substantially less than the MPE, then there is usually no need for further calculation since Eq B2 provides the maximum power density that can exist on the axis of the beam of an antenna that is focused at infinity, in the absence of reflections. (An antenna focused at a lesser distance could produce a higher power density in the region of its focal point, but this condition is unusual.)

If the computation from Eq B2 reveals a power density value that is equal to or greater than the MPE, then it must be assumed that this value may exist at any point in the radiating near-field region and attention should be directed to the exposure fields in the far-field region.

Equations B1 and B2 do not include the effect of ground reflections. Values of power density that exceed the free-space value by a factor of four times can result when the main beam is directed toward a planar ground or reflecting surface. If the shape of the reflecting surface is such that it produces focusing effects, even greater values may result. After considering the sources of error cited above one may calculate the distance to the boundary of the potentially- hazardous zone (in the presence of reflections) as follows.

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$$r = \sqrt{GP / \pi W} \tag{Eq B3}$$

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The use of a cylindrical model has been found useful for evaluating the RF fields near vertical collinear dipole antennas similar to those used for cellular, personal communications services, paging and two-way radio services [Tell, B, C]. In this model, spatially averaged plane wave equivalent power densities (*S*) parallel to the antenna aperture may be estimated by dividing the net antenna input power ( $P_{net}$ ) by the surface area of an imaginary cylinder surrounding the length (*h*) of the radiating aperture. Thus

$$S = \frac{P_{net}}{2\mathbf{p}Rh}$$
(Eq B4)

where *R* is the radius of the cylinder (distance to the antenna).

While the actual power density will vary along the height of the antenna, the average value along its aperture will be close to the value predicted by B4.

Tell [B] found this simple model compared very favorably with a more exact calculation based on the method of moments. Within the aperture of an element of a vertical collinear dipole array antenna operating at 881.5 MHz, he found that for distances closer than 16 ft. from the aperture, the cylindrical model predicted power densities within 20% of the actual spatially-averaged power density.

For sector type antennas, power densities can be estimated by dividing the net antenna input power by that portion of the cylindrical surface corresponding to the angular beamwidth of the antenna. For example, in the case of a 120 degree azimuthal beam width, the surface area should correspond to 1/3 of that of a full cylinder; this will increase the power density near the antenna by three times that of a purely omnidirectional antenna. This is expressed as:

$$S = \left(\frac{180}{\Theta_{BW}}\right) \frac{P_{net}}{pRh}$$
(Eq B5)

where  $\Theta_{BW}$  is the 3 dB beamwidth of the antenna. Use of the 3 dB azimuthal beamwidth of the antenna will generally result in a conservative estimate of the power density for sector antennas since the power is radiated into a sector larger than the reported 3 dB beamwidth. The cylindrical model is also useful for estimating the power density adjacent to FM and TV broadcast antennas where workers may be located during tower work.

### B3. Antennas -- Off Axis

It is more difficult to calculate the power density off the axis of the main beam, and requires the solution of complex mathematical equations. One approach reveals that the collimated beam in the radiating near field falls off with increasing distance approximately 12 dB per unit of antenna radius. Many antennas do not have simple shapes or illumination tapers. In such cases, the approximate formula above will not apply directly, and a more complex analysis is indicated. The U.S. Department of the Air Force

Electromagnetic Radiation Hazards Technical Manual [B3] provides the results of such computations in the form of normalized curves. However, a high-order of precision is not warranted when computing the expected power density because of the many physical parameters in the environment that create significant variations in the values predicted by idealized computations.

# **B4. Scanning Correction**

In the case of scanning antennas, the average power density at a fixed point will be reduced by the value of the effective antenna-pattern beamwidth divided by the scanning angle (the number of degrees of antenna rotation during a scan). This assumes that a constant rotational velocity is used, and that the antenna rotates in one direction, rather than stopping after a scan, and reversing direction. Accordingly, the potentially-hazardous distance is decreased by at least the square root of this ratio (if the period of rotation is less than the averaging time specified in the MPE). The antenna's effective beamwidth in the far field will, in general, be somewhat different from the 3 dB beamwidth. The exact value depends upon the form factor of the radiation pattern and associated sidelobes.

In the Fresnel Zone, the effective angle of the beamwidth will vary with distance. Here, the average power density W of the scanning antenna is given approximately by the following relationship;

|      | W = (aP / A) | $(a/2\pi r)(360/\theta)$                | $q > 360(a/2\pi r)$ | (Eq B5) |
|------|--------------|---|---------------------|---------|
| and, | W = 4P/A     | <b>q</b> > 360( <i>a</i> /2π <i>r</i> ) |                     | (Eq B6) |

where:

q = the scanned angle, in degrees P = the average power transmitted

A = the effective area of the antenna

a = antenna diameter or width

r = distance from the antenna

# **B5. Bibliography**

[B1] Bickmore, R. W. and R. C. Hansen, "Antenna Power Densities in the Fresnel Region," *IRE Proceedings*, no. 47, pp. 2119 [E] 2120, Dec. 1959.
[B2] Mumford, W. W., "Some Technical Aspects of Microwave Radiation Hazards," *IRE Proceedings*, no. 49, pp. 427 [E] 447, Feb. 1961.

[B3] USAF, Electromagnetic Radiation Hazards Technical Manual, TO 31Z-10-4, Aug. 1966, US Dept. of the Air Force, 1966.

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# Appendix C

#### **Additional Techniques for SAR Measurement**

#### C1. Thermography

A method for rapid evaluation of the distribution of SAR throughout an entire planar surface of a biological object or a model that is composed of heterogeneous dielectric materials is described in [C3, C10]. The method involves the use of a thermographic camera for recording the rate of rise of temperature in a plane that bisects an entire model (phantom) or a cadaver of a biological subject under study. The temperature distribution before, and immediately after brief, high-power irradiation, is observed on the

precut surface on each half of the bisected model. This is done to prevent cooling by evaporation or flow of the wet synthetic tissue out of its shell (usually composed of synthetic fat or rigid plastic foam).

The technique for using the phantom model to determine SAR distribution throughout a planar surface within a phantom is as follows [C2, C9, C10]:

The model is first exposed to the same electromagnetic source that will be used to expose actual tissue. The power applied to the model will be considerably greater, however, in order to heat it in the shortest time possible. After a short exposure duration, the model is quickly disassembled and the temperature pattern over the surface of separation is observed and recorded by means of a thermographic camera. The exposure duration is typically 5-60 s, depending on the source. After about a 5 s delay (for separating the two halves of the model), the recording is carried out within a 1 s period. Since the thermal conductivity of the model is low, the difference in the measured temperature distribution before and after heating will closely approximate the heating distribution over the planar surface, except in regions of high- temperature gradients where errors may occur due to appreciable diffusion of heat, e.g., at the bone-muscle or fat-muscle interface.

It should be noted that actual animal cadavers can be frozen, bisected (in a plane parallel to the E-field), and each half covered with a thin polyethylene sheet. Then the two halves can be placed tightly together so that the bisected surfaces are in tight contact (with no residual air gap). RF irradiation and subsequent thermographic measurements can proceed, but the precautions of thermodynamic degradation of heating patterns associated with excessive  $\Delta T$ , etc. (see 5.5.1 and C1) should be considered.

As with temperature probes, thermodynamic factors and the imprecise knowledge of the value of the specific heat capacity of the tissue or phantom material tend to limit the accuracy and precision of SAR measurements made via thermography. In addition, a large source of error exists unless the persons performing the measurements posses a great deal of experience. The preparation and use of bisected test subjects (particularly cadavers) for evaluation by thermography affects the accuracy of SAR measurements significantly. In particular, when attempting to provide a continuous, electrical path across the plane of bisection in the subject under test using a plastic or silkscreen membrane, air gaps at the interface frequently cause errors. Also, it is difficult to irradiate a subject, split it open, and observe the heating pattern in less than 5 s after irradiation ceases. This time delay creates large, thermodynamically-induced errors, particularly at the boundaries of dissimilar tissues, and at the air/subject interface. Therefore, uncertainties of at least  $\pm 1$  dB to 2 dB should always be expected and should be cited along with SAR data obtained via thermography.

#### C2. Body Current Measurement

SAR may be assessed by measuring the RF current flowing in an exposed object. In humans, measurement of the induced currents flowing in the legs to ground have been studied at RF frequencies below about 50 MHz [C5, C7, C12]. The induced current is determined by standing the individual on a conductive plate electrode and measuring either the current flowing to ground with a RF-milli-ammeter or measuring the RF voltage drop across a known resistance connected between the plate electrode and ground. The

value of the resistor should be small enough so as not to perturb the body current, i.e., small compared with the capacitive reactance of the plate and standing subject to ground. From the voltage drop *V* across the resistance *R*, the current *I* may be computed from the expression I = V/R. SAR is then obtained by using the expression

$$SAR = rJ^2/s$$
 W/kg

(Eq C1)

where:

r = tissue resistivity ( $\Omega \cdot m$ ) J = current density (A/m<sup>2</sup>) s = mass density (kg/m<sup>3</sup>)

*J* is determined by dividing the measured current by the cross sectional area of the conductive tissues in the region of interest. Measurements of *SAR* derived from current measurements are most accurately related to anatomical areas where the effective conduction cross-section and current are best known. At frequencies of 50 MHz and below, this is typically the legs and ankles, or arms and wrists.

Localized SAR can be derived from current measurements in extremities. A simplified "effective cross-section" related to realistic anatomical features can be used to determine current density. For many years, a 9.5 sq. cm effective cross-section for the adult human ankle was assumed. That value represents about 15% of total area. However, recent experimental evidence points to an effective cross-section of 60% of the total. [Olsen and Van Matre]

#### C3. Fabrication of Simulated Tissues

Phantom models are often used together with temperature probes, E-field probes and thermographic cameras. They are composed of materials with dielectric, thermal and geometric properties identical to the biological subject they represent. Phantom materials have been developed that simulate human fat, muscle, brain and bone. The dielectric properties of the phantom can be varied over a wide range by varying the percentage of constituent materials. In one formulation that was developed by [C6], the relative amounts of polyethylene powder and the salinity of the water used for making the simulated muscle material can be varied to simulate specific tissues of high water content. This particular form of phantom material is very viscous, while pliable and putty-like, and is well suited for the construction of bisected phantoms for thermographic evaluation. Mixing techniques and the exact grain size of the polyethylene powder affect the homogeneity and dielectric properties significantly. Once poured into a mold, the gel has a tendency to entrap air pockets (see Tables 5.4 and 5.5).

A less viscous gel has been developed using hydroxethylcellulose (HEC) gelling agent and salt water (saline Polyethylene powder or sugar should be added for use above 100 MHz [C8]. The HEC gel does not retain permanent voids and air bubbles as readily as the putty-like formulations [C1]. Visual inspection for air bubbles and the position of implanted probes can be performed simply in the optically-transparent HEC gel.

Preservation with a bactericide, and refrigeration are necessary when making either of these simulated soft-tissue gel formulations. Also, evaporation of the water should be prevented by proper sealing of the containers, or the molds or shells that contain the gel.

Phantom models of various tissue geometries and gross anatomies of whole animals or humans can be fabricated. They include stratified layers of muscle and fat of various thickness, circular and irregular cylindrical structures consisting of fat, muscle and bone and spheres of synthetic brain tissue to simulate various parts of the anatomy. Various liquids have been developed and used to simulate internal body tissues. They are poured to fill rigid, outer shells of simulated subcutaneous fat in the shape of a full-size human [C11]. Implantable E-field probes can be easily moved with mechanical scanners in a continuous path throughout these liquid- filled phantoms. This provides an efficient means for the mapping of local SAR distributions throughout models with complex shapes such as the extremities of the human body.

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### **Appendix D**

#### **Theoretical Calculations of SAR**

#### D1. General

The rate of RF energy absorption per unit mass at a point in an absorbing body is called SAR. The derivation of the term is discussed in [D16]. Both "dose rate" and "density of absorbed power" (often called absorbed power density) are equivalent to SAR. Each term refers to the amount of energy absorbed per unit time, in a unit mass. For steady-state sinusoidal fields, *SAR* is given by:

$$SAR = \frac{1}{2\rho} \omega \varepsilon_0 \varepsilon'' E_{int}^2 \text{ (W/kg)}$$
$$= \frac{\sigma}{2\rho} E_{int}^2$$
(Eq D1)

where:

- r = mass density of the body (kg/m<sup>3</sup>)
- $\varepsilon_0$  = permittivity of free space (F/m)
- $\varepsilon''$  = imaginary part of the complex relative permittivity
- w = radian frequency given by  $w = 2\pi f$
- s =conductivity (S/m)
- $E_{int}$  = peak electric field strength in V/m at the point in the body, with the subscript "int" used to emphasize the fact that the field inside the body is not the same as the external field strength surrounding an exposed object.

There are three principal ways in which energy is transferred from an electric field to an absorbing object. One is the transfer of kinetic energy to electrons that are not tightly bound to any one atom, and hence are called free electrons. The DC conductivity of a material is associated with the free electrons. The other two forms of energy transfer are friction associated with the alignment of electric dipoles with  $E_{int}$  and friction associated with ionic and molecular vibrational and rotational motion associated with

 $E_{int}$ . Since  $\varepsilon''$  represents the loss of energy in the E-field, we say a substance is "lossy" if  $\varepsilon''$  is relatively large compared with  $\varepsilon'$ ; thus, a lossy substance that is exposed to RF energy absorbs a relatively greater amount of electromagnetic energy. In most cases, the more water or other "polar" molecules a substance contains, per unit volume, the more lossy it is; and the drier a substance is, the less lossy. For example, a dry piece of paper placed in a microwave oven will not get hot; but a wet piece of paper will heat until the paper is dry, and then will heat no further. Fat has a lower water content than muscle tissue, so fat is less absorbing. It is important to remember that these relative statements are true only if  $E_{int}$  within the various materials is the same in each case. However, as explained below, the permittivity of the material also affects the coupling of external fields, thus influencing the magnitude of  $E_{int}$ .

The dielectric properties of materials are usually listed as dielectric constant and either loss tangent or conductivity. However, use of the term dielectric constant for material exhibiting the frequency-dependence of a material containing water is not accurate since the dielectric properties of water and biomaterials are not "constant" with frequency. The complex permittivity is defined as:

$$\boldsymbol{\varepsilon} = \boldsymbol{\varepsilon}_0 \left( \boldsymbol{\varepsilon}' - \boldsymbol{j} \boldsymbol{\varepsilon}'' \right) \tag{Eq D2}$$

where  $\varepsilon_0 = 8.854 \times 10^{-12}$  farads/m and is the permittivity of free space,  $\varepsilon'$  is the dielectric constant or real part of the complex relative permittivity,  $\varepsilon''$  is the imaginary part of the complex relative permittivity. The loss tangent is defined as:

$$\tan \delta = \frac{\varepsilon'}{\varepsilon''}$$
(Eq D3)

which is a measure of how "lossy" or energy-absorbing a material is. The conductivity  $\sigma$  is related to  $\epsilon''$  by:

$$\varepsilon'' = \sigma / \omega \varepsilon_0$$
 (Eq D4)

It is very important to note that the SAR is not a measure of temperature rise; it is a measure only of the rate of energy absorption per unit mass that depends upon the conductivity of the material [D9]. The temperature rise is a function of the SAR, but it is also a function of the thermal characteristics of the absorber (i.e., the size, shape, and

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thermal conductivity to the surroundings). The term "hot spots" is often used to refer to areas of locally high SAR, but this term is not meant to describe areas of localized high temperature. Since biological systems have various thermo-regulatory mechanisms that can automatically adjust the temperature in various regions of the system, more work needs to be done in the calculation of temperature distribution in living biological systems.

#### **D2. Factors Affecting the Internal Fields**

At each location inside a body, E and H should also satisfy Maxwell's equations. In addition, at any boundary between different dielectric media, (such as bone and muscle tissue) E and H should satisfy certain boundary conditions; that is, at the boundary, the fields in the one medium should be specifically related to the fields in the other medium. The boundary conditions are usually expressed in terms of two vector field components; one component parallel to the boundary, and one normal to the boundary. The boundary conditions for the electric field in dielectrics are:

$$\varepsilon_1 E_{1n} = \varepsilon_2 E_{2n}$$
(Eq D5)  
$$E_{1p} = E_{2p}$$
(Eq D6)

where subscript 1 represents quantities in one medium, and subscript 2 represents quantities in the other medium. Also,  $E_n$  is the component of the electric field normal to the boundary, and  $E_p$  is the component parallel to the boundary. Similar relations hold for the magnetic fields.

$$\mu_1 H_{1n} = \mu_2 H_{2n}$$
(Eq D7)  

$$H_{1p} = H_{2p}$$
(Eq D8)

where **m** is the permeability of the medium. For nonmagnetic materials,  $\mathbf{m} = \mathbf{m} = \mathbf{m}$  and both components of *H* are continuous across a boundary. (It should be noted that the magnetic permeability of biological tissue is equal to that of free space.) The boundary conditions are important because they allow us to make significant qualitative statements about the internal fields.

#### **D3. Low-Frequency Internal Fields**

At very low frequencies, where the wavelength is very long with respect to that of the biological system, most exposures are in the near-field [D12]. At these frequencies the electric and magnetic near-fields are approximately independent of each other and, therefore, are similar to static fields. Such fields are called quasi-static fields. The characteristics of quasi-static electric fields can be demonstrated by considering the following example. Suppose a dielectric object is placed in a quasi-static field, E<sub>inc</sub>, as shown in Fig D1. Einc will cause the dielectric to polarize and produce a secondary field  $E_{\rm s}$  both inside and outside the dielectric. The combination of the secondary and the incident fields inside the dielectric is called the internal field. At the boundary, the sum of the incident and secondary fields should satisfy the boundary conditions with the internal field. Often, the secondary field is less important in a qualitative interpretation, in which case we can talk about the boundary conditions in terms of the incident field and the internal field. Near the ends of the object, the incident field in Fig D1(a) is primarily normal to the boundary. Therefore, the boundary conditions require that the internal field be very weak at the ends because it will be almost equal to  $(\varepsilon_0 / \varepsilon_1)(E_{inc} + E_s)$ , and  $\varepsilon_1 > \varepsilon_0$ . On the other hand, at the sides of the object, the incident field is primarily parallel to the boundary, and the internal field will be almost equal to  $E_{inc} + E_s$ . Since  $E_{inc}$  is essentially

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parallel to the boundary over a much greater region than the region where it is mostly normal to the boundary, the internal field will be determined more by the boundary conditions on the parallel field components; hence, we say that the electric-field coupling in Fig D1(a) is strong. The electric-field coupling is weaker in Fig D1(b), because Einc is mostly parallel to the boundary over a much smaller percentage of the object than in Fig D1(a). Thus, in Fig D1(a) where  $E_{inc}$  is aligned with the long axis of the object, strong electric fields are coupled into the object; if the object is lossy, it will absorb a proportionally large amount of energy. Conversely, in Fig D1(b), where  $E_{inc}$  is normal to the long axis of the object, weaker electric fields are coupled into the object and it will, thus, absorb a proportionally smaller amount of energy.

#### Fig D1 Dielectric Object Placed in a Quasi-static Electric Field, $E_{inc}$

In a lossy object at low frequencies, the magnetic-field coupling can be thought of in terms of induced eddy currents that circulate in closed paths around the incident magnetic-field vector. The circulating currents represent energy transfer from the incident magnetic field to the object. In some sense, the magnitude of the energy transferred is proportional to the cross-sectional area intercepting  $H_{inc}$ . In Fig D2, the magnitude of the energy transferred is greater in (b) than in (a) because the cross-sectional area perpendicular to  $H_{inc}$  is greater in (b) than in (a). Consequently, the magnetic field coupling is stronger in Fig D2(b) than in D2(a). Thus, for low frequencies, we can explain the relative differences in SAR for E-, H-, and K-polarizations, as shown by the dosimetric data. (The K vector defines the direction of propagation of a plane wave.) For E-polarization (where *E* is parallel to the long axis of the object), both E and H coupling are strong. For H-polarization (where *H* is parallel to the long axis of the object),

both E and H coupling are weak. For K-polarization (where the wave propagation vector, k, is parallel to the long axis of the object), E coupling is weak, but H coupling is strong. Thus, for frequencies below resonance, the SAR is greatest for E-polarization, least for H-polarization, and intermediate for K-polarization.

#### Fig D2 Prolate Spheroid Placed in a Quasi-static Magnetic Field, Hinc

Another important characteristic is related to interdependence of the fields in different parts of an absorber. At low frequencies (long wavelength with respect to the irradiated object), the fields in one part of a body are strongly affected by the other parts of the body, and field concentrations can be produced by combinations of size, shape, and orientation of the body in the incident field. At intermediate frequencies, the fields inside can be thought of as propagating waves that are partially transmitted and partially reflected at the boundaries, with constructive and destructive interference causing significant variations in field strength inside the body. Also, at these intermediate frequencies, *E* and *H* cannot exist separately and are strongly coupled together. At very high frequencies, however, the internal fields in different parts of an absorber can be essentially uncoupled, as explained below.

#### D4. High-Frequency Behavior

At very high frequencies, where the wavelength of the incident wave is very small compared with the size of the object, geometrical optics concepts apply. That is, the EM waves can be thought of as rays. In this case, for lossy objects, the effects are mostly surface effects because the depth of penetration decreases rapidly as frequency increases. This is the so-called skin effect. The penetration depth depends on both conductivity and frequency. At 10 GHz, for example, the skin depth in tissue is less than 5 mm. At these very high frequencies, the fields in one part of the body are affected only very slightly by other parts of the body.

#### D5. Methods of Calculation

In principle, the internal fields in any object irradiated by EM-fields can be calculated by solving Maxwell's equations. In practice, this is very difficult and, until recently, could be done only for a few very special cases (idealized models, such as spheres or infinitely long cylinders. Because of the mathematical complexities involved in calculating *SAR*, a combination of techniques has been used to obtain *SAR* for various models as functions of frequency. Each of these techniques provides information over a limited range of parameters. Combining the information thus obtained, gives a reasonably good description of *SAR* as a function of frequency over a wide range of frequencies for a number of useful models.

Up to frequencies of about 30 MHz, a method called the long-wavelength approximation has been used with spheroidal models of human-sized objects. The extendedboundary-condition method (EBCM) has been used to make calculations for spheroidal models of man up to approximately resonance (80 MHz). The iterative extendedboundary-condition method (IEBCM), an extension of the EBCM, has been used for calculation up to 400 MHz for spheroidal models. The classical solution of Maxwell's equations for cylindrical models (for human-sized objects or limbs) has been used to obtain useful average *SAR* data for E-polarization from about 500 MHz to 7 GHz, and for H-polarization from about 100 MHz to about 7 GHz. An approximately 7 GHz. The moment-method solution of a Green's-function integral equation for the electric field is used up to about 400 MHz (for human- sized models). For K-polarization, a technique called the surface-integral-equation technique (SIE) is used up to about 400 MHz with a model consisting of a truncated cylinder capped on each end by hemispheres.

Numerical simulation techniques are now available for determining SAR and current distributions in highly sophisticated millimeter-resolution anatomically-based models exposed to a wide variety of far-field and near-field sources [Gandhi, 1990; Gandhi, 1995]. From a wide array of methods, including the moment method (MM) [Chen and Guru, 1977; Hagmann, et al., 1979; Spiegel, 1984; Livesay and Chen, 1974], finite element method (FEM) [Chiba, et al., 1984; Yamashita and Takahashi, 1984; Morgan, 1981; Lynch, et al., 1985], finite-element time-domain method (FETD) [Brauer, et al., 1989: Chen and Lien, 1979], generalized multipole technique (GMT) [Hafner and Kuster, 1991; Hafner, 1990; Leutchmann and Bomholt, 1990], volume-surface integral equation method (VSIE) [Shankar et al., 1989] admittance [Armitage, et al., 1983] and impedance [Gandhi, et al., 1984] methods and the finite-difference time-domain (FDTD) method [Gandhi, 1990; Lin and Gandhi, 1995; Sumunic, 1995; Gandhi, et al., 1992; Chen, et al., 1991; Chen and Gandhi, 1991] has become the most widely used method of choice for bioelectromagnetic applications in the range of a few MHz to several GHz. An extension of the FDTD method, the frequency-dependent finite-difference time-domain method ((FD)<sup>2</sup>TD) [Luebbers, et al., 1990; Bui, et al., 1991; Joseph, 1991; Kunz and Luebbers, 1993; Taflove, 1995; Lee, et al., 1991, Sullivan, 1992; Sullivan, 1992a; Gandhi, et al., 1993; Furse, et al., 1994], enables broad-band bioelectromagnetic simulations by including the effect of the frequency dispersion of the tissues. This method has been used to calculate SAR and current distributions in the body from ultra-short plane wave pulses with bandwidths of the order of 1 GHz [Sullivan, 1992a; Gandhi, et al., 1993; Furse, et al., 1994]. Several of these techniques are briefly described below.

#### D5.1 Long-Wavelength Approximation.

In the frequency range where the length of the irradiated object is approximately twotenths or less of a free-space wavelength, *SAR* calculations are made by an approximation based on the first-order term of a power series expansion in g of the electric and magnetic fields, where g is the free-space propagation coefficient [D5]. This is called a perturbation method because it is based on the fact that the resulting fields are only a small change from the static fields. Equations for SAR have been derived for homogeneous spheroidal and ellipsoidal models of humans and animals [D10, D13].

#### D5.2 Extended-Boundary-Condition Method (EBCM).

The EBCM is a matrix formulation based on an integral equation and expansion of the EM-fields in spherical harmonics. This method was developed by Waterman [D21] and has subsequently been used to calculate the SAR in prolate spheroidal models of humans and animals [D1]. The EBCM is exact within the limits of numerical computation capabilities, but numerical problems presently limit the method to frequencies below about 80 MHz for prolate spheroidal models of humans. In SAR calculations for prolate spheroidal models of humans, the long-wavelength approximation and the EBCM give identical results up to about 30 MHz, where the long-wavelength approximation begins to become inaccurate.

#### D5.3 Iterative Extended-Boundary-Condition Method (IEBCM).

The EBCM has been extended [D11] to a technique called the IEBCM, that is capable of *SAR* calculations in prolate spheroidal models of man up to at least 400 MHz. The IEBCM is different from the EBCM in two main respects. It makes use of more than one spheroidal harmonic expansion, which allows better convergence for elongated bodies at higher frequencies, and it uses iteration, beginning with an approximate solution, to converge to the solution. These two features of the IEBCM have significantly extended the range of calculations over that of the EBCM.

#### D5.4 The Cylindrical Approximation.

The *SAR* calculated for an appropriately long section of an infinitely long cylinder is a good approximation to the *SAR* of spheroids in the frequency range where the wavelength is very short compared with the length of the spheroid. The lowest frequency at which the approximation is useful depends both on the length of the spheroid and on the ratio of the major axis to the minor axis. For man-sized spheroids, the lower frequency limit occurs for E-polarization when the wavelength is about four-tenths of the length of the spheroid [D14].

#### D5.5 Moment-Method Solution.

A moment-method solution of a Green's-function integral equation for the electric field has been used to calculate the internal electric field in block models, so-called because the mathematical cells of which the model is composed are cubes [D4, D6]. Wholebody average SARs calculated by this method are very close to those calculated for spheroidal models. Although the block model has the advantage that it resembles the human body better than a spheroid because it has simulated arms, legs, and head, the calculations of the spatial distribution of the internal fields made with the above method have been found to be of varying accuracy depending on the location of the cell in question [D15]. Apparently, the calculations are of limited accuracy because the electric field in each mathematical cell is approximated by a constant, and this approximate field cannot satisfy the boundary conditions for cells that touch the curved surface of interfaces between two dissimilar dielectric materials, or the surface of the body (airtissue boundary). An improved moment-method uses tetrahedra as mathematical cells [D18]. Special basis functions are defined within the tetrahedral volume elements to insure that the normal electric field satisfies the correct conditions at interfaces between different dielectric media.

Hagmann [D7] has developed improved means for predicting regionally averaged *SAR* using the block model of man, with the moment-method technique. Since individual

humans vary significantly in their physical size and configuration, regional averaging may be more meaningful than detailed values of SAR for individuals "blocks" within an idealized human model. These techniques insure convergence with no numerical instabilities, and also increase modeling flexibility.

#### D5.6 Surface-Integral-Equation (SIE) Technique.

An SIE method based on a formulation of the EM-field equations in terms of integrals over induced currents on the surface of an object [D8, D22] is used to calculate average *SAR*, principally for K-polarization, and mostly for models consisting of a truncated cylinder capped on each end by hemispheres. Average *SAR* for this model are close to those for a spheroid, depending on how the dimensions of the cylinder- hemispheres model are chosen relative to the spheroid.

#### D5.7 Finite Difference Time-Domain Method (FDTD).

FDTD [Yee, 1966; Kunz and Luebbers, 1993; Taflove, 1995] is a numerical method for the solution of electromagnetic field interaction problems. It uses a geometry mesh, usually of rectangular box-shaped cells (voxels), which is readily developed from CT or MRI scans of real human beings or animals. The constitutive parameters for each cell edge may be set independently so that objects having irregular geometries and inhomogeneous dielectric composition can be analyzed. The FDTD method has been used for a myriad of applications including calculating SARs and induced currents in the human body for plane-wave exposures [Gandhi, et al., 1992], exposure to leakage fields of parallel-plate dielectric heaters [Chen, et al., 1991], exposure to EMP [Chen and Gandhi, 1991], annular phased arrays of aperture and dipole antennas for hyperthermia [Chen and Gandhi, 1992], coupling of cellular telephones to the head [Gandhi, et al., 1996; Gandhi and Chen, 1995; Jensen and Rahmat-Samii, 1995; Dimbylow and Mann, 1994; Luebbers et al., 1992], exposure to RF magnetic fields in magnetic resonance imaging (MRI) machines [Gandhi, et al., 1994] and exposure to power lines [Gandhi and Chen, 1992].

FDTD calculations have been extensively validated for both far- and near field sources. For far-field sources, simulation results have been compared with analytical results for square [Umashankar and Taflove, 1982] and circular cylinders [Umashankar and Taflove, 1982; Furse, et al., 1990; Taflove and Brodwin, 1975; Borup, et al., 1987], spheres [Holland, et al., 1980; Gao and Gandhi, 1992; Gandhi and Chen, 1992], plates [Taflove, et al., 1985], layered half-spaces [Oristaglio and Hohmann, 1984] and even complicated geometries such as airplanes [Kunz and Lee, 1978]. Calculations of currents induced in a standing human have compared well with measurements [Gandhi and Chen, 1992; Furse, et al., (submitted to *Bioelectromagnetics*); Chen and Gandhi, 1989]. In the example shown in Figure D3,  $E_{inc} = 10$  kV/m (vertically polarized, frontally incident) and  $B_{inc} = 33.42 \,\mu\text{T}$  from side to side of the model. The data is in excellent agreement with the data of Deno [Deno, 1977] shown as a dot at the bottom of the curve.

For near-field sources, simulation results have been compared with analytical, measured or method of moment results for dipole antennas in front of layered half-spaces, layered boxes and homogeneous spheres [Furse and Gandhi, (submitted to WTR)]. In addition, the FDTD method has been validated for near-field testing of realistic cellular telephones next to the human head [Gandhi and Chen, 1995; Jensen and Rahmat-Samii, 1995; Dimbylow and Mann, 1994; Luebbers et al., 1992]. The analytical results are in excellent agreement with corresponding experimental results.

# Fig D3 The calculated vertical currents passing through the various sections of a $6 \times 6 \times 6$ mm MRI-based grounded model of the human body exposed to a 60 Hz electromagnetic field.

Depending on the application, human body models may be crude approximations or detailed meshes based on actual anatomy. Several suitable models have been developed from CT and MRI scans of humans. One of the first, which was taken from published anatomical cross sections, had a resolution of 6.55 mm [Gandhi and DeFord, 1988]. Models with resolutions of 2 x 2 x 2 mm [Dimbylow and Mann, 1994], 1.974 x 1.974 x 3 mm [Gandhi, 1995; Gandhi and Furse, 1995], 0.9 x 0.9 x 1.5 mm [Olley and Excell, 1995], 1.7 to 5 mm using subgridding in some regions [Stuchly, et al., 1995] and 3 x 3 x 3 mm, based on MRI scans [Jensen and Rahmat-Samii, 1995; Luebbers, 1996] have been reported. A popular source of anatomical data suitable as the basis of an FDTD biological mesh is the *Visible Human Project* [Visible Human Project] of the National Library of Medicine. Various types of data are available, with the most useful perhaps being the cross sections. These are 1 mm slices for the male and 0.33 mm slices for the female. Both have a cross-sectional resolution of 0.33 mm. FDTD meshing of these data still requires considerable effort, especially in assigning the colors of the slices to particular tissue types.

The Armstrong Laboratory, Radiofrequency Radiation Division, Brooks Air Force Base, has a database of various animals that it uses as input for FDTD calculations. This database consists of magnetic resonance imaging (MRI) scans of the Sprague-Dawley rat, Rhesus monkey, and pigmy goat. In addition, the U.S. Naval Medical Research Detachment at Brooks AFB has obtained MRI scans of a phantom Rhesus monkey that was filled with a solution containing TX-150, polyethylene powder, sodium chloride and water. The phantom was scanned in squatting and sitting positions. Each MRI scan was converted to a TIF image for use with Adobe Photoshop<sup>™</sup> software on the PC. All images for each animal were aligned so as to eliminate some of the scanning artifacts previously reported [Mason et al., 1995]. The phantom Rhesus monkey was assigned a single Red-Green-Blue (RGB) value, whereas in the other models, each tissue type was assigned a specific RGB color value as shown in Table D 1.

All 65,536 pixels on each image (81 images for the rat, 184 images for the monkey, and 276 images for the goat) were then "painted" manually with the RGB colors representing the appropriate tissue types. Atlases and skeletons of the rat, monkey and goat were used to identify the location of each tissue type [Barrett et al., 1994; Hopkins et al., 1972, 1973; Popesko et al., 1992]. Throughout the "painting" process, images were

reconstructed 3-dimensionally using Spyglass Slicer<sup>TM</sup> software on the PC and VoxBlast<sup>TM</sup> software on the Silicon Graphics workstation to ensure correct anatomical placement of each tissue type. Each image was also checked for unknown pixel values using Wave Advantag<sup>eTM</sup> software. To predict SAR values, the anatomical information within each TIF image file was converted to PGM files and imported into a FDTD mathematical model that used a look-up-table (LUT) containing permittivity values for each tissue type.

Development of models suitable for FDTD dosimetric calculations, while straight forward, is not trivial. MRI and CT scans provide voxel maps of density, but many of the tissues have the same or similar densities, and many of the regions outside of the major organs require a detailed understanding of anatomy to determine what tissue are present, e.g., fat, fluid, air. Even when some automatic tissue definition can be used, the bulk of the task falls on manual tissue recognition by a trained anatomist. Moreover, since FDTD calculations use the tissue dielectric properties, i.e., dielectric constant and conductivity, and mass density at each voxel location, knowledge of these properties for **Table D1 RGB colors for 42 tissue types** 

| Tissue Type           | RGB Value     | Tissue Type            | RGB Value     |
|-----------------------|---------------|------------------------|---------------|
| Air (External)        | 9, 24, 135    | Intestine (Large)      | 255, 0, 134   |
| Air (Internal)        | 0, 0, 0       | Intestine (Small)      | 0, 255, 254   |
| Bile                  | 105, 25, 132  | Kidneys                | 180, 0, 255   |
| Bladder               | 0, 128, 0     | Ligaments              | 50, 100, 200  |
| Blood                 | 0, 255, 146   | Liver                  | 73, 0, 255    |
| Blood Vessel          | 200, 100, 200 | Lung(Inner)            | 200, 112, 50  |
| Body Fluid            | 100, 255, 0   | Lung (Outer)           | 0, 117, 255   |
| Bone (Cancellous)     | 170, 148, 12  | Lymph                  | 230, 99, 74   |
| Bone (Cortical)       | 255, 0, 217   | Mucous Membrane        | 182, 255, 0   |
| Bone (Marrow)         | 128, 128, 128 | Muscle                 | 70. 255, 0    |
| Cartilage             | 255, 255, 128 | Nails (Toe and Finger) | 60, 200, 0    |
| Cerebral Spinal Fluid | 208, 255, 0   | Nerve (Spine)          | 0, 0, 255     |
| Eye (Cornea)          | 65, 63, 120   | Nerve (Brain)          | 245, 150, 160 |
| Eye (Lens)            | 30, 50, 70    | Pancreas               | 255, 0, 109   |
| Eye (Retina)          | 80, 100, 70   | Skin/Dermis            | 255, 0, 0     |
| Eye (Sclera/Wall)     | 255, 220, 0   | Spleen                 | 255, 82, 0    |
| Eye (Aqueous Humor)   | 159, 0, 255   | Stomach                | 140, 70, 20   |
| Fat                   | 255, 112, 0   | Testicles              | 113, 11, 11   |
| Gall Bladder          | 200, 60, 70   | Tooth                  | 10, 180, 80   |
| Glands                | 0, 87, 255    | Perfect Conductor      | 150, 125, 0   |
| Heart                 | 0, 128, 128   | 2/3 Muscle             | 180, 170, 160 |

each voxel is also required. Properties of many of the tissues has been measured over a wide frequency range [Gabriel, 1996; Stuchly and Stuchly, 1980; Geddes and Baker, 1967; Durney, et al., 1986]. All tissues are highly frequency dispersive, and some tissues, e.g., bone, heart, skeletal muscle, demonstrate anisotropic properties at frequencies below 1 MHz [Epstein and Foster, 1983]. Additional measurements of tissue properties are on-going as there remains questions about variations in tissue properties with individuals, age, health status, temperature, *in vivo* versus *in vitro*, etc.

The FDTD method solves Maxwell's differential equations at each cell edge at discrete time steps. Since no matrix solution is involved, electrically large geometries can be analyzed. FDTD solutions for three dimensional complex biological geometries involving millions of cells have become routine, e.g., studies of *SAR* distributions associated with exposure to hand-held radio transceivers [Gandhi, et al.,1996; Gandhi and Chen, 1995; Jensen and Rahmat-Samii, 1995; Dimbylow and Mann, 1994; Luebbers, et al., 1992]. FDTD may be used for both open region calculations, e.g., SAR and current distributions induced in the human body under plane-wave exposure conditions [Gandhi, et al., 1992], or closed regions, such as within a TEM cell. Commercial FDTD software is available from several sources, with some of these also offering FDTD meshes for human heads and bodies. These commercial packages provide a graphical user interface for viewing the FDTD mesh. Some provide interactive mesh editing while others allow for import of objects from CAD programs.

The actual FDTD calculations may be excited in different ways. Most commonly the electric fields on one or more mesh edges are determined by an analytical function of time, such as a gaussian pulse or sine wave. This then acts as a driven voltage source which, for example, may be used to excite an antenna such as a short monopole on a metal box to approximate a hand-held radio transceiver. This monopole antenna could be driven by a driven voltage source located on the mesh edge at the monopole base next to the top of the box. Both [Kunz and Luebbers, 1993] and [Taflove, 1995] describe methods for modeling RF sources. A variety of FDTD sources, including current sources, are described in [Picket-May, et al., 1994]. Alternatively a plane wave may be incident on the object as the excitation source.

The time variation of the excitation may be either pulsed or sinusoidal. The advantage of the pulse is that the response for a wide frequency range can be obtained. For accurate results, however, the frequency dependent behavior of the biological materials must be included in the calculations. Methods for doing this are well known [Kunz and Luebbers, 1993; Taflove, 1995] so that transient electromagnetic field amplitudes for pulse excitation can be calculated, e.g., (FD)<sup>2</sup>TD [Furse, et al., 1994]. When the results at a single frequency or at a few frequencies are all that is desired, sine wave excitation is preferred. This is especially true if results for the entire body are needed, such as the SAR distribution, since storing the transient results for the entire body mesh and applying Fast Fourier transformation (FFT) to calculate the SAR versus frequency requires extremely large amounts of computer storage.

**D5.7.1. Cell size and time-step size requirements.** The choice of cell size is critical in applying FDTD. It must be small enough to provide accurate results at the highest frequency of interest, and yet be large enough to keep resource requirements manageable. Cell size is directly affected by the dielectric properties of the materials present. The greater the permittivity and/or conductivity the shorter the wavelength at a given frequency and the smaller the cell size required. Once a cell size is selected, the maximum time step is determined by the Courant stability condition. After the cell size is determined, a problem space large enough to encompass the scattering object, plus space between the object and the outer absorbing boundary, is determined. From the number of Yee cells needed and the number of time steps required, resource requirements can be estimated (see D5.7.2).

The fundamental constraint is that the cell size must be much less than the smallest wavelength for which accurate results are required. An often guoted constraint is "10 cells per wavelength," meaning that the side of each cell should be 1/10 of the wavelength at the highest frequency (shortest wavelength) of interest. Since FDTD is a volumetric computational method, if some portion of the computational space is filled with penetrable material, one must use the wavelength in the material to determine the maximum cell size. For problems containing biological materials this results in cells in the material that are much smaller than if only free space and perfect conductors were being considered. For example, consider a human body meshed with 5 mm cubical cells. At 10 cells per free space wavelength this would correspond to a maximum frequency of 6 GHz. However, since the relative dielectric constant of biological tissues is high, the wavelength in the tissue is reduced (by the square root of the relative dielectric constant). For example, if the maximum relative dielectric constant of the biological tissue in the model is 49, the wavelength in that tissue and, hence, the maximum frequency would be reduced to about 857 MHz, i.e., 1/7 of the free space value. If results at higher frequencies are needed the cell size must be reduced. (See Figure D4 for the cell size requirement, as a function of frequency, for muscle tissue for I/4 and I/10 cell size criteria.)

# Fig D4 FDTD cell-size requirements as a function of frequency for muscle, a high water-content tissue.

Another cell size consideration is that the important characteristics of the problem geometry must be accurately modeled. This will normally be met automatically by making the cells smaller than *I*/10 unless some special geometry features smaller than this are factors in determining the response of interest. In some situations there is a specific region of the object where smaller FDTD cells are needed. For example, a region of high dielectric material, or fine geometry features such as eyes. But if uniform FDTD cells are used throughout for the computation, then these small cells must be used even in regions where they are not needed. One approach to reduce the total number of FDTD cells for these situations is to mesh local regions with smaller cells than in the main mesh [Kim and Hoefer, 1990; Zivanovic, et al., 1991]. Commercially available software has this local grid capability.

The other basic constraint on FDTD calculations is the time step size. For a threedimensional grid with cell edges of length  $\Delta x$ ,  $\Delta y$ ,  $\Delta z$ , with *v* the maximum velocity of propagation in any medium in the problem, usually the speed of light in free space, the time step size  $\Delta t$  is limited by

$$v\Delta t \le \frac{1}{\sqrt{\frac{1}{\left(\Delta x\right)^2 + \left(\Delta y\right)^2 + \left(\Delta z\right)^2}}}$$
(Eq D9)

**D5.7.2 Estimation of the computer resources required.** Given the shortest wavelength of interest, the cell dimensions are determined as 1/10 of this wavelength (or less if greater accuracy is required). From this and the physical size of the problem geometry the total number of cells in the problem space (denoted as *NC*) can be determined. Assume that the material information for each cell edge is stored in 1 byte (INTEGER\*1) arrays with only dielectric materials considered. An estimate of the computer storage required, in bytes (assuming single-precision FORTRAN field variables), can be obtained from

$$storage = NC \times \left( 6 \frac{components}{cell} \times 4 \frac{bytes}{component} + 3 \frac{edges}{cell} \times 1 \frac{byte}{edge} \right)$$
(Eq D10)

where components indicate the vector electric and magnetic field components. If magnetic materials are included, six edges must be considered for the material arrays. The relatively small number of auxiliary variables needed for the computation process have been ignored.

The computational cost in terms of the number of floating point operations required can be estimated from

Operations =  $6 \times (components / cell) \times (15 operations / component) \times N \dots (Eq D11)$ 

where 15 operations is the approximation based on experience and N is the total number of time steps. N is typically of the order of five to ten times the number of cells on one side of the problem space. N will be larger for resonant objects and smaller for lossy objects.

For a human body that fits into a box  $63 \times 36 \times 183$  cm, with a 15 cell border around the body to separate it from the outer boundary, the problem space is about  $160 \times 100 \times 400$  or 6.4 million cells. Using Eq D10, the computer random access memory (RAM) necessary to do this calculation is approximately 172 MBytes. Since this does not allow for storage of instructions and other arrays, and since the operating system will take some computer memory, a machine with about 256 MBytes of RAM should be sufficient.

A conservative estimate of the number of time steps needed is 10 times the longest dimension in cells, or 4000 time steps. Using the above equation an estimate of 2.3 x

1012 operations results. Typical MFLOPS (Million FLoating Point Operations per Second) ratings for computers are 80 for a Pentium PC or typical new work station, 120 for a fast work station, and several hundred for a super computer. If we use 300 MFLOPS for the super computer, then the calculation times for the human body are 8 hours for the PC or typical work station, 5 hours for the fast work station, and 2 hours for the super computer.

While the high frequency limitations of FDTD calculations are based on the size of the object in wavelengths, the low frequency limitations are usually determined by a combination of the geometry features and time step. For example, consider applying FDTD for a 60 Hz calculation for a human body. Based on the wavelength the FDTD cells could be huge, but then the body shape would be unrecognizable. If FDTD cells of 10 cm were chosen to make at least a crude body shape, the maximum time step would be  $1.92 \times 10^{-10}$  seconds. If calculations are to be made for at least 1 period of the sine wave in order to read some semblance of steady state, 86 million time steps would be required. This is not feasible on current computers, which illustrates the difficulty of using FDTD for extremely low frequencies. Other methods, such as finite elements, are preferred for these very low frequencies.

**D5.7.3.** Averaging volume for comparison with spatial peak SAR limits. Millimeter resolution and anatomically-based models of the head and torso of humans has, in some cases, led to questions of interpretation of the appropriate volume over which spatial peak SARs should be averaged. Modern safety standards and guidelines specify time-averaged whole-body-averaged SARs and spatial peak SARs, neither of which should be exceeded. The spatial peak SAR is usually averaged over a specified volume, e.g., 1 gram of tissue in the shape of a cube [IEEE, 1991]. In some near-field exposure situations, e.g., hand-held cellular telephones, the peak SAR values are generally observed at or near the surface of the ear, which is irregular in shape and is made up of skin, fat and cartilage with skull bone and brain behind this region. Finding a precise cube of *tissue* around the peak values is often not possible due to irregularities in shape. In addition, some tissues (particularly bone) are heavier than others, so even cubical subvolumes of the same size will have different mass, depending where in the head they are located.

Further complicating this issue is the fact that the FDTD voxel size is not always (in fact, rarely) divisible into exactly 1- gram cubes. While it is possible to use interpolation or extrapolation of data to reduce the problem, the difficulties of handling surface structure and heterogeneous tissue masses remains. Since the fields from near-field sources decay rapidly away from the source, the size of the volume and the percentage of tissue encompassed can have a significant effect on the results. This was demonstrated by [Gandhi, et al., 1996] who calculated the spatial peak SARs associated with hand-held cellular and personal communication services (PCS) transceivers using the FDTD method with a resolution of 1.974 x 1.974 x 3 mm. The results obtained using subvolumes of cells 5 x 5 x 4 were significantly different from the results obtained using 6 x 6 x3 cells to obtain subvolumes of 1 cm<sup>3</sup>, where each of the subvolumes selected were close to and around the regions of high SARs. The absorbed powers were divided by the mass calculated for the individual subvolumes to obtain the 1-gram SARs.

It is important to note that this difficulty in interpreting spatial peak SAR criteria in the context of a shaped, possibly heterogeneous model, is *not unique to numerical simulation or to FDTD simulations in particular*. Some measurement systems also

provide a discrete set of values, at a given distance apart (resolution), and hence, the calculation of 1-gram averaged SARs leads to the same difficulty, particularly if anatomically-detailed phantoms are used. If regularly-shaped models are used the problem of calculating the 1-gram SAR is lessened, but then the effects of the shape and properties of the ear and other features, which definitely affect the SAR distribution, are not considered.

#### D5.8 Generalized Multipole Technique (GMT).

During the 1980's several groups developed what were later unified under the name *generalized multipole technique* (GMT) [Ludwig, 1989]. GMT refers to methods which approximate the unknown field in each domain by several sets of functions which, in contrast to the method of moments, do not have singularities within their respective domains or their boundaries.

The expansions are matched at discrete points on the boundary of the domains resulting in an overdetermined system of equations with a dense matrix. The overdetermination factor is typically between 2 and 10. The system is solved in the least squares sense, usually with QR-factorization methods [Golub, et al., 1989].

Since the global expansion functions of the GMT are very smooth at the boundaries, the accuracy close to the boundaries is very high, which is important for dosimetry applications. The greatest advantage of the GMT, however, lies in the fact that the residual errors resulting from the least squares technique can be employed to validate the quality of the results [Kuster, 1992a]. Since the largest errors usually occur at the boundaries, the accuracy of the entire solution can be precisely determined [Regli, 1993]. The GMT, therefore, leads to very reliable dosimetric assessments. Since the method is closely related to other analytical methods, accurate simulation of scattering problems ranging over many orders of magnitude in fields strength are possible.

The severe limitation of the GMT is the difficulty involved in simulating real-world applications. In contrast to the method of moments, in which sequential basis functions are equivalent to a compact current, a GMT expansion is equivalent to a current distribution over the whole boundary of the domain. For geometrically complex bodies, the selection and location of the origin of the expansion functions is not quite obvious and requires considerable expertise.

The method is described in detail in [Hafner,1990]. Commercial software based on the GMT is commercially available, including a graphic interface for the PC. The code has been successfully applied to dosimetric studies [Kuster, 1992a, and Kuster, 1993], and to antenna design [Tay and Kuster, 1994].

#### D5.9 Impedance Method.

To obtain a detailed view of the power deposition pattern resulting from time-varying magnetic fields used in hyperthermia, a method of modeling portions of the human body using an impedance network has been developed [Gandhi, et al., 1984]. The region of interest is subdivided into a number of cells, each of which is then replaced by an equivalent impedance, and currents induced in the resulting network due to the prescribed magnetic filed are found by the application of circuit theory. This approach allows very fine modeling of inhomogeneities in the human body, with cell sizes of 0.5 cm or smaller possible. In addition, the individual cells are assumed to have anisotropic electrical properties, and this allows accurate modeling of interfaces.

# D6. Theoretical Considerations for the Determination of SAR Associated with Near-Field Exposures

The results of detailed mathematical analyses of the SAR distribution at different points within a person's body, due to exposure to near-field sources, aid in estimating the order of magnitude of the corresponding induced *SAR*. The corresponding exposure may or may not be in the reactive near field of a radiator. The following considerations can aid estimating the type of exposure (near- or far-field), and in the subsequent assessment of the feasibility of estimating the internal *SAR* distributions in exposed personnel.

# D6.1 Theoretical Considerations for Estimating RF Coupling and SAR Associated with Reactive Near-Field Exposure.

Either reduced or enhanced absorption situations exist (see 5.6) for personnel exposed to nonradiating, reactive near fields of a nearby source. One principal factor that determines the relative magnitude of the induced *SAR* is the type and degree of coupling between the RF source (an active radiator or passive radiator) and the exposed object. For the case where the distance between the object (person) and the source is much less than one wavelength, *E* and *H* exhibit a rapid decrease in amplitude with increasing separation distance.

The characteristics of the reactive near-field region are described in 3.1.4. For a small electric dipole antenna, the field strength at a point *r* can be expressed mathematically as follows [D17]:

$$E_{r} = \frac{I_{0}h}{4\pi} e^{-jkr} \left( \frac{2\eta}{d^{2}} + \frac{2}{j\omega\varepsilon d^{3}} \right) \cos\theta$$

$$E_{\theta} = \frac{I_{0}h}{4\pi} e^{-jkr} \left( \frac{j\omega\mu}{d} + \frac{1}{j\omega d^{3}} + \frac{\eta}{d^{2}} \right) \sin\theta \qquad (Eq D11)$$

$$H_{\phi} = \frac{I_{0}h}{4\pi} e^{-jkr} \left( \frac{jk}{d} + \frac{1}{d^{2}} \right) \sin\theta$$

where:

$$k = \frac{2\pi}{\lambda}$$
 (1/m)

 $h = \text{ impedance of free space (377 } \Omega)$ 

- e = permittivity of free space (F/m)
- m = permeability of free space (H/m)
- h =length of the dipole (m)
- $I_o$  = antenna current (A)
- w = angular frequency (radians/second)
- l = wavelength (m)
- d = distance from the center of the dipole to the location of interest (m)
- q = angle between the axis of the dipole and the unit direction vector from the center of the dipole to the point *r*.

When the distance *d* is much smaller than *I*, the term  $(1/d^3)$  is dominant. Here, reactive near-field conditions produce a majority of the energy deposition in lossy dielectric objects. The coupling in a reactive near-field situation may be analyzed as a quasi-static, electromagnetic field problem. Capacitive (electric field) coupling provides the primary means of inducing energy from the electric fields surrounding the radiator into the exposed object. Magnetic fields can also induce RF energy (via internal currents), thereby producing additional RF absorption.

#### D6.2 Theoretical Studies of Induced SAR: Near-Field vs Plane Wave Exposure.

For the near- field exposure situation, the induced spatially-averaged or whole-bodyaveraged *SAR* is almost always much less than the whole-body average *SAR* induced by a normalized plane-wave exposure. (See 5.6.2.2 for a discussion of the normalized exposure field concept.) The only exception is when direct contact with a radiator or reradiator induces a large RF current in the body which subsequently flows through the body to RF ground. The degree of coupling between the object (person) and the passive reradiator will dictate whether an enhanced or a reduced absorption situation occurs with respect to spatially localized SAR.

The induced *SAR* associated with a generalized model of an active, near-field radiation source, has been determined [D3]. Here, human exposure to a near-field source is compared with exposure to a plane-wave field. Both exposure situations were normalized to their respective spatial-maximum external E-fields for the reduced absorption case. In particular, mathematical analyses were performed using a block model of a person, and experimental verification was accomplished with human "phantom" models using implantable, E-field probes. The study involved a standing person immersed in a vertically polarized, spatially non-uniform E-field. The E-field distribution was defined as a half-cosine shape along both the vertical and horizontal axes. The field was further defined as existing in a lateral plane located just in front of the person's feet, simulating exposure to the emission from an RF heat-sealer. The plane containing the field was normal to an imaginary line drawn between the center of the radiator and the center of the person. Finally, the near field was assumed to decay rapidly with increasing distance from the source, so that a two-dimensional analysis was feasible.

The resulting induced, whole-body averaged SAR in the above studies was shown to be significantly lower than the whole-body SAR induced by a plane wave. The general relationship between the SAR induced by near-field exposures and far-field exposures is:

$$SAR_{n} = \frac{SAR_{f}}{\left[1 + \left(\frac{A_{v}}{d_{v}}\right)^{2}\right]\left[1 + \left(\frac{A_{h}}{d_{h}}\right)^{2}\right]}$$

(Eq D 12 )

where:

 $A_v = \text{constant}$ 

- $A_h = \text{constant}$
- $d_v$  = the dimension (in wavelengths) of the vertical extent of the field in the plane containing the object
- $d_h$  = the dimension (in wavelengths) of the horizontal extent of the field in the plane containing the object.

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The E-field in the abdominal region of the body was one third, or less, at the corresponding abdominal E-field induced by a plane-wave. This implies a decrease in local *SAR* by a factor of at least 9 times, for the near-field case. However, near-field exposure situations are capable of inducing high values of *SAR* in some parts of the body. At 27 MHz, the above study found an *SAR* in the legs that was 3 times higher than the whole-body-averaged *SAR*. It should be noted that the above studies were clearly reduced absorption cases, with weak coupling to the RF source.

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### Appendix E

#### Calibration of SAR-Measuring Temperature Probes

#### E1. General

For accurate determination of *SAR* with non-perturbing temperature probes, the precision with which small temperature changes can be measured must be optimized. Calibrations to assure maximum precision must be performed at regular time intervals, the length of which is dependent on the type of temperature probe to be calibrated. Some must be calibrated daily (especially certain fiber optic probes) whereas some can be calibrated monthly with minimal loss in accuracy.

#### **E2. Calibration Techniques**

Two approaches may be used in performing these calibrations. First, absolute calibrations can be performed at regular time intervals at several temperature points over the range of interest. This is accomplished by comparing the temperature probe being calibrated with a standard thermometer, usually a platinum or thermistor type. The platinum resistance thermometer has good long-term stability, but the thermistor type has a smaller sensing element and thus has a smaller thermal mass. This smaller thermal mass minimizes the creation of temperature gradients in the calibrating water bath and thermal lag problems caused by use of a large metallic temperature probe with its inherently large thermal mass and long time constant. If a dynamic temperature calibration technique is used, a well-stirred water bath should be used. Also, the rate of rise of the temperature of the water must be adjusted so that it is slower than the time constants of either of the probes. The unknown and the reference probes should be placed in intimate contact at their sensor-areas (usually the first few cm near the tip). It is best to use an automated data acquisition system to ensure reliable, instantaneous, and simultaneous readings from all temperature probes. If absolute calibrations of the unknown thermometer can be performed with accuracy approaching the resolution of the instrument being calibrated, this type of calibration is preferred. In some cases an instrument under calibration will not accurately indicate absolute temperature but will correctly measure a change in temperature  $\Delta T$ . In this case, the ratio  $(\Delta T$ -absolute)/ $(\Delta T$ measured) is determined over the range of interest. If this value is constant over the range of temperatures to be studied and over a reasonable period of time, e.g., from week to week, accurate SAR data can be obtained even when absolute temperature can not be precisely defined.

When using thermometers to measure *SAR* in electromagnetic fields, one must be aware of the possibility of RF interference in the thermometer's sensor, leads or electronics. Several methods can be used to determine the magnitude of RFI. One is to note changes at the instant that the RF power is switched on and off. If large, immediate changes are found, the probes should only be used to measure temperature prior to and just after RF exposure. Artifacts may also occur due to the interaction of the RF field with the electrical leads attached to the temperature sensing element of the probe. This interaction can excite a thermally induced voltage at the junction of two dissimilar materials (thermoelectric effect), including high resistance (carbon-loaded Teflon) leads joined to metallic wires. Since this phenomenon is due to the heating of a junction, shielding of these areas should be used to minimize this source of measurement error. Metal foil or RF absorber can be used to shield these junctions. Details of these problems and additional solutions may be found in Hochuli [E1].

With either the dynamic or static calibration technique, it is useful to perform two calibrations of  $\Delta T$ . The first was described above; a temperature rise in a water bath. The second calibration should be performed using a temperature decrease in a water bath. If a difference in the calibration factors exists for the unknown probe when determined by the two methods, the thermal time constant of one or both of the probes may have been longer than the time period that elapsed between the thermometer readings taken during one or both of the calibrations. To correct this, the calibration process should be repeated, using a slower rate of temperature increase or decrease in the bath.

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# Appendix F

#### **Power Density Curves**

#### F1. Normalized On-Axis Power Density Curves

Fig F1 Normalized On-Axis Power-Density Curves for Circular-Aperture (1-q)<sup>r</sup> Tapers

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