Gain and Power Parameter Measurements Using Planar Near-Field Techniques

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Abstract—Equations are derived and measurement techniques described for obtaining gain, effective radiated power, and saturating flux density using planar near-field measurements. These are compared with conventional far-field techniques, and a number of parallels are evident. These give insight to the theory and help to identify the critical measurement parameters. Application of the techniques to the INTELSAT VI satellite are described.

INTRODUCTION

It is generally well known that the relative far-field pattern of an antenna is readily available from near-field measurements using either a planar, cylindrical, or spherical scan surface. It need not be seen to be nearly as clear, however, that gain and other power-related parameters are included in the basic theory and that relatively few additional measurements are required to obtain them. There is also a persistent misunderstanding that since the near-to-far transformation involves an integration, directly rather than gain is obtained, and is in error due to lack of information about far-out and back lobes. In the following, we will present the theory, measurements, and results for planar near-field measurements to help clarify these points and describe how gain and other power-related parameters are obtained. When correctly formulated, it is instructive to see how closely the working equations resemble similar far-field relations. We will describe three methods for obtaining gain, along with techniques for determining the effective isotropic radiated power (EIRP) of an antenna-transmitter system, and saturating flux density for an antenna-receiver system.

Theoretical Summary

We begin a discussion of gain and power-parameter measurements by reviewing the basic equations used in the plane-wave scattering matrix theory of antennas [1]. The transmission equations for the situation where the probes are receiving are

\[ b_x'(P) = F' a_0 \int t_{10}(K) \cdot s_{1x'}(K)e^{j\phi K \cdot P} dK \]

\[ b_y'(P) = F' a_0 \int t_{10}(K) \cdot s_{1y'}(K)e^{j\phi K \cdot P} dK \] (1)

\[ b_x''(P) = F'' a_0 \int t_{10}(K) \cdot s_{2x''}(K)e^{j\phi K \cdot P} dK \]

\[ b_y''(P) = F'' a_0 \int t_{10}(K) \cdot s_{2y''}(K)e^{j\phi K \cdot P} dK. \] (2)

In these equations, \( b_x'(P) \) and \( b_y'(P) \) represent the complex outputs of two probes that have nominally orthogonal polarizations, where \( P \) is the \( x, y \) position vector on the plane \( z = d \), and \( a_0 \) is the input to the antenna under test (AUT) as illustrated schematically in Fig. 1. \( F' \) and \( F'' \) are impedance mismatch factors between the respective probes and the "load" port connected to the probes,

\[ F' = \frac{1}{1 - \Gamma_{x}^{s} / \Gamma_{x}^{r}} \]

\[ F'' = \frac{1}{1 - \Gamma_{y}^{s} / \Gamma_{y}^{r}}. \] (3)

where \( \Gamma_{x}, \Gamma_{x}^{r}, \Gamma_{y}^{s}, \Gamma_{y}^{r} \) are, respectively, the reflection coefficients of the load, the first probe, and the second probe. The transmitting properties of the AUT are specified by the plane-wave transmitting coefficients (PWTs) \( t_{10}(K) \), which are a function of the transverse part of the propagation vector \( k \),

\[ K = k_{x} \hat{x} + k_{y} \hat{y}, \] (4)

and the magnitude of the \( z \) component of \( k \) is denoted by \( \gamma \). In a similar way, \( s_{1x'}(K) \) and \( s_{1y'}(K) \) represent the plane-wave receiving coefficients (PWRCs) of the two probes. Consistent with the notation employed by Kerns in the original development, the lower case \( s \)'s and \( t \)'s represent the complete vectors, not merely the \( x, y \) components.

The solution for the AUT parameters is accomplished by a combination of a Fourier transform of the measured data, and then a probe correction of the resulting angular spectra \( D'(K) \) and \( D''(K) \),

\[ D'(K) = \frac{e^{-j\phi K}}{4\pi F'A'} \int B_x'(P)e^{-j\phi K \cdot P} dP \]

\[ D''(K) = \frac{e^{-j\phi K}}{4\pi F'A''} \int B_y''(P)e^{-j\phi K \cdot P} dP \] (5)

\[ D'(K) = \frac{e^{-j\phi K}}{4\pi F'A'} \int B_x''(P)e^{-j\phi K \cdot P} dP \]

\[ D''(K) = \frac{e^{-j\phi K}}{4\pi F'A''} \int B_y''(P)e^{-j\phi K \cdot P} dP \] (6)

\[ \frac{D'(K)}{D''(K)} = \frac{\sigma_{0x}^{n}(K) - \sigma_{0y}^{n}(K) \rho_{y}^{n}(K)}{1 - \rho_{x}^{n}(K) / \rho_{y}^{n}(K)}. \]

\[ \frac{D'(K)}{D''(K)} = \frac{\sigma_{0x}^{n}(K) - \sigma_{0y}^{n}(K) \rho_{y}^{n}(K)}{1 - \rho_{x}^{n}(K) / \rho_{y}^{n}(K)} \] (7)

\[ \frac{D'(K)}{D''(K)} = \frac{\sigma_{0x}^{n}(K) - \sigma_{0y}^{n}(K) \rho_{y}^{n}(K)}{1 - \rho_{x}^{n}(K) / \rho_{y}^{n}(K)} \]

\[ \frac{D'(K)}{D''(K)} = \frac{\sigma_{0x}^{n}(K) - \sigma_{0y}^{n}(K) \rho_{y}^{n}(K)}{1 - \rho_{x}^{n}(K) / \rho_{y}^{n}(K)}. \] (8)

In these equations, some conventions have been employed for compactness. The first is to express the ratio \( b_x'(P)/a_0 \) as the...
admittance in the space medium, \( y_0 \) is the characteristic admittance for propagated waves in the transmission line connected to the antennas, and \( \Gamma_x \) is the AUT's reflection coefficient.

**Gain Measurements**

Three basic methods exist for obtaining absolute power gain from planar near-field measurements. They are referred to as 1) the direct gain measurement, 2) the gain comparison technique, and 3) the three-antenna measurement. In all three methods, three quantities enter into the calculation of the AUT’s gain. These are one or more sets of relative near-field data such as \( B_1'(P) \), the gain of a standard antenna, and one or more power ratio or insertion loss measurements such as \( A^+ \) or \( A^- \) in (5) and (6). The three techniques differ in how the gain standard is employed, and in the number and type of near-field and power-ratio measurements.

**Direct Gain Measurement**

The first method is called a direct gain measurement because it is based on a direct application of the basic equations and the probe is used as the gain standard. In (11) and (12) both \( t_{0x}(K) \) and \( s_{0x}'(K) \) are absolute parameters, with dimensions of length, and are not simply quantities normalized to some maximum value. Therefore, if \( t_{0x}(K) \) is obtained, the gain follows directly from (11). Since \( t_{0x}(K) \) is found from the probe correction (7) and (8), it follows that \( s_{0x}'(K) \), \( s_{0x}'(K) \), \( D'(K) \), and \( D''(K) \) must also be absolute rather than relative parameters. We will now describe the steps involved in determining each of these quantities. The details of obtaining the PWRC for both probes are discussed elsewhere [2]. It is sufficient to note here that the correct magnitude and dimension of \( s_{0x}'(K) \) are achieved through knowing the probe gain and using (12). From measurements of the probe's gain \( G_p'(0) \), polarization ratio \( \rho_p'(0) \), and reflection coefficient \( \Gamma_p' \), and using the relation \( \rho_p'(0) = \frac{\sqrt{y_0(1 - |\Gamma_p'|^2)} G_p'(0)}{\sqrt{\frac{y_0}{4\pi k^2} (1 + |\rho_p'(0)|^2)}} \), which holds for reciprocal antennas,

\[
|s_{0x}'(0)| = \frac{1}{\frac{\sqrt{y_0(1 - |\Gamma_p'|^2)} G_p'(0)}{\sqrt{\frac{y_0}{4\pi k^2} (1 + |\rho_p'(0)|^2)}}}^{1/2}
\]

Here, we assume that the \( m \) component is the main component for the first probe. From similar information about the second probe, we obtain \( |s_{0x}'(0)| \), again assuming that the \( c \) component is the main component for the second probe. In the special, but frequent case when the "second" probe is obtained by a 90° rotation of a linearly polarized probe, \( s_{0x}'(0) = s_{0x}'(0) \).

The measured near-field data are referenced to the input amplitude and phase as outlined in (1), (7), (5), and (6) by the normalization constants \( A' \) and \( A'' \) as defined by (10). These constants are obtained from amplitude and phase insertion loss measurements. The usual approach to obtain \( A' \) is to place the probe at \( P_0 \), set the receiver output to 0 dB (or some convenient level), connect the "generator" and "load" ports directly together as illustrated in Fig. 2, and observe the change in amplitude and phase. We denote the resulting
amplitude and phase changes by \( a'_e \) and \( \psi'_e \), respectively; then

\[
A' = \frac{a_e}{b_e} \cdot \frac{a'_e e^{i\psi'_e}}{b'_e} = \frac{\Gamma_e}{\Gamma'_e} \cdot \frac{a'_e e^{i\psi'_e}}{b'_e},
\]

where \( \Gamma_e \) and \( \Gamma'_e \) are the reflection coefficients for the generator and load ports. The amplitude change is most accurately measured with a calibrated attenuator as shown in Fig. 2 to reproduce the original level when generator and load are connected together.

The magnitude of this insertion loss is approximately equal to the ratio of the AUT maximum gain to probe gain,

\[
(a'_e)^2 = \frac{G_e(K_0)}{G_e(K_0)}
\]

and for typical antennas and probes is on the order of 25-40 dB. Since any error in this insertion loss causes an identical error in the gain of the AUT, it is important to measure \( a'_e \) as accurately as possible. The source output \( P_0 \) must remain constant during the measurement, and the input must be well isolated and stable. To improve accuracy, the measurement is generally repeated a few times to make sure that connectors are repeatable and measurements consistent to a few hundredths of a decibel. Another approach to improving gain accuracy is to use as large a probe as possible without increasing multiple reflection errors and filtering desired parts of the wide angle pattern. The larger probe will increase signal-to-noise ratio, reduce the magnitude of \( a'_e \) and therefore improve its accuracy, and also reduce probe-gain error.

The parameter \( A'' \) can be obtained from a similar process with the second probe at \( P_1 \). If a single probe is used with either rotation or polarization switching, \( A'' \) is also conveniently measured from the ratio of the probe outputs due to the combined changes in "probes" and position from \( P_0 \) to \( P_1 \):

\[
A'' = A' \frac{b'_e(P_0)}{b'_e(P_1)}.
\]

This approach eliminates the movement of the generator and load ports a second time, and the observed amplitude change will generally be less, thereby improving its accuracy. That is,

\[
\frac{b'_e(P_0)}{b'_e(P_1)} < |a'_e|.
\]

The quantities \( b'_e \) and \( b'_e \) are the outputs at the probe terminals. The insertion losses must be measured at these terminals and calibration of the probes must include adapters, connectors, etc., up to these terminals. If a dual-mode probe and switching are used, the transmission line to the switch must either be considered a part of the probe or properly accounted for in determining \( A' \) and \( A'' \).

The insertion loss measurement usually requires either moving and flexing the transmission line, inserting a substitute section of line with known properties, or removing the AUT.
and probe and translating of the support tower for direct connection. The antenna type and frequency will generally dictate the best method. At frequencies where flexible line or rotary joints can be used without introducing significant signal changes, this is an attractive choice, for then the alignment and mounting of the AUT are not disturbed and the measurement can be repeated quickly to improve accuracy. If the insertion loss technique is not feasible, power measurements at the generator and load ports with the appropriate impedance mismatch correction will give the magnitude of $A^*$ and its phase is arbitrary. $A^*$ can then be found from the ratio technique, (15).

To understand the factors affecting gain measurements, we note the form of the gain equation for a special, and yet very frequently encountered, example of the direct measurement. The special case assumes that the probe is polarization matched to the AUT in the direction of the AUT main beam. Specifically, if the main beam is in the direction $K_0$ and the probe is polarization matched to the AUT for $K_0$, (5), (7), (11), (13), and (14) reduce to

$$G_s(K_0) = \left( \frac{4\pi}{\lambda^3} \right)^2 \frac{1}{\left(1 - \left| \Gamma_{s} \right|^2 \right)^2} \left| \frac{b^{\prime}_{s}(P_0)}{a_{s}} \right|^2$$

and

$$\left| \frac{\delta_{s}b_{s}B_{s}^{\prime}(P_{s})e^{-i\theta_{0}-\theta_{P}}}{\mid a_{s} \mid^2 G_{s}(K_0)} \right|^2$$

for the partial gain of the main component. Five factors determine the gain: 1) a constant involving the wavelength; 2) an impedance mismatch factor involving the various reflection coefficients which is typically on the order of 0.01–0.5 dB; 3) the complex sum of the relative near-field data, which can be thought of as approximately the “volume” under the near-field amplitude pattern for a collimated beam; 4) the insertion loss $|a_{s}|^2$, and 5) the gain of the probe. Of these, the last three are by far the most important since their accuracy will ultimately determine the accuracy of the gain measurement.

For later applications, we compare the gain equation for near-field measurements to one for a hypothetical far-field measurement involving the same two antennas and the same assumption of polarization match. Assume that the probe is placed at a distance $d_f$ in the far field of the AUT in a reflection-free environment. Further, assume that the input to the AUT is $a_f$ and that by some means the far-field insertion loss

$$|A_f|^2 = \left| \frac{a_f}{b_f(K_0)} \right|^2$$

is measured. The AUT gain is then obtained from the Friis transmission equation written in the following form to show its similarity to (17):

$$G_s(K_0) = \left( \frac{4\pi}{\lambda^3} \right)^2 \frac{M_f}{a_f} \left( \frac{\lambda d_f}{\lambda d_s} \right)^2$$

The impedance mismatch factor $M_f$ will depend on how the insertion loss is measured. If it is measured in the same way as the near field, that is by connecting generator and load together, $M_f$ will be identical to the mismatch factor in (17a). The equation for this measurement has four of the five factors noted for the near-field measurement, the probe gain and insertion loss have identical roles, and the product $\lambda d_f$ replaces the product of $\delta_{s}d_{s}$ times the sum of the near-field data. This comparison makes the near-field gain measurement less obscure and also provides two additional results that are useful in other applications. The first will be used in deriving the relation for near-field saturating flux density measurement and is obtained by equating (17a) and (19), since both represent the gain of the AUT. Canceling common factors gives the first result,

$$\frac{b^{\prime}_{s}(P_0)a_f}{b_f(K_0)a_0} = \frac{\lambda d_f}{\lambda d_s} \left[ \frac{\delta_{s}b_{s}B_{s}^{\prime}(P_{s})e^{-i\theta_{0}-\theta_{P}}}{\mid a_{s} \mid^2 G_{s}(K_0)} \right]$$

The second equation gives a new version or derivation of the near-zone correction for two arbitrary antennas. If we let the “far-field” distance $d_f$ be the same as the near-field distance, the apparent gain calculated by the Friis transmission in (19) is no longer the actual gain. If we add a factor $N(d)$ to (19) to represent the near-zone correction, the modified relation now gives the true gain. Since $d_f = d$, the subscript can now be dropped, and since

$$a_f = a_0 \quad b_f(K_0) = b^{\prime}(P_0)$$

if we choose $P_0$ to be along the direction $K_0$. Equating the modified (19) and (17a) and solving for $N(d)$ gives

$$N(d) = \left( \frac{\left[ b^{\prime}_{s}(P_0)e^{-i\theta_{0}-\theta_{P}} \right]}{\lambda d_f} \right)^2$$

Sample Results and Comparison with Other Measurement Techniques

The direct measurement technique has been used extensively on a wide variety of antennas with very good success. It has been used on reflectors, phased arrays, shaped-beam offset reflectors, and horn antennas. To verify consistency with other measurement techniques, a number of comparison tests have been conducted using the near-field technique and far-field methods on the same antenna. Examples of three such comparisons are shown in Table 1.

Gain Comparison

The second method, a gain comparison technique, is useful in cases where neither the insertion loss nor the power
measurements to determine $A'$ and $A''$ are practical, or when the probe gain is not known. In the special case where the probe and AUT are polarization matched, the development of the gain equation is simplified by using (17). An abbreviated form of this equation for the probe AUT combinations is

$$U_s(B_0) = C_M \frac{|B_0 \sum_{i} B_i^* (P_i)e^{-j\Phi_i}P_i|^2}{|A'_m(P_0)|^2 G_s(K_0)}.$$  \hspace{1cm} (22)

A similar relation for the same probe measuring a standard antenna whose gain in the direction $K_0$ is known is

$$G_s(K_0) = C_M \frac{|B_0 \sum_{i} B_i^* (P_i)e^{-j\Phi_i}P_i|^2}{|A'_m(P_0)|^2 G_s(K_0)}.$$  \hspace{1cm} (23)

Here, $C$ is the constant $(4\pi/\lambda)^2$, $M$ is the mismatch factor, the $ap$ and $sp$ subscripts denote, respectively, the quantities for the AUT-probe and standard-probe measurement. We have made the normalization points $P_0$ and $P_1$ explicit in the reference to the normalization constants to show that they do not have to be the same point for the two antennas. The z-separation distance in the two measurements may also be different since it does not appear in the gain equations. Taking the ratio of (22) and (23) and solving for $G_s(K_0)$ gives

$$G_s(K_0) = \frac{(1 - |\Gamma_0|^2)|B_0 \sum_{i} B_i^* (P_i)e^{-j\Phi_i}P_i|^2}{(1 - |\Gamma_0|^2)|B_0 \sum_{i} B_i^* (P_i)e^{-j\Phi_i}P_i|^2} \frac{|a_0 b_0 b_0^* (P_0)|^2}{|a_0 b_0 b_0^* (P_0)|^2} G_s(K_0).$$  \hspace{1cm} (24)

We note for comparison the gain equation for a far-field gain comparison measurement,

$$G_s(K) = \frac{(1 - |\Gamma_0|^2)(\lambda d_{ap})^2}{(1 - |\Gamma_0|^2)(\lambda d_{ap})^2} \frac{|a_0 b_0 b_0^*(P)|^2}{|a_0 b_0 b_0^*(P)|^2} G_s(0).$$  \hspace{1cm} (25)

Here, $d$ is the separation distance, $a_0$ the input, and $b_0$ the output signals. The $ap$, $a$, and $sp$ subscripts denote, respectively, the AUT-probe and standard-probe combinations. The similarity to (24) is again striking with $\lambda d$ replacing the near-field data sum, lattice area product.

In the more general case where both sets of near-field data are required for accurate gain calculation, such as when a linear probe is used to measure a circularly polarized test antenna, a more general treatment is required. The details for this situation can be found in [3]. In summary, gain comparison requires relative near-field measurements of the AUT and standard using the same probe, the gain of the standard, and the relative amplitude measurement implied by (24). To be successful, a series of steps must be carried out in a relatively short time. The inputs between the AUT and the standard must be switched, the two antennas accurately positioned and aligned, and the probe moved between $P_0$ and $P_1$. Simultaneously mounting both the AUT and standard on the positioning tower, either in opposite $z$ directions or side by side, facilitates these changes.

Although this approach has certain advantages, it can increase some error sources. For example, if the standard is much smaller than the AUT, multiple reflection and truncation errors in $a_0 B_0^* (P)$ can be significant. This is illustrated in the results of a comparison measurement where the AUT was a 1.2-m reflector, the standard a pyramidal horn, and the probe an open-ended waveguide. The AUT and the horn were mounted on opposite ends of a rotator to allow for rapid positioning and alignment. The measurement was repeated at a sequence of $z$ distances to evaluate the effect of multiple reflections with the results shown in Fig. 3. As shown, a direct gain measurement was also carried out to check on consistency and accuracy. Table II shows an estimate of errors for the two types of measurements.

### Three-Antenna Measurement

The third method is a three-antenna measurement where two different probes are used and near-field scans are made for all three antenna combinations. If insertion loss measurements are made with each combination, the three gain equations are

$$G_s(K) G_{p1}(0) = G_{ap2}(K)$$  \hspace{1cm} (26a)

$$G_s(K) G_{p2}(0) = G_{ap3}(K)$$  \hspace{1cm} (26b)

$$G_{p1}(0) G_{p2}(0) = G_{p1p2}(0).$$  \hspace{1cm} (26c)

The gains on the right of (26) are obtained by using the fictitious value of $1$ for the probe gain in the data processing, or in the case of polarization match, solving for the gain product of the AUT and probe in (17). The subscripts $ap1$, $ap2$, and $p1p2$ denote, respectively, the three configurations, AUT-probe 1, AUT-probe 2, and probe 1–probe 2. From these three equations, $G_s(K)$, $G_{p1}(0)$, and $G_{p2}(0)$ are found. However, if the insertion loss is measured for the first pair and a relative amplitude measured between the second and third pairs as in the comparison measurement, the resulting equations are

$$G_s(K) G_{p1}(0) = G_{ap1}(K)$$  \hspace{1cm} (27)

$$G_s(K) = G_{ap2} (K) = G_{p1p2}(0) = G_{p1p2}(0).$$  \hspace{1cm} (28)

and only $G_s(K)$ and $G_{p1}(0)$ are obtainable. This technique is seldom used because it requires three near-field scans, and the probe to probe measurement produces relatively large errors due to multiple reflections and truncation.

### System Power Parameters

In addition to antenna gain, two other system power parameters associated with communication satellites are also easily obtained from the near-field measurements. These are the effective isotropic radiated power (EIRP) and saturating

<table>
<thead>
<tr>
<th>Test Antenna</th>
<th>Frequency (GHz)</th>
<th>Measured Gain Using</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scalar horn</td>
<td>8.0</td>
<td>22.03</td>
</tr>
<tr>
<td>Cassegrain reflector</td>
<td>60.0</td>
<td>46.30</td>
</tr>
<tr>
<td>Microstrip array</td>
<td>1.28</td>
<td>30.50</td>
</tr>
</tbody>
</table>

**TABLE I**

GAIN COMPARISON TESTS
To obtain the near-field relations for measurement without access to the AUT input port, we note that the net input power to the AUT is
\[
\Phi_j = \frac{\eta_0 |\Phi_0|^2 (1 - |\Gamma_2|^2)}{2}
\]
and the net output power from the probe, when it is at the reference position, is
\[
\Phi_0(P_0) = \frac{\eta_0 |b'_0(\Phi_0)|^2 (1 - |\Gamma_1|^2)}{2}.
\]

The EIRP in the direction denoted by \( \Phi_0 \) is then obtained from the product of \( G_s(\Phi) \) and \( \Phi_0 \) in a way similar to the derivation of (17):
\[
EIRP(\Phi_0) = (4\pi)^2 \left( \frac{|1 - \Gamma_1 \Gamma_2|^2 \Phi_0(\Phi_0)}{(1 - |\Gamma_2|^2)(1 - |\Gamma_1|^2) \Phi_0(\Phi_0)} \right)
\]
\[
\left( \frac{\delta_2 \delta_1 \delta_3 \delta_4 \Phi^2(\Phi_0) e^{-ik_0 P_0 \Phi_0}}{\lambda^4} \right).
\]

EIRP is then obtained from the relative near-field data \( B_0^2(\Phi) \) used to obtain the patterns and a measurement of the probe output power at the reference point.

The flux density required to saturate the system receiver is normally measured on a far-field range and is given in terms of the net input power \( \Phi_i \), and the gain \( G_i \) of the source antenna by the relation
\[
S_o = \frac{\Phi_i G_i}{4\pi d^2}.
\]
The power input to the source is that sufficient to saturate the receiver.

The development of a similar near-field equation begins with the concept of placing the probe on a far-field range in place of the source antenna, as shown in Fig. 4. In the general case, the main beam of the AUT is at $K_0$ and may not be along the antenna's $z$ axis; the probe is also rotated by the same amount so the gain of the probe appearing in (33) is $G_p(K_0)$. Using (30) and (33), the flux density for this hypothetical situation is

$$S_0 = \frac{\delta \gamma G_p(K_0)}{4 \pi d^2} = \frac{\eta_0 |a|}{8 \pi d^2} (1 - |\Gamma|^2) G_p(K_0).$$

In principle, the gain of the antenna could be obtained from this far-field measurement using the Friis transmission equation,

$$G_e(K_0) = \left(\frac{4 \pi}{\lambda^2}\right)^2 \frac{M}{a_f} \frac{b_e(K_0)}{a_f} \frac{(\lambda d)^2}{G_p(K_0)}.$$

Here, $a_f$ is the input to the probe that produces the output at the AUT $b_e(K_0)$ which is sufficient to saturate the receiver. The $f$ subscript denotes the far-field measurements and $K_0$ denotes the direction of the main beam for the AUT.

If the AUT and the probe are now placed on a near-field range in the usual orientation and data obtained for the near-field gain equation, the antenna gain from (17a) is

$$G_e(K_0) = \left(\frac{4 \pi}{\lambda^2}\right)^2 \frac{M}{a_f} \frac{b_e(P_0)}{a_f} \frac{|\delta \gamma \Sigma B_0^e(P_0) e^{-ik_0P}|^2}{G_p(K_0)}.$$

If the input to the probe $a_e$ is adjusted so that the receiver is again saturated,

$$|b_e(P_0)|^2 = |b_e(K_0)|^2$$

and therefore

$$|a_e|^2 = |a_e|^2 \frac{(\lambda d)^2}{|\delta \gamma \Sigma B_0^e(P_0) e^{-ik_0P}|^2}.$$

Substituting this value of $|a_e|^2$ into (34) gives the equation for near-field measurement of saturating flux density:

$$S_0 = \frac{\eta_0 |a_e|^2 (1 - |\Gamma|^2) G_p(K_0)}{8 \pi} \frac{(\lambda d)^2}{|\delta \gamma \Sigma B_0^e(P_0) e^{-ik_0P}|^2}.$$

This equation has been obtained by considering hypothetical far-field and near-field gain measurements which require power ratio measurements such as $|b_e(P_0)|/|a_e|^2$. The final equation for flux density does not require such power ratios, however, or access to the output port of the AUT. Only the usual near-field data and the probe input power measurements are required.

The first application of near-field techniques to the measurement of RFD and saturating flux density was accomplished at Hughes Aircraft Company in El Segundo, CA. The measurement facility was designed for the INTELSAT V1 satellite.

1 Reference to Hughes Aircraft does not imply endorsement by the National Bureau of Standards.

2 The INTELSAT spacecraft is being developed by an international team of contractors for the International Telecommunications Satellite Organization (INTELSAT). The views expressed herein are not necessarily those of INTELSAT.
and has been used on a number of other satellites as well. Located at the Space and Communication Group headquarters, it provides an excellent environment for satellite testing. Environmental factors are carefully controlled, and the antennas are supported in a fixed position during the testing. Additionally, it is adjacent to the subsystem and system integration facilities, and the antenna tests can be accomplished without the transportation and handling required on a far-field range.

The planar measurement system covers an area of 6.4 m on each side, with scan speeds of 3.8 and 8.9 m/min, respectively, in the x and y directions. The probe position is measured with a laser interferometer with resolution of 1 μm. In its present configuration, the system will perform measurements in the frequency bands from 1.5 to 14.5 GHz.

The INTELSAT VI satellite shown in Fig. 5 is the world's largest and most complex commercial communication satellite. It is spin stabilized and designed to operate in one of three geosynchronous orbital regions. The RF payload is mounted on the despun section of the satellite, and consists of 50 communication transponders covering C band (6 GHz uplink/4 GHz downlink) and Ku band (14 GHz uplink/11 GHz downlink). The RF system is visible below the antennas as the system is configured for near-field testing in Fig. 6. The antenna subsystem consists of two spotbeam antennas, two global communication antennas, and two large hemi/zone feed networks along with their offset reflectors. The hemi/zone feed arrays simultaneously form six beams. These include four zone beams shaped for the particular land areas covered, and two hemispheric beams. The networks are reconfigured for each of the three geosynchronous orbital locations to produce beams appropriate for the earth coverage region. This reconfi-
The direct gain measurement technique was used when only the antennas (without the RF system) were being tested, and access was available at the input or output of the feeds. Fig. 7 is a schematic of the test setup. The results of one measurement are shown in Table III, along with comparison to a similar far-field test.

Figs. 8 and 9 show the test setup for the EIRP and flux density measurements. Relative near field data \( R^2 \) were obtained in the usual manner using the location of the maximum amplitude as the reference point \( R_0 \). The probe was then placed at this point for the power measurements. In saturating flux density measurements, the power input to the probe was gradually increased until the transponder under test
Fig. 8. Measurement system schematic for EIRP measurement.

Fig. 9. Measurement system schematic for saturating flux density measurement.
TABLE IV
SATURATING FLUX DENSITY MEASUREMENT RESULTS FOR INTELSAT VI ANTENNAS

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Channel</th>
<th>Measured SFD (dBW/m²)</th>
<th>Predicted SFD (dBW/m²)</th>
<th>Delta</th>
</tr>
</thead>
<tbody>
<tr>
<td>Global A</td>
<td>9</td>
<td>-81.49</td>
<td>-81.88</td>
<td>+0.39</td>
</tr>
<tr>
<td>Global A</td>
<td>10</td>
<td>-81.15</td>
<td>-81.41</td>
<td>+0.26</td>
</tr>
<tr>
<td>Global A</td>
<td>11</td>
<td>-81.29</td>
<td>-81.6</td>
<td>+0.31</td>
</tr>
<tr>
<td>Global A</td>
<td>12</td>
<td>-81.83</td>
<td>-81.76</td>
<td>-0.07</td>
</tr>
<tr>
<td>Zone 1</td>
<td>1-2</td>
<td>-81.39</td>
<td>-81.35</td>
<td>-0.04</td>
</tr>
<tr>
<td>Zone 1</td>
<td>3-4</td>
<td>-81.41</td>
<td>-81.10</td>
<td>-0.31</td>
</tr>
<tr>
<td>Zone 1</td>
<td>5-6</td>
<td>-81.30</td>
<td>-81.50</td>
<td>+0.20</td>
</tr>
<tr>
<td>Zone 1</td>
<td>7-8</td>
<td>-81.80</td>
<td>-82.00</td>
<td>+0.20</td>
</tr>
<tr>
<td>Zone 1</td>
<td>9</td>
<td>-81.92</td>
<td>-82.20</td>
<td>+0.28</td>
</tr>
</tbody>
</table>

TABLE V
SAMPLE RESULTS FOR INTELSAT VI EIRP MEASUREMENTS

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Channel</th>
<th>Measured EIRP (dB W)</th>
<th>Predicted EIRP (dB W)</th>
<th>Delta</th>
</tr>
</thead>
<tbody>
<tr>
<td>East spot</td>
<td>1, 2</td>
<td>43.05</td>
<td>43.30</td>
<td>-0.25</td>
</tr>
<tr>
<td>West spot</td>
<td>5, 6</td>
<td>48.25</td>
<td>48.90</td>
<td>-0.65</td>
</tr>
<tr>
<td>East hemi</td>
<td>3, 4</td>
<td>35.37</td>
<td>35.40</td>
<td>-0.03</td>
</tr>
<tr>
<td>West hemi</td>
<td>7, 8</td>
<td>34.70</td>
<td>34.50</td>
<td>+0.20</td>
</tr>
<tr>
<td>Zone 1</td>
<td>3, 4</td>
<td>35.90</td>
<td>36.20</td>
<td>-0.30</td>
</tr>
<tr>
<td>Zone 4</td>
<td>1, 2</td>
<td>35.28</td>
<td>35.25</td>
<td>+0.03</td>
</tr>
</tbody>
</table>

TABLE VI
COMPARISON OF ESTIMATED AND MEASURED POWER LEVELS IN EIRP AND SATURATING FLUX DENSITY MEASUREMENTS

<table>
<thead>
<tr>
<th>Freq (GHz)</th>
<th>AUT Gain (dB)</th>
<th>Probe Gain (dB)</th>
<th>Flux Density Power (dBm/W)</th>
<th>EIRP Power (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Estimated</td>
<td>Measured</td>
<td>Estimated</td>
<td>Measured</td>
</tr>
<tr>
<td>3.905</td>
<td>34.4</td>
<td>12.8</td>
<td>-36.4</td>
<td>-29.7</td>
</tr>
<tr>
<td>6.13</td>
<td>32.5</td>
<td>12.2</td>
<td>-36.4</td>
<td>-29.7</td>
</tr>
<tr>
<td>10.5</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>6.0</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

in the satellite reached saturation. Saturation was detected by monitoring output power at the repeater test coupler. When saturation was observed, the power input to the probe was measured and used in (39b) along with the relative near-field data to calculate saturating flux density. A sample of results along with expected performance as predicted from design parameters and component tests are itemized in Table IV.

Measurement of the system EIRP is similar to the saturating flux density. After completing the relative near-field measurements, the probe was placed at the reference point. Power input to the transponder receive test coupler was gradually increased causing the transmit signal from the satellite to increase. When saturation of the repeater under test was indicated by the probe receive power remaining constant, the receive power was measured with the power meter. This value, along with the relative near-field data, was then used in (32) to obtain EIRP. Sample results are itemized in Table V. The estimates of uncertainty for both of these parameters are similar to those for gain, except that power measurement replaces the insertion loss or normalization constant. The uncertainty in the power measurements depends partly on the power level being measured. This can be estimated using (32) and (39b) and the approximation

\[
|\delta_i \delta_j \Sigma_i \Sigma_j B_j (P_j) e^{-\beta_0} \delta_j| = A_i^2,
\]

where \(A_i^2\) is the AUT's effective area. The approximate power input to the probe for saturating flux density measurements is then

\[
\lambda^2 \Sigma_i G_i^2 \frac{4\pi U_i}{\Sigma_i}
\]

and the approximate power received by the probe in EIRP measurements is

\[
\frac{\delta_0 \Sigma_i (EIRP) \sigma_i}{G_i^2}
\]

where \(G_i\) is the probe gain and \(G_i^2\) is the AUT gain. For the present measurements the gains, along with the estimated and actual power levels are noted in Table VI.

For these power levels, the estimated uncertainty in each of these measurements is itemized in Table VII.
TABLE VII
ESTIMATE OF ERRORS FOR SATURATING FLUX DENSITY AND EIRP MEASUREMENTS

<table>
<thead>
<tr>
<th>Source of Error</th>
<th>Estimated Error (dB) for SFD Measurements</th>
<th>Estimated Error (dB) for EIRP Measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Probe gain</td>
<td>0.10</td>
<td>0.10</td>
</tr>
<tr>
<td>Power measurement</td>
<td>0.20</td>
<td>0.20</td>
</tr>
<tr>
<td>Receiver nonlinearity</td>
<td>0.05</td>
<td>0.05</td>
</tr>
<tr>
<td>Multiple reflections</td>
<td>0.05</td>
<td>0.05</td>
</tr>
<tr>
<td>Position errors</td>
<td>0.01</td>
<td>0.01</td>
</tr>
<tr>
<td>Measurement area truncation</td>
<td>0.05</td>
<td>0.05</td>
</tr>
<tr>
<td>Phase errors</td>
<td>0.01</td>
<td>0.01</td>
</tr>
<tr>
<td>RSS Combination</td>
<td>0.24</td>
<td>0.24</td>
</tr>
</tbody>
</table>

Summary

Planar near-field techniques can be used to obtain absolute gain, EIRP, and saturating flux density. The errors in such measurements are determined by the same parameters that affect far-field measurements, the gain of the standard, various insertion loss quantities, and power levels. When used on a system where these quantities are well-known, the uncertainties can be less than for similar far-field tests.

REFERENCES


Edward J. McFarlane was born in Pittsburgh, PA, in 1959. He received the B.S.E.E. degree from Bucknell University, Lewisburg, PA, and the M.S.E.E. degree from the University of Southern California in Los Angeles, in 1982 and 1984, respectively.

He joined the Hughes Aircraft Company, El Segundo, CA, in 1982, where he has worked in the RF test section of the System Test Engineering Laboratory. As a Project Engineer he is presently responsible for the RF systems test program for the INTELSAT VI communications satellite. In this capacity he is involved in communications and antenna subsystem tests as well as work in the systems test planar near-field facility.

Robert D. Ward was born in Knoxville, TN. He received the B.S.E.E. and M.S.E.E. degrees from the University of Illinois, Urbana, in 1970 and 1971, respectively.

From 1971 to 1976 he was with Hughes Aircraft Company El Segundo, CA, where he was a Project Engineer in the Systems Test Department. He was involved in communications testing on the INTELSAT IV, HS 333, and GMS satellite programs. From 1976 to 1982 he was a Senior Project Engineer with the Antenna Department. He was involved with the design of antennas for communications satellites. In 1982 he joined the System Test Engineering Laboratory where he headed a project to engineer a planar near-field facility for the INTELSAT VI satellite program. He is currently head of the RF Engineering Section and is involved with testing of satellite antennas and communications systems.