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Agenda Item:	**x**
Source:	**Motorola PCS Research Labs (Eric Krenz, Paul Moller, Jim Phillips)**
Title:	**Methodology for Characterizing Real-World Radiated Performance of Mobile Phones**
Document for:	**Discussion**

Overview

During the TSGR4#13 meeting a study item was established titled "Feasibility Study of UE antenna efficiency test methods performance requirements" as requested by the operators. The current specification in TS25.101 does not place any requirements for the antenna performance

This document is presented as a contribution for this study item. In this document we present a proposal for the methodology for the radiated performance of mobile terminals. It is expected this document would need some time for review, however we consider based on comments to this document it can be used as a baseline for the study report on this subject.

In the U.S, the CTIA is already adopting a very similar test method. A working group is formed and has drafted a complete test method titled: Method of Measurement for Radiated RF Power and Receiver Performance; draft revision 0.7-B. The CTIA is in the process of requiring test data per their test method as part of their certification process. Although the Method of Measurement for Radiated RF Power and Receiver Performance draft does not yet specify performance criteria, the CTIA will also begin requiring test data to be submitted for an interim period before performance criteria are determined. Being members of CTIA, the system operators have given strong support for this test method.

Introduction

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Radiated performance of mobile phones is a parameter that is critical to the user's experience, but is not accounted for or tested in a very systematic way under present-day standards. All aspects of a mobile phone's performance are specified and tested rather thoroughly via a direct connection (conducted test), but actual radiated performance is simply based on the hope that the antenna works in some way that could be considered acceptable. The critical link establishing actual radiated performance is often verified only by empirical field trials performed when a phone is being put on the market.

Mobile phone system standards around the world appear to specify conducted power & sensitivity quantities, and assume mobile phone antenna efficiencies or "gains" that are wholly unrealistic for small portable devices used on or near the human body. This often leads to dissatisfaction on the part of the carrier and the user when small portable devices are used in systems having been designed based on these antenna performance assumptions.

This document describes a method for characterizing the radiated performance of mobile phones in a manner that is systematic, affordable, and relevant to how the phone is used in the field.

Rationale for Spherical Measurements

Measurement of conducted power that a phone delivers to a 50-ohm load is well understood and reasonably uncontroversial. It is desirable to make measurement of the radiated power and sensitivity of a phone an equally uncontroversial process.

Historically, radiated antenna performance has been evaluated by performing a planar cut of the vertical polarization radiation pattern of the phone in free space. Either a peak value or an averaged value of this single scan is then used to determine the effectiveness of the phone's radiation. While this measurement is certainly easiest to perform, it is not particularly relevant to how a phone is actually used.

Most users hold their cellular phone at angles other than directly vertical or horizontal when held against the head, typically about 30 degrees inclined. In order to capture the radiation towards the horizon from both the left and right side talk positions, at least two planar scans (each at different inclinations) would be required. Additionally, a third scan would need to be taken to capture the on-horizon pattern of the phone when oriented vertically in a shirt-pocket or belt-clip position (as when the phone is being used with a headset accessory). These various phone orientations also imply that there are significant differences in which polarization of a phone's antenna system is likely to participate in the link to the base station.

Furthermore, studies have shown that most cellular phones are used in a highly scattered environment where the radiation at the horizon (theta = 90 degrees) is frequently not the dominant path from the phone to the base station or vice versa. These environments have no line-of-sight to the base station, but instead have energy incident on the phone from a relatively wide variety of angles in the vertical and horizontal planes. In outdoor urban environments, the elevation angle of arrival distribution ranges up to 50 degrees away from the horizon[1], and in 3-dimensional scattered environments (as inside a car or building), this angular distribution can cover the entire sphere around the user. Scattered local environments also depolarize[2] the incident wave, so that both polarizations of the phone's antenna are potentially utilized.

A single azimuthal (on-horizon) pattern scan of one polarization, then, does not adequately deal with different phone use angles or scattering environments, both of which are clear realities. What is needed is a repeatable and realistic method of assessing cellular phone radiated power without the distortions and limitations typical of planar scans. Various alternatives have been proposed and used in the past, ranging from methods that utilize multiple planar scans in order to approximate actual "in-use" conditions, to artificially scattered environments. All of these methods have been used with moderate success, however they all suffer from poor inter-site repeatability or correlation problems to actual in-use conditions.

The spherical scan method has been used with great success in addressing these issues. The spherical scan method simply stated measures the Total Radiated Power (TRP) into the surface of a sphere that completely surrounds the cellular phone and user. This is accomplished by

[1] T. Taga, "Analysis for Mean Effective Gain of Mobile Antennas in Land Mobile Radio Environments," IEEE Trans. On Vehicular Technology, Vol. 39, No. 2, May, 1990.
[2] D. Cox et al, "Cross-Polarization Coupling Measured for 800 MHz Radio Transmission In and Around Houses and Large Buildings," IEEE AP-34, No. 1, January, 1986.

measuring the pattern quantity of interest (Effective Isotropic Radiated Power, EiRP) at each angle and polarization in a complete sphere around the phone, and performing a proper integration to derive the TRP (see appendix A1). If the antenna were enclosed in a perfectly absorbing sphere, the TRP would be the power that would pass through the surface of the sphere. A similar type of measurement can be performed for radiated sensitivity at each angle, which can be integrated to yield a sensible radiated sensitivity figure of merit (see appendix A2). Both of these measurements can readily be related to the analogous conducted parameters of the phone, thus simplifying the task of setting physically sensible criteria (see appendix sections A1.4 and A2.4). While radiated power and sensitivity measurements are appropriate figures of merit for use in certifying the performance of a complete, working phone, radiated efficiency is an analogous measurement that can be used to analyze the performance of an antenna system during the development phases. The same spherical measurement procedure can be used to characterize efficiency (see appendix A3). Finally, this type of measurement is relatively easy to perform using modern motor control and positioning techniques.

Spherical Measurement Procedures

The spherical measurement requires motion in two axes for the device under test, or one axis for the device and one axis for the measurement antenna or probe. Either method may be employed, although moving the measurement probe in one axis is generally preferred largely due to the mechanics involved. This is referred to as the "conic" cut method. The basic procedure is similar to that done with planar scans, only with the addition of one axis of rotation and then repeating the scan in a different plane.

The conic cut method has the advantage that the device under test has to be rotated only in one axis, therefore commonly available positioners can be used. Also, when a phantom human is employed (see later discussion), the device and phantom combination need be rotated about only one axis. The probe antenna is then moved about an orthogonal axis, thus inscribing the full sphere. A commonly used procedure is to set the probe antenna at a location, perform a phi rotation of the phone, then repeat the entire process several times with the probe antenna being at a different theta angle each time. In the end nearly the entire sphere is inscribed and the data need only then be weighted and summed to give the final result.

Most practical implementations use linearly polarized probe antennas and sum the individual results, apply weighting to compensate for non-uniform spatial distribution of the sample points (a step necessary with all spherical methods), and then integrate to determine Total Radiated Power. The final result is a measure in absolute Watts of power that has crossed the surface of the sphere and is available for communications purposes. This power can be converted into an efficiency or figure of merit if the absolute transmitter power is known, or can be directly compared to other devices to give an absolute measure of likely field performance.

For a given positioner system, the test time required for these scans is primarily a function of the angular sampling increment in theta and phi. Empirical studies using conic-cut data at 800/900 MHz and 1800/1900 MHz indicate that theta and phi increments of 15 degrees are more than sufficient to provide consistent results for integrated quantities as defined in appendix A. Some initial studies show that acceptable accuracy is retained even if this is relaxed to 30 degree increments at the low frequency band; further data will be provided when these studies are complete.

To give some idea of test times required for typical tests, some timing data for Motorola's spherical scanning system is provided here. For efficiency scans (i.e., development testing where we connect via a cable to the phone's antenna), with 10-degree increments in theta and phi, an entire spherical scan (including both polarizations and up to 4 frequencies simultaneously) can be completed in approximately 6.5 minutes. For TRP scans of CW-like signals (e.g. AMPS or a phone transmitting in test mode), the same scan for a single channel is also completed in 6.5 minutes. For TRP of digital systems, the complete scan (15-degree increments in theta and phi) for one channel is completed in about 40 minutes. It is believed that these times are reasonably representative of other spherical measurement systems that are currently becoming available.

Use Positions and Human Emulation

The use of a phantom human is an area of ongoing research and has emerged as the best practical way to simulate actual in use conditions in a laboratory environment. Throughout the years several different phantoms have been proposed and implemented. Recently the IEEE sub-committee Scc34-sc2 has developed a phantom for the purposes of evaluating Specific Absorption Ratio in the head of a user of a cellular phone. This phantom is called the Specific Anthropomorphic Mannequin, or SAM for short. The SAM phantom represents the dielectric loading and RF shadowing of an actual human head as best as can be done with single dielectric phantoms. It is based on US army data of actual human heads, and is scaled to the 95th percentile so as to represent a near-worst case human head in close proximity to the cellular phone.

Other phantom representations have been used, such as the salt water column, dielectric cubes and the like. Although all phantoms are not actual humans, the SAM phantom is uniquely tailored to the cellular phone being held against the human head, and is emerging as the standard phantom for all cellular phone evaluations. Using the SAM phantom has the further advantage of maintaining a consistent measurement platform for both the design and evaluation of cellular phones. This is a point that is critical to not only the design of cellular phones, but also to the comparison of laboratory data to actual field data. The importance of using a state-of-the-art phantom cannot be over emphasized when attempting to replicate actual field performance in a laboratory environment.

Ever present is the question of the hand. Except in body worn (belt clip, holster and the like) conditions, nearly all users will be holding the cellular phone with their hand. The hand is a lossy dielectric object that is actually larger than many modern day cellular phone. The impact of such a lossy dielectric object on the field performance of the cellular phone is not insignificant. Some work has been done to develop an appropriate phantom hand for use with cellular phone testing, with varying success. All such examples to date suffer from large repeatability issues, not to mention fabrication and availability problems. As the technology evolves in the area of appropriate lossy dielectric materials, a phantom hand will likely be available and thus should be employed in cellular phone radiated performance evaluation. Until such a phantom is available, the use of hand phantoms has been problematic and has been shown only to increase the measurement uncertainty more than it improves the measurement accuracy. Thus a hand phantom is not widely used at this time for cellular phone radiated performance evaluation.

Test Chamber Performance Criteria

For the most part conventional, well-established techniques can be used in the construction of an antenna chamber for performing spherical assessments of cellular phones. Where there can be significant differences lie in the realization that the cellular phone has a radiation that much more resembles an omni-directional pattern in the 3-D pattern than do many other antennas. It is often the case where some relaxation in chamber performance is possible by utilizing the fact that a particular test antenna has directivity. In such a case the directivity of the test antenna serves to improve the performance of the chamber in regions "behind" the antenna. This is not the case when the radiating structure is nearly omni-directional, and thus many chambers currently in use will have to be evaluated for their suitability. This would be true whether the cellular phone is being evaluated using planar or spherical scans. In fact, due to the integration characteristics of the spherical scan, it has somewhat relaxed reflectivity requirements than is required of simpler planar scans for the same accuracy.

Appendix B describes an efficient method for evaluating a chamber for its performance relative to the desired accuracy. As one would expect, this method uses a nearly perfect omni-directional source that is moved about the axis of rotation of the central positioner. The resultant pattern is then compared to the theoretical ideal pattern and serves to access the suitability of the chamber for omni-directional pattern measurements in general.

Summary

The goal of this test method should be to replicate the actual in-use conditions of cellular phones as much as is practical in a laboratory environment. It must be recognized that perfect replication is not possible, largely due to the huge variabilities present with the usage and deployment of a product into the general population. To that end, this test method proposes to perform a spherical scan of a cellular phone that is in a head-adjacent position of a head (or head and torso) phantom. Such a test method has been shown to be superior to the historical planar scans conducted without the presence of any human or phantom.

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Appendix A: Physical Basis for Integrated Quantities

A1: Radiated Power

Background: Measurement of conducted power that a phone delivers to a 50-ohm load is well-understood and reasonably uncontroversial. It is desirable to make measurement of power radiated by the phone an equally uncontroversial process. The two measurements are represented in figure 1 below.

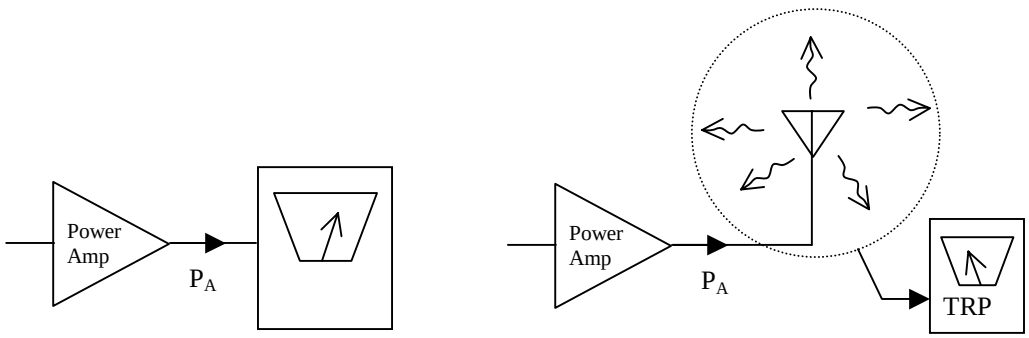


Figure 1: Measurement of conducted power (left) and total radiated power.

P_A = Conducted power (properly, the power available to a 50-ohm load), in W

TRP = Total Radiated Power, the power that is actually radiated by the antenna, in W

The TRP is the sum of all power radiated by the antenna, regardless of direction or polarization. If the antenna were enclosed in a perfectly absorbing sphere, the TRP would be the power that would be absorbed by that sphere. TRP can be related to P_A in this fashion:

$$TRP = P_A \cdot L_m \cdot eff \qquad (1)$$

Where

L_m = Mismatch loss of antenna (relative to 50 ohms)

eff = Radiation efficiency of the antenna

The radiation efficiency, *eff*, is defined in any antenna textbook as the ratio of the power radiated by an antenna to the power delivered to the antenna. It can be seen that the maximum attainable value of TRP is simply P_A, and that this maximum would be obtained if there were no mismatch at the antenna and if the antenna were 100% efficient.

A1.1: Derivation of Total Radiated Power

The Total Radiated Power of a given antenna and source is (see, for example, Stutzman & Thiele, *Antenna Theory and Design*, first edition, 1981, page 33, equation 1-131):

$$TRP = \iint U(\theta,\phi)d\Omega$$

Where $U(\theta,\phi) =$ radiation intensity at each angle in Watts/steradian

Expanding this integral,

$$TRP = \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} U(\theta,\phi)\sin(\theta)d\theta d\phi$$

It is seen that the sin(θ) term results simply from the mathematical expansion of the element of solid angle, dΩ:

$$d\Omega = \sin(\theta)d\theta d\phi$$

The effective isotropic radiated power, EiRP, is defined as (Stutzman & Thiele, page 62, equations 1-226 and 1-227):

$$EiRP(\theta,\phi) = P_T G_T(\theta,\phi) = 4\pi U(\theta,\phi)$$

Where $P_T G_T$ is the product of the power delivered to the antenna and the antenna's power gain. (The equation cited in the reference is actually for the specific case of peak EiRP at the angle of maximum gain, but the reasoning used in the reference produces the above equation for the more general EiRP vs. angle function.)

Then we have

$$U(\theta,\phi) = \frac{EiRP(\theta,\phi)}{4\pi}$$

And the integral for TRP becomes

$$TRP = \frac{1}{4\pi} \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} EiRP(\theta,\phi)\sin(\theta)d\theta d\phi \qquad (2)$$

Thus, if the complete spherical pattern of the EiRP of the phone is integrated with the sin(θ) weighing as described in this equation, the result will be the total power the phone is radiating. It should be noted here that this integration would be modified to yield the same total radiated power if the pattern measurement is expressed in terms of ERP (effective radiated power referenced to a half-wave dipole) rather than EiRP. Specifically, ERP is numerically 2.14 dB less than EiRP:

$$ERP(\theta,\phi) \cong \frac{EiRP(\theta,\phi)}{1.64}$$

so that

$$TRP \cong \frac{1.64}{4\pi} \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} ERP(\theta,\phi)\sin(\theta)d\theta d\phi$$

It must be emphasized that, whether the pattern data itself is taken in the form of ERP or EiRP, use of the appropriate integration will yield numerically the same TRP (as well it should—the phone is radiating the same power in either case).

A1.2: Conversion to Summations of Discretely Sampled Pattern Data

For simplicity, the summations will be derived separately for the two cases where the EiRP data are taken using conic cuts and great-circle cuts. It is assumed in both cases that the measurement points are distributed uniformly in theta and phi.

Conic Cuts: For reference, the Z axis (theta=0 axis) is the long axis of the phone in a free-space test or points straight up out of the top of the phantom's head in a phantom test. A conic cut is defined as a scan of phi from 0 to 360 degrees while theta is fixed at a given value. A series of conic cuts from theta = 0 (probe antenna at zenith) to 180 degrees (probe antenna at nadir) captures an entire spherical pattern.

$M =$ number of samples per conic cut
$N =$ number of conic cuts to form the spherical pattern

$i =$ index for each conic cut, i ranges from 1 to N
$j =$ index for each sample in a conic cut, j ranges from 1 to M

Then the theta and phi intervals are

$$\Delta\theta = \theta_i - \theta_{i-1} = \frac{\pi}{N}$$
$$\Delta\phi = \phi_j - \phi_{j-1} = \frac{2\pi}{M}$$

At this point, a choice must be made as to how samples taken at the edges of intervals are to correspond to the intervals themselves in approximating the integration. For simplicity in the present discussion, we will choose that the EiRP measured at the beginning of a phi interval will represent that entire interval. This will have the effect of discarding the redundant measurement taken at phi=360 degrees. The most correct way to do this would probably be to utilize a trapezoidal rule for approximating the integration, but previous tests done with conic cut data have indicated that the difference in results is not significant.

Substituting the appropriate differentials into equation 2, the summation that approximates the TRP integration in this case is then

$$TRP = \frac{1}{4\pi}\sum_{i=1}^{N}\sum_{j=1}^{M} EiRP(\theta_i,\phi_j)\sin(\theta_i)\frac{\pi}{N}\frac{2\pi}{M}$$

or
$$TRP = \frac{\pi}{2NM}\sum_{i=1}^{N}\sum_{j=1}^{M} EiRP(\theta_i,\phi_j)\sin(\theta_i) \qquad (3)$$

Great-Circle Cuts: For reference, the Z axis (theta=0 axis) is the long axis of the phone in a free-space test or points straight up out of the top of the phantom's head in a phantom test. A great-circle cut is defined here as a scan of theta from 0 to 360 degrees while phi is fixed at a given value. A series of such great-circle cuts from phi= 0 to 180 degrees captures an entire spherical pattern. Note that this coordinate system is defined with respect to the device under test (or the phantom), and may not necessarily correspond with the coordinate system of the specific positioning equipment used to obtain the pattern data.

$M =$ number of great-circle cuts to form the spherical pattern
$N =$ number of samples per great-circle cut

$i =$ index for each sample in a great-circle cut, i ranges from 1 to N
$j =$ index for each great-circle cut, j ranges from 1 to M

Then the theta and phi intervals are

$$\theta_i - \theta_{i-1} = \frac{2\pi}{N}$$
$$\phi_j - \phi_{j-1} = \frac{\pi}{M}$$

Again, a choice must be made as to how samples taken at the edges of intervals are to correspond to the intervals themselves in approximating the integration; we will choose for the present discussion that the EiRP measured at the beginning of a theta interval will represent that entire interval. This will have the effect of discarding the redundant measurement taken at theta = 360 degrees.

The summation that approximates the TRP integration in this case is then

$$TRP = \frac{1}{4\pi}\sum_{i=1}^{N}\sum_{j=1}^{M} EiRP(\theta_i,\phi_j)\left|\sin(\theta_i)\right|\frac{2\pi}{N}\frac{\pi}{M}$$

or
$$TRP = \frac{\pi}{2NM}\sum_{i=1}^{N}\sum_{j=1}^{M} EiRP(\theta_i,\phi_j)\left|\sin(\theta_i)\right| \qquad (4)$$

The absolute value of sin(θ) must be used in this case, because we have performed the unnatural mathematical act of sweeping theta through 360 degrees. Equation 4 can be seen to apply also to the case of conic cuts (cf. equation 3), so can be used for both measurement schemes.

A1.3: Dual Polarization Measurements

In practice, the complete EiRP will be measured at each sample point by measuring its two orthogonally polarized components, $EiRP_\theta(\theta,\phi)$ and $EiRP_\phi(\theta,\phi)$. To accommodate this measurement practicality, we can split the radiation intensity at each angle into two contributions, one from each polarization (power in independent components simply adds):

$U_\theta(\theta,\phi) =$ radiation intensity due to theta component of E-field
$U_\phi(\theta,\phi) =$ radiation intensity due to phi component of E-field

Then equation 2 can be re-derived as

$$TRP = \frac{1}{4\pi}\int_{\theta=0}^{\pi}\int_{\phi=0}^{2\pi}(EiRP_\theta(\theta_i,\phi_j) + EiRP_\phi(\theta_i,\phi_j))\sin(\theta)d\theta d\phi$$

and equation 4 would be rearranged as

$$TRP = \frac{\pi}{2NM} \sum_{i=1}^{N} \sum_{j=1}^{M} (EiRP_{\theta}(\theta_i, \phi_j) + EiRP_{\phi}(\theta_i, \phi_j)) \left| \sin(\theta_i) \right| \quad (5)$$

Equation 5 is proposed as the method for reducing the EiRP pattern data to a single figure of merit. It is applicable for uniform angular sampling in theta and phi. A different formula would need to be derived for non-uniform sampling schemes.

A1.4: TRP Criteria

Since the weighted averaging (total radiated power) expressed in equation 5 is physically related to the power available from the phone and the efficiency of the antenna as expressed in equation 1, a criterion for acceptable TRP can be set based on estimates of these quantities. For example, assume a given MA specifies a mobile station power of 2 W, so that the conducted power P_A in equation 1 is 2 W. Physically, the highest the TRP criterion could be set (and this would require a perfectly matched and 100% efficient MS antenna) would be 2 W.

It is essential that appropriate criteria be established for both free-space and phantom tests, because a given phone that performs very well in free space may perform comparatively poorly on the phantom, and vice-versa. Realistically, candy-bar style phones with fixed antennas achieve roughly 70 to 85% efficiency in free space and 10 to 20% efficiency on the phantom. Typical clamshell style phones with extendable antennas achieve 55 to 75% efficiency in free space and 20 to 30% efficiency on the phantom. Thus, ignoring mismatch loss, sensible criteria for the example MA given above might be a TRP of 2 X 0.55 = 1.1 W in free space and 2 X 0.15 = 0.3 W on the phantom (± the appropriate tolerance according to the MA specification).

It should be clear that, regardless whether the pattern data is recorded in terms of EiRP or ERP, if the appropriate summation (as described in preceding sections) is performed to obtain TRP, the criterion is strictly an absolute power measurement and is not affected. That is to say, adding an extra 2.14 dB to the TRP criteria beyond that obtained from equation 1 is non-physical and, potentially, impossible to satisfy.

The numerical criteria discussed above are broad estimates and meant primarily as an example of how to rationally establish these thresholds. Selecting final criteria that are both attainable and satisfactory is a very challenging task that can only be addressed by agreement within the group. It should be recognized, though, that setting this limit either too high or too low will have negative consequences in system performance or in terms of what products are available to the marketplace. Obviously, setting the criterion too low will allow crummy phones onto the market, resulting in dissatisfied customers and/or higher equipment costs for operators. Setting the criterion too high will drastically change product portfolios that manufacturers can provide to the market. For example, setting a criterion for on-phantom performance that requires antenna efficiency of 75% on the phantom would result in only products of a size and configuration similar to the classic brick phone (e.g. DynaTAC) to pass the standard. More subtly, setting a free-space criterion that is high enough to preclude clamshell-style phones would probably reduce real-world system performance, because these phones (though often having only moderate free-space efficiency) usually have comparatively good efficiency in the actual-use position at the user's head.

If existing cell phones are considered to have acceptable performance in the field, perhaps the best way to establish the TRP criteria is to base them on measured TRPs for several different phones currently on the market. This would ensure that they are high enough to provide sufficient system performance, but not so high as to be out of reach of the manufacturers for phone form factors that are acceptable to the customer.

A potentially viable alternative is not to set fixed criteria at all, but only to establish the measurement standard and make the TRP data for the phones available to the operators. An individual operator could then decide whether to buy a given phone model based on this performance measure together with other factors (such as size, features, battery life, cost, etc.) that that carrier may consider important in marketing

the phone to its own customers. (Or, obviously, individual carriers may want to establish their own criteria for minimum acceptable TRP performance in their systems.)

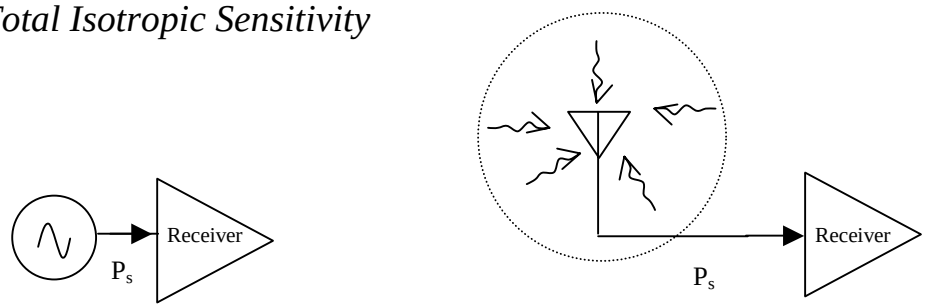
A2: Radiated Sensitivity

Purpose and Disclaimer:

The purpose of this section is to develop a sensible way of reducing a complete spherical pattern of receive-sensitivity data to a single figure of merit, and to give some meaningful examples of this process. The basic principal applied is to compare the DUT's performance to that of a receiver with a perfect (100% efficient) antenna.

We believe this derivation yields the correct result to apply to real chamber measurements of receive sensitivity, but we are not fully satisfied with the mathematical and physical rigor of the underlying arguments. (The present development does not have the same firm theoretical grounding as the development of the Total Radiated Power integrations, for example.) We believe a more refined development will yield the same answer, and plan to address this in a future submission.

A2.1: Total Isotropic Sensitivity



Conducted sensitivity measurement (left) and TIS (right)

Definitions:

We will define a total radiated sensitivity term analogous to Total Radiated Power, referred to as Total Isotropic Sensitivity. Assume plane waves of equal power and equal phase incident on the DUT (device under test) from every direction, and further assume that at each direction, plane waves of equal power in each of the two polarizations (E_θ and E_ϕ) are incident. Now assume the uniform power in all of these waves is simultaneously adjusted so that the power available to the DUT's receiver from the DUT's antenna when immersed in them is the power required for the receiver to operate at its threshold of sensitivity (e.g., a specifiec bit error rate). If the DUT is now replaced with an ideal isotropic antenna with equal gain in each polarization in every direction, the power available from the ideal isotropic antenna from this same uniform incident field is the Total Isotropic Sensitivity[3], TIS.

Define the Effective Isotropic Sensitivities, EIS, as follows:

$EIS_\theta(\theta,\phi) =$ Power available from an ideal isotropic, theta-polarized antenna generated by the theta-polarized plane wave incident from direction (θ,ϕ) which, when incident on the DUT, yields the threshold of sensitivity performance.

[3] This is an intuitively sensible definition, because it compares the DUT's antenna/receiver system to a perfect, 100% efficient antenna that responds equally to either polarization.

$EIS_\phi(\theta,\phi) =$ Power available from an ideal isotropic, phi-polarized antenna generated by a phi-polarized plane wave incident from direction (θ,ϕ) which, when incident on the DUT, yields the threshold of sensitivity performance.

EIS is the pattern quantity that is actually measured in the chamber, by recording power required at each angle and polarization to achieve sensitivity. It is measured by including the same path-loss factor that is used in the chamber to yield EiRP for a transmitting antenna at the same frequency. Note that the EIS terms are defined with respect to a single-polarized ideal isotropic antenna, but the TIS is defined with respect to a dual-polarized ideal isotropic antenna. This is a convenience to make calibration in the chamber correspond with the calibration done for EiRP. That is to say, the same type of path loss terms that are generated when calibrating a chamber to yield EiRP patterns for a transmit test will yield EIS patterns for a receive test as defined here (based on single-polarized isotropic references). TIS is based on a dual-polarized isotropic comparison, because real-world DUTs and propagation are dual-polarized. Proper choice of integration kernels will be seen to reconcile this apparent difference.

In general,

$$EIS_x(\theta,\phi) = \frac{P_S}{G_x(\theta,\phi)} \qquad (6)$$

where P_s is the conducted sensitivity of the DUT's receiver and $G_x(\theta,\phi)$ is the gain relative isotropic (in polarization x) of the AUT's antenna (in this case, including mismatch and ohmic losses) in the direction (θ,ϕ).

A2.2: Derivation of Received Power

Assume a spherical surface centered on the DUT, and calculate the incoming power in the uniform spherical wave described in the definition of TIS. In general, the power flowing into any closed surface can be calculated by integrating the real part of the Poynting vector[4] over that surface:

$$P_{available} = \frac{1}{2}\oiint_S \mathrm{Re}(\vec{E}\times\vec{H})\cdot d\vec{s}$$

where *S* is the spherical surface on which the electric and magnetic fields are evaluated. For purposes of this discussion, the sign convention is chosen so that a positive power indicates a net power flow into the closed spherical surface.

Assume that the spherical surface *S* has a sufficiently large radius r that the far-field approximation can be applied. Then, upon separating the integration kernel into terms for each of two orthogonal linearly polarized components of incoming wave, we have[5]

$$P_{available} = \frac{1}{2\eta_0}\oiint_S (E_\theta^2(r,\theta,\phi) + E_\phi^2(r,\theta,\phi))ds$$

[4] See, e.g., Stutzman & Thiele, *Antenna Theory and Design*, first edition, 1981, page 9, equation 1-34; or Balanis, *Antennas*, x edition, 19xx, page 36, equation 2-9.
[5] This is an analogous development to equations 2-12 and 2-12a on page 38 of Balanis.

where r is the radius of the spherical surface of integration, η_0 is the intrinsic impedance of free space, and $E_x(r,\theta,\phi)$ are the magnitudes of the two components of electric field on the surface *S*. Finally, substituting for the differential element of area, ds, we have

$$P_{available} = \frac{1}{2\eta_0}\oiint_S (E_\theta^2(r,\theta,\phi) + E_\phi^2(r,\theta,\phi))r^2 \sin(\theta)d\theta d\phi \qquad (7)$$

As defined above, the TIS is the power that an ideal isotropic radiator would receive from an incoming spherical wave with equal power in each polarization from every direction, such that the same incoming wave would cause the DUT to operate at sensitivity. Define E_{TIS} to be the magnitude of each of the linearly polarized components of this wave,

$$E_\theta(r,\theta,\phi) = E_\phi(r,\theta,\phi) = E_{TIS}$$

Then equation 7 becomes, for this case,

$$P_{available} = \frac{1}{2\eta_0}\oiint_S 2E_{TIS}^2 r^2 \sin(\theta)d\theta d\phi = \frac{4\pi E_{TIS}^2 r^2}{\eta_0}$$

This is the total power carried in such an incoming wave. However, the power actually received from this same incoming wave by any antenna is

$$P_{received} = \frac{1}{2\eta_0}\oiint_S (G_\theta(\theta,\phi)E_{TIS}^2 + G_\phi(\theta,\phi)E_{TIS}^2)r^2 \sin(\theta)d\theta d\phi$$

where $G_x(\theta,\phi)$ are the antenna's component gains in each polarization as in equation 6.

This can be further simplified to

$$P_{received} = \frac{E_{TIS}^2 r^2}{2\eta_0}\oiint_S (G_\theta(\theta,\phi) + G_\phi(\theta,\phi))\sin(\theta)d\theta d\phi$$

The ideal isotropic dual-polarized antenna envisioned in the above definition of TIS would have a total gain in every direction of 1 (that is, 0 dBi). Therefore, its component gains in each polarization in every direction are 0.5 (that is, -3 dBi), and the power it would receive from this incoming wave (by definition, the TIS) is

$$P_{received} = TIS = \frac{E_{TIS}^2 r^2}{2\eta_0}\oiint_S (\frac{1}{2} + \frac{1}{2})\sin(\theta)d\theta d\phi = \frac{2\pi E_{TIS}^2 r^2}{\eta_0} \qquad (8)$$

For the specific case of the DUT, the power delivered by its antenna to its receiver when immersed in this incoming wave is, *by definition*, the receiver's sensitivity power, P_S, so that

$$P_S = \frac{E_{TIS}^2 r^2}{2\eta_0}\oiint_S (G_\theta(\theta,\phi) + G_\phi(\theta,\phi))\sin(\theta)d\theta d\phi \qquad (9)$$

Furthermore, we can rearrange equation 6 so that

$$G_x(\theta,\phi) = \frac{P_S}{EIS_x(\theta,\phi)}$$

Substituting into equation 9 yields

$$P_S = \frac{E_{TIS}^2 r^2}{2\eta_0} \oiint_S \left[\frac{P_S}{EIS_\theta(\theta,\phi)} + \frac{P_S}{EIS_\phi(\theta,\phi)} \right] \sin(\theta) d\theta d\phi$$

This can be rearranged to yield

$$\frac{E_{TIS}^2 r^2}{\eta_0} = \frac{2}{\oiint_S \left[\frac{1}{EIS_\theta(\theta,\phi)} + \frac{1}{EIS_\phi(\theta,\phi)} \right] \sin(\theta) d\theta d\phi}$$

Substituting this into equation 8 yields

$$TIS = \frac{4\pi}{\oiint_S \left[\frac{1}{EIS_\theta(\theta,\phi)} + \frac{1}{EIS_\phi(\theta,\phi)} \right] \sin(\theta) d\theta d\phi} \qquad (10)$$

A2.3: Conversion to Summations of Discretely Sampled Pattern Data

Using similar notation and mathematics as was used in the Appendix A1 to derive the discrete summation for TRP, it is easily shown that the discrete-sampled version of equation 10 is

$$TIS = \frac{2NM}{\pi \sum_{i=1}^{N} \sum_{j=1}^{M} \left[\frac{1}{EIS_\theta(\theta,\phi)} + \frac{1}{EIS_\phi(\theta,\phi)} \right] \left| \sin(\theta_i) \right|} \qquad (11)$$

Again, the absolute value of sin(θ) must be used in this case, because we have performed the unnatural mathematical act of sweeping theta through 360 degrees. Equation 11 can be seen to apply to both the cases of conic cuts and great-circle cuts.

A2.4: Results for a few Special Cases & Radiated Sensitivity Criteria

Assume that the receiver, environment, and antenna are all at the same temperature, e.g., 290K.

Case 1: The DUT employs a 100% efficient, single-polarized, ideal isotropic radiator: For example, assume the DUT's antenna is an ideal, theta-polarized isotropic antenna. By definition, $EIS_\theta(\theta,\phi)$ is then P_s for every angle, and $EIS\phi(\theta,\phi)$ is infinite at every angle. Then equation 10 becomes

$$TIS = \frac{4\pi}{\oint\!\!\oint_S \left[\frac{1}{P_S} + \frac{1}{\infty}\right]\sin(\theta)d\theta d\phi} = \frac{4\pi}{\oint\!\!\oint_S \left[\frac{1}{P_S} + 0\right]\sin(\theta)d\theta d\phi} = \frac{P_S 4\pi}{\oint\!\!\oint_S \sin(\theta)d\theta d\phi} = P_S$$

In other words, TIS of a 100% efficient, ideal isotropic, single-polarized antenna is just the sensitivity power, P_s.

Case 2: The DUT employs a 100% efficient, dual-polarized, ideal isotropic radiator: $EIS_\theta(\theta,\phi)$ is then $2P_s$ for every angle, and $EIS\phi(\theta,\phi)$ is also $2P_s$ at every angle. (Recall that the EIS is defined with respect to a single-polarized isotropic, and a dual-polarized isotropic antenna must have half the gain in each polarization of a corresponding single-polarized isotropic antenna.) Then equation 10 becomes

$$TIS = \frac{4\pi}{\oint\!\!\oint_S \left[\frac{1}{2P_S} + \frac{1}{2P_S}\right]\sin(\theta)d\theta d\phi} = \frac{P_S 4\pi}{\oint\!\!\oint_S \sin(\theta)d\theta d\phi} = P_S$$

Again, a 100% efficient, ideal isotropic antenna has a TIS that is equal to the conducted sensitivity of the receiver, P_s.

Case 3: The DUT employs a 50% efficient but otherwise ideal, single-polarized isotropic antenna. $EIS_\theta(\theta,\phi)$ is then $2P_s$ for every angle (the antenna is a 3-dB attenuator, degrading the receiver noise figure by 3 dB, so twice the power is required to get the same performance), and $EIS\phi(\theta,\phi)$ is infinite at every angle. Then equation 10 becomes

$$TIS = \frac{4\pi}{\oint\!\!\oint_S \left[\frac{1}{2P_S} + \frac{1}{\infty}\right]\sin(\theta)d\theta d\phi} = \frac{4\pi}{\oint\!\!\oint_S \left[\frac{1}{2P_S} + 0\right]\sin(\theta)d\theta d\phi} = \frac{2P_S 4\pi}{\oint\!\!\oint_S \sin(\theta)d\theta d\phi} = 2P_S$$

So a 50% efficient antenna has a TIS equal to the conducted sensitivity degraded by 3 dB (i.e., twice as large).

This supports two general conclusions. First, the lower limit (best achievable value) for TIS is simply the conducted sensitivity of the DUT's receiver, P_s. This TIS is achieved with a perfectly matched, 100% efficient antenna. Second, the TIS of a real antenna will be the conducted sensitivity of its receiver degraded by the mismatch/efficiency loss of the antenna. Given this, the same examples and guidelines on how to set criteria as discussed at the end of section A2 apply to setting criteria for radiated sensitivity.

A3: Efficiency

It is worthwhile noting that both TRP and TIS, which are measures of actual performance of the phone in the system, reflect the efficiency of the antenna, together with some quantity related to the transceiver (e.g., conducted power or conducted sensitivity). For this reason, efficiency is a useful figure of merit to use to

characterize the antenna on its own. If this is desired, relatively simple mathematics shows that an integration of an antenna's 3-dimensional pattern of gain in dBi, similar in form to equation 2 in appendix A1, yields the radiation efficiency for a phone antenna that is fed via a cable.

Appendix B: Anechoic Chamber Validation

There are many terms in the complete error budget that contribute to the final overall accuracy of the radiated performance quantities measured in an anechoic chamber environment. A few examples of these are the ultimate accuracy of the instrument used to measure absolute power during the calibration process (e.g., a power meter), known gain accuracy of the calibration reference antenna, quiet zone accuracy (i.e., ripple or reflection in the chamber), and connector repeatability. Many of these error contributions can be made quite small at the frequency ranges of interest here. For example, connector repeatability is controlled by proper maintenance and torque of the connectors. Three-dimensional pattern integration of the calibration antenna can yield very precise directivity references. In the case of measuring highly non-directive devices, the two largest single contributors to the error budget are usually the ultimate accuracy of the power measurement instrument, and the quiet zone accuracy (ripple) in the anechoic chamber. This appendix concentrates on the issue of characterizing this critical quiet zone accuracy term. It presents a simple and effective means of qualifying an anechoic chamber's total reflectivity level, which is particularly appropriate to the type of device being measured (i.e., substantially non-directional handset antenna systems).

B1: Chamber Certification

The primary and most essential step in certification of a test site for accurate through-the-air measurements of wireless equipment is to characterize the level of signal reflections at the location that the device under test (DUT) is measured. Even very low levels of reflection introduce large errors in the results. For example, signal reflections of –25 dB will cause a +/- 0.5dB measurement error that is equivalent to a +/- 12% error. In comparing data from two different devices an uncertainty of 24% exists. Diligence and effort can reduce the total reflections to -30 dB and thus reduce the uncertainty to 12%. Determination of this reflection level must be done periodically by measurement because test areas typically degrade as they become cluttered with extraneous objects. Certification and the verification of the test area reflection level can be determined using a simple and economical circular scanning method.

B2: Positioning Equipment.

Certification of the Chamber will require a determination that the reflection levels are within the specification that ensures the expected accuracy. This determination is made by scanning the test area and thus will require a method of moving a test antenna or probe in a path through the area where the DUT will be located. The classical method requires a planar scanner that moves a directional probe in a plane perpendicular to the direct signal path. This is equipment that must be installed for the scan and then removed for testing. An alternative is to use a vertical-axis rotating positioning equipment that is typically found as part of the test system at all through-the-air test sites. Because this equipment is already present, no large delays or expenses are incurred.

B3: Test Area Size and Surfaces.

The test area must provide at least three wavelengths free space between the test area and any surface or any piece of equipment other than a mechanical support for the phone. This support may be either a dielectric support of minimum material or a human test mannequin especially designed for RF simulation of a human. All surfaces, other than the support, exposed to illumination by the antenna on the phone and by the receive antenna must be covered with RF absorbing material designed for the particular frequency of the test. The reflection level of the absorber must be adequate to pass the certification test. Typically, pyramidal absorber one or more wavelengths thick is adequate.

B4: Certification of Integrity

Measurement accuracy is often degraded by reflections of the test signal that can randomly add to and subtract from the direct signal and introduce errors. The expected peak-to-peak error of any particular

environment can be measured directly by scanning the intended location (either outdoor or within an anechoic chamber) of the intended test area for the device under test (DUT). This scan must be done with a suitable measurement probe so that it transverses the maxima and minima of the combinations of the direct and reflected signals . This measurement probe should be a canonical reference design that most closely approximates the radiation characteristics of the intended DUT so that it illuminates the measurement area in approximately the same manner as the DUT. Electric and magnetic dipoles most closely approximate the characteristics of the antennas of portable cellular phones and are therefore very well suited for this task.

B4.1 Circular Path Test Area Ripple Test Description

Vertical electric and magnetic dipoles have omni-directional patterns and therefore their performance is invariant of the azimuth angle. This allows a very economical method of scanning the test area. The dipole may be mounted vertically and displaced from the center of rotation of a conventional azimuth (vertical axis) rotator. As the antenna is rotated, it circumnavigates a circular path that will pass through locations of addition and subtraction of the signals. This circular-path approach is advantageous in that only the typically already-present rotation equipment is needed. The more traditional method of planar scanning of the test area requires additional positioning equipment and is best suited for certification of areas for testing directional antennas. The additional cost and delay needed to use planar scanning equipment is thus avoided if the method described here is employed.

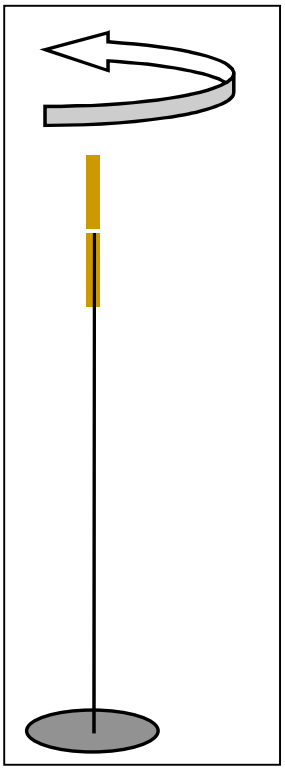
For complete characterization, Different heights and different radii (displacements for the center of rotation) may be measured. The recommended set is two heights and three radii. The heights should be separated by a distance that is more than one-quarter wavelength and less than one-half wavelength. The radii should be in steps of one, two and three quarter wavelengths.

B4.2: Ripple Test – using an orbiting Electric Dipole for Vertical Polarization

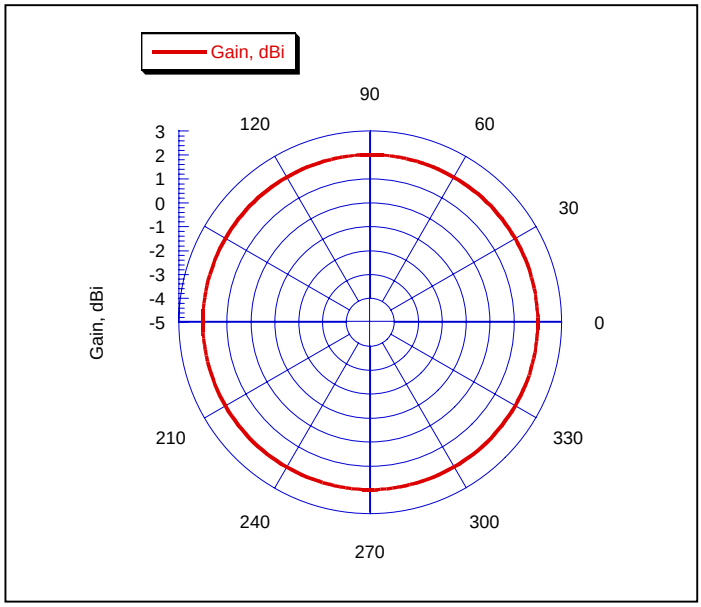
Use a vertically polarized, omni-directional source antenna for this test. An end-fed, sleeve-decoupled, half-wave dipole works well, because there are no cables distorting the azimuth-plane pattern. Precise gain is not important for this chamber validation. The important considerations are that the antenna has an omni-directional pattern in azimuth and that it is decoupled from the attached cable so that there will be no radiation from the cable.

The first step is to ensure that the dipole with the cable is truly omni-directional. Position the dipole precisely on the center positioner's axis of rotation, ensuring that it is exactly vertical. Perform an azimuth scan of the vertical polarization.

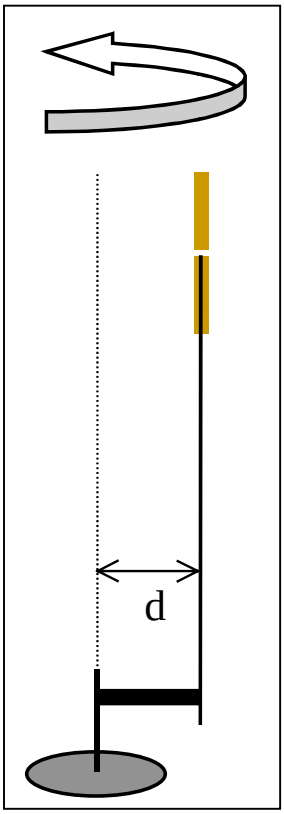
Vertical Decoupled Dipole

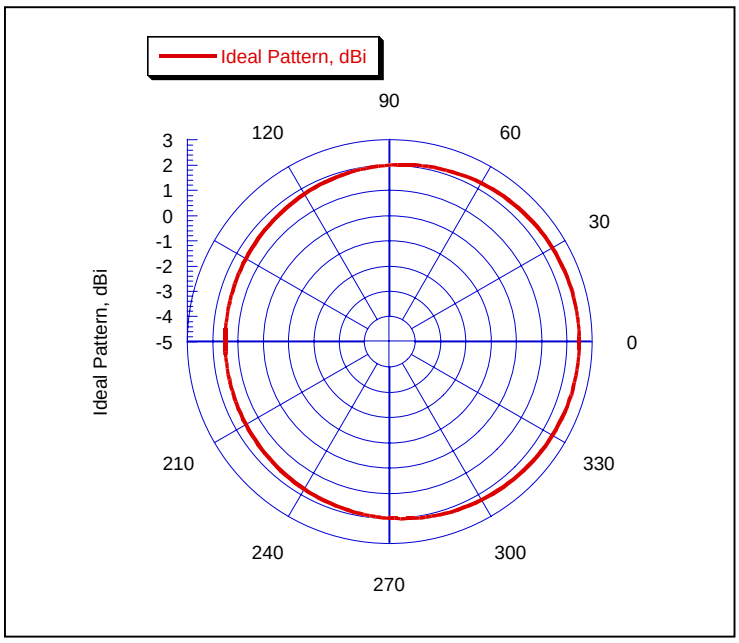


This will give a circular pattern with no ripple, regardless of reflections in the chamber. If pattern is not circular within +/- 0.1 dB, correct source of problem (e.g., dipole not parallel to axis of rotation, reflective dipole mounting means, poor cable decoupling, etc.).

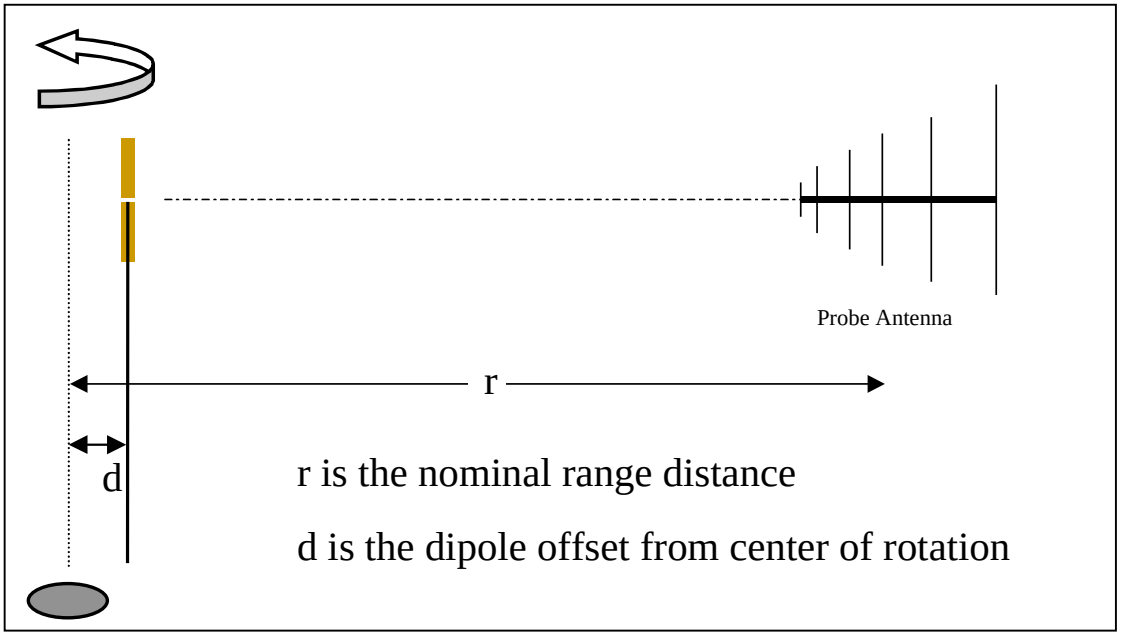


Next, offset the dipole from the center positioner's axis of rotation by distance d, ensuring it is exactly vertical. Perform an azimuth scan of the vertical polarization. In an ideal (no reflection) chamber, this will give an oblong but smoothly changing pattern. The deviation from a perfect circle is due to the different path length from dipole to the probe as the dipole transverses its orbit. (Here, 0 degrees is toward the probe antenna.)

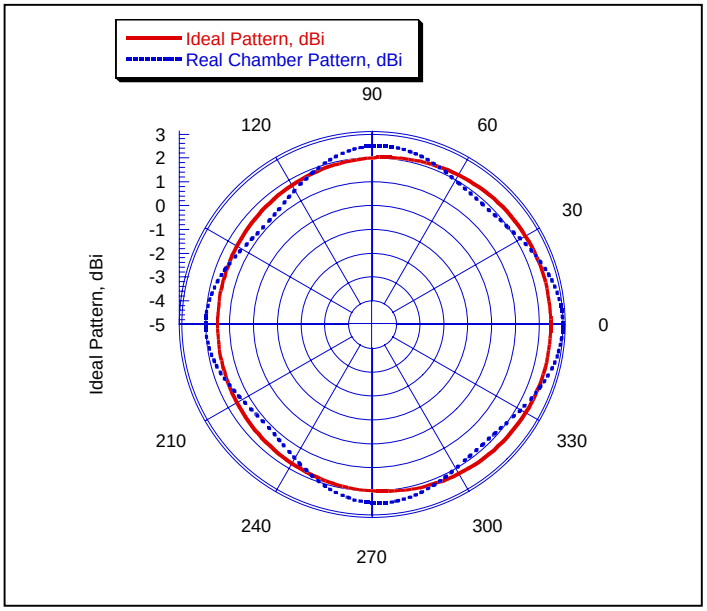




In an ideal, no-ripple chamber, the measured gain at phi=0 (closest distance) is higher than the nominal gain at 90 and 270 degrees by about 20 log (r/(r - d)), and at phi= 180 degrees, it is lower by 20 log ((r + d)/r). This is just from the Friis equation, assuming far field distances for the chamber. This can be used to plot the ideal, oblong pattern response for a given d as in the last chart. There can also be a pattern distortion from the taper of the probe antenna pattern, but this can be ignored for small values of d. For larger values of d or highly directive probe antennas, this contribution can be compensated for through use of the known pattern of the probe antenna.



With a real chamber, reflections will cause ripple in the pattern, compared to the ideal oblong shape. For example, the pattern below shows a peak-to-peak ripple of about 1 dB (+/- 0.5 dB). This shows a figure of merit for the chamber's total reflectivity of about –25 dB (not just reflectivity in a given direction), which is most relevant to measurements of cell-phone-like DUTs (roughly dipole-like patterns). The goal is to make

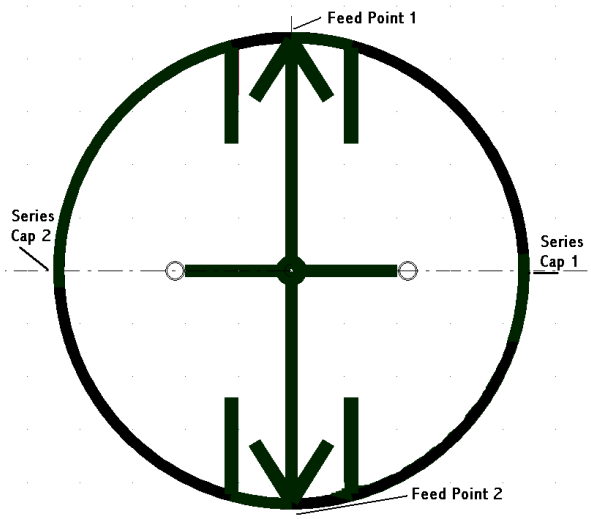


this distortion as small as possible. A practical goal is typically 0.5 dB peak-to-peak ripple that corresponds to a total chamber reflectivity of –30 dB.

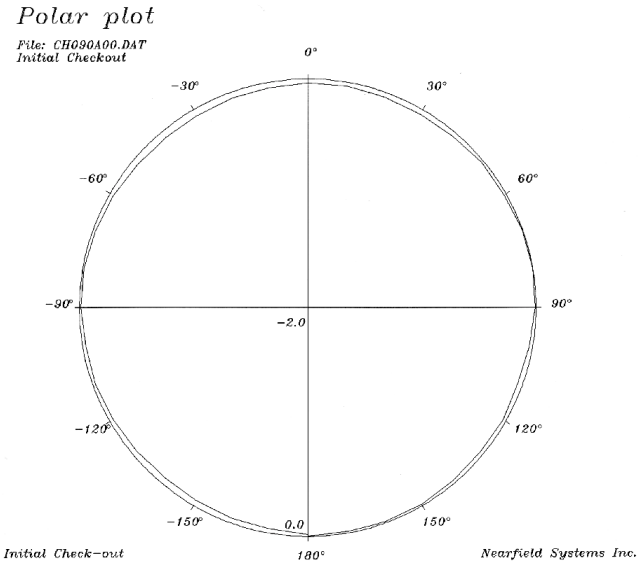
Finally, this experiment should be repeated for several choices of d, different frequencies, and different antenna heights, to establish the dimensions of the chamber's usable quiet zone. Similar experiments using an omni-directional, horizontally polarized antenna (eg, uniform-current loop), can be performed to establish horizontal polarization reflectivity.

B4.3: Ripple Test – using an orbiting Magnetic Dipole for Horizontal Polarization

The Horizontally polarized certification procedure is performed in the same manner as the vertically polarized certification procedure. However, it is much more difficult to construct a truly omni-directional magnetic dipole probe antenna. The best candidate to date is a dual fed, dual capacitively loaded loop antenna. This structure is more complex than the electric dipole but it can be constructed on printed circuit material, as shown below.



The variance in the azimuth radiation pattern can be quite small as shown in the radiation pattern shown below. The radial scale is 2 dB total.



This omnidirectional, horizontally polarized loop antenna is employed in the same procedure as described above for the vertical polarization to complete the chamber validation process.

B5: Comments on Limits for Chamber Ripple

If we expect to obtain on the order of +/-1 dB overall accuracy for radiated performance measurements, we must have a substantially better quiet zone uncertainty, on the order of +/-0.5 dB or less (since it is only one term of several in the error budget). Then, to quantify the quiet zone accuracy, we need a verification tool (i.e., the omnidirectional dipole or loop source) that is substantially better than this +/- 0.5 dB ripple level that we hope to discern. Thus, the omnidirectional test antennas used for chamber validation (not necessarily for chamber calibration) are required to have patterns that are symmetric to +/- 0.1 dB. This level of pattern symmetry is achieved rather easily with an electric dipole rotated about its axis, and can be achieved with some care with the uniform-current (Alford) loops described above[6].

[6] Also refer to, e.g., "Loop Antennas with Uniform Current," Proc. IRE, vol. 32, pp. 603-607, October, 1944.)

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