Accurate Characterization of the Interaction Between Coupling Slots and Waveguide Bends in Waveguide Slot Arrays

Giuseppe Mazzarella and Giorgio Montisci

Abstract—In a waveguide slot array, it is sometimes required to introduce bent short-circuit terminations in the feeding network. This significantly affects the behavior of the coupling slots used in this network, with a very large variation in the coupling coefficient with respect to the standard case. A procedure to accurately evaluate the effect of such bends is presented, thus allowing to include them without any loss in the overall design accuracy. It is based on the method of moments, using a magnetic-field integral equation expressed in terms of the vector potential, which appears to be the most efficient way for waveguide problems. The development is aimed at a very effective implementation, which allows to include it in design tools for waveguide slot arrays without increasing the total computational load, and has been assessed through comparison with experimental results.

Index Terms—Slot array, slot coupler.

I. INTRODUCTION

Through their use dates back to the 1940’s, slot arrays are still quite popular as high-performance antennas, mainly in the higher part of the microwave band. Such antennas guarantee a very high efficiency, usually a very low crosspolar level and a compact realization. Moreover, their design can be done with a great accuracy.

The basic array (or sub-array) structure consists of a main waveguide which feeds, through a rotated slot [1], a number of crossed branch waveguides containing the radiating slots. The coupling slot is a series one, thus, the feeding guide must be terminated with a short-circuit a half-wavelength beyond the center of the last coupling slot and, therefore, juts out of the radiating waveguides [see Fig. 1(a)]. This causes an enlargement of the array that cannot be always tolerated. Moreover, when a slot array is used as a monopulse radar antenna, with the four quadrants separately fed, there is absolutely no room to accommodate this long short-circuit section. In all these cases, the best solution is to bend the waveguide end with a 90° or a 180° curve, as in the transverse section of Fig. 1(b) and (d) (see also Fig. 1(c) for a three-dimensional (3-D) view of the single-bent-short-circuit termination). Both an isolated waveguide bend [2] and an isolated coupling slot [3] can be analyzed by known techniques, but those techniques cannot be used in our case since the slot is not isolated, except inside the curved section. Therefore, it interacts with the bend through its near field. In this paper, we describe an accurate modeling technique for such a near-field interaction, fully taking into account also the waveguide wall thickness. This is obtained with a full-wave method of moment (MoM) procedure. A set of coupled integral equations are obtained by forcing the continuity of the tangential magnetic field at both slot apertures and at the input section of each bent waveguide stub [see Fig. 1(b) and (d)]. The magnetic field in the waveguide is expressed through the Fitzgerald vector potential $F$ computed using its Green function expressed as modal series. Discretization of the unknown magnetic currents and the Galerkin procedure then lead to a linear system whose solutions give the magnetic currents and the scattering matrix of the junction.

This strategy is surely able to get a very accurate analysis of the coupler, but its use must be carefully assessed in an array design procedure. As a matter of fact, array design procedures usually require a very large number of slot analysis and, therefore, the computational weight of a full-wave approach can be too large to prevent its use. Therefore, an analysis procedure that must be both accurate and effective for use in array design, is needed to overcome this problem. To this end, the strategy proposed here have been developed into an analysis procedure in which all steps have been carefully devised and implemented to reach the goal of a very effective and accurate characterization of a slot in a bent termination. First, an entire domain basis function representation for the unknown currents has been used. As discussed, e.g., in [4], this requires a very small number of unknowns. The resulting linear system is, therefore, small and can also be well conditioned. Moreover, the matrix elements, which are no more than field–current reactions computed over surfaces whose size is comparable with the wavelength, varies very smoothly with the frequency or the geometrical parameters. This suggests that a polynomial interpolation of the matrix elements [5] will significantly reduce the matrix fill time by more than an order of magnitude, and the same reduction is obtained in the total analysis time since the system solution requires a very small fraction of the total computational time. The interpolation error involved is negligible even for three-point Lagrange interpolation used here (for details, see [6]).

II. PROBLEM FORMULATION

According to the equivalence theorem, all apertures in the structures shown in Fig. 1(b) and (d) can be replaced by the suitable equivalent
magnetic current $M_i$ on $\Sigma_i$, $M_s$ on $\Sigma_s$, and $M_{sl}$ on $\Sigma_{sl}$. These currents are unknowns and can be computed by enforcing the continuity of the tangential magnetic field at the apertures themselves. As a matter of fact, taking the wall thickness into account requires to use an equivalence theorem on both slot apertures and, therefore, two different unknowns and equations are needed. Including an incident $TE_{10}$ in the radiating guide field as forcing term, we get a set of magnetic-field integral equations (MFIE’s) whose solution allows to compute all entries of the scattering matrix. There is a continuity equation for each aperture ($\Sigma_s$, $\Sigma_i$, $\Sigma_{sl}$, etc.). For the 90° bend, these equations are

$$H_S = H_{RG} + H_{inc}, \quad H_{FG} = H_S, \quad H_B = H_{FG},$$

where

$H_{inc}$ Incident $TE_{10}$.

$H_{FG}$ Magnetic field in the feeding guide region, which depends on $M_i$ and $M_{sl}$.

$H_S$ Magnetic field in the slot region, which depends on $M_i$ and $M_s$.

$H_{RG}$ Magnetic field in the radiating guide region, which depends on $M_i$.

$H_B$ Magnetic field in the waveguide bend region, which depends on $M_{sl}$.

For the 180° bend, a further equation on $\Sigma_{sl}$ is required. Since $H_{RG}$, $H_S$, $H_{RG}$, and $H_B$ depend upon the unknown currents, (1) can be seen as a set of integral equations, which can be transformed by the MoM into a set of linear algebraic ones.

All unknown currents are expressed as a linear combination of suitable basis functions. The slot is usually narrow enough to neglect the longitudinal component of the electric field on it [3]. Therefore, only the axial-directed magnetic current is used as an unknown.

Since all computational surfaces have a rectangular shape, entire domain basis functions have been used everywhere. The slot currents are expressed as truncated Fourier series with respect to the axial coordinate $\xi$ (centered on the slot), as

$$M = \sum_{p=1}^{N} a_p \sin \left[ \frac{p\pi}{L} (\xi + \frac{L}{2}) \right] i_c = \sum_{p=1}^{N} a_p f_p(\xi)$$

where $a_p$ are the expansion coefficients, which are different on the internal and external surfaces of the slot, $f_p$ are the expansion functions, and $L$ is the slot length. The bent termination can be considered as a “stub” waveguide. The magnetic current upon its aperture section $\Sigma_{sl}$ is discretized in a way similar to (2), but using as basis functions the currents corresponding to the well-known [7] tangential electric field of the stub waveguide modes. The section $\Sigma_{sl}$, if present, is dealt with in the same way.

To compute the unknown currents, (2) is substituted into (1) and the resulting MFIE equations are scaled multiplied by the test functions and integrated over the corresponding aperture $\Sigma$ to get a linear system in the unknown coefficients. Since the basis functions form a complete set on each aperture, they are also the best choice for test functions (Galerkin procedure) because the resulting linear equations are the most accurate approximation, in the mean square norm, of the integral equations for a given number of test functions. A few terms in (2) are needed to compute the response over a quite wide frequency range around the resonant frequency of the slot so that the resulting linear system is small. Moreover, since we have chosen orthogonal basis functions, the system is also quite well conditioned.

Now the $B$-field Green function has a Dirac delta singularity at the source location [7], whatever the magnetic current direction, while its modal series expansion contains only an axial impulsive term. The other singular components are hidden into the modal series and this causes a poorer convergence of the series and, in some cases, even a divergent behavior. As a consequence, the MoM matrix elements computation is quite difficult, whatever method is used. To simplify the computation, the waveguide magnetic field can be expressed through the vector potential $F$, which has only a mild singularity at the source location.

$$H = j \omega \epsilon F + 1/(j \omega \mu) \nabla \cdot F,$$

wherein $\omega$ is the angular frequency and $\epsilon, \mu$ are the permittivity and permeability of the medium in which $H$ is computed, we need both the $F$ potential and its divergence. In a waveguide, these can be computed in terms of the magnetic current $M$ as a modal series (see [8]), in the same way as the $H$-field Green function can [7]. Use of those Green functions for $F$ and $\nabla \cdot F$, therefore, lead to a modal series representation for the matrix term. These series are truncated, retaining all modes whose cutoff wavenumber is smaller than a given value $K$.

III. EXPERIMENTAL ASSESSMENT AND RESULTS

In order to assess the procedure described in the previous sections, we have compared for both structures in Fig. 1(b) and (d) the results of
between a coupler with a standard short circuit and a  

it can be observed that  

two different configurations of these parameters are considered. From  

of our simulations with some experimental data. Of course, the accuracy  

are enough to get an accurate response.  

These values will, therefore, be used for all subsequent results. Larger  

that such a few expansion terms are needed not only at the resonance, as  

The experimental comparison presented in Fig. 2(a) (coupler with a  

due to the near-field interaction and partly to the bend itself. In order  

A reasonable assumption, supported by our test, is that the unfolded  

A full wave MoM procedure to take into account the near-field inter-

our simulations with some experimental data. Of course, the accuracy  

on each section and on the truncation level of the modal series. In Fig. 2(a),  

of the modal series. In Fig. 2(a), two different configurations of these parameters are considered. From  

these values are required only for highly accurate design. It is worth noting  

ulated near-field interaction between the slot and bend. Of course, it is also important to quantitatively evaluate  

IV. C ONCLUSION  

A full wave MoM procedure to take into account the near-field interaction between a series slot and a bent waveguide short-circuit has  

of the bent short circuit must be the same as the length of the standard short circuit,  

A typical case is shown in Fig. 3(a) for a single bend and in Fig. 3(b)  

is considered. The slot length allows the slot to resonate at the chosen test frequency (28.4 GHz) for  

IV. C ONCLUSION  

A full wave MoM procedure to take into account the near-field interaction between a series slot and a bent waveguide short-circuit has  

All the procedure steps have been carefully devised to get a very  

not vary very much with the tilt angle for our standard guides [3], we  

REFERENCES  


Fig. 3. Difference in $S_{21}$ between a coupler with a standard short circuit and a  

coupler with: (a) a single-bent or a (b) double-bent short circuit for $B_{21} = B_{23}$  

Fig. 4. Variation in $S_{31}$ with respect to the values for a slot isolated from the  

termination. All data as in Fig. 3(a).  

Fig. 4. Variation in $S_{31}$ with respect to the values for a slot isolated from the  

termination. All data as in Fig. 3(a).
A Varactor-Tuned RF Filter

Andrew R. Brown and Gabriel M. Rebeiz

Abstract—An electronically tunable filter at 1 GHz is presented in this paper. The filter uses a suspended substrate design and commercially available varactors for filter tuning. The filter has a 60% tuning range from 700 MHz to 1.33 GHz with a low insertion loss (better than 3 dB from 1 to 1.33 GHz). This paper discusses the effects of the varactor series resistance and the electrical length of the distributed resonator on the overall resonator quality factor and filter insertion loss. The input third-order intermodulation product intercept point was measured to be better than 17 dBm across the entire tuning range.

Index Terms—Frequency control, tunable filters.

I. INTRODUCTION

Low-loss tunable frequency filters are often used as tracking filters for multiband telecommunication systems, radiometers, and wide-band radar systems. Typically, tracking filters are mechanically tuned by adjusting the cavity dimensions of the resonators or magnetically altering the resonant frequency of a ferromagnetic YIG element [1], [2]. Neither of these approaches can easily be miniaturized or produced in large volumes for wireless communication products. The filters must be custom machined, carefully assembled, tuned, and calibrated. An alternative to the mechanically tuned and YIG filters is based on solid-state varactor diodes. Varactor filters have previously been developed using two- to three-pole filters [3]–[5]. However, the effects of the varactor series resistance and electrical length of the distributed portion of the resonator have not been investigated.

An electrically tunable capacitively loaded interdigital filter is presented in this paper. The tuning element is a reverse-biased varactor diode. The resonators of the tunable filter are shortened interdigital fingers with varactor diodes at the ends. The coupling is carefully controlled by the geometry of the fingers and the tuning is performed by changing the bias on the varactor diodes. Since both the interdigital fingers and diodes are carefully controlled and fabricated in batch, this filter can be easily produced in large quantities. The varactor-controlled tunable filter is based on a suspended substrate stripline technology. The suspended substrate allows for a very low effective dielectric constant, resulting in very wide low-loss transmission lines.

II. CENTER FREQUENCY OF A VARACTOR-LOADED TRANSMISSION-LINE RESONATOR

The design of a varactor-loaded interdigital filter is similar to the capacitively loaded combline filter presented in [1], but is adapted for the interdigital topology. The interdigital filter is a symmetric filter of coupled resonators. The first finger at the input and output ports is a shorted line that acts as an impedance transformer for the filter. This is the only line with a fixed termination. The interior coupled lines are shorted at one end and loaded with varactor diodes at the other end (Fig. 1). To allow for biasing, large capacitors are added ($C_{bias}$). When the bias voltage is changed, the thickness of the depletion region of the varactor diodes changes. This alters the capacitance of the varactor tuning the resonant length of fingers. The width and separation of the interior lines are determined only by the bandwidth of the normalized filter response function, and is independent of the center frequency. The center frequency of the filter is determined by the resonant lengths of the lines, which is tuned by the varactors. The tuning range for the filter is limited by the fixed lengths of the input and output finger lengths, internal impedance of the filter, range of capacitance of the varactor diodes, and electrical length of the fingers.

The electrical length of a single finger $\theta$ without the capacitive loading is given by

$$\theta(V_{bias}) = 2\pi f L \sqrt{\varepsilon_{eff}} / c$$

(1)

Fig. 1. (a) Topology of the varactor loaded interdigital bandpass filter. (b) Cross section of suspended substrate stripline.