An Efficient Approach for Characterizing Power Combining Dividing Network of Distributed Phased Array Antenna

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Abstract — **An approach that mixes numerical and analytical techniques is proposed for modeling the circuit discontinuities and the power-dividing networks that feed the T/R modules in Synthetic Aperture Radar. This approach is based on a combination of mode matching and boundary integral techniques, and suitable for modeling a wide range of N-port microwave networks made with waveguide, coaxial and microstrip components. While it offers the possibility of handling complex shaped junctions, it converges much faster than other numerical methods implemented in commercial software package such as FEM, FD, MoM and others.**

I. INTRODUCTION

RF subsystems of space borne Synthetic Aperture Radar (SAR) generate high power microwave transmitter pulses and amplify the Radar echoes received through the antenna for digitization and processing. Distributed phased array antenna employs multiple Transmit/Receive (T/R) modules distributed across the physical apertures of the antennas. In this approach, the total transmit power is shared among all the modules, and each only needs to transmit at relatively lower peak power. The key technology issues, in this approach, are reduction in the antenna weight and significant improvement in their power efficiency. The power efficiency improvement can lead to the reduction in requirement on the space craft power subsystem with corresponding saving in mission cost. This requires formidable design efforts to characterize the extensive passive network circuitry, which contains various kinds of discontinuities and multiport junctions.

A well-designed feed network contributes significantly in the reduction of return loss and in the improvement of power efficiency. In this paper, we describe a theoretical approach based on a combination of mode matching and boundary integral techniques for analyzing various discontinuities and multiport configurations. The proposed approach represents a relevant alternative over the time consuming numerical 3D standard numerical techniques used in commercial packages.

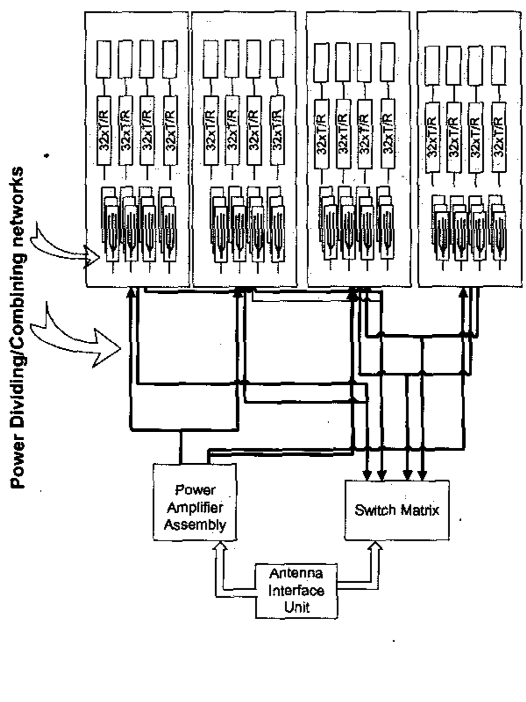

Fig.1 Block Diagram of SAR System

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II. MODAL EXPANSION

One example of circuits to be studied is shown in figure 2. In the case of a waveguide junction, the widths of ports are denoted W_1, W_2 , .. Wn. For a microstrip junction, the concept of planar waveguide model is used, which permits the use of effective values ε_{eff}, $W_{eff,1}$, $W_{eff,2}$, $W_{eff,n}$. In this work, the analytical formulation proposed by Kobayashi [1] is used for calculating the dynamic effective values. The dominant mode is assumed incident from port 1. At the reference plane of i^{th} port (i=1,2,.. n), local coordinates are defined so that the origin is on the port wall (Fig. 2). The electric field and its derivative at $z_i = 0$ are then expressed in the following way

$$E_y = \delta_{1,i} + \sum_{m=1}^{\infty} R_m^i \, f_{mi} \tag{1}$$

$$\frac{\partial E_y}{\partial n} = \gamma_{1,i}\, \delta_{1,i} - \sum_{m=1}^{\infty} R_m^i \, \gamma_{1,i}\, f_{mi} \tag{2}$$

with

$$f_{mi} = \cos\left[\frac{(m-1)\pi}{W_{eff,i}} x_i\right] \tag{3}$$

$$\gamma_{mi} = \sqrt{\left[\frac{(m-1)\pi}{W_{eff,i}} x_i\right]^2 - K_i^2} \tag{4}$$

where δ_{1i} is the Kronecker delta and
$K_i^2 = K_0^2\, \varepsilon_{eff,i}$. K_0 is the wave-number in free space.

III. BOUNDARY INTEGRAL FORMULATION

The discontinuity region Ω in Fig. 2 is delimited by a contour Γ which is divided in many sub-contours: $\Gamma_{w,i}$ (i=1,2,…,n) represent interfaces (at $z_i = 0$) between Ω and the uniform waveguide or microstripline; Γ_c coincides with the electric or magnetic walls. In the discontinuity region, the electric field u = Ey satisfies the Hemholtz equation:

$$\left(\nabla^2 + K^2\right) u(\vec{r}) = 0 \tag{5}$$

Boundary conditions on Γ have the general form:

$$u = \bar{u} \qquad \hat{n}\cdot\left(\vec{\nabla} u\right) \equiv q = \bar{q} \tag{6}$$

n is the outward normal to Γ, and u and q are prescribed functions on Γ. Using the fundamental solution u^*_p and Green identity, the following equation for the field E_y can be obtained

$$\alpha\, E_y(\vec{r}_p) = \int_\Gamma \left[\frac{\partial E_y(\vec{r})}{\partial n} u_p^* - E_y(\vec{r}_p) q_p^*\right] d\Gamma \tag{7}$$

The functions u^*_p and q^*_p are the Green's function, for equation (5) and its normal derivative, respectively. They are given as [2]

$$u^* = -\frac{j}{4} H_0^{(2)}\left(K\left|\vec{r} - \vec{r}_p\right|\right) \tag{8}$$

$$q^* = \frac{\partial u^n}{\partial n} = -\frac{j}{4} K\, H_1^{(2)}\left(K\left|\vec{r} - \vec{r}_p\right|\right) \cos(\beta) \tag{9}$$

where β is the angle between $(r - r_p)$ and n, and $H_0^{(2)}$ and $H_1^{(2)}$ denote the second-kind Hankel function of zero and first order, respectively.

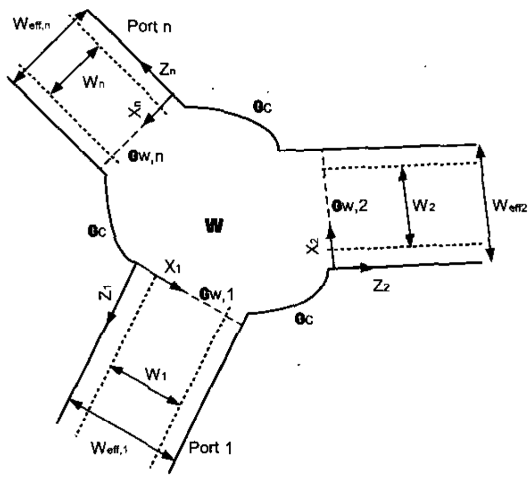

Fig. 2: n-port Junction

IV. MATRIX FORMULATION

When we apply the boundary conditions on the magnetic and electric walls and truncate the infinite series in equation (1) and equation (2) to a finite number of modes N_i, equation (7) leads to

$$\frac{1}{2}E_y^p+\sum_{i=1}^{Nw}I_{w,i}+I_e+I_m=0 \tag{10}$$

Where

$$I_{w,i}=\delta_{ij}\int_0^{weff,i}(q^*-\gamma_{0,i}u^*)\,dx_i+\sum_{m=o}^{Ni}R_m^{ij}\times\int_0^{weff,i}\phi_{m,i}(q^*+\gamma_{m,i}u^*)\,dx_i \tag{11}$$

$$I_e=-\sum_{i=1}^{Ne}\int_{\Gamma l}\frac{\partial E_y}{\partial n}u^*\,d\Gamma_l \tag{12}$$

$$I_m=-\sum_{k=1}^{Nm}\int_{\Gamma k}E_y\,q^*\,d\Gamma_k \tag{13}$$

In these equations, the incident, transmitted and reflected coefficients R_m^{ij} are independent coefficients, this is why they were taken outside the integrals. Therefore, there is no need to use interpolation approximation on $\Gamma_{w,i}$ which permits to extract directly the S parameters. This is one of the advantages of this approach.

In order to obtain a matrix formulation suitable for numerical analysis, a discretisation technique has been applied on contours Γ_c which are divided into N_c elements, respectively. E_y and $\partial E_y/\partial n$ within each element are defined only at nodal points. In this way, equations 10 turn into the following algebraic system:

$$\begin{bmatrix}[M^{11}] & [M^{12}] & [M^{13}] & [M^{14}]\\ [M^{21}] & [M^{22}] & [M^{23}] & [M^{24}]\\ [M^{31}] & [M^{32}] & [M^{33}] & [M^{34}]\\ [M^{41}] & [M^{42}] & [M^{43}] & [M^{44}]\end{bmatrix}\begin{Bmatrix}\{R_\Gamma\}_{w1}\\ \{R_\Gamma\}_{w2}\\ \{R_\Gamma\}_{w3}\\ \{R_\Gamma\}_{w4}\end{Bmatrix}=\begin{Bmatrix}\{V^1\}\\ \{V^2\}\\ \{V^3\}\\ \{V^4\}\end{Bmatrix} \tag{14}$$

For a =1,2,3,4 and b= 1,2,3 the element of the sub-matrixes are given by

$$M_{\alpha\beta}^{ab}=\frac{1}{2}f_{\beta b}\delta_{ab}+\int_0^{weffb}f_{\beta b}\left[q_\alpha{}^*+\gamma_\beta u_\alpha{}^*\right]dx_b \tag{15}$$

$$M_{\alpha\beta}^{a4}=-\int_{el\beta}u_\alpha{}^*d\Gamma \tag{16}$$

The elements of the vectors $\{V^a\}$ are given by

$$V_\alpha^a=-\frac{1}{2}\delta_{al}-\int_0^{weff1}\left[q_\alpha{}^*-\gamma_1 u_\alpha{}^*\right]dx_1 \tag{17}$$

The solution of the linear system (14) gives the complex amplitudes of higher modes at the reference planes in all lines

$$\{R_\Gamma\}_{wl}=\begin{Bmatrix}R_0^{ij}\\ R_1^{ij}\\ \vdots\\ R_N^{ij}\end{Bmatrix} \tag{18}$$

As the incident wave is normalized, S parameters are defined as

$$S_{ij}=R_0^{ij}\sqrt{\left(\frac{W_{eff,j}}{W_{eff,i}}\right)} \tag{19}$$

R_1^{ij}, R_2^{ij}, … R_n^{ij} are the amplitudes of higher modes ate the reference plan of the ports.

V. RESULTS

The proposed approach is based on the mode matching technique together with boundary integral formulation, its numerical accuracy is essentially related to approximation and applicability of the mode matching as well as discretisation of the boundary contour. As the mode matching is widely analytical, its accuracy is superior to pure numerical techniques. Thus, this combined technique requires fewer nodal points than either the finite element or boundary element method, and proves to be efficient in reducing calculation time.

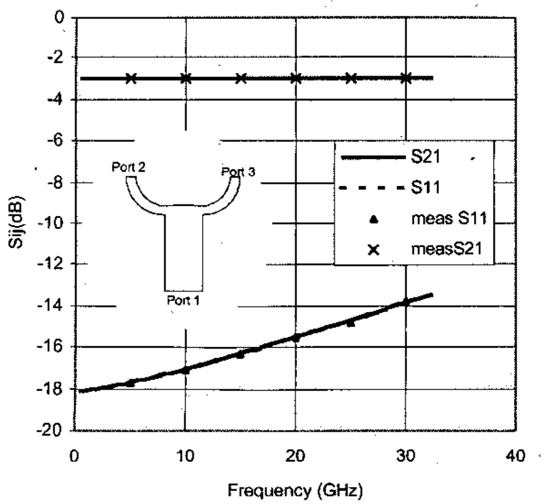

Fig. 3 Curved Y power divider junction

Our attention focuses on the effectiveness of the proposed approach for application to a wide range of planar circuits with multiport and arbitrarily shaped discontinuity such as the T/R feed networks in SAR system. Figure 3 shows that scattering characteristics of three-port curved planar Y-junction is well simulated by this approach. In this example, $w_1 = 0.236$ mm, $w_2 = w_3 = 0.032$mm, the radius of the circular junction is 1.016 mm and the thickness of the substrate is 0.254mm.

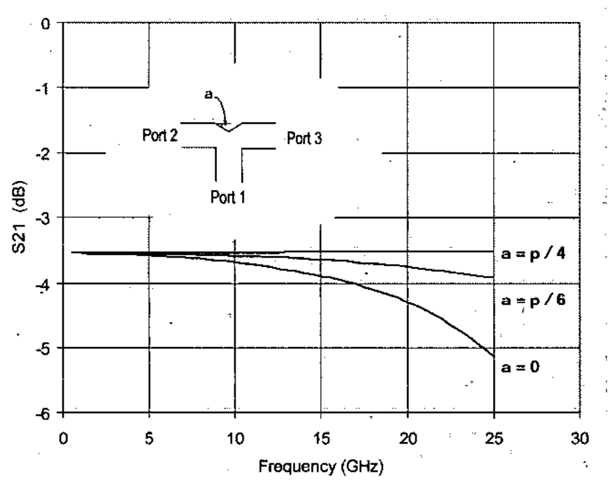

Figure 4: Mitered T junction

Figure 4 shows S parameters of a right-angle four-port junction. Figure 5 shows S parameters of mitered T junction with different angles. In these example a RT/duroid 6010 substrat ($\varepsilon r = 10.2$, $\tan(\delta) = 0.002$) has been used.. The efficiency of this approach is confirmed by the excellent agreement with the experimental results obtained over the frequency range up to 30 GHz

VI. CONCLUSION

A novel design and modeling approach has been presented for the analysis of arbitrary discontinuities and multiport RF/microwave circuit. This has been essentially achieved by combining a boundary integral technique with a mode matching method. Circuit with irregular contour and curved multiport could be accurately analyzed using this approach. This proposed technique will be very useful for the analysis of a variety of multiport junction, particularly in SAR distributed feed system where extensive passive dividing combining network is needed.

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