A New Expression of Near-Field Gain Correction Using Photonic Sensor and Planar Near-Field Measurements

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Abstract

We propose a new expression of the near-field gain correction to calculate the on-axis far-field gain from the on-axis near-field gain for a directive antenna. The new expression is represented by transversal vectorial transmitting characteristics of two antennas that are measured by planar near-field equipment. Due to the advantages of the photonic sensor, the utilization of the new expression realizes the measurements of the on-axis far-field gains for two kinds of double ridged waveguide horn antennas within 0.1 dB deviation from 1 GHz to 6 GHz without calibrating the photonic sensor system.

Key words: Near-Field Gain, On-Axis Gain, Photonic Sensor, Planar Near-Field Measurements.

1. Introduction

The accurate measurements of the on-axis far-field gains of antennas are of primary importance in antenna measurements. The conventional method for the on-axis far-field gains uses the transmission measurements between antennas whose distance are far enough to fulfill the far-field conditions [1], or the near-field gain corrections to the transmission measurements between antennas at finite distances less than the distances of the far-field conditions [2]~[5]. As the most accurate method, the extrapolation method is used in national metrology laboratories [6].

The conventional near-field gain correction is calculated using the electric and magnetic fields at the apertures of antennas. Usually the electric and magnetic field distributions are calculated at the apertures of pyramidal horn antennas from some theoretical models [2], [4], [5]. For some complicated antennas such as real double-ridged waveguide horns, it is difficult to calculate the field distributions theoretically.

Ludwig [3] obtained the electric and magnetic fields for the near-field gain correction using spherical wave expansions determined from the far-field pattern of an antenna. This method is equivalent to the combination of the near-field gain correction and spherical near-field antenna measurements by the present point of view.

We propose a new expression of the near-field gain correction combined with the planar near-field antenna measurements [7]. From the theory of the planar near-

field measurements, the transmission between antennas at any distance can be expressed by the transversal vectorial transmitting characteristics (omitting "transversal vectorial" below) of the antennas. Using the transmitting characteristics, the far-field gain can be exactly obtained. Therefore the near-field gain correction at any distance is expressed by the ratio of the far-field gain to the near-field gain in terms of the transmitting characteristics of two antennas. This is a new expression of the near-field gain correction.

Generally, it is known that the far-field gain can be obtained by the planar near-field antenna measurements if the characteristic of the probe used in the measurements is known. However, in real measurements, there are serious problems such as truncation errors and multiple reflections between the antenna and the probe [8], especially for broad beam-width antennas such as double-ridged waveguide horn antennas.

In contrast, if the photonic sensor is used as the probe, the problems are overcome because the photonic sensor is a small dielectric object [9]. However the measurement system using the photonic sensor is difficult to be calibrated in order to obtain absolute gains of antennas.

Combining the new expression and the planar near-field measurement using the photonic sensor, we can obtain the on-axis far-field gain without calibrating the photonic sensor system.

In the following sections, the near-field gain correction is defined and expressed in a new form. To validate

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the new expression, some simulation results are shown. Finally the measurement results of double-ridged waveguide horn antennas using the photonic sensor demonstrate the usefulness of the new expression.

II. Near-Field Gain Correction

2-1 Conventional Expression

To obtain the on-axis far-field gain of an antenna that is defined on the far-field condition, the correction to the on-axis near-field gain of the antenna is required. The near-field gain correction (NGC) is defined by the ratio of the far-field gain G to the near-field gain G_{NF} as

$$NGC = \frac{G}{G_{NF}}. \quad (1)$$

Assuming that two antennas designated as 1 and 2 are separated at a distance z , the product of the near-field gains for each antenna is defined as

$$G_{NF1}(z)G_{NF2}(z) = \frac{1}{M_1 M_2} \left(\frac{4\pi z}{\lambda} \right)^2 \frac{P_{rec1}}{P_{in2}} \quad (2)$$

using the Friis transmission formula [10] where M_i ($i=1, 2$) is the mismatch power loss of antenna i , λ is wavelength, P_{rec1} and P_{in2} are the received power by antenna 1 and the input power (not the transmitted power) into antenna 2 respectively. Throughout the paper, we assume that the antennas are reciprocal and multiple reflections between the antennas can be neglected using a filtering method of the transmission data [3]. The far-field gain G_i of the antenna i is related to the near-field gain as

$$G_i = \lim_{z \rightarrow +\infty} G_{NF_i}(z). \quad (3)$$

The power transmission formula is given as [2]~[5]

$$\frac{P_{rec}}{P_{in}} = \frac{\left| \int_S (\mathbf{H}_2 \times \mathbf{E}_1 + \mathbf{E}_2 \times \mathbf{H}_1) \cdot \hat{\mathbf{n}} ds \right|^2}{4 \left\{ \text{Re} \int_{S_1} \mathbf{E}_1 \times \mathbf{H}_1^* \cdot \hat{\mathbf{n}}_1 ds_1 \right\} \left\{ \text{Re} \int_{S_2} \mathbf{E}_2 \times \mathbf{H}_2^* \cdot \hat{\mathbf{n}}_2 ds_2 \right\}} \quad (4)$$

where $\mathbf{E}_1$ and $\mathbf{H}_1$ are the fields when antenna 1 is transmitting, $\mathbf{E}_2$ and $\mathbf{H}_2$ are the fields when antenna 2 is transmitting, S_1 and S_2 are surfaces enclosing antennas 1 and 2 respectively, S is a surface enclosing either antenna, $\hat{\mathbf{n}}$, $\hat{\mathbf{n}}_1$, and $\hat{\mathbf{n}}_2$ are the unit normals to the surfaces, and $*$ means taking the complex conjugate.

To calculate (4), we must obtain the fields $\mathbf{E}_i$ and $\mathbf{H}_i$ radiated by antenna i on each surface. They are calculated by the assumed field distributions at the antenna apertures [2], [4], [5] or obtained by the spherical wave expansions whose coefficients are determined by the

far-field pattern of the antenna i [3].

We define a ratio

$$C_{12}(z) = \frac{\lim_{z' \rightarrow +\infty} z' \left| \int_{S'} (\mathbf{H}'_2 \times \mathbf{E}'_1 + \mathbf{E}'_2 \times \mathbf{H}'_1) \cdot \hat{\mathbf{n}} ds' \right|}{z \left| \int_S (\mathbf{H}_2 \times \mathbf{E}_1 + \mathbf{E}_2 \times \mathbf{H}_1) \cdot \hat{\mathbf{n}} ds \right|} \quad (5)$$

where quantities with primes correspond the ones in the case where the antenna distance is z' . Using (2) to (5), we can get

$$C_{12}(z) = \sqrt{\frac{(G_1 G_2)}{\{G_{NF1}(z)G_{NF2}(z)\}}} \quad (6)$$

because M_1 , M_2 , and the denominator of (4) are independent on z . Combining the three antenna method [1], we finally obtain the NGC for antenna 1 as

$$\begin{aligned} NGC_1(z) &= \sqrt{\frac{(G_1 G_2)}{\{G_{NF1}(z)G_{NF2}(z)\}}} \frac{(G_1 G_3)}{\{G_{NF1}(z)G_{NF3}(z)\}} \frac{\{G_{NF2}(z)G_{NF3}(z)\}}{(G_2 G_3)} \\ &= \frac{C_{12}(z)C_{13}(z)}{C_{23}(z)} \end{aligned} \quad (7)$$

where subscript 3 represents quantities for antenna 3. If antenna 1 and antenna 2 have completely the same characteristics, (7) reduces to (1) as

$$NGC_1(z) = \sqrt{\frac{(G_1 G_1)}{\{G_{NF1}(z)G_{NF1}(z)\}}} = C_{11}(z). \quad (8)$$

This means that $C_{ij}(z)$ ($i, j=1, 2, 3$) can be considered as an extended NGC.

Therefore the NGC is expressed by the electric and magnetic fields radiated by each antenna in the measurements.

2-2 New Expression

A new expression is derived from quantities closely related to antenna far-field patterns, that is, the transmission characteristics of antennas obtained by the planar near-field antenna measurements [7].

The extended NGC between antenna 1 and 2 is expressed as [11]

$$C_{12}(z) = \frac{\lim_{z' \rightarrow +\infty} |z' S_{12}(z') e^{jkz'}|}{|z S_{12}(z) e^{jkz}|} \quad (9)$$

using (5) and the power transmission formula given as

$$\frac{P_{rec}}{P_{in}} = |S_{12}(z)|^2 \quad (10)$$

where $S_{12}(z)$ is the S parameter between antennas 1 and 2. The phase factors e^{jkz} in (9) is included for the numerical integration of the denominator to be done easily

explained later.

Owing to [7], the S parameter between antennas 1 and 2 is given as

$$S_{12}(z) = \int_{-\infty}^{+\infty} \mathbf{S}_{01}(\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) e^{-j\gamma z} d\mathbf{K} \quad (11)$$

where $\mathbf{S}_{01}$ is the transversal vectorial receiving characteristic of the antenna 1, $\mathbf{S}_{20}$ is the transmitting characteristic of the antenna 2, $\mathbf{R}$ and z are the transversal vector, the z component of the position vector of the antenna 2 relative to the antenna 1 respectively, $\mathbf{K}$ and γ are the transversal vector and the z component of the wave-number vector $\mathbf{k}$ respectively. $\mathbf{S}_{i0}$ is related to the transversal electric field $\mathbf{E}_{it}(\mathbf{R}, z_i)$ at the position $(\mathbf{R}, z_i)$, radiated by antenna i as

$$\mathbf{S}_{i0}(\mathbf{K}) = \frac{e^{j\gamma z_i}}{2\pi a_0} \int \mathbf{E}_{it}(\mathbf{R}, z_i) e^{+j\mathbf{K} \cdot \mathbf{R}} d\mathbf{K} \quad (12)$$

where a_0 is the incoming wave amplitude to antenna i . Inserting (11) into (9), (9) reduces to

$$C_{12}(z) = \frac{\lim_{z' \rightarrow +\infty} \left| z' \int \mathbf{S}_{01}(\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) e^{jz'(k-\gamma)} d\mathbf{K} \right|}{\left| z \int \mathbf{S}_{01}(\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) e^{jz(k-\gamma)} d\mathbf{K} \right|} = \frac{|2\pi k \mathbf{S}_{01}(\mathbf{0}) \cdot \mathbf{S}_{20}(\mathbf{0})|}{\left| z \int \mathbf{S}_{01}(\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) e^{jz(k-\gamma)} d\mathbf{K} \right|}. \quad (13)$$

The integral of the numerator is evaluated by the stationary phase method [10].

Since antennas are assumed to be reciprocal and used in the transmitting mode for the planar near-field measurements, it is convenient to express (13) in only the transmitting characteristics. From the reciprocity [7], $\mathbf{S}_{10}(\mathbf{K})$ is related to $\mathbf{S}_{01}(\mathbf{K})$ as

$$\begin{aligned} \eta_0 \mathbf{S}_{01}(\mathbf{K}) \cdot \hat{\mathbf{\kappa}}_1 &= \frac{\omega \varepsilon}{\gamma} \mathbf{S}_{10}(-\mathbf{K}) \cdot \hat{\mathbf{\kappa}}_1 \\ \eta_0 \mathbf{S}_{01}(\mathbf{K}) \cdot \hat{\mathbf{\kappa}}_2 &= \frac{\gamma}{\omega \mu} \mathbf{S}_{10}(-\mathbf{K}) \cdot \hat{\mathbf{\kappa}}_2 \end{aligned} \quad (14)$$

where η_0 is the characteristic admittance of the measurement system and $\hat{\mathbf{\kappa}}_1 = \mathbf{K}/|\mathbf{K}|$, $\hat{\mathbf{\kappa}}_2 = \hat{\mathbf{z}} \times \hat{\mathbf{\kappa}}_1$.

Inserting (14) into (13), we obtain the new expression as

$$C_{12}(z) = \frac{|2\pi k \mathbf{S}_{10}(\mathbf{0}) \cdot \mathbf{S}_{20}(\mathbf{0})|}{\left| z \int_{k\gamma}^{\frac{1}{k\gamma}} [\gamma^2 \mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) + \{\mathbf{K} \cdot \mathbf{S}_{10}(-\mathbf{K})\} \{\mathbf{K} \cdot \mathbf{S}_{20}(\mathbf{K})\}] e^{jz(k-\gamma)} d\mathbf{K} \right|}. \quad (15)$$

This expression has only transmitting characteristics that are closely related to the far-field pattern [7] and explicitly shows the dependence on the separation z between antennas. We clearly find how the denominator

approaches to the numerator as z increases to infinity, that is, how the on-axis near-field gain converges to the on-axis far-field gain.

In practical point of view, it is difficult to perform the two-dimensional numerical integration of the denominator in (15) because the phase variation of the phase term $\exp\{jz(k-\gamma)\}$ becomes rapid when $|\mathbf{K}|$ approaches k in the case where z is a few meters even at 1 GHz. However the phase variation of $\mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K})$ is slow. Therefore it is convenient to transform the integral shown below.

Since the phase term is only depend on the magnitude of $\mathbf{K}$ (kK_n where K_n is K normalized by k), we first integrate along the angle ϕ of $\mathbf{K}$ and define a function as

$$F(K_n) = \int_0^{2\pi} \left[\gamma^2 \mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) + \{\mathbf{K} \cdot \mathbf{S}_{10}(-\mathbf{K})\} \{\mathbf{K} \cdot \mathbf{S}_{20}(\mathbf{K})\} \right] d\phi. \quad (16)$$

Next integrating along K_n and changing it to two variables depending on the integration intervals, we have the form suitable to the numerical integration of the denominator as

$$\begin{aligned} & \left| z \int_{k\gamma}^{\frac{1}{k\gamma}} [\gamma^2 \mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K}) + \{\mathbf{K} \cdot \mathbf{S}_{10}(-\mathbf{K})\} \{\mathbf{K} \cdot \mathbf{S}_{20}(\mathbf{K})\}] e^{jz(k-\gamma)} d\mathbf{K} \right| \\ &= \left| kz \left\{ \int_0^1 F(\sqrt{\tau(2-\tau)}) e^{jk\tau} d\tau + j \int_0^{+\infty} F(\sqrt{1+\varepsilon^2}) e^{jk\varepsilon - k\varepsilon} d\varepsilon \right\} \right| \end{aligned} \quad (17)$$

where

$$\begin{aligned} \tau &= 1 - \sqrt{1 - K_n^2} & 0 \leq K_n \leq 1 \\ \varepsilon &= \sqrt{K_n^2 - 1} & 1 \leq K_n \leq \infty. \end{aligned} \quad (18)$$

The first and second terms in (17) correspond to the propagation and evanescent modes [7]. Since the phase variation of the phase term in the first term is constant and the second term attenuates exponentially, it is easy to integrate (17) numerically.

III. Numerical Simulation

To validate the new expression, a numerical simulation using R-band standard pyramidal horn antennas is performed as shown in Fig. 1.

The antenna models the R-band standard pyramidal horn antenna (MI 12-1.7 produced by MI Technologies) whose aperture has dimensions of 272 mm ($2a$, E plane) $\times$ 367.5 mm ($2b$, H plane) $\times$ 271 mm (l , Height). All symbols for dimensions follow [5]. The waveguide part has a cross section of dimensions 109.2 mm ($2s_1$) $\times$ 54.6 mm ($2s_2$) and a length of 150 mm. The E plane, the H plane, and the on-axis line of the antennas are coincident to the $y=0$ plane, the $x=0$ plane, the z axis respectively. The waveguides are excited and received by only TE₁₀ mode.

Fig. 1. Simulation model of R-band standard pyramidal horn antennas at the distance of 0.5 m.

Fig. 2. Electric field distribution at $z=60$ mm plane of dimensions $2\text{ m}\times 2\text{ m}$ and 2.45 GHz in linear scale.

The simulation was done using FEKO [12] that uses the method of moments. The calculated frequency is 2.45 GHz. The edge size of all parts except excitation ports is 10 mm and the excitation ports are meshed in 5 mm. The electric and magnetic symmetries are used for the H plane and the E plane respectively.

3-1 Transmitting Characteristics of Antenna

Fig. 3. Transmitting characteristic $S_{10}(\mathbf{K})$ in $\mathbf{K}$ space whose magnitude is normalized by k at 2.45 GHz in linear scale.

The transmitting characteristic $S_{10}(\mathbf{K})$ of the antenna is calculated using the electric near-field data on the plane of dimensions $2\text{ m}\times 2\text{ m}$ at $z=60$ mm with 20 mm steps by the planar near-field antenna method [7]. Fig. 2 shows the x and y components of the electric near-field.

Fig. 3 shows the x and y components of $S_{10}(\mathbf{K})$ that is the two-dimensional Fourier transform of the electric field.

In Fig. 2, the near-field distribution is concentrated around the antenna aperture and the y component is negligibly small. Therefore the x component of $S_{10}(\mathbf{K})$ has no ripples due to the truncation error [8] and the y component is negligible in Fig. 3. Indeed there is no difference between $S_{10}(\mathbf{K})$ calculated by the near-field data on the two planes: $2\text{ m}\times 2\text{ m}$ and $4\text{ m}\times 4\text{ m}$. The far-field pattern calculated by FEKO agreed with that of $S_{10}(\mathbf{K})$ within the magnitude of 0.005 .

3-2 Near-Field Gain Correction

The magnitude and phase of $S_{10}(-\mathbf{K}) \cdot S_{10}(\mathbf{K})$ are shown in Fig. 4. As expected, the amplitude is more

 (a) Magnitude of $S_{10}(-\mathbf{K}) \cdot S_{10}(\mathbf{K})$

 (b) Phase of $S_{10}(-\mathbf{K}) \cdot S_{10}(\mathbf{K})$

 (c) Magnitude and phase of $S_{10}(-\mathbf{K}) \cdot S_{10}(\mathbf{K})$ at the $k_x=0$ and $k_y=0$ cut-planes

 Fig. 4. Distribution of $S_{10}(-\mathbf{K}) \cdot S_{10}(\mathbf{K})$ in $\mathbf{K}$ space normalized to k .

concentrated around the origin than that of $S_{10}(\mathbf{K})$ and the phase variation is slow as in Fig. 4. Therefore the phase variation of the integrand in (16) is slow and the numerical integration of (16) can be done as follows:

- 1) The sampling intervals of K_n and ϕ are taken to be 0.02 and 5 degrees,
- 2) Interpolation at the sampling points in (K_n, ϕ) from the original square grid points in $\mathbf{K}$ space is done using two dimension splines,
- 3) And the summation around ϕ at the constant K_n is taken as the integration by ϕ .

 Fig. 5. Magnitude and phase of $F(K_n)$ normalized to the value at $K_n=0$.

 Fig. 6. Magnitude of $zS_{21}(z)$ and the moving average values in the interval of N points.

 Fig. 7. Comparison of the NGC in (1) and the new NGC in (15) using the averaged values ($N=21$) of $zS_{21}(z)$ by FEKO.

Following the above steps, we obtained the magnitude and phase of $F(K_n)$ as shown in Fig. 5. The variation of the magnitude and the phase are similar to the ones in Fig. 4(c) as expected.

To avoid the multiple reflection in $S_{21}(z)$ by FEKO, the moving average of $zS_{21}(z) \exp(+jkz)$ is done in the interval of 120 mm (21 points) around the center. An example demonstrating the effect for $z=0.5$ m (worst case) is shown in Fig. 6 and shows that the effect of the

multiple reflection is reduced within 0.02 dB. The main component is the term including $\exp(-j2kz)$ and $N=11$ (the interval of 60 mm) is the best for cancelling it ($0.5 \lambda = 61.2$ mm) among $N=8$ to 12. However we adopted $N=21$ since it is a simple average using all points.

Using $F(K_n)$ and (17), we can calculate the NGC at any antenna separation. To validate the NGC by (15) or (16) and (17), we calculated the near-field gains in (2) using the averaged values ($N=21$) of $zS_{21}(z)$ simulated by FEKO at the five separations ($z=0.5$ m, 1 m, 2 m, 3 m, 4 m, and 10 m) and the far-field gain by FEKO. Then we compare the NGC in (1) by FEKO with the new NGC in (15) as shown in Fig. 7.

In Fig. 7, the NGC calculated by (1) and the one by (15) agrees within 0.04 dB whose difference comes from mainly the interpolation error in (16). Therefore the new NGC in (15) is valid.

IV. Measurement

To demonstrate the effectiveness of the new NGC, we have applied it to determine the gains of two kinds of double-ridged waveguide horn antennas from 1 GHz to 6 GHz with 0.5 GHz steps. The antennas are often used in Electromagnetic Compatibility (EMC) measurements and have the same product name (ETS-Lindgren 3115), whereas the structures and the antenna characteristics are different. In the following, we distinguish between those by specifying the old one (discontinued) and the new one (available at present).

4-1 Measurement Setup

We used the planar near-field antenna equipment shown in Fig. 8 and a photonic sensor shown in Fig. 9.

The gain of the amplifier (HP8348A) is about 35 dB. The output power and the IF of the vector network analyzer (VNA, Agilent E8363C) were 0 dBm and 100 Hz respectively. The scanned range was $1.8 \text{ m} \times 1.8 \text{ m}$ with

Fig. 8. Planar near-field measurement system using the photonic sensor as the probe.

(a) Geometry

(b) Picture

Fig. 9. Photonic sensor as the probe.

10 mm steps in the planar scanner (NSI 300 V-8×8). The distance between the aperture of each antenna and the photonic sensor was 58 mm. The principle of the operation of the photonic sensor is explained minutely in [9].

4-2 Results of Planer Near-Field Measurements

Because the photonic sensor can measure the transversal electric field (except proportional constant) radiated by an antenna, the new NGC can be obtained in the same procedure as in the simulation by FEKO.

$S_{21i}(\mathbf{R}, z)$, S -parameter measured by the VNA, is proportional to the electric field $\mathbf{E}_i(\mathbf{R}, z)$ transmitted by the antenna i at the sensor position $(\mathbf{R}, z)$ as

$$S_{21i}(\mathbf{R}, z) = \alpha_i \hat{\mathbf{p}} \cdot \mathbf{E}_i(\mathbf{R}, z) \quad (19)$$

where α_i is a constant including the error terms of the measurement system and the reflection coefficient of the antenna i , $\hat{\mathbf{p}}$ is the unit vector parallel to the dipole moment vector of the photonic sensor [9], [11].

As typical characteristics, the data at 3.5 GHz (the center frequency in the measurements) are shown below.

Fig. 10 shows the x components of $\mathbf{S}_{10}(\mathbf{K})$ of the old type ($i=1$, antenna 1) and the new type ($i=2$, antenna 2). Since the antennas are linearly polarized in the x direction, the y components are negligible and not shown here.

Fig. 11 shows the magnitude and the phase of $\mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K})$. The ones of $\mathbf{S}_{20}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K})$ are almost the same as those. $F(K_n)$ for $\mathbf{S}_{10}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K})$ and $\mathbf{S}_{20}(-\mathbf{K}) \cdot \mathbf{S}_{20}(\mathbf{K})$ are depicted in Fig. 12.

As seen in Fig. 10, $\mathbf{S}_{10}(\mathbf{K})$ ($i=1, 2$) at 3.5 GHz are dif-

 (a) x component of $S_{10}(\mathbf{K})$ (old type)

 (b) x component of $S_{20}(\mathbf{K})$ (new type)

 Fig. 10. $S_{10}(\mathbf{K})$ (old type) and $S_{20}(\mathbf{K})$ (new type) in $\mathbf{K}$ space whose magnitude is normalized by k at 3.5 GHz.

ferent from each other. This fact is true for other frequencies. Therefore the antenna gains of both antennas have different frequency characteristics as seen later. The phase variations of $F(K_n)$ are slow and similar to the R-band standard horn antenna in Fig. 5. However the magnitude of $F(K_n)$ decreases gradually compared to that in Fig. 5, since the pattern of the double-ridged waveguide horn is wider than the standard horn.

4-3 New NGC and Gain

The near-field gains of both types of antennas were measured at $z=1.5$ m, 3 m, and 3.87 m, using the three antenna method (one in the old type, and two in the new type) [1] and compared with the ones obtained by using the new NGCs that were calculated through (15) to (17) using the measured data in Fig. 12.

Since the two antennas of new type had the same $S_{20}(\mathbf{K})$ in this case, the $C_{22}(z)$ is calculated by (15) where the subscript 1 is replaced to 2. Then the $NGC_2(z)$ is equal to $C_{22}(z)$ as shown by (8). On the other hand, the $NGC_1(z)$ is determined by (7) where the

 (a) Magnitude of $S_{10}(-\mathbf{K}) \cdot S_{20}(\mathbf{K})$

 (b) Phase of $S_{10}(-\mathbf{K}) \cdot S_{20}(\mathbf{K})$

 Fig. 11. Distribution of $S_{10}(-\mathbf{K}) \cdot S_{20}(\mathbf{K})$ in $\mathbf{K}$ space normalized to k at 3.5 GHz.

 Fig. 12. Magnitude and phase of $F(K_n)$ normalized to the value at $K_n=0$ at 3.5 GHz.

subscript 3 is replace to 2 as

$$NGC_1(z) = \frac{C_{12}^2(z)}{C_{22}(z)}. \quad (20)$$

Fig. 13 compares the differences of the near-field gains relative to the near-field gains at $z=1.5$ m and the NGC_i relative to the NGC_i at $z=1.5$ m. The differences obtained by both methods agree within 0.2 dB except 2

Fig. 13. Differences of the near-field gains by the three antenna method and those of the NGC_i ($i=1, 2$) by the new method. Each reference is the value at $z=1.5$ m.

GHz for the antenna of the new type. This is due to the large truncation error of $S_{20}(\mathbf{K})$ that is not shown here. Overall, Fig. 13 demonstrates that the new method to calculate the NGC is effective even for broad beam-width antennas.

Finally we show the far-field gains obtained by the new NGC in Fig. 14. The far-field gain converted from the near-field gain at each z is calculated by $NGC_1(z)$ in (1) or $NGC_2(z) (= C_{22}(z))$ in (20). The far-field gain of the same antenna must be the same because $NGC_i(z)$ converts the near-field gain at any z to the same gain at infinity. Fig. 14 shows each gain agrees within 0.1 dB except 2 GHz for both types and 1 GHz for the old type. Therefore Fig. 14 demonstrates the usefulness of the new expression of the NGC combined with the planar near-field measurements using the photonic sensor, in order to measure the on-axis far-field gain and the on-axis near-field gain at any distance. In Fig. 14, the average far-field gain is calculated by averaging the far-field gains corrected by the corresponding NGCs from the near-field gains at 1.5 m, 3 m, and 3.87 m.

V. Conclusion

We have proposed a new expression of the near-field

Fig. 14. Far-field gains (specified by Inf.) of antennas of the old and new types obtained by the new NGC at $z=1.5$ m, 3 m, 3.87 m. The deviation of each gain from the average far-field gain (Inf. and Avg.) is also shown.

gain correction to calculate the on-axis “far-field” gain from the on-axis “near-field” gain for a directive antenna.

The expression uses the transversal vectorial transmitting characteristic of the antenna that is closely related to the antenna pattern and is the Fourier transform of the radiated electric field on a plane measured by the planar near-field equipment.

The new expression has the advantages:

1. There is no need to correct S_{21} measured by the VNA by using error terms representing the measurement system,
2. There is no need to measure the reflection coefficients of the antennas, and
3. The error terms representing the measurement system for each antenna measurement can be different, that is, the measurement condition can be changed for each antenna measurement.

Merits of using the photonic sensor as the probe are

4. It is easy to calculate the new NGC because the photonic sensor can detect directly the electric field on the plane,

5. The photonic sensor can measure the electric field

without influence of unnecessary scattered fields because the sensor can be very close (a few centimeters) to the aperture of antennas.

We have validated the new NGC through the numerical simulation and demonstrated the usefulness of the new NGC in the measurements using the photonic sensor and the planar near-field equipment. The new expression realizes the measurements of the on-axis far-field gain for two kinds of double ridged waveguide horn antennas within 0.1 dB deviation at 1 GHz to 6 GHz.

To improve the reliability of the measurements using the new NGC, we are now evaluating the uncertainty of the measurements, mainly, that of the photonic sensor system and will be submitted in the near future.

References

- [1] G. E. Evans, *Antenna Measurement Techniques*, Artech House, pp. 19-22 and p. 66, 1990.
- [2] T. S. Chu, R. A. Semplak, "Gain of electromagnetic horns," *Bell System Technical Journal*, vol. 44, no. 3, pp. 527-537, Mar. 1965.
- [3] A. C. Ludwig, R. A. Norman, "A new method for calculating correction factors for near-field gain measurements," *IEEE Trans. Antennas and Propagation*, vol. 21, no. 5, pp. 623-628, Sep. 1973.
- [4] T. Uno, S. Adachi, "Range distance requirements for large antenna measurements," *IEEE Trans. Antennas and Propagation*, vol. 37, no. 6, pp. 707-720, Jun. 1989.
- [5] G. Mayhew-Ridgers, J. W. Odendaal, and J. Joubert, "Horn antenna analysis as applied to the evaluation of the gain-transfer method," *IEEE Trans. Instrumentation and Measurement*, vol. 49, no. 5, pp. 949-958, Oct. 2000.
- [6] A. C. Newell, R. C. Baird, and P. F. Wacker, "Accurate measurement of antenna gain and polarization at reduced distances by an extrapolation technique," *IEEE Trans. Antennas and Propagation*, vol. 21, no. 4, pp. 418-431, Jul. 1973.
- [7] D. M. Kern, *Plane-Wave Scattering-Matrix Theory of Antennas and Antenna-Antenna Interactions*, National Bureau of Standards Monograph 162, Ch. 2, 1981.
- [8] A. C. Newell, "Error analysis techniques for planar near-field measurements," *IEEE Trans. Antennas and Propagation*, vol. 36, no. 6, pp. 754-768, Jun. 1988.
- [9] M. Hirose, T. Ishizone, and K. Komiyama, "Antenna pattern measurements using photonic sensor for planar near-field measurement at X band," *IEICE Trans. Communications*, vol. E87-B, no. 3, pp. 727-734, Mar. 2004.
- [10] R. E. Collin, *Antennas and Radiowave Propagation*, McGraw-Hill, pp. 284-286 and pp. 303-304, 1985.
- [11] M. Hirose, S. Kurokawa, "Far-field gain measurement using a new expression of near-field gain correction and photonic sensor," *Proceedings of ISAP 2011*, FrB2-1, SE10_1002, Oct. 2011.
- [12] FEKO, <http://www.feko.info>

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