Near-Field Alignment of Phased-Array Antennas

Willard T. Patton, *Life Fellow, IEEE,* and Leonard H. Yorinks, *Member, IEEE*

***Abstract*—This paper describes the algorithms and equipment used to apply near-field scanning techniques to the phase alignment of phased-array antennas. This procedure achieves a level of precision not previously available. The electronic scanning property of the antenna is used to bring different sections of the antenna spectrum within range of the near-field scanning process. These partial spectra are then merged to define the entire spectrum of the antenna. This process provides the resolution needed to determine the excitation at individual elements by the inverse Fourier transformation operation. The process described here has been used in the production of a very large number of phased-array antennas currently in service.**

I. Introduction

THE alignment of a phased-array antenna is the process of bringing all of its radiating elements into phase alignment so that their radiated power will add coherently in a given direction. The process varies depending on the architecture of the array feed system. In a transmission lens array, for example, the path from the source to each radiating element varies in a predictable fashion. This allows one to compute the correction at each element to convert the spherical phase front to a planar phase front. The alignment of the antenna and its resultant gain are improved when the insertion phase differences due to construction tolerances and other sources are determined and corrected.

The constrained feed architecture presents different and somewhat more complex problems for alignment. A constrained feed system often provides essentially equal path length from the source to each of the radiating elements. Errors at various junctions in the feed system correlate in the array aperture. A phase error at any one junction affects all elements served by that junction. This correlation of errors can produce large and unwanted sidelobes. These sidelobes tend to become more severe as the number of elements connected to the junction increases. The cost of manufacturing junctions with the precision required for a large low sidelobe array becomes prohibitive. The designer must therefore find another way to eliminate these correlated errors. Measurement of phase error at the individual elements, and correction of error using the element phase shifters make the residual error uncorrelated from element to element. Uncorrelated errors do not contribute significantly to peak sidelobes. Indeed, the phase quantization error associated with the phase shifters, together with the errors associated wth the measurement process, combine to the maximum residual of the original error due to all junctions between the array port and the element. Similar corrections can be applied to compensate for relatively large differences in the path lengths to each element.

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The authors are with Lockheed Martin Government Electronics Systems, Moorestown, NJ 08057 USA.

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II. Alignment Measurements

An early technique for measuring the phase at the individual element was to vary its phase to modulate the signal from the array received at a remote point. Modulation by the phase shifter was measured against the background of the aggregate signal from all the other elements. A technique described by Alexander and Gray [1] used only the amplitude of the modulation to detect when the signal from the element was aiding or opposing the signal from the rest of the array. This technique works well for moderate-size arrays of the order of a few thousand elements. It works less well as the array gets larger and the signal modulated approaches the noise level or the limit of the instantaneous dynamic range of the receiving system.

The near-field antenna measurement system [2] offers many advantages for the measurement of the radiating properties of array antennas. Not the least of these is the advantage of indoor testing independent of weather conditions. The original motivation for this technique was greater accuracy in determining the radiating pattern of an antenna. However, it also provided information often not previously available to antenna engineers, namely the phase of the pattern. By very much the same algorithm used to transform the near-field measurement to the far-field pattern, the far-field pattern could be transformed to the aperture of the array. This transformation yields the amplitude and phase distribution in the aperture that led to the near field and, hence, to the far field. However, this transformation is subject to a fundamental limit on resolution in the aperture [3].

Since the far field is the Fourier transform of the aperture field, the far field (and the near field as well) is the superposition of a spectrum of plane waves. Only those plane wave spectral components with wave number less than the intrinsic wave number of free-space $(2\pi/\lambda)$ are within the observable spectrum, i.e., visible space. If the the spectral period of the array is larger than the visible space region, transformation from the far field to the aperture plane will not provide sufficient resolution to align the array at each radiating element. This is indeed the case for very wide scanning arrays with subwavelength spacing, where the spectrum describing the array aperture illumination is substantially larger than the observable spectrum.

An array consisting of a collection of uniformly illuminated subarrays interconnected by a waveguide beam former had previously been successfully phase aligned. That process used the inverse transformation of far field to aperture field of

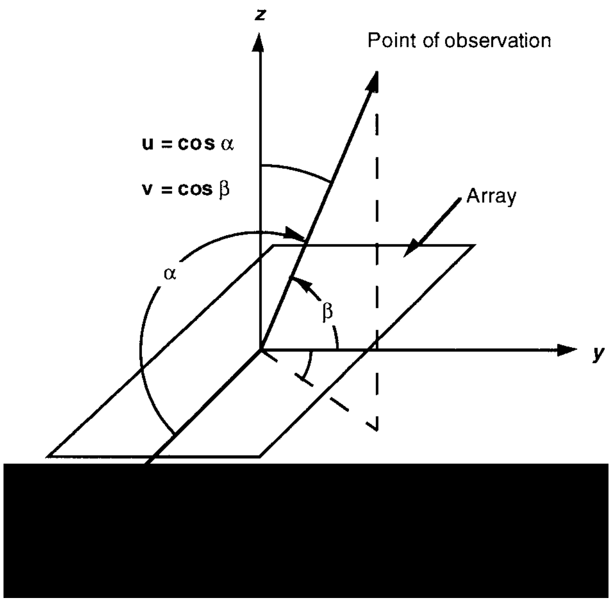


Fig. 1. Array coordinate system.

a subarray. The computer controlling the near-field scanner provided the waveguide shim thickness to be inserted into each branch of the beamformer. In this case, the intrinsic resolution of the process was very much finer than the size of the antenna subarrays, but not as fine as a single element. Using this technique to determine the phase at the individual elements of the array requires a method to overcome the fundamental limit on resolution.

The answer to this resolution problem lies in the properties of the Fourier transform relating a linear phase ramp in one domain to a translation in the transformed domain. This allows us to use a linear phase ramp to translate the array spectrum relative to the observable spectrum. Thus, all of the array spectrum is rendered visible with several scans, each using a different phase ramp. Although the fundamental limit to resolution by a single scan remains intact, we are able to record the entire spectrum through a series of scans, which are then merged to obtain the complete array spectrum [4], [5]. The details of the process are given in Section III.

III. Spectral Merging

The discussion of the spectral merging process below is applicable to arrays with either rectangular or triangular element latices. The specific dicussion relates to the triangular lattice since it is more efficient (i.e., has fewer elements per unit area) than the rectangular lattice for arrays which either cover a conical region or one quarter hemisphere. The antenna geometry for the problem is shown in Fig. 1.

A. The Array Spectrum

The variables of the Fourier transform that map real space to the array spectrum must be defined to relate the dimensions of the array and the coordinates of the array spectrum. It is convenient to relate this transform to the Schelkunoff [6] expression for the directional pattern of an antenna located in the x-y plane

$$S(u,v) = P_e(u,v) \times \sum_m \sum_n a_{n,m} e^{j2\pi(m\frac{d_x}{\lambda}u + n\frac{d_y}{\lambda}v)}$$

where the variables u, v are direction cosines relative to the $\mathbf{x}$ and $\mathbf{y}$ axes, respectively. The double summation was called the space factor of the array by Schelkunoff. Classically, the variables u and v have values less than one. However, if the angle domain is extended to imaginary angles, the range of u and v will each be the entire real number line.

The double sum is a scalar function. All the vector properties of the antenna are represented by the element pattern which is the Fourier transform of the field distribution at the element. This includes, specifically, the polarization characteristics of the antenna.

The antenna spectrum can therefore be considered the product of two factors, especially for large arrays. One is due to the excitation of the array—the array pattern—and the other is due to the configuration of the element of the array—the element pattern $\mathbf{P}_e$. This is true to the extent that all of the elements are identical in configuration and perform in an identical environment. Elements at the edge of the array violate the identical environment assumption, but the assumption is effective for arrays of more than 1000 elements. This is particularly true near the broadside pointing direction since the edge effect is most pronounced at very wide scan angles. (Yorinks [7] showed that edge effects are negligible when the array dimensions in wavelengths are greater than the average sidelobe level in decibels. This is significant for low sidelobe arrays where array alignment precision is especially important.) The element pattern, sometimes referred to as the active element pattern, denotes the performance of the element in an array with all neighbors terminated. It is slowly varying near broadside scan and rapidly varying at large scan angles. It may even have a null at large scan angles if the element/array is blind at that angle. Therefore, as will be shown later, one should choose scans to cover the parts of the array spectrum as close to broadside as possible for the given element spacing.

B. Fourier Analysis

Extending the variables to the entire real number line admits the application of Fourier transform theory. Therefore, the antenna spectrum $S(u,v)$ and the aperture distribution $a(x,y)$ are related by

$$S(u,v) = C_2 \int\int_{\text{Aperture}} a(x,y)e^{jk(ux+vy)}\,dx\,dy$$

and

$$a(x,y) = C_1 \int\int_{\infty} S(u,v)e^{-jk(ux+vy)}\,du\,dv$$

where

$$k = \frac{2\pi}{\lambda}$$

and C_1 and C_2 are constants depending upon units, normalization, and other considerations of computational convenience.

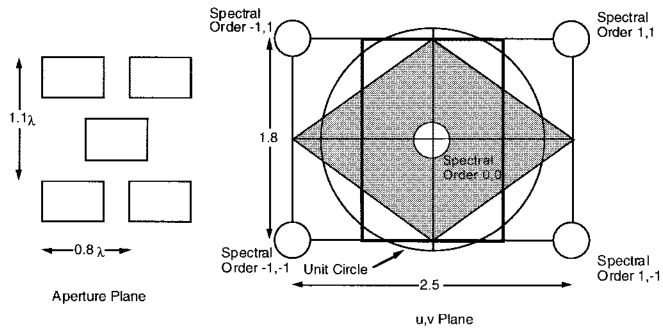


Fig. 2. Triangular array lattice and portion of spectrum showing periodicity and two possible spectral cells (gray-shaded diamond and bold rectangle).

The Fourier transform of a periodic array is a periodic function determined by the aperture excitation function. The spectral period is given by the reciprocal of the array spacing in wavelengths. This periodic function is multiplied by the active element pattern, which is determined by the radiating element and its neighbors in the array. The spectral representation of the space factor or array pattern is periodic. The distribution in a cell of the period is repeated over all other cells in the transform space or u, v space. The element pattern is not periodic in transform space, but it multiplies the space factor or the array pattern to form the spectrum of the antenna. The objective in aligning the antenna is to transform the array pattern to the aperture plane in such a way as to obtain the individual element excitations.

C. Spectral Cell

The spectral cell, which is the period of the transform in (u, v), is that section of the spectrum that when moved to adjacent spectral orders (grating lobes) will completely cover the spectral plane. Although the spectral cell does not have a unique shape, its area is determined by the periodicity of the array. This is illustrated in Fig. 2, which shows two examples of such an area for an array with triangular lattice. Either the diamond-shaped area or the area within the bold rectangle represents the two-dimensional (2-D) spectral period.

D. Transformation of the Array Spectrum to the Aperture Plane with Full Resolution

When transforming the array pattern to the aperture plane using a simple 2-D transform, it is easiest to deal with a rectangular region in the spectral domain. The spectral region shown by the bold rectangle in Fig. 2 corresponds to a lattice with half the elements desired, as shown in Fig. 3(a).

Alternatively, if the spectral cell shown shaded in Fig. 3(b) is transformed to the aperture plane, the resulting data will fall on the rectangular lattice shown. This has twice the number of elements as the original triangular lattice. If this spectral cell is properly constructed, those elements absent from the triangular lattice will transform to have zero excitation. Therefore, the objective is to fill the spectral cell with the proper data.

If spectral cells are centered over each of the spectral orders (grating lobes), they would fill the spectral plane. The motion from one spectral order to the next involves translation in two directions. This is a consequence of the triangular grid. To construct a rectangular spectral cell that accounts for the triangular grid, we reuse the lower right quadrant with respect to the main lobe (region 1 in Fig. 4) as the lower right quadrant with respect to the upper left spectral order and similarly for the other quadrants. The sign given these reused quadrants matches the sign of the spectral order. These signs will alternate if there is an even number of rows and columns in the array aperture, but will be all the same for an odd number of rows and columns. Note that in the even alternating case, there is no element at the phase center of the array. If there were to exist an element at the phase center of the array, the spectral orders would all have the same sign and the quadrants would be reused with the original sign.

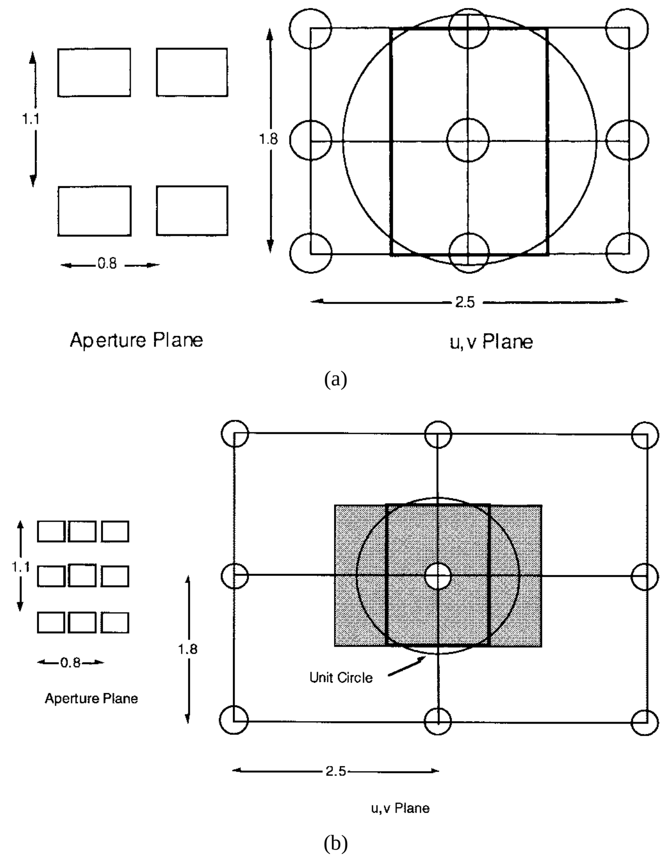


Fig. 3. Alternate spectra formed by rectangular lattices. (a) Alternate columns of original triangular lattice array. (b) Twice the elements of original triangular lattice array.

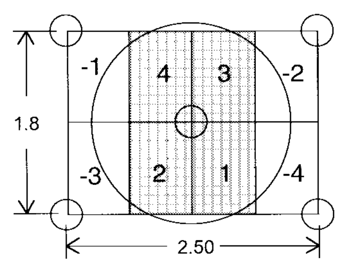


Fig. 4. Spectral cell can be constructed by properly replicating data in half of the cell.

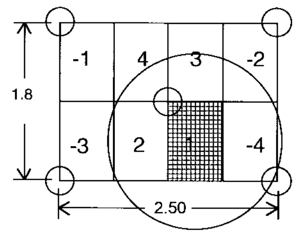


Fig. 5. Array spectrum has been translated to center desired spectral region in visible space.

Inspection of Fig. 2 shows that less than a full spectral cell is contained within the unit circle, i.e., visible space. However, the spectral cell of Fig. 3(b) can be constructed from each of the numbered regions shown in Fig. 4. The angular spectrum can be translated relative to visible space by applying a linear (planar) phase taper to the array excitation. A phase rate of 360° per element will translate the main beam to the next spectral order in that direction. 180° per element will translate half way to the next spectral order. To translate the spectrum so that one quadrant of the spectral cell is centered in visible space as shown in Fig. 5, 90° per row together with 45° per column is required. This allows that part of the spectrum to be centered with respect to the array broadside direction and recovered with minimum variation in the element-probe coupling factor. This approach may be repeated to scan each of the numbered regions in Fig. 4 to the center of the unit circle. The entire spectral cell may then be constructed by piecing together each of the numbered partial spectral regions. When transformed to the aperture plane, the merged spectrum will provide spatial resolution to each element. In merging the four spectral regions about the main beam, the average of the regions has been taken where the regions overlap since there is no basis to consider the principal plane cuts through the main beam in one quadrant better than another. Wittmann *et al*. [8] have independently confirmed the merged spectrum technique for phased-array alignment.

The method just described works effectively since it enforces triangular symmetry in the spectrum. There are fast Fourier transform (FFT) algorithms that deal with the triangular grid and although the one used here is not elegant, it is simple and convenient. Some of the alternate techniques might result in some saving in computation time.

The angular spectrum could, in fact, be obtained with only two scans, thereby saving data acquisition time. However, that would require use of data from less reliable lower levels of the probe-element gain pattern and would also lower the signal to measurement noise power by 3 dB. The decision is basically a trade between test time and accuracy.

E. Active Element Pattern

The active element pattern is taken in the sense of Peter Hannan [9]. It is the gain and directional performance of an element of an array as influenced by the surrounding elements. There are method of moments and finite-element codes for determining the active element pattern mathematically. The probe pattern is generally measured by an outside standards agency such as NIST (Boulder, CO). The element-probe coupling factor [2] can be measured directly with a near-field facility and the active element pattern determined by removing the precalibrated probe factor. The element-probe coupling factor is measured by steering the array to each angle for which element-probe data is required and recording the response in that direction. Note, however, that if only the scalar properties of the array are of interest, i.e., only its phase and amplitude distribution, it is not necessary to separate the element-probe coupling factor.

The result obtained when the near-field measured data is transformed to the far-field is the product of the scalar array spectrum multiplied by the active element pattern and by the pattern of the probe antenna. When a linear phase taper is applied to the array, only the scalar array spectrum is translated in direction cosine space. The element pattern and the probe pattern are invariant under a linear phase taper. When the far-field transform result is divided point by point by the element-probe coupling factor, the result is the array spectrum at those points.

To recover the gain of the array by near field techniques, it may be required to scan the near field with probes of two orthogonal polarizations. Matching the probe polarization to the array-element polarization avoids this requirement. Double scanning would also be required if the polarization distribution of the element were unknown, but of concern. Polarization match is required to recover scalar characteristics of the array only to the extent that it reduces element-probe coupling loss.

F. Fast Fourier Transform (FFT) Measurement Density

FFT's using "power of 2" algorithms are convenient for manipulating the data. These have the same number of data points in array coordinates as in angular coordinates. "Zero padding" the measurement data to extend its size will reduce the spacing between data points in the spectrum and increase angular resolution. It is required to produce an adequate sampling of the spectrum to define the sidelobe structure of the antenna.

The sampling data spacing should be smaller than the element spacing to prevent aliasing of the data. It should be commensurate with the element spacing to ensure that the element excitation data corresponds to the location of the elements. If we use a sampling interval of half the row spacing, r and half the column spacing c the sampling spectral cell will be $2\lambda/r$ by $2\lambda/c$. The resolution of the antenna is λ/Rr by λ/Cc, where R is the number of rows and C is the number of columns. To insure adequate definition of the individual lobes, there should be at least four data points per resolution unit. Therefore, there should be $8R$ by $8C$ data points. If the number of rows or the number of columns is not a "power of 2," the next larger "power of 2" should be substituted. The measurement plane can be "zero padded" beyond the actual measurement space to provide the needed resolution. Zero padding is preferred to actual measured data after the amplitude has fallen to the instrumentation noise level to avoid introducing additional noise into the spectrum.

G. Near Field

Measurements are made three to five wavelengths in front of the aperture. The angular spectra derived from the measured data differs from that derived from the aperture as

$$S_M(u,v) = S_A(u,v)e^{-j2\pi w(u,v)\frac{z}{\lambda}}$$

where

- $S_M(u,v)$ is the measured spectrum;
- $S_A(u,v)$ is spectrum which would be obtained by transforming the excitation at the aperture;
- $w(u,v) = \sqrt{(1-u^2-v^2)}$;
- z is the separation between the measurement plane and the aperture plane.

The angular spectrum at the measurement plane may be transferred to the aperture by correcting the phase distribution of the angular spectrum for the displacement of the phase center from the aperture to the measurement plane. This correction can be continued beyond visible space where it becomes a strong attenuating factor. This is a consequence of the evanescent character of plane waves with wave numbers in excess of 360° per wave length. Further, it implies that very little of the angular spectrum beyond the unit circle, i.e, visible space, will reach the measurement plane, even at only a few wavelengths distance. This is the reason that multiple scans and spectral merging is required when the spectral cell extends beyond the unit circle.

IV. Measurement Errors and Their Reduction

A number of errors arrise during the near-field measurement that affect the array pattern and, therefore, the array excitation determination. These errors and approaches to mitigate them are given below.

A. Inadequate Measurement Plane Size

A measurement plane of inadequate size will distort the angular spectrum and cause an error in the predicted gain of the antenna. A rule of thumb for measurement plane size is to extend a ray from the edge of the array in the direction of the widest angle of interest in the pattern to the point of intersection with the measurement plane. This simple approximation shows that the greater the separation between the aperture and measurement planes, the larger the scan plane must be. A more accurate determination for an antenna of known excitation would be to transform the excitation to the scan plane at the largest scan angle of interest, making certain that the scan plane covers that distribution to at least 30 dB below the peak value.

B. Probe Position Errors

Measurement data is required at points on a rectangular grid. Any displacement of the probe from these grid points at the time of measurement must be phase corrected. This assumes a probe location measurement system independent of the probe placement system. Laser interferometer systems often used in near-field scanners can measure probe position within a spherical error of less than 100 micrometers root mean square (rms) over a scan plane of more than 50 square meters. If the probe position is measured with such precision, errors in actual probe position relative to ideal position may be addressed by k correction [10] as well as other methods [11]. With k correction, phase is corrected as the projection of the error displacement along the direction of the main beam. Correcting phase for displacement along the scan direction assumes that the dominant plane waves are clustered in that direction. The k correction algorithm was compared to several others and was found to be the most cost effective, i.e., it provided the most correction for a given computational effort.

C. Multipath

Multipath errors result from multiple reflections of probe energy reaching the array. One important multipath contribution involves reflections from the array support structure to the probe carriage and back to the array elements. Carefully covering any support structure with absorber will alleviate this contribution to multipath. Another multipath source is multiple bounces between the probe tower and the array. Designing the probe carriage and the tower to minimize the number of surfaces parallel to the array and using absorber on the probe tower to attenuate multiple reflections are two principal methods to reduce this multipath error.

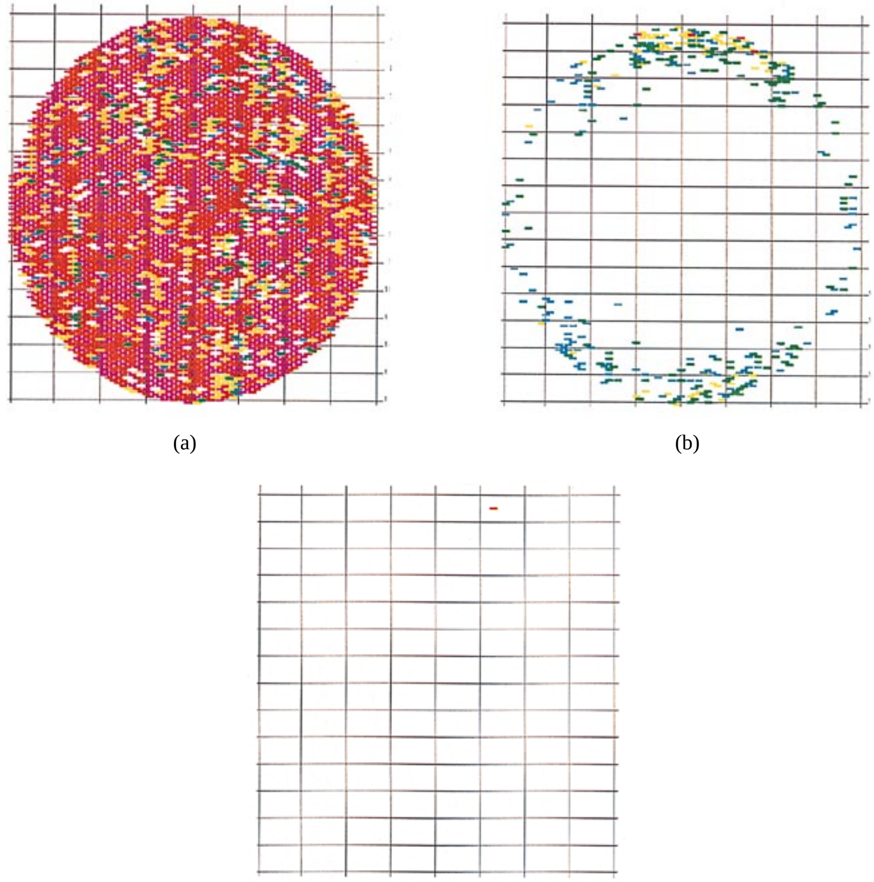
Several methods can be used to aid in diagnosing the source of multipath errors but none are fool-proof. The near field measured for each of the symmetrically positioned beams used for the merged spectrum processing can be compared. These should exhibit mirror or rotational symmetry with respect to each other. Lack of symmetry, especially in one of the beams, can lead to a source of multipath in the near-field facility. Also, if repeated measurements are made with the array at difference displacements along its normal, the multipath components can be extracted from the primary transformed far field or untransformed near-field patterns. This is similar to the sliding load measurement used for calibrating automatic network analyzers.

D. Receiver Settling Time

Maximizing the data collection rate necessitates a wide-band receiving system to respond to high-speed beam port, and frequency switching rates during a scan. However, a narrow-band receiver provides high signal-to-noise ratios (SNR's). This dilemma may be addressed by combining numerical integration, to improve the signal-to-noise ratio, and a relatively wide-band receiver (like a network analyzer) which settles quickly. The average of 16 readings taken at the reciprocal of the instrumentation bandwidth will reduce the noise bandwidth by a factor of 16 giving a 12-dB SNR improvement. Over-sampling the measurement plane also improves SNR in the spectrum. The Fourier transform process will also increase SNR by an amount equal to the directive gain of the array.

E. Time- or Position-Dependent Sensitivity

Stability and repeatability of the measurement system for periods of time greater than the scan duration is essential. In addition, cable drops and motional joints must have very low wow. Relaxation of a cable drop after motion can cause a phase hysteresis. The phase length of a cable can change



(a) (b)

(c)

Fig. 6. Alignment results (measured phase deviation from desired value). (a) Unaligned. (b) After single alignment with uncorrected measurements. (c) After alignment with fully corrected measurements.

when in motion, but relax to its stationary value a few seconds after stopping. Therefore, cables which flex durng the scan must have very good phase stablility and repeatability with motion. If this is the case, the effects of their phase variation with motion can be measured separately and subsequently removed from the antenna measurements. Such errors can be measured by placing a short circuit on the probe and reading the signal from a carefully balanced reflectometer. A second measurement is made and the corrections are applied. When the resulting data is then transformed using the same near-to-far field transformation used for pattern computation, the probe reflection and any other constant reflections will sum in the zeroth spectral line (broadside), whereas the instabilities will contribute to the sidelobe structure. Hence, this diagnostic tool relates the actual errors to their effect on sidelobe measurement accuracy.

F. Probe-Array Polarization Mismatch

The probe-array polarization mismatch is largely a signal-to-noise issue unless gain and polarization characteristics of the antenna are required. If only principal polarization data is desired, the probe polarization should be closely matched to the polarization of the array radiating elements. For example, using a rectangular waveguide probe to measure an array of rectangular waveguide radiators gives comparable data at all angles to measurements made with orthogonally polarized probes from which the principal polarization pattern is computed.

G. Array Induced Errors

A failed phase shifter will produce different errors in each of the merged beams. This error can be removed by rotating the phase of all elements of the array in increments of 45° during the scan. Counter rotating and summing during data analysis removes all influence of the failed phase shifters. By superposition the contribution of the element with the stuck phase shifter will add to zero after the counter rotation, whereas the contribution of elements with operating phase shifters will add coherently. This process also removes the effect of phase-shifter differential phase errors and provides additional signal integration as well.

H. Array Built-in Test

One additional lesson was learned while trying to simultaneously integrate a phased-array antenna and the near-field measurement system intended to measure it. The array must have sufficient built-in test equipment to ascertain that the array is fully operational, at least from an electronic control standpoint. Although the array alignment method described above can accurately measure the amplitude and phase at each element and demonstrate the effectiveness of the array control system, using the near-field scanner for this purpose is very expensive.

V. Applications

The process described above was invented to complement the design of a large phased-array antenna. It was necessary to achieve the required antenna performance and has been used to align a large number of such antennas. The alignment method described here provides the necessary phase precision at a large cost saving, compared to precisely manufacturing and trimming the line lengths within an antenna. The elemental phase errors measured by the system are stored within the antenna beam-steering system to correct phase errors at each operating frequency. The array is subsequently operated with the corrections applied.

The effectiveness of the alignment method is illustrated in Fig. 6(a)–(c). These show measured phase deviation from the desired value of: (a) an unaligned array; (b) the array after a single alignment without all error corrections applied; and (c) after a second alignment with all error corrections (probe position, phase shifter rotation, etc.) applied. The phase scales have been removed to illustrate the alignment progress qualitatively. Fig. 6(a) shows the phase excitation is totally random. The antenna does not, in fact, form a well-defined beam. After the first alignment, there are still some small phase errors around the periphery of the aperture. Finally, after a fully corrected alignment, only one element does not meet the minimum phase-error criterion of the display. In fact, this array had a single failed phase shifter, which could not be repaired. This is a rare but acceptable condition.

Facilities implementing the invention are currently in use. These facilities described in Yorinks *et al.* [12], cost approximately $10 million to build and facilitate to the size and precision necessary. If test and alignment requirements are less stringent, e.g., to provide a simpler test and evaluation tool, the facility design and implementation can be modified substantially in precision to reduce costs. For example, if only an indication of array errors and their location are desired, spectral merging may not be required and a single beam scan might suffice. In any event, the design of such an instrument can best be accomplished with foreknowledge of the antenna to be tested.

VI. Conclusion

This article presented a method for aligning a planar phased array that can resolve aperture excitation to individual elements. The method uses the periodic properties of the radiated far-field spectrum of an array antenna. It merges the data from measurements in different spectral regions into a complete spectral cell, which is subsequently transformed to the aperture plane. The method has been used successfully to align a very large number phased-array antennas.

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Willard T. Patton (S'57–M'59–SM'74–F'79–LF'95) was born in Schenectady, NY, in 1930. He received the B.S.E.E. and M.S. degrees from the University of Tennessee, Knoxville, in 1952 and 1958, respectively, and the Ph.D. degree in electrical engineering from The University of Illinois Urbana-Champaign, in January of 1963.

He served as Instructor in Electrical Engineering at the University of Illinois Urbana-Champaign from 1958 to 1962. He joined the Defense Electronics Division of RCA, Moorestown, NJ, in 1962 rising to Section Manager of Antenna, Microwave, and Transmitter Equipment Design in 1989. From 1989 until he retired from General Electric, Moorestown, in 1993, he served as Staff Scientist for the Engineering Department, Government Electronic Systems Division, GE. He was responsible for the design of antennas on the moon and on Mars, as well as the phased-array antennas for the Aegis Weapon System AN/SPY-1, AN/SPY-1A, and AN/SPY-1B. In the design of the latter array, he personally contributed the algorithms for near-field alignment, which gives that antenna its unique low-sidelobe performance on reception.

Dr. Patton received the David Sarnoff Award for Outstanding Technical Achievement in 1975 citing "...outstanding contributions to the development of multifunction tactical phased array radar..." and in 1985 citing "...for establishing new dimensions in radar antenna technology..." He holds the patent for a method of determining excitation of individual elements of a phased array from near-field data." He served on the Administrative Committee of the Antennas and Propagation Society from 1984 to 1986. He is a member of Eta Kappa Nu, Tau Beta Pi, Phi Kappa Phi, and Sigma Xi.

Leonard H. Yorinks (S'63–M'70) was born in Brooklyn, NY, in 1944. He received the B.E. degree from The Cooper Union, New York, NY, in 1965, and the M.S.E.E. and Ph.D. degrees from the University of Pennsylvania, Philadelphia, in 1970 and 1980, respectively.

He joined the Missile and Surface Radar Division of RCA, Moorestown, NJ, in 1965 and has held increasingly responsible positions through its transition to GE Aerospace, Martin Marietta, and Lockheed Martin. He served as Manager, Advanced Antenna Systems and Manager of the Antenna and Transmitter Center. He is currently Engineering Program Manager for Advanced Naval Radar Antenna Systems. He has been responsible for or played a major role in the design, assembly, and test of a number of phased-array antennas including the AEGIS AN/SPY-1A, and AN/SPY-1B antennas and the counter battery radar (COBRA) antenna—the first fielded ground-based tactical solid-state radar antenna. His current interests are in the design and test of large active phased-array systems. He has presented or published over 24 papers and holds four patents. He has also been a member of the adjunct faculty of Drexel University, Philadelphia, PA.

Dr. Yorinks was an officer of the Philadelphia AP-S/MTT-S Chapter from 1986 to 1989 and received the 1988 Outstanding Chapter Award as Chairman. He was a member of the Technical Program Committee of the 1986 International IEEE Antenna and Propagation Society Symposium and was Chairman of the 1989 Benjamin Franklin Symposium.