Broadband Planar Traveling-Wave Arrays (TWA) with 2-D Elements

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Abstract— *This paper discusses latest design of broadband planar Traveling-Wave Array (TWA), a class of broadband planar phased array antennas using planar 2-dimensional (2-D) broadband traveling-wave antenna elements closely spaced above a conducting ground plane, for low cost and multioctave bandwidth. The TWA is fundamentally different from the other two 2-D broadband phased arrays: Current Sheet Antenna (CSA) and Fragmented Aperture (FA). TWA arose directly from antenna concept with an inherent ground plane, while CSA and FA evolved from scattering technology. TWA is much thinner and lighter than CSA and FA, roughly by a factor o 5 and 15, respectively. This study employs an empirical design approach based on the technique of Active Element Gain Pattern (AEGP). Measured data showed TWA's potential for broad bandwidth and wide-angle scan. Some controversies on AEGP in the literature are also clarified.*

1. INTRODUCTION

Phased arrays are a key subsystem in communications, radars, imaging, etc. Phased arrays of the planar type have been most widely deployed due to their practical advantages such as transportability and relatively low life-cycle costs. However, for over half a century, broadband planar phased arrays have faced an insurmountable technical challenge in broadbanding. Their element spacing must be ≤ $\lambda/2$ or so, where λ denotes wavelength, throughout the operating band; otherwise disruptive grating lobes would emerge. On the other hand, small element spacing leads to strong mutual coupling between adjacent elements, creating complex and rapidly changing impedance which is difficult to match over a broad bandwidth.

2-dimensional (2-D) element antennas are mostly narrowband, such as the microstrip patch antenna, or too large to fit the unit cell of a broadband planar array. Thus ultra-wideband planar array designs have mostly employed 3-dimensional (3-D) element antennas, such as flared slots and horns. However, arrays of 3-D elements have large dimensions perpendicular to the plane of the array and in the E-plane, leading to high production cost and large weight and thickness as well as grating lobes in the E plane. Additionally, at the low end of the operating band, densely packed elements (element spacing «$\lambda/2$) have strong mutual coupling, making broadband impedance matching very difficult.

Broadband planar beam-scan array using 2-D radiating elements had been envisioned by Wheeler over four decades ago by way of a conceptual Current Sheet Antenna (CSA) [1]. Kornbau observed first in 1984 that a closely spaced array of dipoles without a ground plane can be designed to be very broadband [2, 3]. Hansen showed in 1999 [4] that a planar phased array using planar dipoles, without a ground plane, exhibits easy-to-match active resistance and fairly stable element gain pattern, over a wide range of scan angles and bandwidth (over 5:1). He noted, however, that the array failed to work when a ground plane is added to change the array's bidirectional radiation to unidirectional.

Recently, progresses have been made in implementing a ground plane for planar phased arrays of 2-D elements, with results showing multioctave bandwidth, wide-angle scan, low cost, and easy transportability. This paper reviews the state-of-the-art and presents a new design approach with some empirical and numerical data. Some controversial issues on the Active Element Gain Pattern (AEGP) technique, which was employed in the empirical design, were also clarified.

2. STATE-OF-THE-ART OF PLANAR ARRAY WITH 2-D ELEMENTS

Apparently inspired by their success in broadband FSS (Frequency Selective Surface) radome technique using dipole elements, Munk and other CSA researchers [3, 5-9] have been developing planar phased arrays of closely spaced flat dipoles with a ground plane. Bandwidths up to 10:1 have been reported. Their approaches are basically to place an electrical conducting ground plane behind the 2-D radiating elements with an effective $\lambda/4$ spacing, maintained by using layers of dielectric substrate and superstrate as λ would vary with the operating frequency. Some CSA arrays had to employ a lossy ferrite ground plane to dissipate power on the back side of the planar array, resulting in lower efficiency, lower gain and higher noise temperature.

Also apparently evolved from research in scattering, the Fragmented Aperture (FA) array [10-13] is based on placing several layers of metal foil fabricated in some patterns in front of a conducting ground plane to achieve an effective $\lambda/4$ spacing. To achieve wide operating bandwidth, electrical or optical control of the pattern of the middle foils may be used to move the effective ground plane. Some planar phased arrays taking the FA approach also used resistive layers as part of its backing. The practicality and performance of the FA array were questioned in [5] and [13].

The third approach for broadband planar beam-scan array using 2-D radiating elements is the Traveling-Wave Array (TWA), which is the subject of this paper.

3. The Traveling-Wave Array (TWA)

The Traveling-Wave Array (TWA) [14-17] approach evolved from the development of single broadband planar conformal 2-D TW (Traveling Wave) antennas [18-21] as well as the theoretical work in planar arrays [22-23]. These planar TW antennas are comprised of a planar 2-D TW structure backed by a closely spaced conducting ground plane. The success in developing a range of the single TW antennas with ultra-wide bandwidth (up to 10:1 or more), high efficiency, thin thickness, light weight, and platform-conformability has inspired the TWA approach for array applications.

The theory of planar phased array is generally formulated for the case of an infinite array. All individual array elements are of similar planar TW structure in similar unit cells. An example of TWA is characterized by its unit cell, *abcd*, as shown in **Figure 1** for a center element and its immediate vicinity in an infinite planar periodical array structure. The array elements are densely packed, spaced less than $\lambda/2$ between centers of adjacent elements throughout the entire operating frequencies, to suppress grating lobes; this is a common feature of CSA, FA, and TWA.

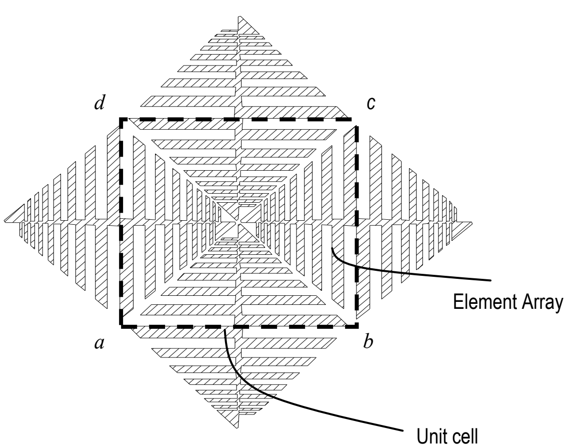


Figure 1 – A TWA unit cell and its immediate vicinity.

Strong coupling between adjacent elements in a TWA structure reduces discontinuities in phase and amplitude in the source aperture. Since the disruptive grating lobes that limit the bandwidth and scan angles of a planar phased array are rooted in the aperture source discontinuities between discrete elements of the array, adjacent elements in TWA are strongly coupled or even connected to smooth out the discontinuity in the aperture.

In Figure 1, the hatch lines denote conducting region, and the blank area are non-conducting. As can be seen, the planar structure is close to a self-complementary geometry, in which conducting and non-conducting regions are similar and not readily differentiable. Broadband impedance matching is thus made easy by this pseudo-self-complementary structure, based on Babinet's principle. It is worth noting here that, for a narrower bandwidth of, say, under 2:1, a non-self-complementary TW structure may be employed to optimize the performance of the phased array.

The cross-sectional view of the array is shown in **Figure 2**, in which dielectric superstrate and substrates further facilitate impedance matching. For circular polarization, the element antenna has four feed points in the center. However, similar to the TW antenna, the dielectric constants of the substrates/superstrates can be, and often preferred to be, close to 1.0 (free space).

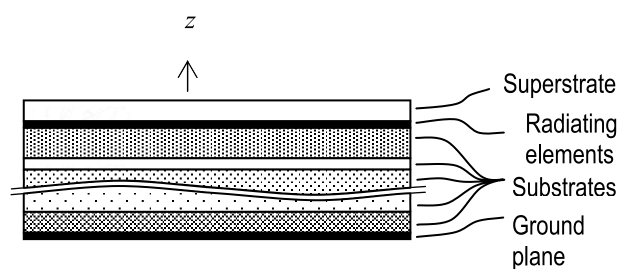


Figure 2 - Cross-sectional view of a planar TWA.

Different from CSA and FA, in TWA there are an inherent conducting ground plane and a TW structure, which also facilitate propagation of a dominant-mode TW transverse to the broadside axis z in the waveguide structure, as shown in **Figure 2**.

The design theory above is formulated for the case of a transmit array from the perspectives of radiation, yet applicable to the case of receive as well by way of reciprocity. However, it is worth pointing out again that one must be careful in applying reciprocity theorem in antennas, even if it is fully linear and isotropic, in both analyses and measurements, in particular for phased arrays, as will be discussed in further details later.

4. TWA Versus CSA and FA

TWA's distinguishing features and advantages, as compared with CSA and FA, are in its thickness, weight, and manufacturing cost. TWA is thinner than CSA and FA roughly by a factor of 3 and 15, respectively. TWA is lighter and less costly than the CSA and FA by a large margin since TWA does not have to resort to extensive use of dielectric substrate/superstrate (such as PTFE substrate).

It was noted that both CSA and FA have evolved largely from research in scattering of planar periodic structures. The scattering problem is generally simpler than the antenna problem, in theory and practice, yet it can lend profound insight to the related antenna theory and application. For example, from Babinet's principle, a fundamental theorem in electromagnetic scattering, the theory of planar frequency-independent antenna was developed: any self-complementary planar conducting structure has a constant impedance of 188.4 Ω (60π Ω) independent of frequency. However, in the case of a slot or magnetic current source, this self-complementary planar conducting structure is merely a scatterer, not a practical antenna.

To realize a practical antenna from a planar scatterer, one must make it radiate only to one side of the planar antenna. To this effect a parallel conducting ground plane and a feed network must be added to the scatterer; this is a key feature of the TW antenna. It took two decades of research for planar frequency-independent self-complementary scatterers to metamorphose into practical antennas — the TW antennas — as discussed by this author [18-21].

5. A TWA MODEL AND ITS PERFORMANCE

To perform empirical design, we employed the Active Element Pattern (AEP) technique, which has been widely used and reported in empirical design of planar arrays since the 1960s. However, the theory of this technique has not been clearly and formally addressed until three decades later by Pozar [24].

Recently, Hansen reported in his book [25] that he had replaced the terminology AEP by Scan Element Pattern (SEP), and emphasized that SEP is an extremely useful design factor. (Hansen did not seem to have formally defined SEP in the book, merely stated that it replaced AEP.) This author feels that the term AEP has been used extensively in the literature, but ought to be called Active Element Gain Pattern (AEGP). Adding the word "Gain" in the terminology is to emphasize that the antenna gain pattern, not merely the antenna pattern which generally refers to an arbitrary relative power intensity, is meant here. The theory of AEGP will be discussed later in further details.

To perform design study using the AEGP technique, a planar 11×11-element TWA was designed and fabricated, as depicted in **Figure 3** in top and cross-sectional A-A′ views. The planar array is comprised of planar 2-D FI (frequency independent) TW antenna elements [17] positioned parallel to a conducting ground plane spaced less than $0.5\lambda_H$, where λ_H is the wavelength at the highest frequency of the operating band under consideration. Adjacent antenna elements are tightly coupled with a coupling ring between them to facilitate continuous propagation of the TW and thus minimize phase granularity in the current distribution. Either linear or circular polarization can be achieved with a broadband feed network.

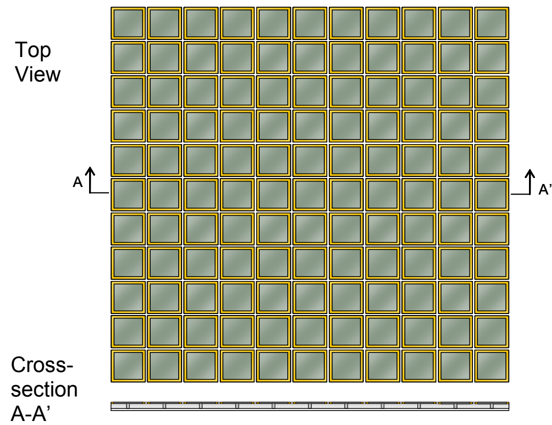
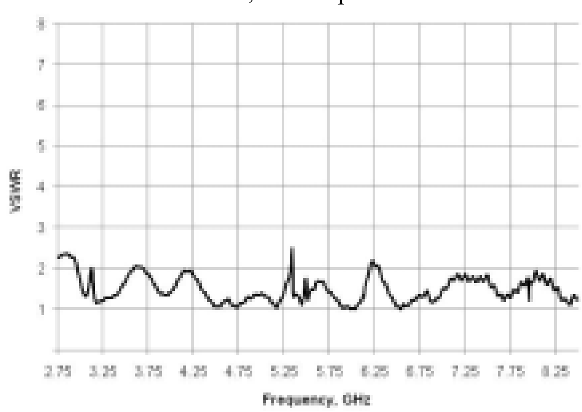


Figure 3 – Top and cross-sectional view of a planar TWA.

As an AEGP model, the center element has a transmitter or receiver, and other elements are terminated each in a 120 Ω load below the ground plane. (An isolated planar self-complementary TW antenna closely spaced from a ground plane has an impedance with a real part around 120 Ω and a very small imaginary part.) **Figure 4** shows measured active-element VSWR of the 11×11-element AEGP array of **Figure 3**. As can be seen, the impedance match is fairly



good over 2.75-8.25 GHz, with VSWR generally < 2:1.

Figure 4 - Measured active element VSWR.

Measured AEGP, which take into account mutual coupling and beam scan of a planar array, are generally well shaped, showing potential for wide-angle scan, as exemplified in **Figure 5** at three frequencies: 2.75, 4.25, and 6.00 GHz. The asymmetry in these patterns is largely due to fabrication and measurement errors. The array gain patterns are then obtained from the AEGP and the array factor. **Figure 6** shows computed array gain patterns with beam scan angles at 0° (broadside), 30°, 45°, and 60° off at these frequencies based on the active element gain pattern, as an approximation for a uniformly distributed phased array.

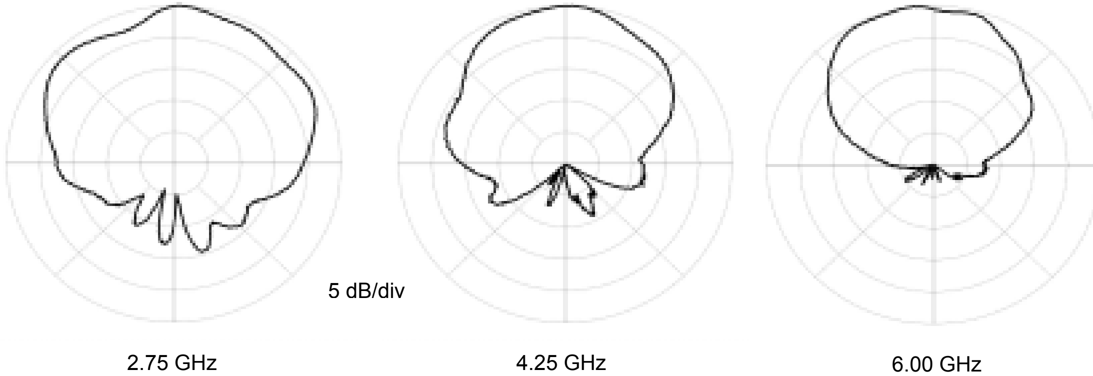


Figure 5 - Measured Active Element Gain Patterns (AEGP).

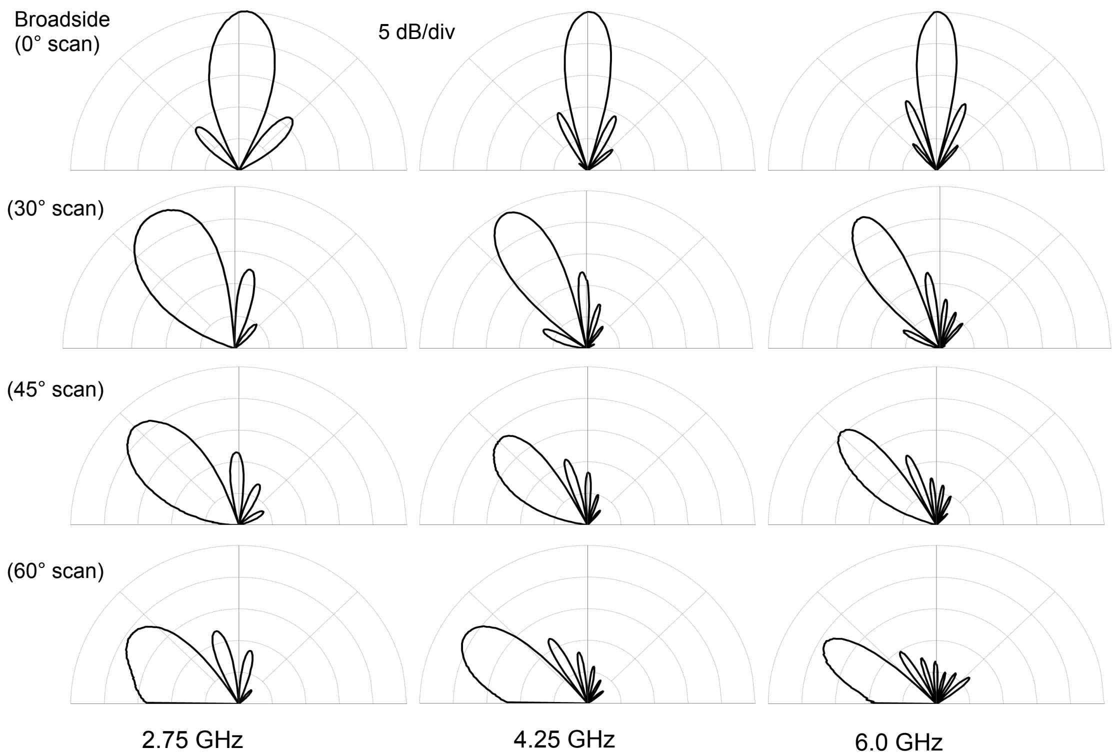
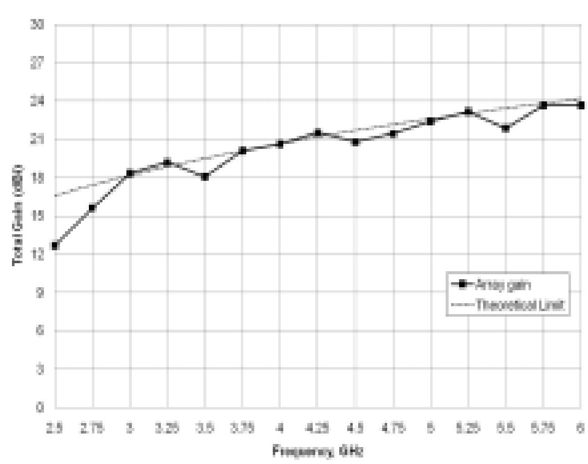


Figure 6 - Computed array gain patterns with beam scan angles at 0° (broadside), 30°, 45°, and 60° off.

Figure 7 shows realizable array broadside gain in comparison with the theoretical limit, in dBi, based on the measured AEGP (Active Element Gain Pattern), of the 11×11-element array of **Figure 3**.



The array gain is displayed in comparison with the theoretical gain limit for the aperture area of the array, given by $4\pi A/\lambda^2$, where A is the area of the array aperture. Similar to the single antenna, the concept of equivalent magnetic current and the principle of duality are essential in developing and understanding the TWA, and also helpful in interpreting the limited data of CSA and FA available in the literature.

Figure 7 - Measured broadside gain vs theoretical limit.

6. Theory and Controversies for Active Element Gain Pattern (AEGP)

Hansen claimed that the Scan Element Pattern (SEP) is significantly different between the transmit case and the receive case [25]. (Although Hansen did not formally define SEP, merely stating that it replaces AEP, his SEP appears to be identical to AEP and AEGP.) He also stated that SEP should be measured with the array in the receive mode, and that the scan impedance cannot be measured with a network analyzer. Such observation is inconsistent with the fundamental reciprocity principle for a linear isotropic electromagnetic system specified in an open or closed space, with the appropriate radiation or boundary conditions, respectively. In the present research, measurements were carried out for both transmit and receive cases, using network analyzers, with nearly identical results, contrary to Hansen's claims.

This author offers the following theory on AEGP not only to resolve this controversy but also to gain a more accurate grasp of the AEGP technique. We will consider the case of transmit. For the receive case, the array should exhibit identical characteristics, such as impedance, gain, and pattern, if the array's RF paths are passive (without amplifier, photonic delay lines, etc.) according to the principle of reciprocity.

In practice, an array element radiator is fed from a transmission line along which an electromagnetic wave propagates from a source, as shown in **Figure 8**, for a planar array of $M \times N$ elements, with only row 1 displayed.

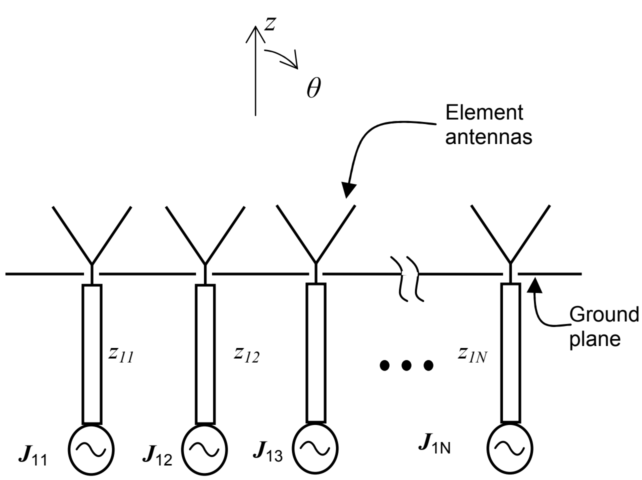


Figure 8 – Side view of a planar array of $M \times N$ elements with only row #1 displayed.

It is noted [e.g., 22-23] that analytical and numerical solutions often treat a source as a spatial delta function of voltage or current, as in [25], ignoring the inherent internal impedance of a real source in the physical problem. When replacing the RF transmission line representation of the source feed network with that of a simple circuit and a source in the form of a spatial delta function, controversial results like those in [25] would arise.

This $M \times N$-element planar array is excited by a current source $\boldsymbol{J}$ represented by a column matrix as

$$\boldsymbol{J} = \{\boldsymbol{J}_{mn}\}, \tag{1}$$

where $\boldsymbol{J}_{mn}$ denotes the current source for element mn, as shown in **Figure 8** for row #1. For a specific source $\boldsymbol{J}$, there are corresponding fields $\boldsymbol{E}(\boldsymbol{r})$ and $\boldsymbol{H}(\boldsymbol{r})$ at a selected arbitrary point $\boldsymbol{r}$ in the far zone of the array antenna.

For convenience of the present discussion, we will first consider the ideal case in which each array element is impedance matched between the radiating element, the feed transmission line, and the source (that is, $z_{mn} = Z_{mn}$ = element impedance = a real number for each element mn.).

Consider an AEGP case, Case k, in which only element pq is excited, as shown in **Figure 9** (with only row p displayed). The array source, according to Eq. (1), denoted by $\boldsymbol{J}^{\boldsymbol{k}}$ for Case k, is given by

$$\boldsymbol{J}^{\boldsymbol{k}} = \begin{pmatrix} 0 \\ \vdots \\ 0 \\ \boldsymbol{J}^{\boldsymbol{k}}_{\boldsymbol{pq}} \\ 0 \\ \vdots \\ 0 \end{pmatrix} \tag{2}$$

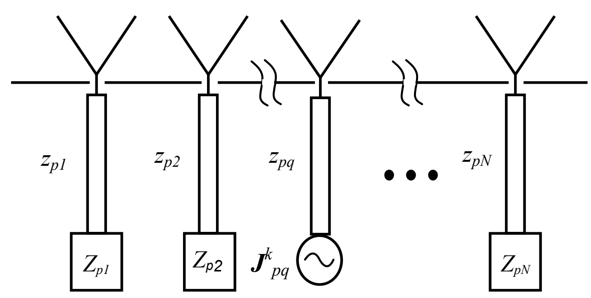


Figure 9 –AEGP Case k with only element pq excited (with only row p displayed).

The corresponding electric and magnetic fields at an arbitrary point $\boldsymbol{r}$ in the far zone, for Case k, are denoted by $\boldsymbol{E}^k(\boldsymbol{r})$ and $\boldsymbol{H}^k(\boldsymbol{r})$, respectively. Note that the AEGP for Case k is proportional to $\boldsymbol{E}^k(\boldsymbol{r})$ or $\boldsymbol{H}^k(\boldsymbol{r})$, and is a complex vector which contains the amplitude, phase, and polarization characteristics of the antenna gain $\boldsymbol{G}$. If element pq is near

the center of the array, the element properties, in particular the AEGP, are fairly stable and can be used as the basis for the array design.
It is worth commenting that, in **Figure 9**, the characteristic impedance of the transmission line, z_{pq}, does not have to be identical to the generator impedances of the array elements, Z_{pq}. They can be different and be impedance matched via a transformer section of adequate bandwidth placed in between.

When all the element sources in the array are on, that is, for full excitation according to Eq. (1), we can invoke the principle of superposition and thus obtaining the corresponding fields $\boldsymbol{E}(\boldsymbol{r})$ and $\boldsymbol{H}(\boldsymbol{r})$ as follows.

$$\boldsymbol{E}(\boldsymbol{r}) = \sum_{k=1}^{M\times N} \boldsymbol{E}^{k}(\boldsymbol{r}) \tag{3}$$

$$\boldsymbol{H}(\boldsymbol{r}) = \sum_{k=1}^{M\times N} \boldsymbol{H}^{k}(\boldsymbol{r}) \tag{3}$$

Now, beam scan of the phased array can be achieved either by generating the desired phase for $\boldsymbol{J}_{mn}$ in Eq. (1), or by varying the length of the individual feed transmission lines. Under the assumption of full impedance match made earlier in this section, the impedance at this scan angle is the same as in the case without scan (with a broadside beam).

If the three impedances (for radiator, feed network, and source) are not matched for the broadside case, beam scan would change the impedance of the element radiator, making broadband impedance matching difficult.

8. Other Design Issues

There is another problem that also contributes to the controversy discussed here. A subtle, complex and very important relationship between scattering and radiation is often overlooked in the study of the single antenna [26]. For the phased array, a similar intricacy exists but, to the author's knowledge, has not been discussed in the open literature. This would also contribute to the apparent discrepancy between receive and transmit cases.

A theory is also being developed independently for the case of receive, from the perspectives of the information sampling theory and electromagnetic scattering theory, leading to results similar to the transmit case and offering useful and clear insights in key design issues such as impedance matching, gain efficiency, and scan angles.

7. Advantages of The Traveling-Wave Array (TWA)

CSA, FA, and TWA are all in varying stages of research. At present, TWA has the following significant advantages:

1. TWA is thinner, by a factor of about 5, than CSA, and 15 times thinner than FA.
2. TWA is much lighter in weight since it is much thinner and does not resort to dissipative loading or substantial dielectric loading,
3. TWA has an inherent ground plane.

8. Conclusions

The planar Traveling-Wave Array (TWA), as a class of broadband planar phased array antennas using planar 2-dimensional (2-D) broadband traveling-wave antenna elements, is shown to be fundamentally different from the other two 2-D broadband phased arrays: Current Sheet Antenna (CSA) and Fragmented Aperture (FA). TWA arose directly from antenna concept with an inherent ground plane, while CSA and FA evolved from scattering technology. TWA is much thinner and lighter than CSA and FA, roughly by a factor of 5 and 15, respectively. Measured data based on the technique of Active Element Gain Pattern (AEGP) showed TWA's potential for broad bandwidth and wide-angle scan. Some controversies on AEGP in the literature are clarified.

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