Impedance Matching of Open–Ended Waveguide Radiating Elements†

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ABSTRACT

Impedance matching of open–ended waveguide radiators in a planar phased–array environment is a non–trivial problem that is aggravated by the constraints placed on the element size by antenna needs. The approach discussed here achieves an efficient matching structure, based on the use of a waveguide simulator approximating the free–space impedance presented to the element in the array environment

I. INTRODUCTION

Electronically scanning planar phased–array antennas usually have a configuration in which radiating elements at the aperture plane are fed separately by transmission lines with individually controllable phase. These radiating elements must ordinarily satisfy the following constraints and requirements:

a. Physical size must be compatible with array element spacing lattice dimensions.
b. Element pattern must be broad enough to permit scanning over the required angular limits without excessive reduction of antenna gain.
c. Mutual coupling between elements must be low enough to avoid excessive variation of the driven impedance level with scan angle.
d. Good impedance match and low conductive, dielectric, and magnetic losses are required for efficient operation.
e. For certain applications it may be necessary to accommodate circular polarization as well as linear, with minimal generation of cross–polarization.

One candidate geometry for the radiating element is the small, dielectrically loaded, flush circular or square waveguide recommended by Wheeler [1]. With sufficiently small element size, this approach is known to provide low mutual coupling and broad element patterns consistent with wide angle scanning free from "blind spots". The major drawback to this approach is the difficulty of realizing an impedance transformer with even moderately wide bandwidth. This paper explores considerations and proposes a very simple approach to the design of impedance transformers for such elements in a planar phased–array environment.

A key assumption for the following discussion is that the radiation impedance presented to the element in the array environment can be adequately approximated by the impedance presented by a simulator whose axis represents an extension of the element waveguide. The aperture plane is then represented by a junction between the small element waveguide and the larger, air–filled simulator waveguide. It is not the purpose of this paper to delve into techniques for choosing an appropriate simulator waveguide, since a body of literature already exists on this subject, including the basic report of Hannan and Balfour [2]. In general, various simulators are needed to achieve field patterns at the element surface corresponding to particular scan angle excitations, and very few of these comprise only a single element located at the axis of the simulator waveguide. The approximation made here does not insist on exact field configuration, but only that the on–axis simulator should present a good representation of the typical impedance seen by the element over the important range of scan angles. Thus, great simplification of calculation is achieved at the expense of some loss of generality.

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II. EQUIVALENT CIRCUIT MODEL

Following Marcuvitz [3], it can be recognized that the junction of the element and simulator waveguides may be represented by an equivalent circuit consisting of an N:1 transformer, a shunt susceptance, and a shift of reference plane in one of the guides (see Figure 1). Because the element

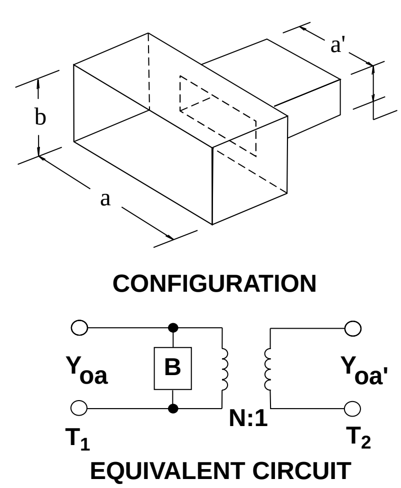

Fig. 1. Equivalent circuit for the junction of simulator and element waveguides

guide is somewhat smaller in cross–sectional dimensions than the simulator guide, the following experimentally confirmed observations have been made:

a. The equivalent shunt susceptance at the junction between the element waveguide and the larger simulator waveguide is essentially independent of the dielectric filling of the element waveguide, because virtually all the stored energy is in evanescent modes entirely within the larger waveguide.

b. A unique value of the N:1 transformer ratio exists that is independent of the (uniform) dielectric filling of the waveguides.

c. The shift of reference plane is negligible for most cases of interest.

To arrive at a quick and easy closed–form approximation to the conductance and susceptance relationships at the waveguide junction, consider the superposition of the formulas and curves given in Sections 5.24a and 5.26a of Marcuvitz[3], for symmetric changes in width and height of rectangular waveguides. It is found that for a wide range of cases, very good agreement is obtained between experimental and analytical results under the following rules:

1. In case the element waveguide is circular, it is replaced in the computations by a square waveguide of about 6.5 percent smaller cross–sectional area.
2. The N:1 transformer turns ratio is calculated using the width–change formulas. Figure 2 shows the computed turns ratio versus width change.
3. A shunt capacitive susceptance is assumed, equal to that calculated by the height–change formulas.
4. A shunt inductive susceptance is assumed, equal to that calculated by the width–change formulas, and with the result increased in proportion to the height ratio.

Although this approximation may seem questionable, it has been used as the basis for a general–purpose analysis program that has yielded excellent results in predicting hardware behavior for many cases. The closed–form equations are especially convenient for allowing rapid presentation of results over a range of frequencies and for various iterations of parameter values.

III. RADIATING ELEMENT MATCHING APPROACH

The equivalent–circuit model provides some insights into the difficulties associated with matching of the small radiating element. In the unconstrained case where two different waveguides are to be matched, it is customary to introduce two or more quarter wavelength sections of intermediate size with optimally chosen characteristic impedance values. Step discontinuity susceptances are usually small and occur at quarter–wavelength spacing, so that frequency slope compensation occurs. A minor adjustment of length or transverse dimensions then generally suffices to achieve a good impedance match over a bandwidth approaching the theoretical limit.

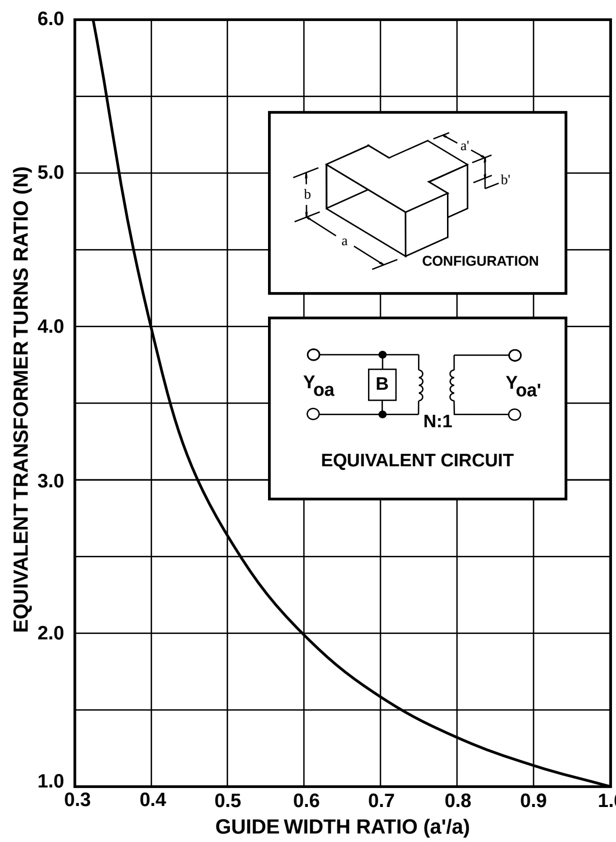

Fig. 2. Equivalent circuit transformer ratio at a change of waveguide width

In contrast, the radiating element case introduces the severe constraints that no intermediate waveguide sizes are possible; that the element waveguide dimensions are fixed at small values by the demands of array lattice spacing, wide element patterns, and low mutual coupling; and that the element guide is probably too small to propagate unless dielectrically loaded. As a consequence of these constraints:

a. A fairly large shunt inductance is likely to exist at the interface between the element waveguide and the simulator waveguide. Figure 3 shows a normalized plot of susceptance versus element size.

b. The large jump in transverse dimensions between the element and simulator waveguides translates into a large value for the N:1 transformer of the equivalent circuit. Consequently, a large real part ratio must be accommodated by the matching circuitry.

c. The characteristic impedance needed for the initial quarter wavelength section in the element waveguide is likely to be very high and attainable only near cutoff if homogeneous, uniform dielectric filling is used. This condition implies a very rapid change of impedance with frequency, and correspondingly small bandwidth for good impedance matching.

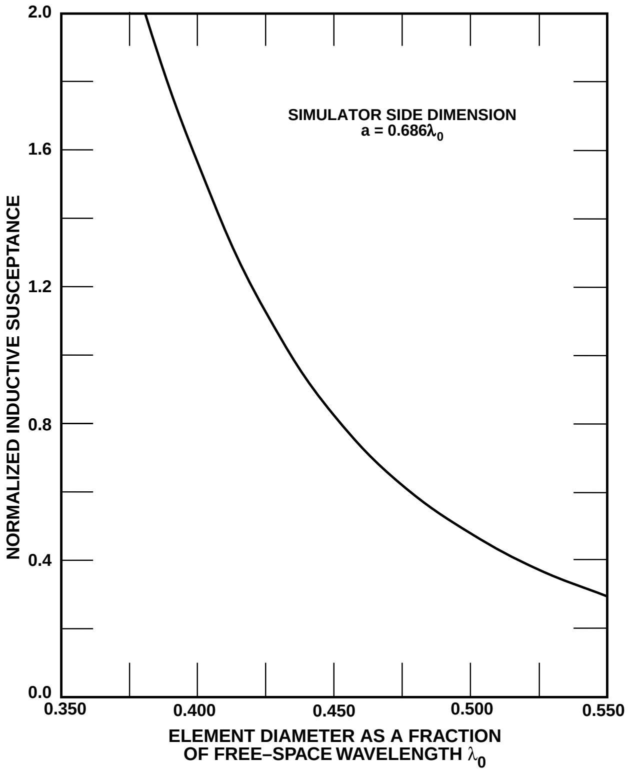

Fig. 3. Circular element inductive susceptance (normalized by free–space simulator admittance) vs. normalized element diameter.

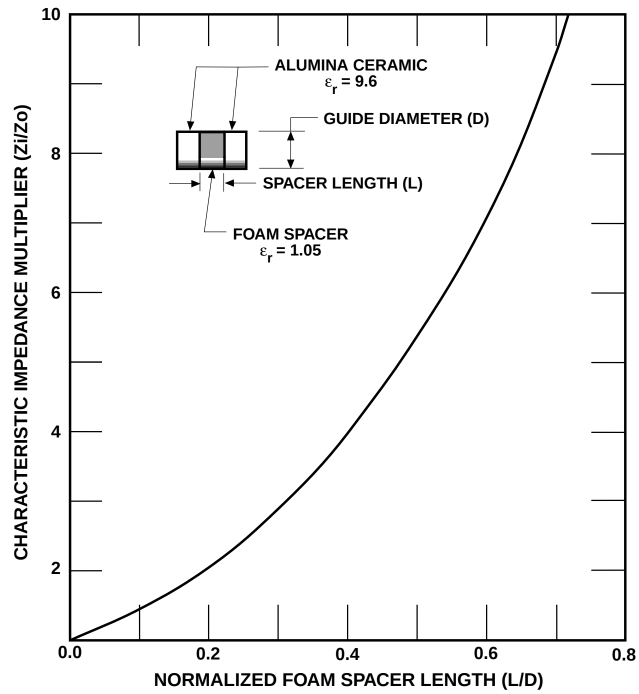

Fig. 4. Increase of characteristic impedance of a circular dielectric–filled guide of $\lambda_g/4$ length, resulting from insertion of a high impedance center section.

Condition (a.) generally requires that a quasi–lumped shunt capacitance, such as a ceramic wafer, be used at the end of the element waveguide to resonate the aperture–plane shunt inductive susceptance, with frequency slope compensation introduced as discussed below. Condition (b.) simply suggests that a multiple quarter wavelength tranformer be realized within the limits of the element waveguide transverse dimensions, and that bandwidth limitations will exist because of the large change of real part. That is, the N:1 transformer effect usually causes the simulator waveguide impedance to increase by a multiplier of three to five times when viewed from the element waveguide. Characteristic impedance values for the matching transformer must take into account this phenomenon. The Z_0 values may also be adjusted to provide the frequency slope compensation of the resonated aperture plane susceptance. For example, a compensating negative imaginary–part slope can be produced, over a limited frequency range, in a two–section transformer by increasing the two Z_0 values in the same ratio. This approach tends to aggravate further condition (c.) above, but fortunately a simple technique is available to overcome the difficulty.

This technique is to use a quasi–artificial transmission line arrangement in which a quarter wavelength section is formed from two ordinary propagating guide sections separated by a high impedance section, possibly a cutoff waveguide. Figure 4 shows a plot of the relative increase of apparent characteristic impedance for a circular waveguide $\lambda_g/4$ section consisting of two equal lengths of alumina ceramic (ε_r = 9.6) separated by a variable length of foam–filled guide (ε_r = 1.05). It is clear that a very significant increase of characteristic impedance can be achieved (over a limited bandwidth) by this technique. For a multiple section transformer, the structure takes on the appearance of a filter–like arrangement, and resembles practical element–matching structures that have been reported in the literature [1].

IV. Examples

Figures 5 and 6 show examples of open–ended waveguide radiator matching structures with calculated and measured data. In the case of Figure 5, the element waveguide is square with a side dimension of 0.435 λ_0, where λ_0 is the free–space wavelength at the center frequency of the plot. A square air–filled simulator waveguide of side dimension 0.672 λ_0 is assumed. A thin ceramic wafer of $\varepsilon_r \approx 6.3$ is used to resonate the aperture inductive susceptance. The source waveguide feeding the element is completely filled with a material of dielectric constant $\varepsilon_r \approx 2.2$, and the two quarter wavelength sections are indicated by the regions containing the foam–filled volumes. The design achieves a return loss of about 20dB. or greater over a 12.9 percent bandwidth with good correspondence between the measured and computed results.

In Figure 6, a circular element waveguide is shown with diameter of 0.391 λ_0, where λ_0 is again the free–space wavelength at the center frequency of the plot. In this case the simulator is a square waveguide of side dimension 0.679 λ_0. The element matching consists of two quarter wavelength transformers, in which the second section is formed in a smaller circular waveguide coupling directly to the ferrite rod of a dual–mode reciprocal phase shifter. Because of the great change in size between the simulator and the ferrite rod, the real–part ratio that must be matched is on the order of 23:1. A 17dB. minimum return loss is achieved over a bandwidth slightly greater than 6 percent. The measured data shows additional frequency fluctuations introduced by the matching transformer at the feeding end of the ferrite phase shifter.

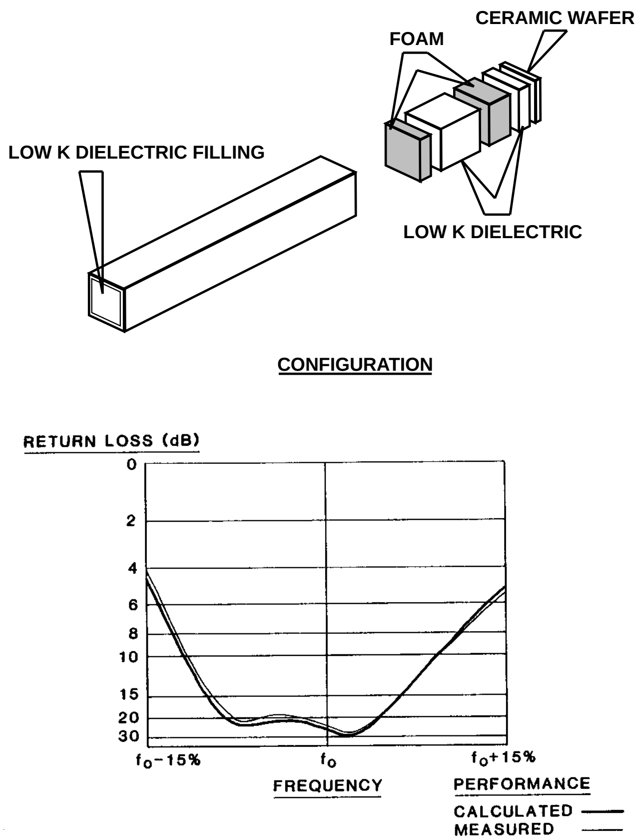

Fig. 5. Square element matching transformer.

V. References

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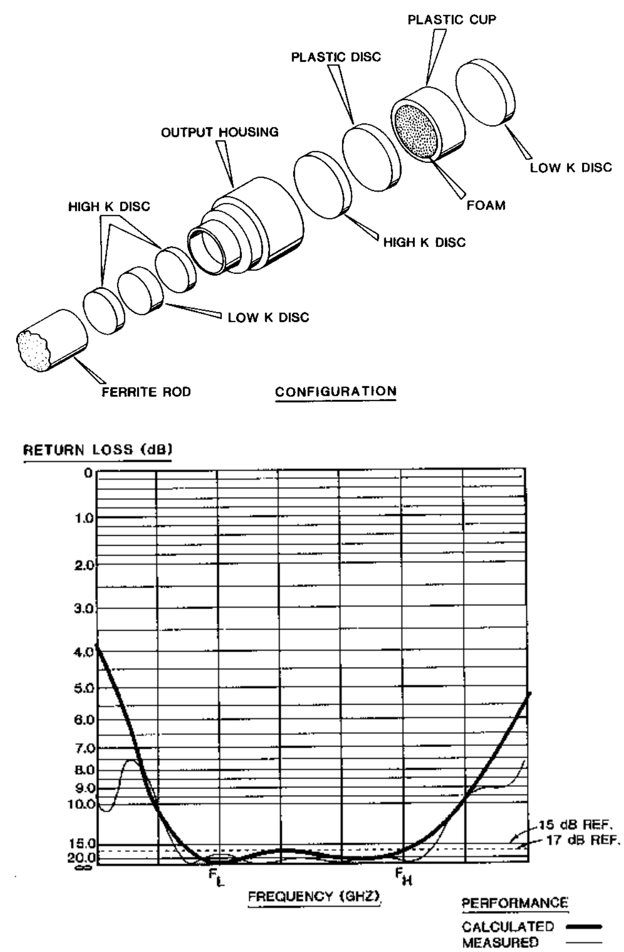

Fig. 6. Radiating element transformer for ferrite phase shifter.