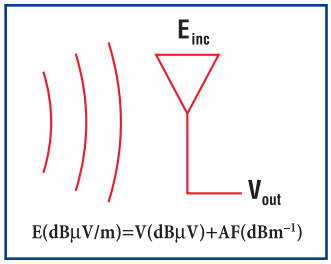
Common EMC Measurement Terms

According to Krause[1], "A *radio* antenna may be defined as the structure associated with the region of transition between a guided wave and free space, or vice versa". The EMC test community uses specialized versions of *radio* antennas for measuring electric and magnetic field amplitudes over a very broad frequency spectrum.

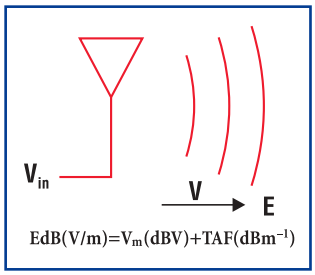
EMC test engineers must perform very precise measurements. Sometimes the terms used in describing these measurements can be confusing. This section describes the meaning of some of these terms.

Antenna Factor



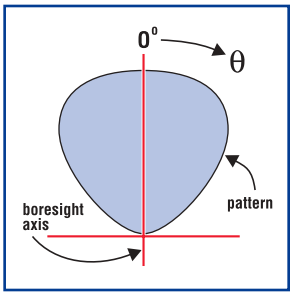
This is the parameter of an EMC antenna that is used in the calculations of field strength during radiated emissions measurement. It relates the voltage output of a measurement antenna to the value of the incident field producing that voltage. The units are volts output per volt/meter incident field or reciprocal meters. As can be seen in the derivations, the analytical expression for Antenna Factor (AF) has the equivalent of frequency in the numerator, thus AF will typically increase with increasing frequency. EMCO antennas used for radiated emissions testing are individually calibrated (the AF is directly measured) at all appropriate measurement distances. The calibrations produce values that are defined as the "equivalent free space antenna factor". This antenna factor is measured by EMCO using the three antenna measurement method over a ground plane. The calibration procedure corrects for the presence of the reflection of the antenna in the ground plane, giving the value that would be measured if the antenna were in "free space". The typical antennas used for measurements are broadband antennas such as BiConiLog™ and log periodic antennas. In case of any disagreement, a "reference antenna"—a tuned resonant dipole—is considered to be the arbiter for measurement purposes.

Transmit Antenna Factor



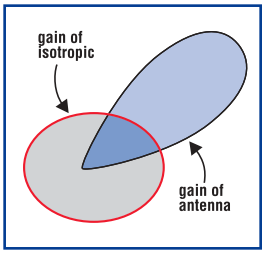
The Transmit Antenna Factor (TAF) is similar to the AF in that it describes the performance of an EMC antenna over its frequency range of operation. This parameter relates the E-field produced by an antenna, at a given distance, to the input voltage at the input terminals of the antenna. The units are volts/meter produced per volt of input, so the final units are reciprocal meters just as the AF. The AF and TAF are not the same nor are they reciprocal, though the AF and TAF can be computed from each other. The expressions for these computations may be found in the section "Antenna Terms and Calculations". The TAF is **not** a direct function of frequency, but it is a function of distance.

Antenna Pattern



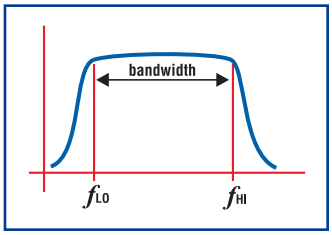
The antenna pattern is typically a polar plot of the relative response of an antenna as a function of viewing angle. The "on-axis" viewing angle where the response of the antenna is a maximum is called "the boresight axis". The pattern shows the response of the antenna as the viewing angle is varied. Typically, simple antennas exhibit a "dipole response" where the pattern of the antenna is donut shaped with the dipole on the common axis of the donut. More complex antenna patterns are approximately pear shape with the bottom of the pear facing away from the antenna.

Gain



The gain of the antenna is a parameter that describes the directional response of the antenna compared to an isotropic source, a theoretical antenna that radiates the same amount in all directions. The higher the gain, the better the antenna concentrates its beam in a specific direction. The simple example is a light bulb compared to a flashlight. The flashlight, with its reflector, concentrates the light in one direction, where the light bulb produces light in all directions.

Bandwidth



The bandwidth of an antenna is the operating frequency range of the antenna and is expressed in MHz.

Typical EMC antennas will have a ratio of upper useful frequency to lowest useful frequency on the order of 5 to 1. Some unique designs provide ratios of as much as 25 to 1. This ratio is a dimensionless ratio, i.e., it has no units.

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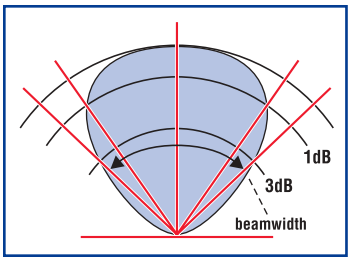
Beamwidth



This parameter is the descriptor of the viewing angle of the antenna, in the plane of measurement, typically where the response has fallen to one half the power received when the antenna is perfectly aligned. EMC antennas are normally designed to provide a viewing angle in the primary plane, approaching 60° between the half power points where the response of the incident is 3 dB down, if at all possible. The only specific requirements are found in IEC 801 - 3 and IEC 1000 - 4 - 3, which, if the analysis of the requirement is performed, translates into a minimum viewing angle of 28 ° (± 14 °) at the 1 dB down points. The 1 dB down response is usually compatible with the 3 dB down value of 60 °.

Reflection Coefficient

The voltage reflection coefficient, typically ρ, is the ratio of the voltage reflected from the load of a transmission line to the voltage imposed on the load of the same transmission line. Its values vary from zero to one. When the load impedance is "matched" to the source impedance (has the same value), and the characteristic impedance of the transmission line, then the reflection coefficient is quite small (no reflection), approaching zero. When there is "mismatch" (the load impedance differs from the characteristic impedance) the reflection coefficient can approach one (almost all incident power is reflected).

The reflection coefficient is usually determined by a measurement of the Voltage Standing Wave Ratio.
It is then computed from:

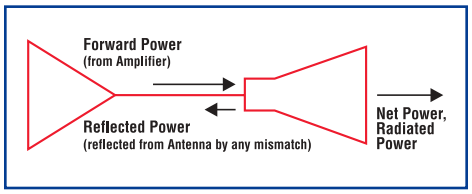
$$|\rho| = \frac{(VSWR-1)}{(VSWR+1)}$$

VSWR

The Voltage Standing Wave Ratio (VSWR) is a measure of the mismatch between the source and load impedances. Numerically, it is the ratio of the maximum value of the voltage measured on a transmission line divided by the minimum value.

When the value is high, most of the power delivered from a generator is reflected from the antenna (load) and returned to the generator. The amount of power not reflected is radiated from the antenna. This means that the generator rating for an antenna with a high VSWR can be quite high. Users should choose an antenna with a low VSWR when possible. As a practical matter, this may not always be possible, particularly at lower frequencies.

RF Power Terms As Applied To Antennas



There are several ways of discussing RF power as used to excite antennas for the generation of electromagnetic fields. This set of related definitions is provided to understand the definitions as used in this catalog.

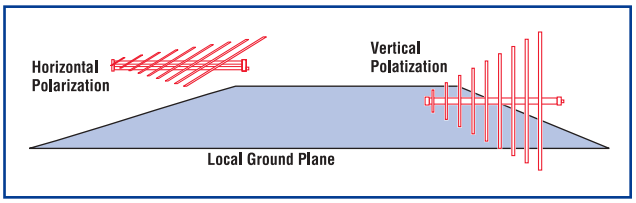
Forward Power — The output from an amplifier that is applied to an antenna input to generate an electromagnetic field.

Reflected Power — When mismatch exists at the antenna port, a fraction of the power applied (the forward power), is reflected from the antenna back toward the amplifier. This is termed the reflected power.

Net Power — The power applied to the antenna that is actually radiated is called the net power or radiated power. It is the difference between the forward power and the reflected power. Usually, this value cannot be directly measured, but is computed from the direct measurement of forward and reflected power by taking the difference:

$$P_{net}(W) = P_{forward}(W) - P_{reflected}(W)$$

In this catalog the Net Power is the power value that is required as an input to an antenna to generate a specific E-field level.



Polarization

The orientation of the measurement axis of a linearly polarized antenna with respect to the local ground plane. Vertical polarization occurs when the measurement axis of the antenna is perpendicular to the local ground plane. Horizontal polarization occurs when the measurement axis is parallel to the local ground plane. Most EMC test specifications require measurements in both vertical and horizontal polarizations of the measurement antenna.

[1] John D. Krause, Ph.D., Antennas, McGraw-Hill, New York, 1950, p. 1.

Antenna Calculations

1 Definition of Antenna Factor

This term is traditionally tied to the recieve antenna factor, the ratio of field strength at the location of the antenna to the output voltage across the load connected to the antenna.

.❶ $$AF = \frac{E}{V}$$

where:

AF = Antenna Factor, meters $^{-1}$
E = Field Strength, V/m or µV/m
V = Load Voltage, V or µV

Converting to dB (decibel) notation gives:

.❷ $$AF_{dB(m^{-1})} = 20 \log \left(\frac{E}{V}\right)$$

or:

.❸ $$AF_{dB(m^{-1})} = E_{dB(V/m)} - V_{dB(V)}$$

The antenna factor is directly computed from:

.❹ $$AF = \frac{9.73}{\lambda \sqrt{g}} \ (m^{-1})$$

where:

λ = Wavelength (meters)
g = Numeric Gain

In the same sense, for magnetic fields, as seen by loop antennas:

.❺ $$AF_{H\,dB(S/m)} = H_{dB(A/m)} - V_{dB(V)}$$

In terms of flux density (B-Field):

.❻ $$AF_B = AF_H + 20 \log (\mu)$$

.❼ $$AF_B = AF_H - 118, \ T / V$$

Loop antennas are sometimes calibrated in terms of equivalent electric field, where:

.❽ $$AF_{E\,dB(m^{-1})} = AF_{H\,dB(S/m)} + 20 \log \eta$$

.❾ $$AF_{E\,dB(m^{-1})} = AF_{H\,dB(S/m)} + 20 \log (120\pi), \text{ or}$$

$$= AF_{H\,dB(S/m)} + 51.5 \text{ dB}$$

where:

η = the Impedance of Free Space
= 120 π Ω

2 Conversion of Signal Levels from mW to µV in a 50Ω System

Voltage and power are equivalent methods of stating a signal level in a system where there is a constant impedance. Thus:

.❶ $$P = \frac{V^2}{R}$$

where:

P = Power in Watts
V = Voltage Level in Volts
R = Resistance Ω

For power in milliwatts (10^{-3} W), and voltage in microvolts (10^{-6} V),

.❷ $$V_{dB(\mu V)} = P_{dBm} + 107$$

3 Power Density to Field Strength

An alternative measure of field strength to electric field is power density:

.❶ $$P_d = \frac{E^2}{120\pi}$$

where:

E = Field Strength (V/m)
P = Power Density (W/m²)

Common Values:

E		P_D	
200	V/m	10.60	mW/cm²
100	V/m	2.65	mW/cm²
10	V/m	26.50	µW/cm²
1	V/m	0.265	µW/cm²

4 Power Density at a Point

.❶ $$P_d = \frac{P_t G_t}{4\pi r^2}$$

In the far field, where the electric and magnetic fields are related by the impedance of free space:

P_d = Power Density (W / m²)
P_t = Power Transmitted (W)
G_t = Gain of Transmitting Antenna
r = Distance from Antenna (meters)

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5

Friis Transmission Formula

The Friis Transmission formula describes Power received by an antenna in terms of power transmitted by another antenna:

.❶ $$P_r = \frac{P_t G_t G_r \lambda^2}{(4\pi r)^2}$$

where:

P_r = Power Received (W)

P_t = Power Transmitted (W)

G_r = Numeric Gain of Receiving Antenna

G_t = Numeric Gain of Transmitting Antenna

r = Separation Between Antennas (meters)

λ = Wavelength (meters).

6

Electric Field vs Power Transmitted (Far Field)

The electric field strength at a distance from a transmitting antenna such that the electric and magnetic field values are related by the impedance of free space is:

.❶ $$E_{V/m} = \frac{\sqrt{30 P_t G_t}}{r}$$

where the terms are as defined above.

For simple radiating devices having low gain, far field conditions exist when:

.❷ $$r \geq \frac{\lambda}{2\pi}$$

where:

λ = Wavelength (meters)

For more complex antennas having higher gain values, far field conditions exist when:

.❸ $$r \geq \frac{2D^2}{\lambda}$$

where:

D = Maximum Dimension of the Antenna (m)

7

Relationship of Antenna Factor and Gain in a 50Ω System

.❶ $$G_{dB} = 20 \log \left(f_{MHz}\right) - AF_{dB(m^{-1})} - 29.79$$

8

Power Required to Generate a Desired Field Strength at a Given Distance when Antenna Factors are Known

.❶ $$P_{dB(W)} = 20 \log_{10} \left(E_{desired\ (V/m)}\right) + 20 \log_{10} \left(d_m\right) - 20 \log_{10} \left(f_{MHz}\right) + AF_{dB(m^{-1})} + 15$$

9

Relationship Between Frequency and Wavelength in Free Space

.❶ $$f\lambda = c$$

where:

f = Frequency (Hz)

λ = Wavelength (meters)

c = Velocity of Light

= 3×10^8 m/s

a simpler relationship is:

.❷ $$\lambda = \frac{300}{f_{MHz}}$$

10

Decibel Formulas

A decibel is one tenth of a Bel, and is a ratio measure of relative amplitude. In terms of power, the number of decibels is ten times the logarithm to the base 10 of the ratio.

In terms of power:

.❶ $$dB = 10 \log_{10} \left(P_1 / P_2\right)$$

where P_1 and P_2 are in watts.

In a constant impedance system, power references can be made between different measurement points. They can also be related to voltage or current measurements:

.❷

$$\text{Power Ratio} = 10 \log_{10} \left(P_1 / P_2 \right)$$

$$= 10 \log_{10} \left(\frac{V_1^2 / R_1}{V_2^2 / R_2} \right)$$

$$= 20 \log_{10} \left(\frac{V_1}{V_2} \right)$$

$$- 10 \log_{10} \left(\frac{R_1}{R_2} \right)$$

for a constant impedance system.

.❸ $R_1 = R_2$

and:

.❹

$$\text{Pwr Ratio (dB)} = 20 \log_{10} \left(\frac{V_1}{V_2} \right)$$

Also:

.❺ $G_{dB} = 10 \log \left(g \right)$

.❻ $V_{dB(reference)} = 20 \log \left(V / V_{reference} \right)$

.❼ $P_{dB(reference)} = 10 \log \left(P / P_{reference} \right)$

where a typical reference for voltage is microvolts and a typical reference for power is milliwatts.

The reverse relationships are:

.❽ $g = 10^{G_{dB}/10}$

.❾ $V = 10^{V_{dB(reference)}/20}$

.❿ $P = 10^{P_{dB(reference)}/10}$

11 Transmit Antenna Factor

The transmit antenna factor of an antenna is computed from the gain or receive antenna factor, and is a measure of the transmitting capabilities of that antenna. It is valid under the conditions of measurement of the receive antenna factor, in a 50 Ω system.

.❶ $TAF_{dB(m^{-1})} = G_{dB} - 2.22 - 20 \log_{10} \left(d_m \right)$

Where:

$TAF_{dB(m^{-1})}$ = Transmit Antenna Factor

G_{dB} = Antenna Gain of Transmitting Antenna

d_m = Distance (m)

Alternatively:

.❷ $TAF_{dB} = 20 \log \left(f_{MHz} \right) - AF_{dBm^{-1}}$

$- 32.0 - 20 \log \left(r_m \right)$

12 Computing Power Required for a Specific Field Intensity Given Power Required to Generate 1 Volt/meter

Antenna transmitting capabilities are often given in terms of the input power to an antenna to generate 1V/m at one or more distances. The input power required to develop a different electric field level value is found by:

.❶ $P_{dB(w)} = P_{dB(w)\,(1\,V/m)} + 20 \log_{10} \left(E_{desired\,V/m} \right)$

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SI Prefixes, Multipliers, & Abbreviations

PREFIX	MULTIPLIER	SYMBOL
pico	10^{-12}	p
nano	10^{-9}	n
micro	10^{-6}	μ
milli	10^{-3}	m
kilo	10^{3}	k
Mega	10^{6}	M
Giga	10^{9}	G

SI Base Units

QUANTITY	NAME	SYMBOL
length	meter	m
mass	kilogram	kg
time	second	s
electric current	ampere	A

SI Prefixes, Multipliers, & Abbreviations

CONSTANT	COMPUTATIONAL VALUE
Speed of light in a vacuum:	$c = 2.99792458 \times 10^{8}$ m/s $\approx 3.00 \times 10^{8}$ m/s
Permittivity constant:	$\epsilon_0 = 1/(\mu c^2)$ F/m $\approx 8.85 \times 10^{-12}$ F/m
Permeability constant:	$\mu_0 = 4\pi \times 10^{-7}$ H/m $\approx 1.26 \times 10^{-6}$ H/m

SI Derived Units

QUANTITY	NAME	SYMBOL	
area	square meter	m^2	
volume	cubic meter	m^3	
frequency	hertz	Hz	s^{-1}
mass density	kilogram per cubic meter	kg/m^3	
speed, velocity	meter per second	m/s	
angular velocity	radian per second	rad/s	
acceleration	meter per second squared	m/s^2	
angular acceleration	radian per second squared	rad/s^2	
force	newton	N	$kg \cdot m/s^2$
pressure	pascal	Pa	N/m^2
work, quantity of heat	joule	J	N•m
power	watt	W	J/s
quantity of electricity	coulomb	C	A•s
potential difference	volt	V	W/A
electric field strength	volt per meter	V/m	
electric resistance	ohm	Ω	V/A
capaticitance	farad	F	A•s/V
magnetic flux	weber	Wb	V•s
inductance	henry	H	V•s/A
magnetic flux density	tesla	T	Wb/m^2
magnetic field strength	ampere per meter	A/m	
admittance	siemens	S	1/Ω
electricantenna factor	per meter	1/m	
magnetic antenna factor	siemens per meter	s/m	

Conversion Table for Magnetic Units

		TO CONVERT FROM:				
	Unit System:	SI (MKS)		CGS		
	Magnetic Qty.:	B	H	B	H	B
	Units:	**tesla**	**amp-turn/m**	**gauss**	**oersted**	**gamma**
TO:	**tesla**	1	$4\pi \times 10^{-7\dagger}$	10^{-4}	$10^{-4\dagger}$	10^{-9}
	amp-turn/m	$7.96 \times 10^{5\ddagger}$	1	$79.57747^{\ddagger}$	79.57747	$7.96 \times 10^{-4\ddagger}$
	gauss	10^{4}	$4\pi \times 10^{-3\dagger}$	1	$1^{\dagger}$	10^{-5}
	oersted	$10^{4\ddagger}$	$4\pi \times 10^{-3}$	$1^{\ddagger}$	1	$10^{-5\ddagger}$
	gamma	10^{9}	$4\pi \times 10^{2\dagger}$	10^{5}	$10^{5\dagger}$	1
		MULTIPLY BY ABOVE VALUE				

† Assumes $\mu = 1$; if $\mu \neq 1$, multiply by value of μ to convert from *H* to *B*.
‡ Assumes $\mu = 1$; if $\mu \neq 1$, divide by value of μ to convert from *B* to *H*.
1 tesla ≡ 1 weber/m^2.

For example, 1 tesla = 10^4 gauss.
1 gauss = 79.6 ampere-turns/m in an unloaded coil ($\mu = 1$).
If $\mu = 2.50$, 1 tesla = $7.96 \times 10^5 / 2.50 = 3.18 \times 10^5$ amp-turns/m.

Understanding Radiated Emissions Testing

In a radiated emissions test, electromagnetic emissions emanating from the equipment under test (EUT) are measured. The purpose of the test is to verify the EUT's ability to remain below specified electromagnetic emissions levels during operation. A receive antenna is located either 3 or 10 meters from the EUT. In accordance with ANSI C63.4, the receive antenna must scan from 1 to 4 meters in height. The scanning helps to locate the EUT's worst-case emissions level.

Figure 1 shows a block diagram of an emissions test system such as might be used for ANSI C63.4 testing. The test set-up is composed of a receive antenna, a first interconnecting cable, a preamplifier, a second interconnecting cable, and a radio noise meter (receiver or spectrum analyzer).

The purpose of each of the components of the radiated emissions test setup are:

The Receive Antenna

The performance measure of this antenna in relating the value of the incident E-field to the voltage output of the antenna is the Antenna Factor. This is usually provided by the manufacturer in dB with units of inverse meters. A variety of antennas can be used for these measurements. Typically, a combination of two antennas is used to cover the frequency range from 30 MHz to 1000 MHz—a biconical covering the frequency range of 30 to 200 MHz and a log periodic covering the frequency range of 200 to 1000 MHz. New antenna technology, such as EMCO's BiConiLog™ antenna, can cover the complete frequency range. This Antenna Factor is shown at A.

The First Interconnecting Cable

This cable connects the antenna output to the preamplifier input. There is a reduction in measured signal amplitude due to losses in the cable. To increase accuracy, these losses need to be added to the measured value of the voltage out of the antenna to compensate for the losses. Cable loss is shown at B.

The Preamplifier

The preamplifier is typically used with spectrum analyzers to compensate for the high input noise figure typical of such devices. Receivers may not need this device. The amplifier makes the measured signal larger, thus the final answer must be corrected by subtracting the gain of the preamplifier. Preamplifier gain is shown at C.

The Second Interconnecting Cable

This cable connects the output of the preamplifier to the radio noise meter. There is a reduction in measured signal amplitude due to losses in the cable. To increase accuracy, these losses need to be added to the measured value of the voltage out of the antenna to compensate for the losses. Cable loss is shown at D.

The Radio Noise Meter

Typically, the radio noise meter is either a receiver or a spectrum analyzer. Either is essentially a 120 kHz bandwidth, tunable, RF microVolt meter calibrated in dB µV. A signal response is shown in Figure E.

The calculation of the measured E-field signal level is then given by:

$$E(dB\mu V/m) = V(dB\mu V) + CL_1(dB) - PAG(dB) + CL_2(dB) + AF(dBm^{-1})$$

where:

$E(dB\mu V/m)$	=	Measured E-field
$V(dB\mu V)$	=	Radio Noise Meter Value
$CL_1(dB)$	=	Loss in Cable 1
$PAG(dB)$	=	Preamplifier gain
$CL_2(dB)$	=	Loss in Cable 2
$AF(dBm^{-1})$	=	Antenna Factor

This computed value can be compared to the published specification limit for determination of whether the measured value is less than the specification limit, thus showing compliance with the requirement.

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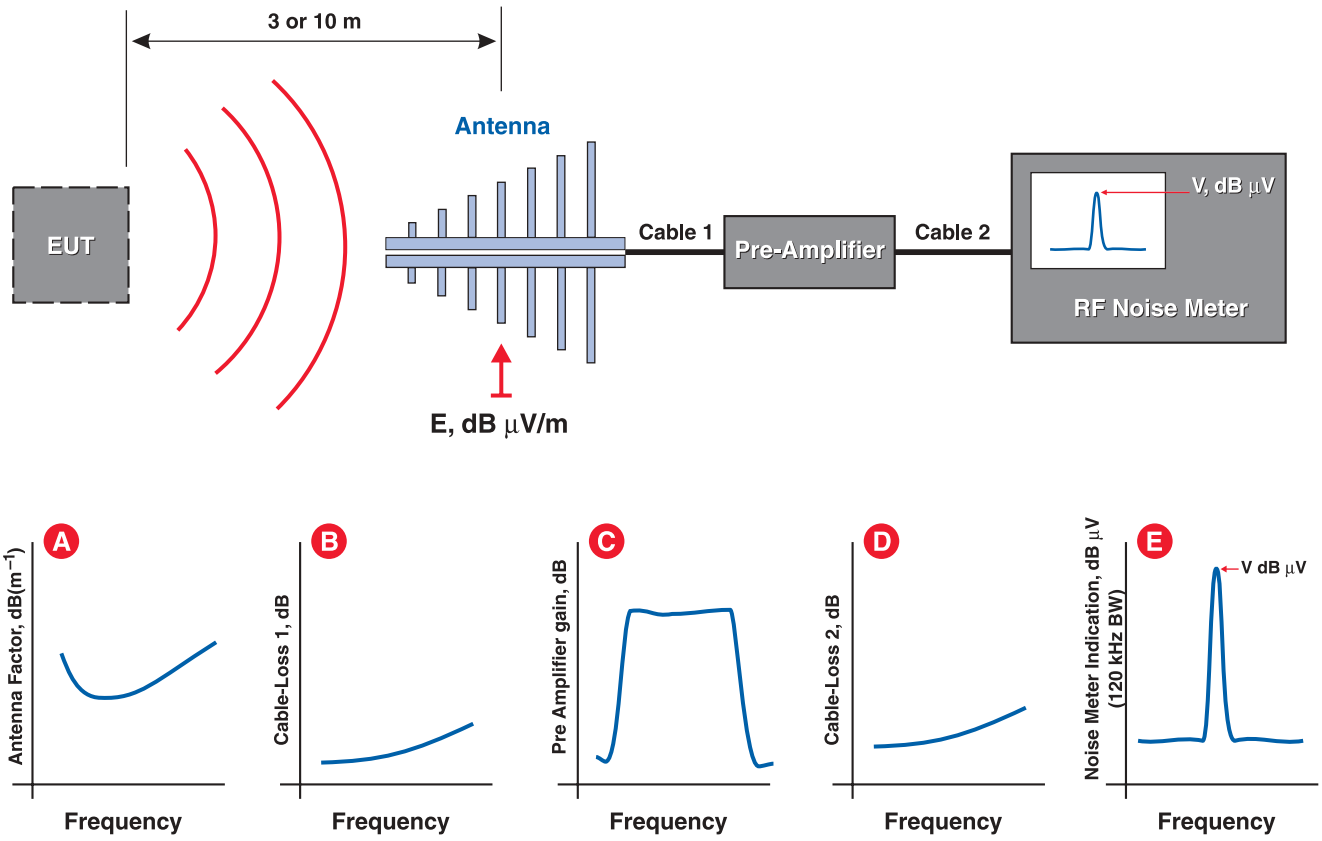

Figure 1. Radiated Emissions Test

$$E(dB\,\mu\,V/m) = V(dB\,\mu\,V) + CL_1(dB) - PAG(dB) + CL_2(dB) + AF(dBm^{-1})$$

Figure 1. Radiated Emissions Test

Notes on Figure 1

A is the Antenna Factor (AF) versus frequency. The units are meters^{-1}. The AF relates the RF voltage output of an antenna to the E-Field causing the voltage to appear.

B, D are the reduction in signal that is caused by losses in the interconnecting coaxial cables.

C is the low noise figure preamplifier gain, which can vary with frequency.

E is the value measured at some frequency by the 120 kHz bandwidth radio noise meter.

Understanding Radiated Immunity Testing

In a radiated immunity test, a test signal of RF energy, typically three or ten volts per meter, is directed at the equipment under test (EUT), and the EUT's reaction to this test signal is analyzed. The purpose of the test is to demonstrate the ability of the EUT to withstand the excitation of the signal, without showing degraded performance or failure. The more immune a product is to this test signal, the better it should operate when other electronic or electrical equipment is present in its environment.

Figure 1 shows a block diagram of an immunity test system such as might be used for IEC 61000-4-3 testing. The test setup is composed of a signal generator, an amplifier, a forward/reverse power coupler with its associated power meter, a radiating antenna, and an omni-directional E-field probe system.

The purpose of each of the components of the radiated immunity test setup is:

The Signal Generator

The signal generator is used to provide the test signal. It should have adequate output resolution to allow precise setting of the reference level of the E-field to within 1 % of the desired level. The signal generator must be capable of providing the desired 80 % AM with a 1 kHz sine wave for testing. A typical signal generator output is shown at A.

The Amplifier

The amplifier increases the test signal strength to levels, that when applied to the antenna, will produce the desired E-field levels. Note that EMC test amplifiers are specified with a minimum gain. Due to the extremely wide bandwidth, they can show ripple of several dB in the pass band. The amplifier must be operated in a linear mode to assure repeatability. A typical amplifier response is shown at B.

The Forward / Reverse Power Coupler

The forward / reverse power coupler is placed in-line with the amplifier output to the antenna input, as near to the antenna as practical. The difference in the forward and reverse power (the net power) is recorded to determine the input level necessary for developing the desired test signal, and to show that this desired input to the antenna is developed during testing. This is shown at C.

The Antenna

The antenna generates the desired E-field. Its performance in generating the E-field is given by the Transmit Antenna Factor (TAF), as shown at D.

The Omni-directional Probe System

The probe system is used to directly measure the value of the field strength, at E.

Figure 2 shows a graphical display of the signals of the immunity test system. This figure also includes computations of the signal levels at a specific frequency of 100 MHz.

The output level is given by:

$$E(dB\,\mu\,V/m) = SG_{out}(dB\,\mu\,V) + AG(dB) + TAF(dB)\,m^{-1})$$

where:

$E(dB\,\mu\,V/m)$	=	The E-field test level
$SG_{out}(dB\,\mu\,V)$	=	The signal generator output
$AG(dB)$	=	Amplifier gain
$TAF(dB)\,m^{-1})$	=	The transmit antenna factor

The variables and terms in the expression above are used for calibration test setups. They demonstrate how instrumentation and facility factors contribute to meet the typically required E-field uniformity value of – 0.0 dB, + 6.0 dB. Remember that actual testing to demonstrate that the EUT will not malfunction when exposed to the desired level requires the addition of 80 % amplitude modulation with a 1 kHz sine wave to the test signal. This, in turn, requires an additional 5.1 dB $\{10 \times \log_{10}[(1.8)^2]\}$ of linear gain from the amplifier than is found during calibration.

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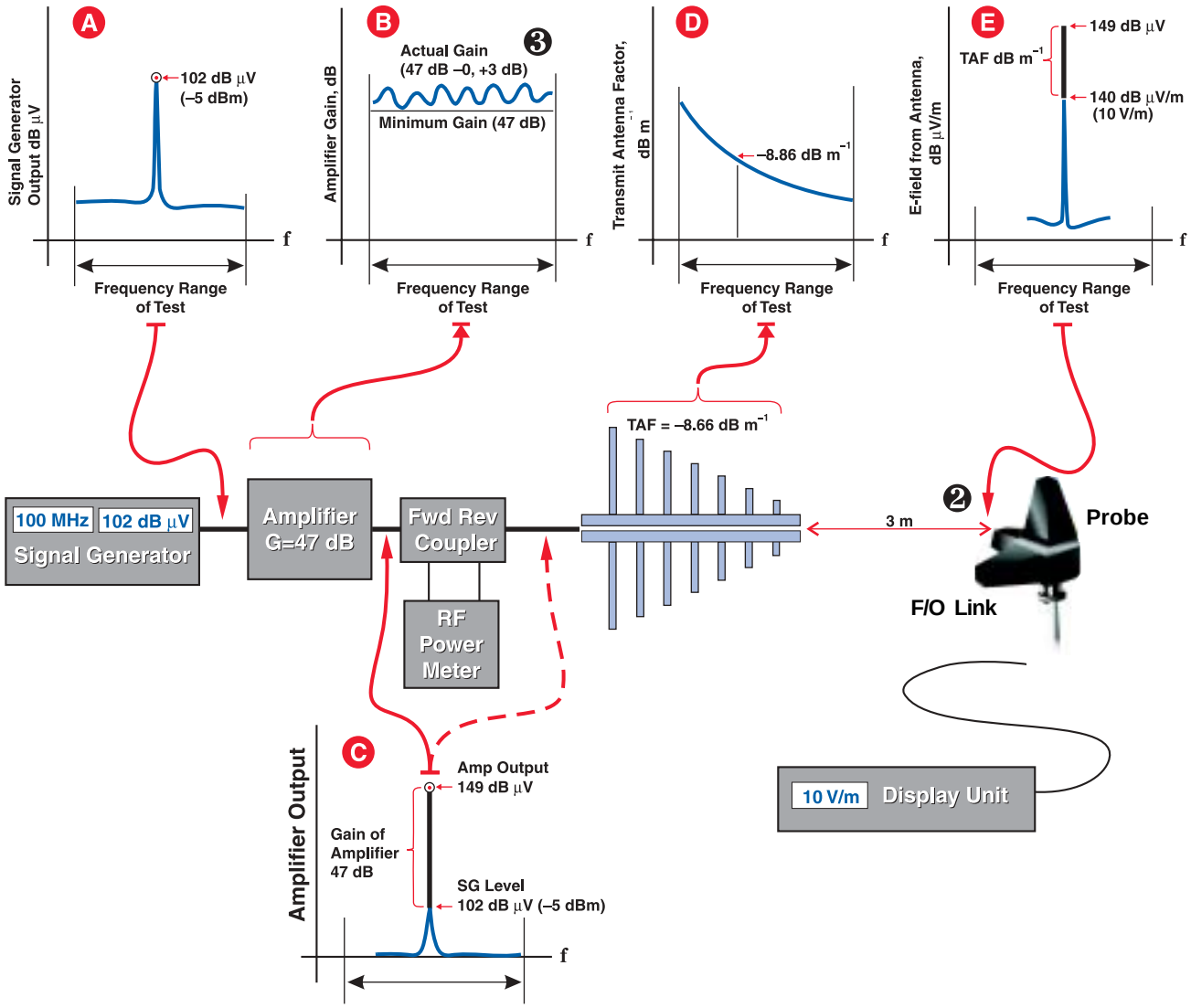

Figure 1. Block Diagram of Typical Immunity Test Setup with Signal Levels & Characteristics Added

Figure 1. Radiated Immunity Test

Notes on Figure 1

❶

A is signal generator output.

B is amplfier gain versus frequency.

C is signal generator output + amplifier gain (= input to antenna.

D is transmit antenna factor (TAF).

E is E-field level generated (= input to antenna + (TAF).

❷ For immunity testing, test distance is from tip of antenna.

❸ Immunity amplifiers are typically specified at minimum gain. Pass band ripple is result of extra wide bandwidths.

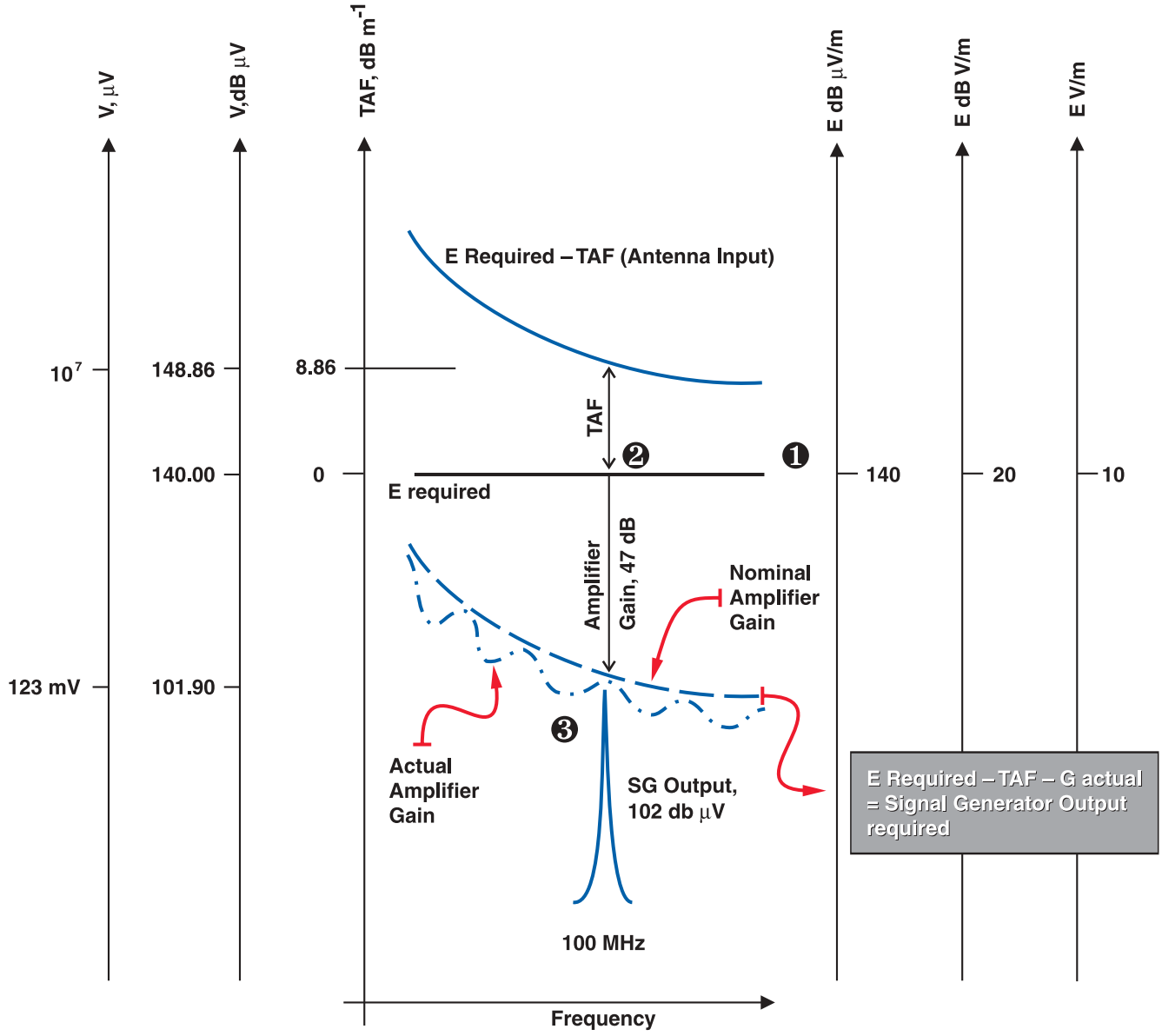

Figure 2. Graphical Representation of Immunity Test System Signal

Figure 2. Radiated Immunity Test

Notes on Figure 2

❶ Required test field is
10 V/m (= 20 dB V/m = 140 dB µ V/m).

❷ At 100 MHz TAF is – 8.66 dB => V_{in} to Antenna is
140 – (–8.66) = 148.66 dB µV.

❸ At 100 MHz Signal Generator Output is
Antenna Input – Amplifier gain
(= 148.86 – 47 = 101.86 dB µV = 123 µV).

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EMC Antenna Parameters and Their Relationships

JOHN D. M. OSBURN

INTRODUCTION

The basics of the EMC profession often get buried under the day-to-day effort of continuous measurement and the volume of test and reporting paperwork. The fundamental parameter of the most common of technical tools, the EMC antenna, is used over and over without thought as to its actual meaning. This parameter is the antenna factor (AF). A review of the basics behind this parameter, and a related parameter, the transmit antenna factor (TAF), provides a basis for the use of the numerical values, and a more fundamental understanding of radiated EMC measurements.

EMC ANTENNAS

EMC antennas are used for EMC measurements in rather rugged environments involving frequent handling, rapid replacement with a different antenna for another frequency band and the normal wear and tear of day-in, day-out usage, two shifts a day, six days a week, in almost all weather conditions.

For all their apparent simplicity, antennas used in an electromagnetic compatibility (EMC) laboratory are as specialized and as sophisticated as antennas for any other application. These antennas are different in that their application makes broad bandwidths the most important design parameter, and gain, efficiency, and low input voltage standing wave ratio (VSWR) become secondary. Broad bandwidths are driven by the broad frequency spectrum covered in the performance of EMC radiated emission and immunity measurements.

The concepts of antenna factor and transmit antenna factor are integral to understanding radiated EMC measurements.

The antenna parameters that are familiar to most antenna designers are then secondary design objectives in the development of these antennas. For all the importance of the bandwidth, however, the antenna parameter most often used, the AF, does relate to the performance of the antenna. The AF is used to quantify the value of incident electric fields, and its companion parameter, the TAF, is used to determine the value of the electric field at a known distance from the generating antenna.

EMC antennas are used for two types of measurements: radiated emissions (RE) and radiated immunity (RI). In the first case, formal calibration of the antenna and the use of traceable standards are required. In the second case, calibration is not required as the calibrations are usually performed as part of a complete EMC test setup.

Each of these types of measurements employs a separate descriptive parameter. For radiated emissions measurements, the parameter is the AF. For radiated immunity or susceptibility measurements, it is the TAF.

These two parameters are illustrated in Figure 1. This figure also illustrates other relationships between parameters used in the following derivations.

ANTENNA FACTOR

The antenna factor is the term applied in radiated emissions testing to convert a voltage level fed by a transducer to the input terminals of an EMI analyzer into the field-strength units of the electromagnetic field producing that voltage.[1] It relates the value of the incident electric or electromagnetic field to the voltage at the output of the antenna. For an electric field antenna, this is expressed as

$$AF = \frac{E}{V_L} \qquad (1)$$

where

AF = antenna factor, m^{-1}
E = electric field, V/m
V_L = voltage at antenna terminals, V

The AF is usually expressed in dB and when used to determine the value of an incident electric field, the expression is

$$E_{[dB(\mu V/m)]} = V_{[dB(\mu V)]} + AF_{[dB(m^{-1})]} \qquad (2)$$

The derivation of the AF is straightforward and is based on several fundamental relationships in antenna theory. The relationships can be stated as: *"The ratio of power in the terminating resistance to the power density of the incident wave is defined as the effective aperture."*[2]

Thus

$$A_e = \frac{P_{out}}{P_d} \qquad (3)$$

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where
A_e = effective receiving antenna, m^2
P_{out} = power delivered by antenna, W
P_d = power density of the incident wave, W/m^2

Following this, then

$$P_{out} = P_d A_e \quad (4)$$

In addition, from the ITT Handbook[3]

$$A_e = \frac{G_r \lambda^2}{4\pi} \quad (5)$$

where:
G_r = gain of the receiving antenna
λ = wavelength, m

The output voltage from the antenna V_L and the output power are related by the impedance seen by the antenna.

$$P_{out} = \frac{V_L^2}{Z} \quad (6)$$

where
P_{out} = power delivered at the output of the antenna, W
V_L = output voltage, V
Z = load impedance of the device connected to the antenna, Ω

Finally, the relationship between electric field strength and power density of the incident and the electric field strength is

$$P_d = \frac{E^2}{120\pi} \quad (7)$$

where
P_d = power density of the incident wave, W/m^2
E = electric field, V/m

Substituting Equations (4), (5) and (6) into Equation (7), gives, for a plane wave

$$\frac{V_L^2}{Z} = \frac{E^2}{120\pi} \times \frac{G_r \lambda^2}{4\pi} \quad (8)$$

Solving for the AF gives

$$AF = \frac{E}{V_L} = \sqrt{\frac{480\pi^2}{Z\lambda^2 G_r}} \quad (9)$$

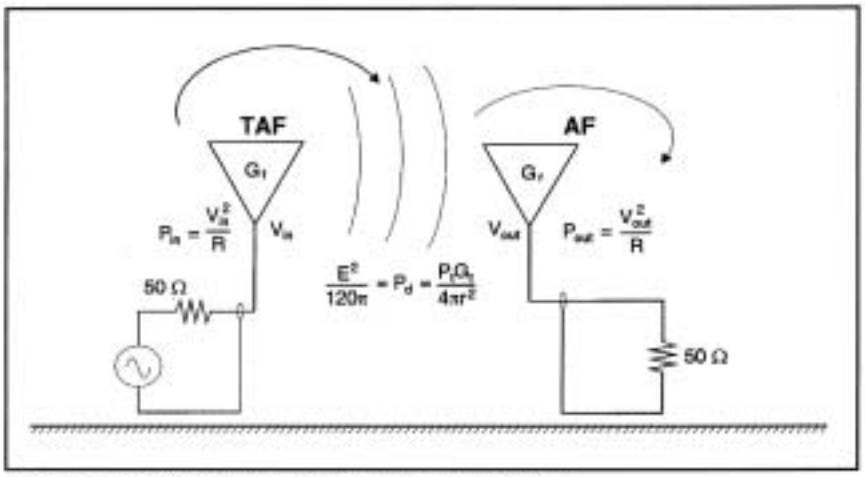

***Figure 1**. The Relationship Between Antenna Parameters.*

In a 50-ohm system this becomes

$$AF = \frac{9.73}{\lambda\sqrt{G_r}} \quad (10)$$

In dBs, this becomes (in units of inverse meters)

$$AF = 19.8 - 20 \times \log(\lambda) - 10 \times \log(G_r) \quad (11)$$

TRANSMIT ANTENNA FACTOR

The TAF provides a means of computation of the input voltage to the antenna to provide a given value of electric or electromagnetic field at a stated distance from the antenna. The transmit antenna factor relates the value of the electric or electromagnetic field generated by an antenna as a function of its input. Thus, the fundamental relationship is

$$TAF = \frac{E_{(dBV/m)}}{V_{(dBV)}} \quad (12)$$

The transmit antenna factor is, then, expressed in terms of dB

$$E_{[dB(V/m)]} = V_{in[dB(V)]} + TAF_{[dB(m^{-1})]} \quad (13)$$

or:

$$V_{in[dB(V)]} = E_{[dB(V/m)]} - TAF_{[dB(m^{-1})]} \quad (14)$$

Derivation of the TAF proceeds from three standard relationships. The first is a variation of the Friss transmission formula[4]

$$P_d = \frac{P_t G_t}{4\pi r^2} \quad (15)$$

where
P_d = radiated power density at distance r from the antenna, W/m^2
P_t = power input to the antenna, W
G_t = numerical gain of the antenna
r = distance from the antenna where the power density is evaluated, m

The second is Ohm's Law[5]

$$P = \frac{V^2}{R} \quad (16)$$

where
P = power dissipated in a load, w
V = voltage across the dissipating element, V
R = resistance (impedance) of dissipating element or load, Ω

The third relationship is Ohm's Law for Free Space[6]

$$P_d = \frac{E^2}{\eta} = \frac{E^2}{120\pi} \quad (17)$$

where
P_d = power density of the incident wave, W/m^2
E = electric field strength at that point in space, V/m
η = impedance of free space, $120\pi\Omega = 377\ \Omega$

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Combining Equations (15) and (16) leads to a familiar expression

$$E = \frac{1}{r}\sqrt{30\,P_t G_t} \qquad (18)$$

which relates the electric field strength at a point r away from the transmitting antenna having input power P_t and gain G_t.

By rearranging Equation (16) we have

$$V_{in} = \sqrt{P_t R} \qquad (19)$$

Recalling the definition of the transmit antenna factor, the ratio of the E-field developed to the input voltage to the antenna, we can find the TAF by taking the ratio of the E-field produced from Equation (17) to the power dissipated in the antenna given in Equation (18)

$$TAF = \frac{E}{V} = \frac{\frac{1}{r}\sqrt{30\,P_t G_t}}{\sqrt{P_{in} R}} \qquad (20)$$

Remembering that the transmitted power P_t is identical to the power dissipated in the load, P_{in}, and R is 50 Ω, Equation (19) simplifies to

$$TAF = \frac{1}{r}\sqrt{0.6 G_t} \qquad (21)$$

This result is reasonable, as the TAF should be an inverse function of distance from the source, and a direct function of the gain of the transmitting antenna, and should be independent of power input. It should be noted that the gain value in Equation (17) is the effective gain of the antenna, calculated from the measured values of the AF. The TAF as used above incorporates antenna efficiency, the effect of antenna mismatch and other losses.

The TAF expression can be converted to dB form by taking 20 x $\log_{10}$ of both sides of Equation (20). This gives

$$TAF_{(dB)} = G_{t(dB)} - 2.22 - 20 \times \log_{10}(r_{(m)}) \qquad (22)$$

Note that the TAF is proportional to the gain of the antenna and inversely proportional to the distance from the antenna. This is rational and suggests that the derivation is correct.

CONVERSION BETWEEN AF AND TAF

As can be seen from the derivations, although the AF and TAF have the same units, m^{-1}, *they are neither identical nor reciprocal.* They are connected by the fact that the gain is identical for both expressions. This fact allows the TAF to be computed from the AF by recalling that $f\lambda = c$, and rewriting Equation (11) as

$$G_{(dB)} = 20 \times \log_{10}(f_{(MHz)}) - AF_{dB(m^{-1})} - 29.79 \qquad (23)$$

Substituting Equation (22) for Equation (23) gives

$$TAF_{(dB)} = 20 \times \log_{10}(f_{(MHz)}) - AF_{dB(m^{-1})} - 32.0 \qquad (24)$$

(at the distance of calibration)

This conversion is valid for the conditions from which either the AF or TAF is measured. If the AF is measured over a ground plane (typical condition), then the TAF computed from the AF is valid for a similar condition.

Remember that the concept of reciprocity, as it applies to antennas, relates to the transmit and receive pattern. As such, the reciprocity does not include the effects of impedance mismatch, efficiency or other factors. These factors are included in the measured AF. Thus, if measured antenna factors are used, the TAF computed from these values will be accurate when the antenna is used under the same conditions, over a ground plane. A semi-anechoic chamber also fulfills the same conditions, subject to the constraint that, over the frequency range of the application of this concept, the RF absorber must be effective.

SUMMARY

The above discussions have provided simple derivations of two parameters of an EMC antenna, the AF and the TAF. These parameters are in daily use by many, but the source of the values is not well-known. It is the purpose of this paper to provide the derivations of these parameters to illustrate the use of antennas and why they work as they do.

ACKNOWLEDGMENTS

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Computing Required Input Power for a Given E-Field Level at a Given Distance

Introduction

This Application Note explains three separate methods for the calculation of input power to an antenna to achieve a specific value of E-field for immunity or susceptibility testing, at a specific distance from the antenna. The estimates of required power agree well between the three methods (See Table 2.), so any of the three methods can be used as a function of information available regarding the antenna.

Calculations

This section describes three completely different, independent methods for calculating the input power required for a given E-field value as a function of frequency, at a given distance from the antenna. Inherent assumptions are that bore-sight alignment exists from the antenna to the point where the E-field is evaluated, and that ideal propagation conditions exist.

The three methods are:

- The Friis Transmission Formula
- The Transmit Antenna Factor
- Using information from published catalog or data sheet values of E-field for a given reference level of E-field.

Input power to an antenna to develop specified E-field value is readily accomplished, given some combination of the following information:

- numerical gain, G_i
- gain in dB, G_i, *(dB)*, and
- antenna factor, $AF\ (dB\ m^{-1})$.

The required information is:

- distance from the transmitting antenna reference point,
- frequency, and
- required E-field level.

Expressions for computing any of these input variables, given a value for one, are shown in Table 1.

Table 1.

Expressions for Computing G_i , G_i *(dB)*, and $AF(dB\ ^{m-1})$, Given A Value for One Parameter

	$AF(m^{-1})$	G_i	$G_i(dB)$
$AF(m^{-1})$		$20\times\log\left(\frac{9.73}{\lambda\sqrt{G_i}}\right)$	$20\times\log[f(MHz)]$ $-29.79-G_i(dB)$
G_i	$\left(\frac{9.73\times f(MHz)}{(300)\times 10^{\frac{AF}{20}}}\right)^2$		$10^{\frac{G_i(dB)}{10}}$
$G_i(dB)$	$20\times\log[f(MHz)]$ $-29.79-AF(m^{-1})$	$10\times\log(G_i)$	

Sample values, at 100 MHz, used in the calculations are:

- numerical gain over an isotropic antenna, G_i, = 2.05,
- gain in dB of the antenna, G_i, *(dB)* = 3.1 dB,
- Antenna Factor , $AF\ (dB\ m^{-1})$ = 7.1 dB m^{-1}

Figure 1 shows the geometry assumed for the calculations, and some of the important variables.

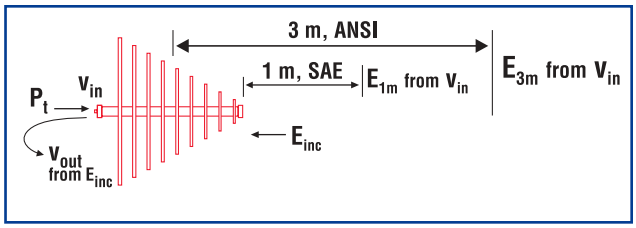

Figure 1.

Geometry for the Calculation of Input Power for a Given E-Field

Note that the reference point for calibration is different for the two standards applied for calibration of E-field generating antennas. The Society of Automotive Engineers *Aerospace Recommended Practice 958* is applied for calibration of antennas used in MIL-STD testing, for a spacing of 1 meter, tip-to-tip. The American National Standards Institute C63.5 is applied from

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feed point for biconical dipole antennas, and, as shown, from the mid-point of all elements of a log periodic antenna's element array. These measurement reference points on the antennas are important because they define the starting point of the distance to the point in free space, where the value of the E-field is defined.

These calculations are valid for estimates of input power when testing will be conducted in an anechoic chamber. The ARP and ANSI calibration methods produce a value for AF that is referred to as the "equivalent free space antenna factor". This means that the effects of the calibration environment are removed from the antenna factor, and that the numerical values are very close to that which would be measure if the antenna had, in fact, been calibrated under true free space conditions.

Method I Using the Relationship for E-field at a Distance Derived from Ohm's Law for Free Space and the Friss Transmission Formula:

A variant of the Friss transmission formula is:

$$E\ (V/m) = \frac{1}{r}\sqrt{30 \times P_t\ (W) \times G_t}$$

It relates the E-field [*E (V/m)*] produced at a distance [*r (m)*] due to net input RF power [P_t *(W)*] being applied to an antenna with known gain[G_t], (see Figure 1).

Solving for input power gives:

$$P_t = \frac{E^2 r^2}{30 \times G_t}$$

As an example calculation, suppose that a 10 V/m field was required at 3 meters for immunity testing. If the antenna chosen is a log periodic antenna with a gain of 2.05 at 100 MHz, the input power required is:

$$P_t = \frac{(10.0)^2 \times (3)^2}{30 \times 2.05} = 14.64\ W$$

Method II Using Transmit Antenna Factor

The transmit antenna factor is a measure of the effectiveness of the given antenna in transmitting electromagnetic power. It relates the RF voltage [V_{in} *(Volts)*] to the E field [$E|_d$ *(V/m)*] at a distance [*d(m)*] from the antenna, as determined at the distance of calibration of the antenna. (See Figure 1.) It is given by:

$$TAF\ (dBm^{-1})\Big|_d = G\ (dB_i)\ - 2.22 - 20 \times \log[d(m)]$$

For the example given, G_t = 3.1 *dB* and *d* = 3*m*

Thus:

$$TAF(dBm^{-1})\Big|_{3m} = 3.1 - 2.22 - 20 \times \log[3m] = -8.66 dB(m^{-1})$$

Remembering that:

$$E(dB\ V/m)\ = V_{in}\ (dBV) + TAF\ (dB\ m^{-1})$$

Then:

$$V_{in}\ (dBV) = E(dB\ V/m)\ - TAF\ (dB\ m^{-1})$$

or:

$$V_{in}\ (dBV) = 20 \times \log\ [V\ (V/m)] - TAF\ (dB\ m^{-1})$$

and:

$$V_{in}\ (dBV) = 20 \times \log\ [10(V/m)] - (-8.66) = 28.66\ (dB\ V)$$

To compute input power requirements, a linear value for voltage is required:

$$V_{in}(V) = 10^{\frac{V_{in}\ (dBV)}{20}} = 10^{\frac{28.66\ (dBV)}{20}} = 27.10\ V$$

Then the required power input is:

$$P_t(W) = \frac{V_{in}\ (V)^2}{R\ (\Omega)} = \frac{(27.1)^2}{50} = 14.64\ W$$

Method III Calculation of Required Input Power Using Published Graphical Data:

Using the graphical data, it is estimated that the input power required for 10 V/m at 3 meters is 0.15 W. A typical plot of input power for a given E-field level is shown in Figure 2.

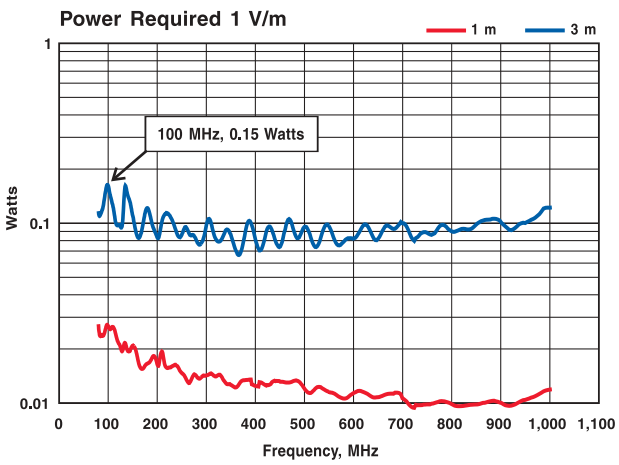

Figure 2.

Typical Plot of the Input power for a Specific E-Field Value

As seen in Figure 2, the 100 MHz value for input power for 1 V/m is approximately 0.15 W.

Computing the input power in dB:

$$P_{in}(dB\ W)\big|_{10V/m\,@\,3m} = 10 \times \log\left[0.15\ (W)\right] = -8.24\ dBW$$

The desired E-field strength is 10 V/m:

$$E(dBV/m) \ = 20 \times \log\left[10(V/m)\right] = 20\,dBV/m$$

Then power for 10 V/m is

$$P_{in}(dB\ W) = P_{in(chart)}(W) + E(dBV/m)$$

Or, substituting the actual values:

$$P_{in}(dB\ W) = 20\,dBV/m\ - (8.24\ dBW) = 11.76\ dBW$$

The linear value of the input power is:

$$P_{in}(W)\big|_{10V/m@3m} = 10^{\frac{11.76}{10}} = 15.00\ W$$

Comments

Three different computations of the desired input power, as shown above, give three similar answers. The results are summarized in Table 2. Examination of these answers reveals that the variance of the three answers is just under 2.5%, using the largest of the answers as a reference.

Table 2.

Summary of Results

Method of Computation	Answer
Friss Transmission Formula	14.63 *W*
Transmit Antenna Factor	14.69 *W*
Using Published Chart Values	15.00 *W*

The larger value obtained from using the Chart Values is likely to be due to inaccuracy in reading the chart. This value is 4.67 % (about 0.2 dB) larger than the other values, with more precise input. Comparing the two values with numerical input data gives a 0.03 dB difference.

Note that the Friss Transmission Formula, used in Method l, does not consider the effects of the ground plane. The answer derived from this formula agrees with the answer derived using Method ll's "equivalent free space antenna factor" value methodology. The results from these first two Methods also agree with Method lll's answer. This is derived using a graphical portrayal of the transmit antenna factor as computed from the measured antenna factor.

Cautions

These computed values are based on the nominal conditions of "free space" testing, *i.e.*, testing in a Anechoic Chamber, to contain the fields generated. They are useful estimates for other conditions, but engineering judgment must be applied to the selection of amplifier in all cases.

Other adjustments for sizing the input amplifier include the antenna input VSWR. At the upper and lower ends of their bandwidth, some antennas will have a VSWR that far exceeds the nominal value. In this case a correction, as shown in Figure 3, should be applied.

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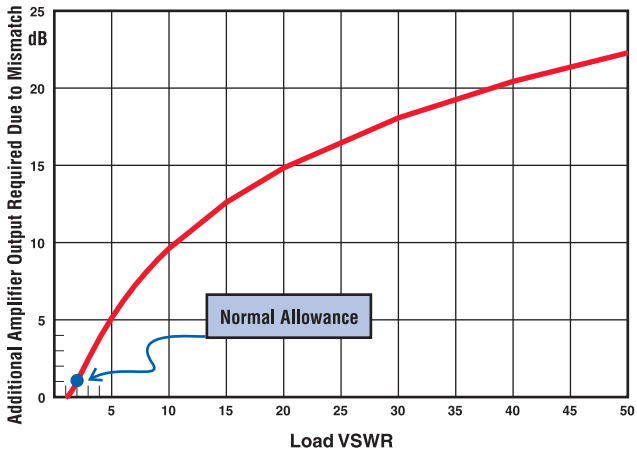

Figure 3.

Correction for Input Power Required, when the Antenna Input VSWR Exceeds 2:1.

If a numerical result is desired, the VSWR correction can be computed from:

$$VSWR\ \ Correction = 20 \times \log\left(\frac{1}{1-\left(\frac{VSWR-1}{VSWR+1}\right)^2}\right)$$

Remember also that the amplifier must be operated in its linear operating range, at least 1 dB below the 1 dB compression point.

In addition, no amplifier is as reliable running close to or at maximum rated power, thus an allowance for rating to assure that the amplifier is running at about 90 % of rated power will produce almost indefinite operation. This adds another approximately 1 dB to the required output power.

Note that the ANSI calibration distance for measurement of the antenna factor (AF) of log periodic antennas (at the center of the element array), is different than the distance where the immunity test calibration is performed (measured from the tip of the antenna). The ratio of these distances, in dB, should be added to the power amplifier rating to more closely estimate the required power input level. For a typical EMCO large log periodic antenna, this difference in distance is just under 1 meter. If the calibration distance is 3 meters from the tip of the antenna, the correction would be 20 x *log* [4m ÷ 3m] = 2.5 dB. Smaller antennas will need a lesser allowance.

With respect to sizing the amplifier for use with a given antenna, remember that most calibration measurements are conducted with continuous wave (CW) excitation of the antenna. When actual testing is accomplished, it is usually accomplished with amplitude (AM) modulation. The most recently published specification requires 80 % AM with a 1 kHz sine wave. This requires an amplifier with 1.8 times the linear voltage output range of the CW signal. This, in turn, requires the amplifier output to be $(1.8)^2$ larger than for the CW case, since power input increases as the square of the input voltage. This means that the power amplifier gain will need to be 10 x *log* $[(1.8)^2]$ = 5.1 dB greater than needed for the CW case.

A summary of these factors is shown in Table 3.

Table 3.

Summary of Input Power Allowances for Sizing an Amplifier

Factor	Allowance (dB)
True Linear Operation	1.0
Calibration Distance	2.5
Modulation Allowance	5.1
TOTAL	8.6

Thus, any amplifier input computed for use may actually need a 8.6 dB higher rating for proper continuous operation.

In addition, if the antenna input VSWR is more than 2:1, additional compensation for the reflected power from the antenna port is required.

NOTE: Cable loss is not considered in this allowance table since it is a factor of cable length.

DERIVATION AND USE OF FORWARD POWER GRAPHS

Forward Power graphs in this catalog are derived from several methods and each chart indicates which method was used. Since SAE and ANSI antenna factors are typically within a few dB of free-space antenna factors, graphs indicating "Forward Power Derived From AF" are valid for free-space environments, such as a fully anechoic chamber or an absorber-treated ground plane. Graphs indicating "Forward Power Measured Over Conducting Ground" are valid only for the specific geometry listed (antenna/field probe height and separation) on an OATS or in a large high quality semi-anechoic chamber. Graphs indicating "Forward Power Measured Over Ferrite Ground" generally fall between the ground plane and free-space values. As with the ground plane numbers, these results are strictly valid for the same geometry on an OATS or a semi-anechoic chamber with comparable ferrite panels on the floor. Small chamber power requirements depend on the chamber's dimensions and antenna location. Typically, small chamber power requirements will fall in the vicinity of the conducting and ferrite-ground results.

Antenna Selection by Test Type

FCC 15 — RADIATED EMISSIONS

20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
200 MHz - 2 GHz	3106 Dbl RdgWaveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns

FCC 18 — RADIATED EMISSIONS

20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
30 MHz - 300 MHz	3124 Calculable Biconical
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
20 Hz - 5 MHz	6511 Loop - Passive

VCCI — RADIATED EMISSIONS

20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3142B BiConiLog™
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic

VCCI — RADIATED IMMUNITY

200 MHz - 1 GHz	3101 Conical Log Spiral
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
1 kHz - 30 MHz	6509 Loop - Active

VDE — RADIATED EMISSIONS

10 kHz - 30 MHz	6502 Loop - Active

CISPR/EC — RADIATED EMISSIONS

30 Hz - 50 MHz	3301B Rod - Active

IEC/CISPR/EN — RADIATED EMISSIONS

20 MHz -200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
20 Hz - 5 MHz	6511 Loop - Passive
10 kHz - 30 MHz	6512 Loop - Passive

IEC/CISPR/EN — RADIATED IMMUNITY

20 MHz - 300 MHz	3109 Dbl Rdg Waveguide
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Log Periodic
960 MHz - 40 GHz	3161 Log Periodic
1 kHz - 30 MHz	6509 Loop - Passive

SAE J551 — RADIATED EMISSIONS

20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
30 MHz - 300 MHz	3110B Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
30 Hz - 50 MHz	3301B Rod - Active
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
10 kHz - 30 MHz	6512 Loop - Passive

SAE J551 — RADIATED IMMUNITY

20 MHz - 300 MHz	3109 Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
960 MHz - 40 GHz	3160 Std Gain Horns
1 kHz - 30 MHz	6509 Loop - Active

SAE J1338 — RADIATED IMMUNITY

960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns

SAE J1113 — RADIATED EMISSIONS

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
30 MHz - 300 MHz	3110B Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active

SAE J1113 — RADIATED IMMUNITY

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 300 MHz	3109 Dbl Rdg Waveguide
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
26 MHz - 2 GHz	3140 BiConiLog
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Log Periodic
960 MHz - 40 GHz	3161 Log Periodic
1 kHz - 30 MHz	6509 Loop - Passive

SAE J1507 — RADIATED IMMUNITY

960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns

SAE J1551 — RADIATED IMMUNITY

960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns

SAE J1816 — RADIATED EMISSIONS

30 Hz - 50 MHz	3301B Rod - Active
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
10 kHz - 30 MHz	6512 Loop - Passive

NSA 65-6 — TRANSMIT

1 kHz - 30 MHz	3303 Rod - Active
1 kHz - 30 MHz	6509 Loop - Passive

NSA 65-6 — RECEIVE

30 Hz - 50 MHz	3301B Rod - Active
1 kHz - 30 MHz	6507 Loop - Active

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MIL-STD-285 TRANSMIT

20 MHz - 200 MHz	3104C Biconical
20 MHz - 300 MHz	3109 Biconical
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
200 MHz - 2 GHz	3148 Log Periodic
1 kHz - 30 MHz	3303 Rod - Passive
1 kHz - 30 MHz	6509 Loop - Passive

MIL-STD-285 RECEIVE

20 MHz - 200 MHz	3104C Biconical
20 MHz - 300 MHz	3109 Biconical
30 MHz - 300 MHz	3110B Biconical
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3142B BiConiLog™
200 MHz - 2 GHz	3148 Log Periodic
30 Hz - 50 MHz	3301B Rod - Active
1 kHz - 30 MHz	6507 Loop - Active

TELECOM

440 MHz - 460 MHz	3125 Dipole (450)
590 MHz - 610 MHz	3125 Dipole (600)
824 MHz - 915 MHz	3125 Dipole (870)
935 MHz - 960 MHz	3125 Dipole (950)
1600 MHz -1620 MHz	3125 Dipole (1610)
1710 MHz -1785 MHz	3125 Dipole (1750)
1805 MHz -1880 MHz	3125 Dipole (1840)
1850 MHz -1910 MHz	3125 Dipole (1880)
2440 MHz -2460 MHz	3125 Dipole (2450)
2990 MHz -3010 MHz	3125 Dipole (3000)
400 MHz - 6 GHz	3164 Diag Dual Polar Horn

MIL-STD-461E RADIATED EMISSIONS

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
400 MHz - 6 GHz	3164 Diag. Dual Polar Horn
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
30 Hz - 50 MHz	3301B Rod - Active
20 Hz - 500 kHz	7604 Coil - Transducer

MIL-STD-461E RADIATED SUSCEPTIBILITY

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 300 MHz	3109 Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
20 Hz - >50 kHz	7603 Magnetic Field Coil
30 Hz - >50 kHz	7605 Magnetic Field Coil
30 Hz - >50 kHz	7606 Magnetic Field Coil

MIL-STD-1541 RADIATED EMISSIONS

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
30 MHz - 1 GHz	3121C Dipole
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3142B BiConiLog™
200 MHz - 2 GHz	3148 Log Periodic
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
20 Hz - 5 MHz	6511 Loop - Passive

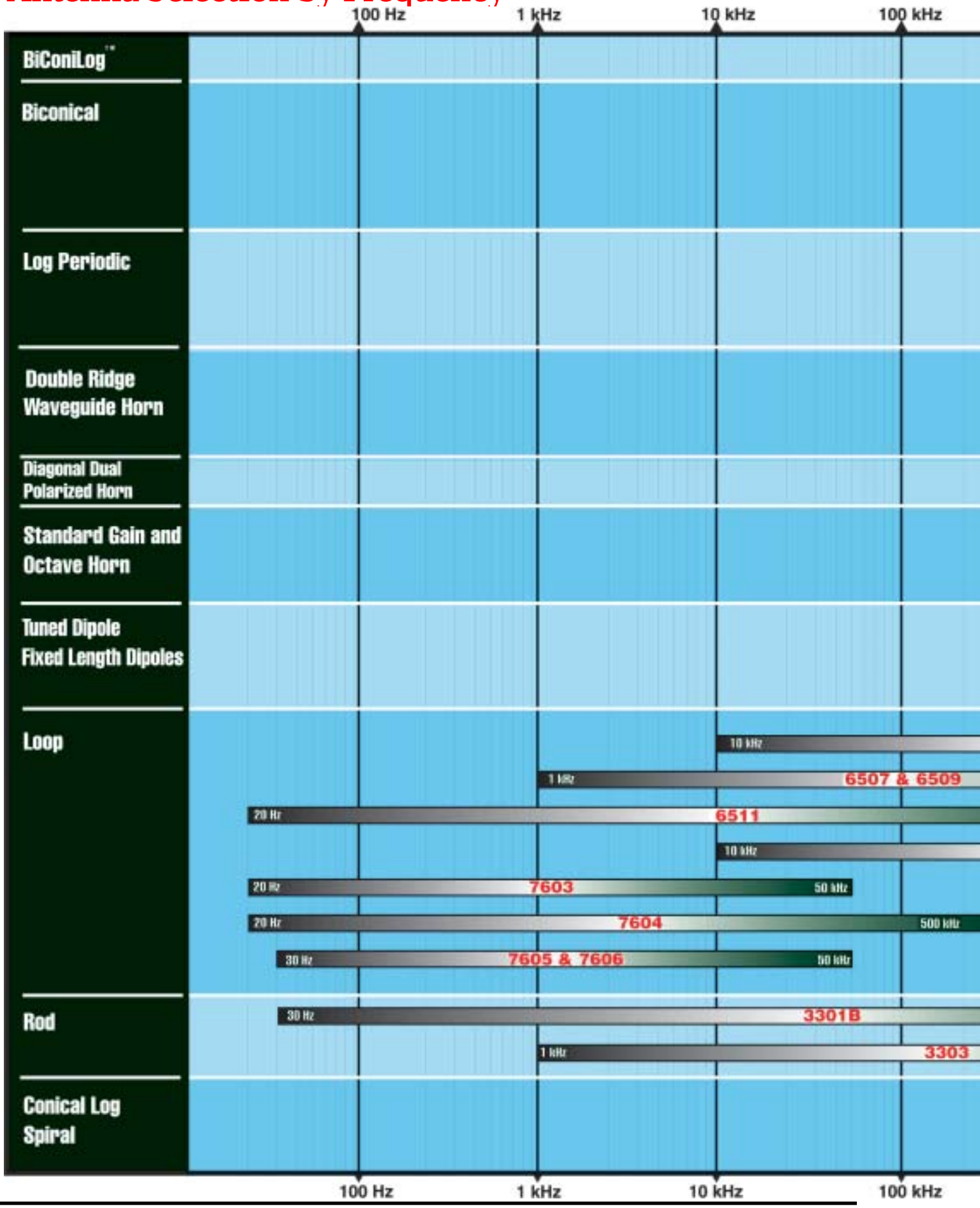
MIL-STD-1541 RADIATED SUSCEPTIBILITY

100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 300 MHz	3109 Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
26 MHz - 2 GHz	3140 BiConiLog™
26 MHz - 2 GHz	3142B BiConiLog™
80 MHz - 2 GHz	3144 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
960 MHz - 40 GHz	3160 Std Gain Horns
1 kHz - 30 MHz	6509 Loop - Passive
20 Hz - 5 MHz	6511 Loop - Passive

NACSIM RADIATED EMISSIONS

200 MHz - 1 GHz	3101 Conical Log Spiral
1 GHz - 10 GHz	3102 Conical Log Spiral
100 MHz - 1 GHz	3103 Conical Log Spiral
20 MHz - 200 MHz	3104C Biconical
30 MHz - 300 MHz	3110B Biconical
30 MHz - 300 MHz	3124 Calculable Biconical
200 MHz - 2 GHz	3106 Dbl Rdg Waveguide
1 GHz - 18 GHz	3115 Dbl Rdg Waveguide
18 GHz - 40 GHz	3116 Dbl Rdg Waveguide
400 MHz - 6 GHz	3164 Diag Dual Polar Horn
960 MHz - 40 GHz	3160 Std Gain Horns
960 MHz - 40 GHz	3161 Std Gain Horns
26 MHz - 2 GHz	3142B BiConiLog™
200 MHz - 5 GHz	3147 Log Periodic
200 MHz - 2 GHz	3148 Log Periodic
30 Hz - 50 MHz	3301B Monopole (Rod)
10 kHz - 30 MHz	6502 Loop - Active
1 kHz - 30 MHz	6507 Loop - Active
20 Hz - 5 MHz	6511 Loop - Passive

Antenna Selection by Frequency



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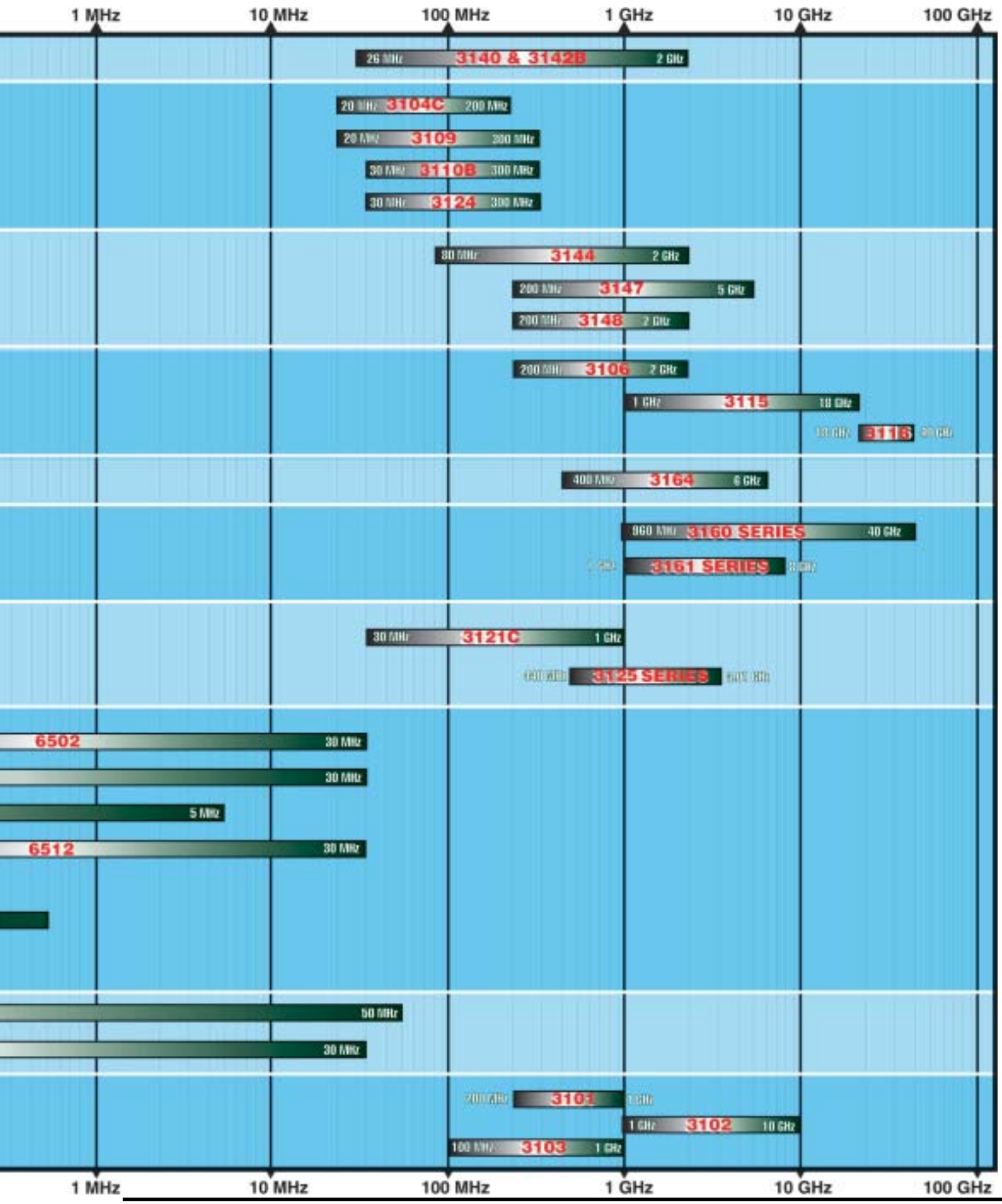

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Expanded Uncertainty Values for Antenna Calibrations (95% Confidence)

TYPE	MODEL NUMBER	FREQUENCY RANGE	SAE, ARP 958 1M	ANSI C63.5 3M	ANSI C63.5 10M	IEEE 291, IEEE 1309, ANSI C63.4
Conical Log Spiral (L&R)	**3101**	200MHz-1GHz	200-300 MHz +/- 2.8 dB 300-850 MHz +/- 0.8 dB 850-1000 MHz +/- 1.4 dB			
Conical Log Spiral (L&R)	**3102**	1GHz-10GHz	1-10 GHz +/- 0.8 dB			
Conical Log Spiral (L&R)	**3103**	100MHz-1GHz	+/- 2.0 dB Type B			
Biconical	**3104C**	20MHz-200MHz	20-200 MHz +/- 1.2 dB	200MHz +/- 0.9 dB	20-30 MHz +/- 1.0 dB 30-200 MHz +/- 0.9 dB	
Biconical	**3109**	20MHz-300MHz	20-30 MHz +/- 1.0 dB 30-200 MHz +/- 1.4 dB 200-300 MHz +/- 2.0 dB	20-30 MHz +/- 0.9 dB 30-300 MHz +/- 0.9 dB	20-30 MHz +/- 1.0 dB 30-300 MHz +/- 0.9 dB	
Biconical	**3110B**	30MHz-300MHz	30-200 MHz +/- 1.2 dB 200-300 MHz +/- 2.2 dB	20-200 MHz +/- 0.8 dB	20-30 MHz +/- 1.0 dB 30-300 MHz +/- 0.9 dB	
BiCal	**3124**	30MHz-300MHz		300MHz - 300 MHz +/-0.50 dB		
Double-Ridged Waveguide Horn	**3106**	200MHz-2GHz	0.2-1.9 GHz +/- 1.0 dB 1.9-2.0 GHz +/- 1.3 dB			
Double-Ridged Waveguide Horn	**3115**	1GHz-18GHz	1-18 GHz +/- 0.3 dB			
Double-Ridged Waveguide Horn	**3116**	18GHz-40GHz	18-30 GHz +/- 0.8 dB 30-40 GHz +/- 1.3 dB			
Diagonal Dual Polarized Horn	**3164**	400 MHz - 6 GHz	.4-6 GHz +/- 1.0 dB			
Octave Horns	**3161-01**	1GHz-2GHz	1-2 GHz +/- 0.9 dB			
Octave Horns	**3161-02**	2GHz-4GHz	2-4 GHz +/- 0.5 dB			
Octave Horns	**3161-03**	4GHz-8GHz	4-4.5 GHz +/- 1.0 dB 4.5-7.5 GHz +/- 0.4 dB 7.5-8 GHz +/- 1.3 dB			
Dipole	**3121C 4 Balun Kit**	30MHz-1GHz				
Dipole	**3121C Balun 1**	30MHz-60MHz		30-60 MHz +/- 0.9 dB	30-60 MHz +/- 1.0 dB	
Dipole	**3121C Balun 2**	60MHz-140MHz		60-140 MHz +/- 0.7 dB	60-140 MHz +/- 0.7 dB	

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Expanded Uncertainty Values for Antenna Calibrations (95% Confidence)

TYPE	MODEL NUMBER	FREQUENCY RANGE	SAE, ARP 958 1M	ANSI C63.5 3M	ANSI C63.5 10M	IEEE 291, IEEE 1309, ANSI C63.4
Dipole	**3121C Balun 3**	140MHz-400MHz		140-375 MHz +/- 1.0 dB 375-400 MHz +/- 1.4 dB	140-375 MHz +/- 1.0 dB 375-400 MHz +/- 1.4 dB	
Dipole	**3121C Balun 4**	400MHz-1GHz		400-700 MHz +/- 1.0 dB 700-1000 MHz +/- 1.4 dB	400-700 MHz +/- 1.6 dB 700-1000 MHz +/- 2.1 dB	
DB3&4 Tuned				140-400 MHz +/- 0.8 dB 400-1000 MHz +/- 1.0 dB	140-400 MHz +/- 0.8 dB 400-1000 MHz +/- 1.0 dB	
BiConiLog	**3142B**	26MHz-2GHz/ 26MHz-1.1GHz	26-30 MHz +/- 1.4 dB 30-1000 MHz +/- 0.8 dB 1-2 GHz +/- 1.2 dB	26-30 MHz +/- 1.5 dB 30-1000 MHz +/- 1.0 dB 1-2 GHz +/- 1.3 dB	26-30 MHz +/- 2.1 dB 30-1000 MHz +/- 1.0 dB 1-2 GHz +/- 1.4 dB	
LPA	**3144**	80MHz-2GHz	+/- 2.0 dB, Type B	80-2000 MHz +/- 0.9 dB	80-2000 MHz +/- 0.8 dB	
LPA	**3147**	200MHz-5GHz	200-2000 MHz +/- 0.5 dB 2-5 GHz +/- 1.6 dB	200-1000 MHz +/- 0.8 dB 1-2 GHz +/- 0.9 dB 2-5 GHz +/- 1.5 dB	200-1000 MHz +/- 0.8 dB 1-2 GHz +/- 1.0 dB 2-5 GHz +/- 1.5 dB	
LPA	**3148**	200MHz-2GHz	200-1000 MHz +/- 0.6 dB 1-2 GHz +/- 1.41 dB	200-1000 MHz +/- 0.9 dB 1-2 GHz +/- 0.9 dB	200-1000 MHz +/- 0.8 dB 1-2 GHz +/- 0.9 dB	
Rod	**3301B 41 inch Rod**	30Hz-50MHz				30Hz-50MHz +/- 0.3 dB
Rod	**3303**	1kHz-30MHz				1kHz-10kHz +/- 1.0 dB 10kHz-30MHz +/- 0.1 dB
Loop	**6502**	10kHz-30MHz				(TBD)
Loop	**6507**	1kHz-30MHz				(TBD)
Loop	**6509**	1kHz-30MHz				(TBD)
Loop	**6511**	20Hz-5MHz				(TBD)
Loop	**6512**	10kHz-30MHz				(TBD)
Coil	**7603**	20Hz-50kHz				No Cal/VSWR Only
Coil	**7604**	20Hz-500kHz				No Cal/VSWR Only
Coil	**7605**	30Hz-50kHz				No Cal Req
Coil	**7606**	30Hz-50kHz				No Cal Req

Data effective 8 June 1998. EMCO Laboratory capabilities include all products produced by EMCO and models of the same technology produced by other manufacturers. Uncertainty values reflect the uncertainty analysis conducted using 1997 data. Values released are valid with a 2 sigma (95%) confidence level and are representative of the measurement quality conducted by the EMCO Laboratory using the industry recognized standards listed above.

Laboratory Standards Compliance List

SAE, ARP 958 - 1997, Society of Automotive Engineers, Aerospace Recommended Practice 958, Electromagnetic Interference Measurement Antennas; Standard Calibration Method.

ANSI C63.4 - 1992, American National Standard, Methods of Measurement of Radio-Noise Emissions from Low-Voltage Electrical and Electronic Equipment in the Range of 9 kHz to 40 GHz.

ANSI C63.5 - 1988, American National Standard, Calibration of Antennas Used for Radiated Emission Measurements in Electromagnetic Interference (EMI) Control.

ANSI C63.6 - 1988, American National Standard, Guide for the Computation of Errors in Open Area Test Site Measurements.

ANSI C63.7 - 1992, American National Standard, Guide for Construction of Open-Area Test Sites for Performing Radiated Emission Measurements.

ANSI Z540-1 - 1994, American National Standard, Calibration Laboratories and Measuring and Test Equipment - General Requirements.

ANSI Q91 - 1994, Quality Systems - Model for Quality Assurance in Design, Development, Production, Installation and Servicing.

ISO Guide 25 - 1990, International Standards Organization, General Requirements for the Competence of Calibration and Testing Laboratories.

IEEE 291 - 1991, Institute of Electrical and Electronics Engineers, Standard Methods for Measuring Electromagnetic Field Strengths of Sinusoidal Continuous Waves, 30 Hz to 30 GHz.

IEEE Std 1309 - 1996, Institute of Electrical and Electronics Engineers, Standard for Calibration of Electromagnetic Field Sensors and Probes, Excluding Antennas, from 9 kHz to 40 GHz.

NIST Technical Note 1297, 1994 edition, National Institute of Standards and Technology, Guidelines for Evaluating and Expressing the Uncertainty of NIST Measurement Results.

NIS 81, Edition 1, May 1994, NAMAS, The Treatment of Uncertainty in EMC Measurements.

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A Statistical Approach to Measurement Uncertainty

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Despite the fact that most authors have skirted the subject of Type A analyses, the method is not so difficult that it should be avoided altogether.

Introduction

Over the past couple of years, the topic of measurement uncertainty has come to the forefront in the international EMC community. In brief, the intention of measurement uncertainty is to take the more traditional terms of precision, accuracy, random error, and systematic error used in scientific circles and replace them with a single term. This term represents the total contribution to the expected deviation of a measurement from the actual value being measured[1-2].

In general, precision is a measure of random error, or how closely repeated attempts hit the same point on a target, while accuracy is a measure of systematic error, or how close those attempts are to the center of the target. It is obvious that both contributions must be accounted for in order to determine the quality of a measurement, although the combination of the two can sometimes be more subtle than might be expected. There has been some discussion over whether the term reproducibility is also replaced by uncertainty since the concept of reproducibility must contain variations in the equipment under test (EUT) and therefore does not represent the same quantity as measurement uncertainty[3].

The methods for determining a measurement uncertainty have been divided into two generic classes:

- Type A represents a statistical uncertainty based on a normal distribution.
- Type B represents uncertainties determined by any other means.

In last year's ITEM, Manfred Stecher wrote an article describing the introduction of uncertainty evaluations into various EMC standards and explained the technique typically used to determine measurement uncertainties for EMC measurements[4]. (A similar paper was also presented at the 1996 IEEE International EMC Symposium in Santa Clara, CA.[5]) The article gives an adequate introduction to the Type B evaluation method, which uses individual measurements, manufacturers specifications, and even educated guesses to determine a combined uncertainty. However, the author is a little too quick to discard the statistical Type A uncertainty measurement as impractical. To be sure, the Type A analysis does suffer from the very pitfalls which Mr. Stecher points out. However, with a bit of care it is possible to obtain a significant amount of useful information from the technique.

The advantage of a Type A uncertainty measurement is that when done correctly, the resulting value is irrefutable since it has been determined from real world measurements. The biggest complaint I hear from engineers being exposed to the Type B uncertainty budget method for the first time is the fact that too many of the terms are either poorly defined by equipment manufacturers or must simply be estimated. In many cases, the chosen values may be too stringent in order to provide a safety margin. On the other hand, the desire for smaller total uncertainties can lead to using smaller estimations than is realistic for some terms that are hard to determine.

Antenna manufacturers have easy access to a vast database of antenna calibrations with which to determine statistical trends. However, as the data shown here will demonstrate, it is not necessary to have an extremely large sample to get acceptable results. The real issue in using a statistical approach is in determining where it fails and using a Type B analysis to fill in the gaps. The document NIS 81, "The Treatment of Uncertainty in EMC Measurements", released by NAMAS[1], recommends this exact approach.

Certainly a Type A analysis of a set of measurements can be expected to include all random errors of the entire measurand and none of its systematic errors. But does that mean that the Type A uncertainty

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contains only the random portion of the individual terms which might be used in a Type B analysis? Certainly not. It is not possible to completely separate random and systematic effects into individual categories[2]. Random effects such as positioning error[6] or cable length and frequency dependence of standing waves can serve to randomize such systematic errors as site imperfections or mismatch errors. Intentionally varying the setup between repeated measurements can do the same.

The discussion presented here will focus primarily on antenna calibration measurements, which have many similarities to radiated emissions measurements. However, many of the techniques demonstrated will be applicable to radiated susceptibility and conducted tests. Despite the fact that most authors have skirted the subject of Type A analyses, the method is not so difficult that it should be avoided altogether. In fact, many of the more difficult terms to determine for a typical Type B analysis are already included in the random error of the total measurement, thus reducing the overall task.

Type A Evaluation of Uncertainty

Random effects cause repeated measurements to vary in an unpredictable manner. The associated uncertainty can be calculated by applying statistical techniques to the repeated measurements. An estimate of the standard deviation, $s(q_k)$, of a series of n readings, q_k, is obtained from

$$s(q_k) = \sqrt{\frac{1}{(n-1)}\sum_{k=1}^{n}(q_k - \bar{q})^2}$$

where $\bar{q}$ is the mean value of measurements. The random component of the uncertainty may be reduced through repeated measurements of the EUT. In this case, the standard deviation of the mean, $s(\bar{q})$, given by:

$$s(\bar{q}) = \frac{s(q_k)}{\sqrt{n}}$$

represents the uncertainty of the resulting mean. This last point has been misinterpreted by some due to a confusing statement in section 3.2.6 of NIS 81, "The standard uncertainty, $u(x_i)$, of an estimate x_i of an input quantity q is therefore $u(x_i) = s(\bar{q})$." This statement is certainly true, but some readers have construed it to mean that the standard uncertainty of a single measurement q_k is given by $s(\bar{q})$.[7-8] This is true if $n = 1$ so that $s(\bar{q}) = s(q_k)$. This is explained more clearly in section 3.2.3 of NIS 81 where the concept of predetermination is discussed.

Time constraints and other practical considerations will often make it unfeasible to perform more than a single measurement on an EUT. However, repeat measurements can be performed on a similar EUT to predetermine $s(q_k)$ as the expected standard uncertainty of an individual measurement. If a smaller uncertainty due to random errors is desired, multiple measurements can be made and the value of $s(\bar{q})$ reduced accordingly. The value of used in this case is the number of measurements made on the EUT, not the number of measurements used in the predetermination.

It should be noted that variations in the EUT as a function of time, as well as that of multiple like EUTs as in the method demonstrated here, will be included in the uncertainty determined through this method. For this reason, and to provide a means to determine some of the systematic contributions to the uncertainty, it is recommended that a stable reference radiating source such as a comb generator or amplified noise source be used as the EUT for uncertainty determinations. This provides a repeatable EUT which, when used in conjunction with "round robin" type testing, can allow determination of even the systematic elements of your measurement uncertainty to within the uncertainty of the round robin test. This type of EUT also provides a broad continuous frequency range for uncertainty determination as opposed to the random spectrum points of a typical EUT. If a reference source is not available, a number of the same benefits may be obtained using a stable signal generator and appropriate radiating antenna. However, transmit cable placement and other effects will add some additional random error and this method does not lend itself to inter-site comparisons.

Uncertainty Example: Antenna Calibration

Antenna calibrations have uncertainty aspects similar to that of both radiated emissions and susceptibility tests. However, the one systematic element missing from the test is the absolute value of the fields (or signal levels) involved. Thus, the test only depends on the linearity of the instrumentation and not its absolute calibration. However, this does not effect the validity of this method for determining Type A uncertainties for EMC tests since the systematic error in field level cannot be determined by this method without inter-site comparisons or other tests using multiple methods to determine the absolute field level.

To obtain the data shown here, a sample of 26 different antennas calibrated over a three-and-a-half month period was used. The antennas were identical log-periodic antennas with a frequency span of 200

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MHz to 1 GHz. The time period spanned from mid-August through the end of October, providing a significant variation in temperature and weather conditions. The data were measured at 1601 frequency points using a vector network analyzer, low loss cables, and a positioning tower with 0.1 cm positioning resolution. The test was performed at a 10-meter separation, 2-meter transmit height, and 1- to 4-meter scan height per ANSI C63.5 on an open area test site (OATS). The vector network analyzer used has over 100 dB of available dynamic range and superior external noise rejection, making it ideal for use on an OATS where ambients would be a problem for traditional spectrum analyzer/tracking generator combinations. Since the network analyzer does not offer a max-hold or other such functionality traditionally seen on a spectrum analyzer, it is necessary to perform the max-hold by transferring individual traces to a controlling PC and allowing the test software to perform the max-hold.

To facilitate this method, it is also necessary to step the tower and take frequency sweeps at discrete heights since the network analyzer cannot sweep the entire frequency band fast enough to get acceptable height resolution when the tower moves continuously. Tests with tuned dipoles have shown that the variation in the antenna factor at 1 GHz for a 1-cm step, a 5-cm step, and a single frequency point continuous motion max-hold is less than ±0.02 dB. If the tower step size was too large, or the sweep time too slow in the case of a continuous motion, the measurement could be expected to introduce a systematic error since missing a peak signal would always result in a measurement lower than the real value. This error exists for scanned height radiated emissions tests as well as antenna calibrations. For an antenna calibration, this error would result in a larger antenna factor than is actually the case.

The standard deviation of the 26 antenna factors and their maximum deviations from the average are shown as a function of frequency in Figure 1. The negative of the standard deviation is also shown for comparison to the negative deviation. The symmetry of the positive and negative deviations is a first indication of how closely the sample approximates a normal distribution. An asymmetrical envelope would be a cause for concern and indicate the need for separate positive and negative uncertainty values at the minimum.

Neglecting, for the moment, any additional contributions to the uncertainty due to systematic errors, these data indicate that the expanded uncertainty ($k = 2$) of an individual calibration is less than ±0.5 dB at all frequencies. That means that an uncertainty of ±0.5 dB with greater than 95% confidence can be claimed for this

Uncertainty Terms from NIS-81

- Estimated standard deviation from a sample of readings:

$$s(q_k) = \sqrt{\frac{1}{(n-1)}\sum_{k=1}^{n}(q_k - \bar{q})^2}$$

- Standard deviation of the mean of readings:

$$s(\bar{q}) = \frac{s(q_k)}{\sqrt{n}}$$

- Standard uncertainty (of the mean of readings) resulting from a type A evaluation:

$$u(x_i) = s(\bar{q})$$

- Standard uncertainty for contributions with a normal probability distribution:

$$u(x_i) = \frac{U}{k}$$

where U represents the expanded uncertainty of the normal distribution (last item below).

- Standard uncertainty for contributions with rectangular probability distribution:

$$u(x_i) = \frac{a_{i+} - a_{i-}}{2\sqrt{3}}$$

for an asymmetrical distribution, where a_{i+} and a_{i-} are the bounds of the rectangular region, or

$$u(x_i) = \frac{a_i}{\sqrt{3}}$$

for a symmetrical region with bounds $\pm a_i$.

- Standard uncertainty for contributions with U shaped probability distribution:

$$u(x_i) = \frac{a_{i+} - a_{i-}}{2\sqrt{2}}$$

for an asymmetrical distribution, where a_{i+} and a_{i-} are the bounds of the U shaped region, or

$$u(x_i) = \frac{a_i}{\sqrt{2}}$$

for a symmetrical region with bounds $\pm a_i$ or where a_i is the larger of a_{i+} or a_{i-}.

- Standard uncertainty in terms of the measured quantity:

$$u_i(y) = c_i \cdot u(x_i)$$

- Combined standard uncertainty:

$$u_c(y) = \sqrt{\sum_{i=1}^{N} u_i^2(y)}$$

- Expanded uncertainty:

$$U = k \cdot u_c(y) \quad \text{or} \quad U = k_p \cdot u_c(y)$$

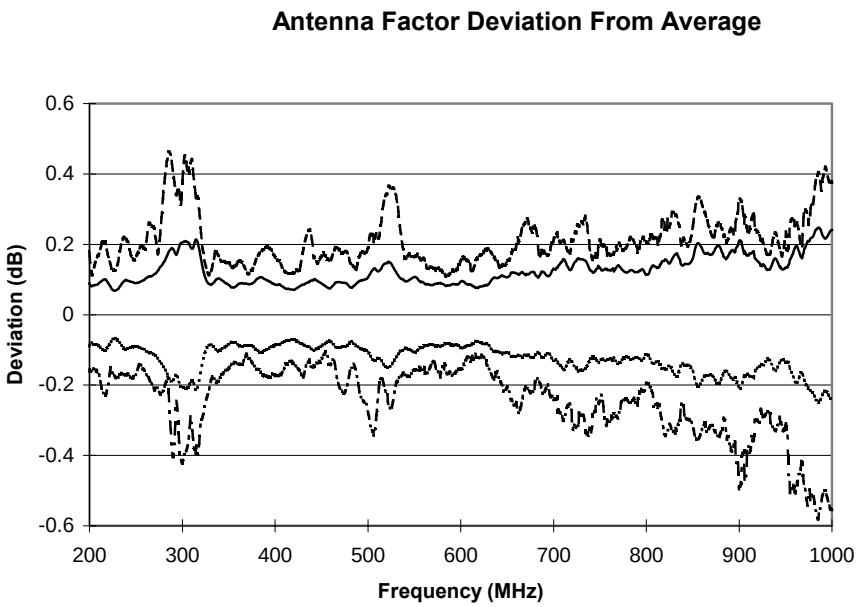

Figure 1. The Standard Deviation, Maximum Minus Average, and Minimum Minus Average of the Antenna Factors from a Set of Twenty-six Identical Antennas. Neglecting any additional systematic errors, the expanded uncertainty (k=2) for an individual antenna factor would then be twice the standard deviation for a value of approximately ±0.5 dB with greater than 95% confidence at all frequencies.

calibration. A second verification of this claim is the fact that, with the exception of the negative deviation above 950 MHz, none of the antennas in the sample had antenna factors which deviated from the average more than ±0.5 dB. This provides an added level of confidence in the quality of this uncertainty value.

It should be noted that this data represents calibrations of 26 different antennas performed at different times. About fifty percent of the antennas were directly off the production line, but the other half were re-calibrations of antennas as many as ten years old. Although one might expect a batch of antennas to be as close to identical as possible, this sample surely must contain some amount of manufacturing uncertainty. If this manufacturing uncertainty is non-zero, then it must also be contained within the above uncertainty. Since the "perfect" antenna, in terms of manufacturing quality, is defined by the average of all antennas, this uncertainty must be totally random.

Although this manufacturing uncertainty may add to the total calibration uncertainty measured by this technique, as long as the resulting uncertainty is within an acceptable range, it does not matter whether or not the contribution is large or small. In this case, there is also the added benefit that the intentional introduction of totally random effects due to

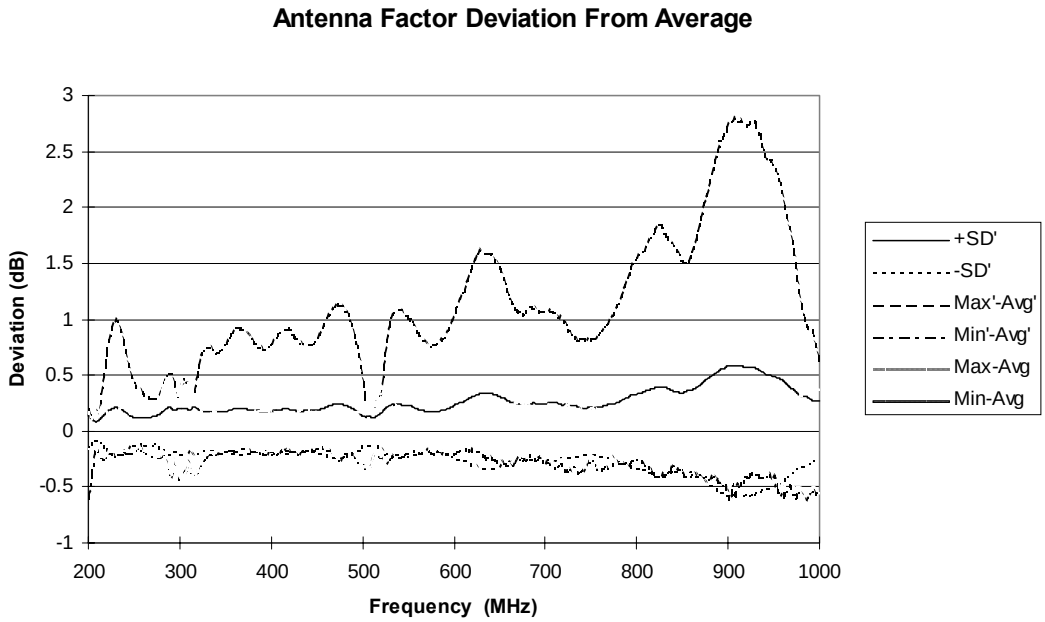

Figure 2. The Effect of EUT Variation on the Standard Deviation. Note the Difference in the Maximum Deviation for the Primed Sample Vs. Un-primed. The effect on the standard deviation is sufficient to equal or surpass the maximum deviation in the un-primed sample.

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the antennas may help to randomize the systematic contributions to the total measurement uncertainty. For emissions measurements, this is equivalent to using multiple identical EUTs to determine the Type A uncertainty. As long as variations in the EUTs do not add excessive contributions to the random error, it is possible to obtain a suitable uncertainty value from tests of multiple EUTs.

This is not to say that multiple EUTs always provide an acceptable method for performing this type of uncertainty calculation. One or two EUTs with significant deviations from the norm can introduce a significant change to the resulting uncertainty, as demonstrated in Figure 2. One additional antenna, a different model which varies significantly from the average at different frequencies, was introduced to the sample to demonstrate the possible effects.

Note that for most of the frequency range, the deviation is below 1 dB, yet the contribution of one additional antenna is sufficient to change the standard deviation such that it is larger than the original maximum deviations at most frequencies. As this example shows, when working with relatively small samples, it is important not to introduce factors which may exaggerate the error in the measurement. Likewise, it is important to avoid arbitrarily discarding data because it makes the result look bad!

Figure 2 also demonstrates the difference between systematic errors and random errors. One sample with a large systematic error (not present in the other samples) can have a significant effect on the uncertainty, since it changes not only the standard deviation, but also the mean of the samples. This causes the negative maximum deviation (minimum - average) to be larger than before, even though the absolute value of the minimum curve is the same.

Note that the standard deviation shown in Figure 2 would still allow the claim of ±0.5 dB for the expanded uncertainty from 200 to 600 MHz. It is perfectly acceptable (and often necessary due to band breaks in equipment) to generate frequency dependent uncertainties. The expanded uncertainty from 600 to 800 MHz could then be set at ±0.8 dB and from 800 MHz to 1 GHz at ±1.2 dB.

Cleanup: Type B Evaluation

In order to develop a final value for the expanded uncertainty, it is necessary to make an evaluation of all of the typical contributions to uncertainty to determine which terms are included in the Type A result. In this case, nearly every effect exerts a random contribution due to the considerable variation in test setup and conditions over the allotted time period. Factors such as temperature and humidity effects, ground plane quality due to ground moisture, ground plane warping with temperature , test cable lengths, cable connection quality, cable calibration, etc. all varied significantly over the sample. Table 1 lists some of the typical entries in a Type B uncertainty budget and suggests which ones are likely to be totally random, systematic, or a combination of the two. The listed distribution type repre-

Contribution	Distribution Type	Random	Systematic	AF Cal	RE	RI
Antenna factor	Normal		X		X	
Cable calibration	Normal	X		X	X	
Coupler calibration	Normal		X			X
Receiver/probe linearity	Rectangular	X	x	X	X	X
Receiver level detection	Rectangular	x	X		X	X
Antenna directivity	Rectangular	X			X	
AF variation with height	Rectangular		X		X	
Antenna phase center	Rectangular		X		X	
Field uniformity	Rectangular	X				X
Frequency interpolation	Rectangular	X	x		X	X
Distance measurement	Rectangular	X	x	X	X	X
Height measurement	Rectangular	X	x	X	X	
Site imperfections	Rectangular	X	x	X	X	
Mismatch	U-shaped	X	x	X	X	X
Temperature effects	Rectangular	X		X	X	X
Setup repeatability	Type A	X		X	X	X
Ambient signals	Rectangular	X	x	X	X	
EUT repeatability	Type A	X			X	X

Table 1. *Various Possible Contributions to Measurement Uncertainty, along with Typical Accepted Distribution Types and Possible Contributions to Both Random and Systematic Errors. (For items which might have both random and systematic contributions, a lowercase X represents the typically smaller or less likely contribution.)*

sents the typical accepted distribution type for that entry. Note that most calculated values are assumed to be rectangular, which may often be a more stringent criteria than is necessary.

For the antenna calibration example given, only a few possible systematic errors may need to be accounted for. In general, a purely systematic error is likely to have a U-shaped distribution since it always represents a deviation to one side of the real value. A common example of this type of contribution is that of cable mismatch. Since a perfect match is represented by zero reflection and a mismatch in any direction results in non-zero reflection, the statistical probability that there is some mismatch causes the peaks of the distribution to occur away from the zero (matched) value. The effect of mismatch is to change the detected level by reflecting the power back towards the source. It can also generate standing waves in cables which then serve to increase or decrease the detected signal based on the frequency and cable length. A standing wave is a systematic error in a fixed system for a given frequency. However, over frequency, the effect is either constructive or destructive, giving a net random contribution. In the case of antenna calibrations, the mismatch at the antenna is a part of the calibration, so standing waves are the only contribution of concern. All other cable contributions are included in the cable calibration, which is part of the random error contribution.

Another systematic error contribution which generates a U-shaped distribution is the max-hold height step. For continuous height scanning, this is a function of sweep time versus tower speed. Since the "correct" value is a maximum, any deviation from the correct value can only be less than the correct value. As mentioned previously, this error was verified to be less than ±0.02 dB. This error is a good example of the difference between performing a Type B analysis of an entire test setup or using a mixed Type A and Type B analysis. The height step has both random and systematic components, but the random components would be measured in a Type A analysis. Thus the contribution to the mixed Type B analysis is not the same as that for the Type B-only analysis.

The final and most troubling contribution is that of site imperfections. Fortunately several factors mitigate this contribution somewhat. Temperature changes over the test period caused dimensional changes in the surface of the metal ground plane which would have randomized the effects somewhat. Also, similar to the effects of standing waves, the imperfections in the ground plane will have different effects at different frequencies. Variations in antenna positioning with respect to site defects will randomize these effects as well[6, 9]. Thus there is a high probability that there will be "worst case" points throughout the frequency band which will capture the site imperfection effects in a Type A analysis. However, there is always the likelihood of a significant systematic effect which is not easily determined. It should also be noted that NIS-81 classifies site imperfections as a rectangular distribution. This is largely due to the fact that existing site verification techniques do not provide for individual determination of the random and systematic contributions from the site.

For EMC measurements, a good pair of antennas may be used to determine the normalized site attenuation (NSA) of a test site and use the deviation of that value from theoretical for the site contribution. However, in the case of antenna calibration, this option is circular. Since the uncertainty of an NSA measurement can be no better than that of the antenna calibration, the uncertainty of the resulting antenna calibration could never be better than the NSA measurement plus its uncertainty!

Two options remain for antenna calibrations. The first is to attempt "round-robin" testing to compare one site to others and use the average value as the perfect site. The second method, to be published as an amendment to CISPR 16-1 in 1998, involves using calculable dipoles to verify that a site matches a perfect theoretical ground plane through an exhaustive sequence of tests[10].

Comparisons between antenna measurements made using the same test system on the NIST (Boulder, CO) OATS and the OATS used for the antenna calibrations given in the examples show the variation between the sites to be on the order of 0.5 dB. This variation is of the same order of magnitude as the random uncertainty contribution and thus an individual test is insufficient to draw a conclusion on the systematic error contribution of the site. It should be reiterated here that the assumption of a "golden site" for comparison purposes is not recommended. Instead the use of round-robin testing for determination of a statistically perfect site, or the new CISPR method for site verification is recommended for validation of an antenna calibration site.

Total Uncertainty

The combined standard uncertainty, $u_c(y)$, of a quantity y is computed from the square root of the sum of squares (RSS) of the individual contributions $u(x_i)$. If an individual

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uncertainty component does not directly correspond to the measurand, it must first be converted into the proper form by applying the appropriate conversion factor or function, $u_i(y) = c_i \cdot u(x_i)$. For example, a positioning uncertainty in centimeters must be converted into its effect on the field in dB before it can be applied to the antenna factor uncertainty. Then, for N uncorrelated individual components, the combined standard uncertainty is:

$$u_c(y) = \sqrt{\sum_{i=1}^{N} u_i^2(y)}\,.$$

The expanded measurement uncertainty, U, is then determined by multiplying the combined standard uncertainty by the desired coverage factor, k, which determines the level of confidence in the uncertainty value. Thus, $U = k \cdot u_c(y)$. For the recommended 95% level of confidence, $k = 2 \cdot$

Using 0.5 dB for the expanded uncertainty of the random contribution and 0.75 dB as a rectangular distribution for the remaining contribution due to site imperfections and any other systematic effects,

$$u_c(y) = \sqrt{\left(\frac{0.5}{2}\right)^2 + \frac{0.75^2}{3}} = 0.5$$

Using k =2 this results in a total combined expanded measurement uncertainty of 1.0 dB. Since the random error contribution is a significant portion of the total $(u_c(y)/u(q_k) < 3)$ it is necessary to use an adjusted value for the coverage factor, k_p. In this case, the value is around 2.01, but in the case of only a few samples, this value could be as large as 3 to 14.

Conclusion

While there are inherent difficulties in performing a Type A analysis of a test setup, it is important not to dismiss the concept altogether. It is a relatively simple matter to obtain sufficient measurement data to produce an acceptable measure of the total random contribution to the uncertainty. This has the advantage of providing a measured value and thus limits the number of assumptions necessary to arrive at a total expanded uncertainty value. The ability to prove uncertainty claims with measured data is likely to become more important as new EMC regulations are put into effect.

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Understanding the measurement uncertainties of the bicon/log hybrid antenna

Measurement uncertainty associated with the bicon/log hybrid antenna for radiated emissions and site validation tests relate to many factors, including height dependency, polarization and loading.

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Since their first introduction in 1994 at the Roma International Symposium on EMC,[1] bicon/log hybrid antennas have become very popular in EMC labs worldwide. Because there are no band breaks in frequency sweep, test time and effort are reduced. EMC engineers have assumed that performance of these antennas is simply that of a biconical antenna at a lower frequency range until it transitions to a regular log periodic dipole array (LPDA) antenna at higher frequencies. Questions have been raised about this assumption, and some have suggested that a higher measurement uncertainty (U) should be used due to the characterization of the phase center position and antenna pattern variation from that of a dipole on which emission and site validation standards are based. Very limited research has been conducted on the uncertainty evaluation of these hybrid antennas, despite the fact that more and more EMC engineers have come to realize that predicting and reducing measurement uncertainty has become an important aspect of EMC testing.

Most antenna manufacturers and calibration labs provide individually calibrated antenna factors (AF) with associated U values. A thorough understanding of these values is essential. EMI and normalized site attenuation (NSA) tests are performed over a conducting ground plane. Calibration labs may be able to provide very accurate calibrations of the free-space AFs, which are intrinsic properties of the antennas. Studies have shown that antenna performance can change by a few decibels over a ground plane, and this effect is antenna type specific. In many cases, the performance of a bicon/log hybrid antenna over a ground plane is different from that of a bicon or a log antenna. A good free-space AF with a low U does not always translate into a low U in the EMI or NSA measurement due to the influence from the conducting ground.

This article will address several aspects of measurement uncertainty related to the bicon/log hybrid antenna application. They are: the height dependency of the hybrid antenna AF above a conducting ground plane; the geometry and polarization-dependent AF and NSA measurement; active phase center variation with frequency; antenna beam pattern; and the comparisons of a bicon/log hybrid with separate bicon and log antennas. Some manufacturers also apply capacitive loading on the bow tie elements to improve the low frequency performance of these antennas. This article also explains how this loading impacts the measurement U.

HEIGHT/POLARIZATION DEPENDENCY ABOVE A CONDUCTING GROUND PLANE

AF is defined as the ratio of the incident electric field over the receive voltage at a 50-ohm load connected to the feed point of the antenna. The free-space AF is obtained when the antenna is in free-space and the incident electromagnetic field is a plane wave. The free-space AF is an intrinsic property of the antenna, just like the physical length of a ruler, and should not vary no matter how the calibration is performed. However, just like heat or cold can change the length of a ruler, the environment in which the antenna is used can also impact the AF. EMI and NSA measurements are performed over a conducting ground plane, and unlike temperature, which does not change a ruler all that much, the ground plane can change the AF by as much as 2 or 3 dB depending on the polarization and height. Different types of antennas also interact with the ground plane differently, causing the effect on the AF to be antenna specific.

Figure 1 shows a traditional bicon/log hybrid antenna, while Figure 2 shows an enhanced model for improved low frequency performance. Let us use the traditional model shown in Figure 1 to illustrate the dependency of the AF on height above a conducting ground plane. Figures 3 and 4 show numerically-calculated AFs at heights of 1 m, 2 m, 3 m and 4 m for a horizontally or vertically-polarized antenna. Note again that the EMI or NSA measurements are typically performed for a height scanning from 1 m to 4 m. If we were to use a free-space AF or an AF calibrated at a fixed height to do a measurement at a different height, no matter how accurate the calibration is, there exists an error.

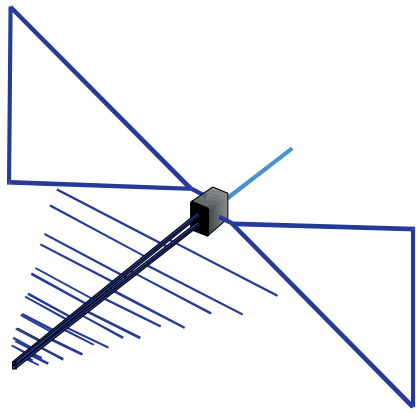

Figure 1. Traditional bicon/log hybrid antenna.

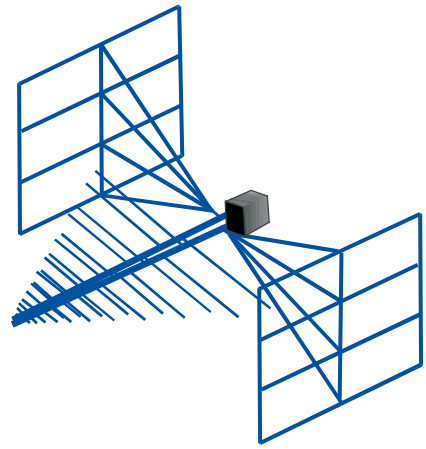

Figure 2. Enhanced model of the bicon/log hybrid antenna.

We may be tempted to say we just need to use a matrix of AFs, so that at each different height we could use a different AF. However, for different frequencies, AFs are different; for different polarizations, AFs are different; for different separation distances, AFs are also different. It becomes a practical issue for calibration in all these different cases and requires applying a complicated multi-dimensional matrix of AFs during an EMI test.

Instead of this complicated web of AFs, is there an acceptable compromise if we are willing to sacrifice a little bit of accuracy? It turns out that the free-space AF provides an acceptable average. As shown in Figure 3, the free-space AF falls right in the middle for most of the frequencies. This is also why the ANSI, CISPR and other international standards have moved toward the use of free-space AF for product EMI test in recent years. However, it is also clear that we may be able to get a near perfect free-space AF, but it would not be perfect for our typical EMI or NSA measurements, simply because a typical measurement condition is not free-space. For total measurement U, in addition to the antenna calibration U obtained from antenna calibration labs, we must assess additional uncertainty values for the antenna and geometry-dependent test setups.

... for different frequencies, AFs are different; for different polarizations, AFs are different; for different separation distances, AFs are also different.

STANDARD SITE METHOD CALIBRATION AND IMPLICATIONS ON NSA MEASUREMENTS

ANSI C63.5 calibration calls for a three-antenna-method calibration over a ground plane, commonly known as the standard site method. In this measurement, the receive antenna is scanned in heights from 1 m to 4 m. NSA measurement as defined in ANSI C63.4 is simply the reverse of the ANSI C63.5 antenna calibration procedure. The only significant difference is that for NSA measurement, the site is the unknown, where for antenna calibration, the AFs are the unknowns.

A common question when following the NSA measurement procedures is "I calibrated my antennas very recently. When I use the AF to do my test, with separate bicons and

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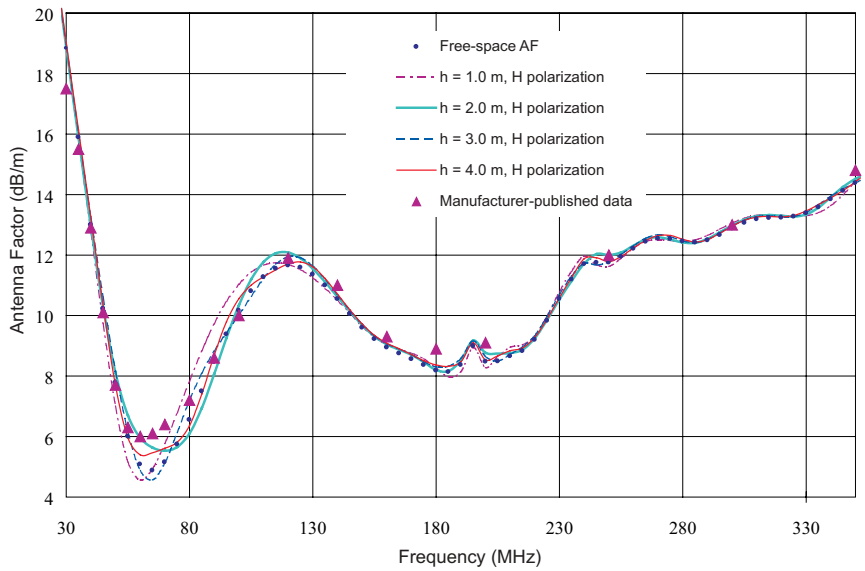

Figure 3. Numerically-calculated bicon/log hybrid AF at different heights for horizontal polarization above a conducting ground plane. Manufacturer-published data are also shown as triangles.

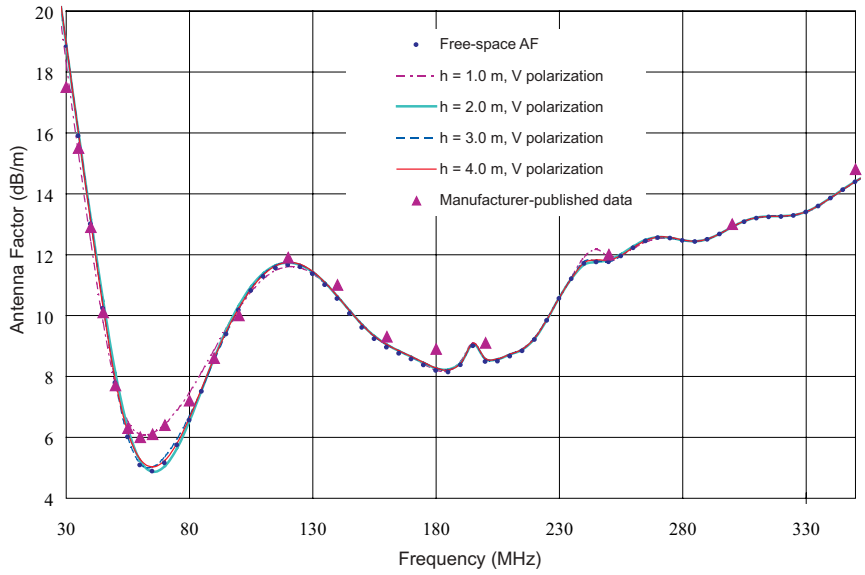

Figure 4. Numerically-calculated bicon/log hybrid AF at different heights for vertical polarization above a conducting ground plane. Manufacturer-published data are also shown as triangles.

log antennas, I can pass the NSA requirement, but when I use the hybrid antennas, I failed the test. Is that due to my antennas or my site?" Other questions are "I need to calibrate my antennas for site validation; which AFs do I need?", and "Do I need free-space AF, 3-m or 10-m calibration, and how about height and polarization?"

It was explained above how AFs change under different geometries and how free-space AF can be used for product EMI test as an acceptable average. To better answer the above questions, we need to quantify exactly how much different geometries affect the performance of a specific antenna. For NSA measurement, since we are dealing with a tighter tolerance, we want to decrease our measurement uncertainties. We will show that the free-space AF approximation becomes inadequate. Let us first look at some antennas calibrated per the ANSI C63.5 standard site method. Figure 5 shows the resulting AFs for a separation distance of 3 m, with the receive antenna scanned from 1 to 4 m in height.

In the standard site method, the discrepancy results not only from the height variation, but also from other factors, such as the non-plane wave illumination of the receive antenna, mutual coupling between the transmit and receive antennas, and the dipole antenna pattern assumption made in the theoretical model.[2] As shown in Figure 5, using a single AF to do an NSA test for all these geometries is a crude approximation. One thing to note is that Figure 5 only shows the difference in the AF for a single antenna under different geometries.

For an NSA measurement, there are two antennas involved, transmit and receive. The resulting difference is the sum of two antennas. For example, at 180 MHz, the free-space AF is different from the AF for the vertical polarization (h1 = 1.5 m) by 2 dB. If a free-space AF were to be used for an NSA site validation measurement, the NSA error just due to the AF difference would be 4 dB (2 dB from the transmit antenna, and 2 dB from the receive antenna). Thus, it is unlikely a site would pass the NSA 4-dB requirement under this condition.

This answers the first question of whether the NSA failure is due to the antenna or site: it is probably neither the site or the antenna calibration that is at fault. Perhaps the answer lies in the method being used, and whether the correct AF is applied. Because the NSA procedure is simply the reverse of the procedure for an ANSI antenna calibration, if the NSA geometry stays the same as the calibration geometry, the errors shown in Figure 5 exactly cancel.

This also answers the second question of which antenna calibration is needed for a site validation test; the geometries for site validation and antenna calibration need to be identical to get the lowest mea-

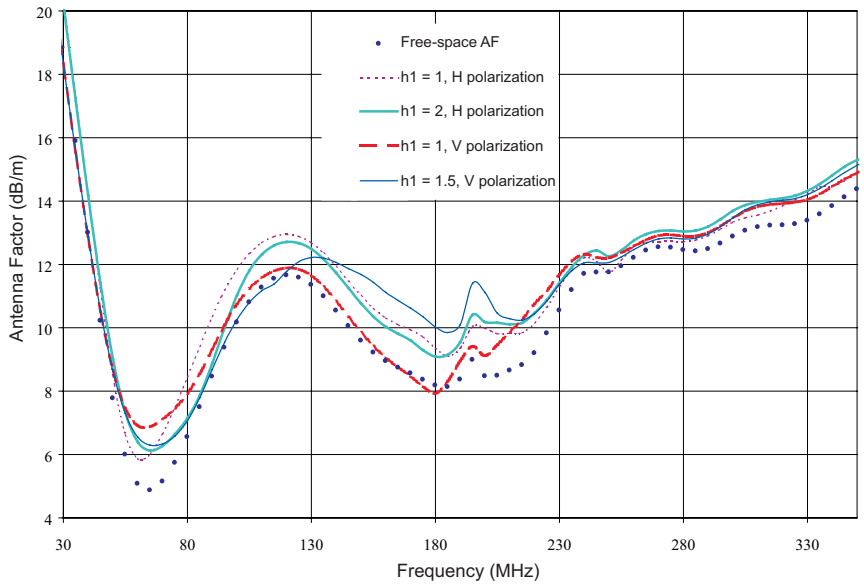

Figure 5. Numerically-calculated bicon/log hybrid AF obtained using the 3-m ANSI C63.5 standard site method. The receive antenna is scanned from 1 to 4 m with a step size of 0.05 m. "h1" is transmit height.

surement uncertainty. However, there is one catch. The antenna calibration site needs to be very good, because any errors generated in the antenna calibration will be transferred to the site validation test.

Let us look at the results in Figure 5 from another perspective. If we assume that the transmit antenna is the equipment under test for an emission measurement, and we use the free-space AF to qualify the EMI from this piece of equipment, the difference between the different geometries and the free-space AF is the error in our measurement. For the example given in Figure 5, this error is 2 dB in some cases. The biconical antenna was also studied for the same circumstances,[2] and the errors were shown to be about 1 dB smaller. For lowest U, a biconical antenna is recommended.

ACTIVE PHASE CENTER VARIATION WITH FREQUENCY

The radiating elements for a bicon/log hybrid antenna move from the bigger elements in the back to the smaller elements in the front as the frequency goes up. The radiating position for a specific frequency is commonly referred to as the active phase center. It appears that the electromagnetic fields are radiated from that center.

Because the phase center moves with frequency, other common questions people ask are: "Where do I measure the distance from the bicon/log hybrid antenna for my test? Should I measure from the tip of the antenna or from the center of the antenna?" A typical answer is to measure from the tip for an immunity test, and from the center for an emission measurement, as specified by the ANSI, CISPR and IEC standards. It is rather clear that this position is just an approximation during the frequency sweep. Thus, uncertainties are introduced in the measurements by assuming a fixed position.

The question arises for bicon/log hybrid antennas from different manufacturers; these antennas may not have the same design or the same length. Are their uncertainties different in an EMI test? The answer is absolutely yes. The next question is whether this error can be estimated. This question may be answered by simply looking at the E^{d}_{max} formulation.*

If we can assume that a bicon/log hybrid antenna acts like a series of dipoles radiating in different positions at different frequencies, E^{d}_{max} should not be calculated for a fixed distance. For example, when we perform a 3-m calibration below 100 MHz, the bow tie elements are active. If the antenna is 1 m long and the reference position is the center of the antenna, we are in fact performing a 4-m test (0.5 m addition for each antenna). Figure 6 shows the E^{d}_{max} values for a horizontally-polarized antenna with the transmit antenna at 1 m height. It shows that more than 2 dB of error can be expected just due to the active elements being different from the reference point.

For antenna calibration, if the active center can be accurately characterized, applying E^{d}_{max} for the correct distances will rectify the error. For a radiated emission test, if the free-space antenna factor is used, this error cannot be amended, and becomes part of the measurement uncertainty. On the other hand, if a biconical or dipole antenna is used, this phase center is well-defined, and a lower measurement uncertainty is achieved. A log antenna can suffer from the same phase center error, but conceivably, a single log antenna is shorter than the hybrid. The phase center error would be smaller. For a critical test where low measurement uncertainty is desired, a simple dipole, bicon and/or log antenna are preferred over the hybrid antenna.

ANTENNA DIRECTIVITY AND BEAM PATTERN

The intent of the ANSI C63.4 NSA and emission measurement is to use a field sensor with a dipole pattern (because the Roberts' Dipole is the undisputed reference). If the antenna pattern is different from a dipole, it would not be an issue if the measurement were performed in a free-space environment as long as we can keep the antenna pointing to the equipment under test at all heights (boresighting).

For a measurement over a conducting ground plane, however, there is a signal reflection from the

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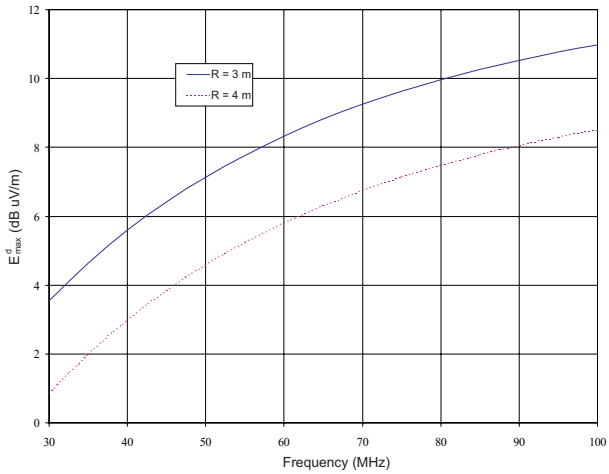

Figure 6. E^d_{max} for 3-m and 4-m separation distance for a horizontally-polarized antenna. The transmit antenna height is 1 m, and the receive antenna height is scanned from 1 to 4 m.

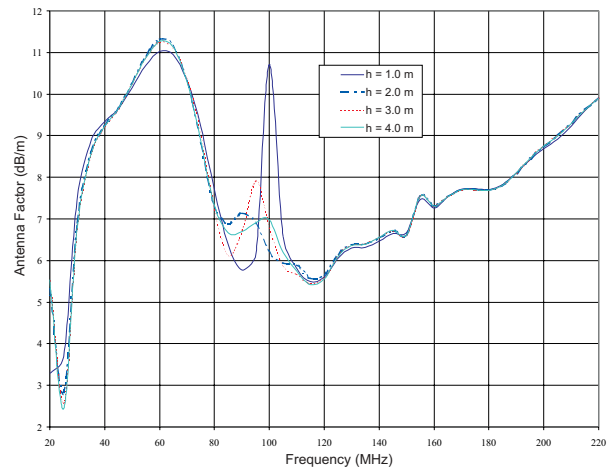

Figure 7. Numerically-modeled AF for a vertically-polarized bicon/log hybrid with an L-shaped end-loading.

ground plane. The reflected field enters the antenna pattern at an angle. The addition of the direct and reflected signal will not add the same way if the antenna pattern is different. There are certain pattern variations from that of a dipole for the hybrid antenna.[1] Any deviation due to the antenna pattern needs to be treated as a source of measurement uncertainty. Bicon antenna patterns have been illustrated to be close to those of dipoles,[4] so, again, this error is smaller for the biconical antennas.

CAPACITIVE-LOADING (LOW-FREQUENCY IMPROVEMENTS) FOR CERTAIN BICON/LOG HYBRID

The VSWR for a hybrid such as the sample shown in Figure 1 is on the order of 20:1 at the 30-MHz range, which means that about 80% of the forward power is reflected back to the source. To generate a certain field level for a radiated immunity test, a huge amplifier is sometimes needed. Several manufacturers introduced capacitive-loading to their antennas, such as shown in Figure 2, to improve the mismatch condition. This improvement is most useful for radiated immunity tests.

For emission tests, the loading elements can couple strongly with the ground when polarized vertically. Figure 7 is an example of the antenna factors at different heights for a vertically-polarized antenna with an L-shaped enhancement. This L-shaped enhancement is a variation of the T-shaped bow tie shown in Figure 2. Even though we could treat such coupling as part of the measurement uncertainty, this value would be unacceptably large, as is the case in Figure 7 (on the order of 5 dB).

The solution for making such an antenna suitable for both radiated emission and radiated immunity testing is to make the end-loadings removable. For immunity testing, the end-loadings are left on to gain the better match (thus requiring a smaller amplifier for a fixed field level). Since the purpose of the immunity test is to generate a given field level, as long as we can measure the generated field, this coupling is not an issue. In addition, most immunity tests are performed in a fully-lined anechoic room, or over a partially absorber-lined ground plane, and this coupling is not as significant.

For an emission measurement, the end-loading should be removed. The antenna in that case would simply perform like a traditional bicon/log hybrid antenna. One thing to note is that the antenna does not need to be calibrated for use in immunity mode, thus saving the cost of calibration for both emission and immunity configurations.

CONCLUSIONS

This article has presented several issues of measurement uncertainties related to the bicon/log hybrid application. Many general measurement uncertainty related issues are not discussed here since they are not particular to this type of antenna. These include cable mismatch uncertainty, site irregularity, site edge diffractions, etc. In an actual measurement, these factors all play important roles in the total measurement uncertainty evaluation. On the other hand, an important issue which tends to be ignored by many EMC engineers is the antenna-specific uncertainties. In most cases, care must be taken when using different antennas and their associated antenna factors. A compromise must be made between the ease of measurement and the accuracy of the measurement.

* *E^d_{max} is a concept introduced by Smith, German, and Pate[3] and later adopted by the ANSI C63 standard site method and NSA formulation. It represents the maximum electric field in a height scan for a tuned dipole with a radiated power of 1 pW.*

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