THEORETICAL AND EXPERIMENTAL CHARACTERIZATION OF NONSYMMETRICALLY SHIELDED COPLANAR WAVEGUIDES FOR MILLIMETER WAVE CIRCUITS

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ABSTRACT

The nonsymmetrically shielded coplanar waveguide (NSCPW) is proposed as a quasi-TEM transmission line with advantageous characteristics for mm-wave circuit applications. The structure combines properties of the finline and the suspended stripline. In addition to a full wave theoretical analysis, an experimental technique has been developed which enables the evaluation of the transmission line spectrum in a wide frequency band (15:1) with a single transmission measurement. Results are shown to be in excellent agreement with theoretical predictions. Design curves are also presented.

1. INTRODUCTION

A number of different transmission lines have been proposed for applications in the millimeter wave range [1]. E-plane and suspended line structures seem to be the most viable candidates for applications in the range from 60 to 100 GHz. These transmission lines belong to the class of quasi-planar circuits. They combine the advantages of the waveguides in terms of low loss and of planar circuits in terms of integration capabilities.

Despite their advantages with respect to passive mm-wave circuits, the above mentioned technologies exhibit some difficulties with respect to active device embedding, e.g. for amplifiers or oscillators. For instance, the finite cut-off frequency of finlines implies reactive loading of active devices at low frequencies, which may cause stability problems. Also the intrinsic isolation between adjacent circuit elements is poor, resulting in unwanted feed-back paths. Suspended line, on the other hand, makes device mounting difficult, often resulting in increased parasitics and degraded performances of both packaged and unpackaged devices.

In this paper we propose a modified structure combining the advantages of both finline and suspended stripline. It consists of a coplanar waveguide inserted into a nonsymmetrical housing. The cross sectional geometry of the nonsymmetrically shielded coplanar waveguide (NSCPW) is shown in Fig. 1a. By simply eliminating the grounded metal fins we obtain the nonsymmetrical suspended substrate stripline (NSSS), Fig. 1b. The unsymmetrical construction of the split-block housing simplifies circuit mounting by soldering or epoxy-bonding it to one half housing. Critical mechanical tolerances and poor electrical contacts at mounting grooves can be circumvented in this way.

The quasi-TEM propagation of both variants allows broad-band matched device embedding, which makes the present structure substantially different from the unsymmetrical finline proposed in [2]. A wide range of characteristic impedances and different field configurations can be achieved by changing between NSCPW and NSSS. The former is well matched to packaged transistors, since most of the electric field is concentrated in the plane of the substrate involving less field distortion in the vicinity of the transistor housing. NSSS has pronounced low loss performance, since most of the field is concentrated in the air gap between the strip and the housing wall.

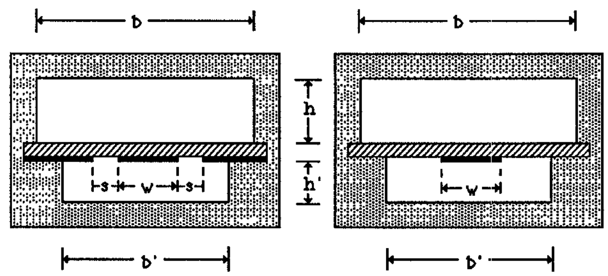

Fig. 1. Cross sectional geometries of the NSCPW and NSSS

NSCPW is best suited for:
- mounting of beam lead and three terminal solid state devices;
- matching elements in both series and shunt configurations;
- biasing circuits;

while NSSS is useful for:
- filters (e. g. end-coupled bandpass, broadside-coupled bandstop filters)
- directional couplers, power splitters
- low loss line sections, transition to waveguides.

Both theoretical and experimental investigations of the NSCPW/ NSSS structures have been carried out and are presented in this paper.

A sophisticated analysis method is required to solve the associated boundary value problem, including the effect of the metallization thickness and optional mounting grooves. One such technique is the transverse resonance method which is adopted here and is briefly recalled in section 2. An experimental method has also been developed which allows the broadband evaluation of the modal spectrum of the structures. The propagation characteristics of the dominant and higher order modes are obtained by a simple, yet very accurate method, described in section 3. An excellent agreement between theory and experiments has been obtained as demonstrated by the results presented in section 4. Based on theoretical computations, design curves in terms of characteristic impedance and usable frequency range are also given.

2. METHOD OF ANALYSIS

The method of analysis is the generalized transverse resonance method already described in a number of papers [3-6]. The symmetry of the cross section can be used to analyze only one half of the structure, by replacing the symmetry plane by a magnetic wall (or an electric wall for odd-symmetric modes). The reduced cross section is shown in Fig. 2a. (The metallization thickness has been exaggerated for illustration purposes.) Furthermore, we insert two electric walls at distance L one from the other so as to form a resonant cavity. As seen in the x-direction normal to the substrate, the structure appears as the connection of four rectangular waveguides (see Fig. 2a) with same width L but different heights.

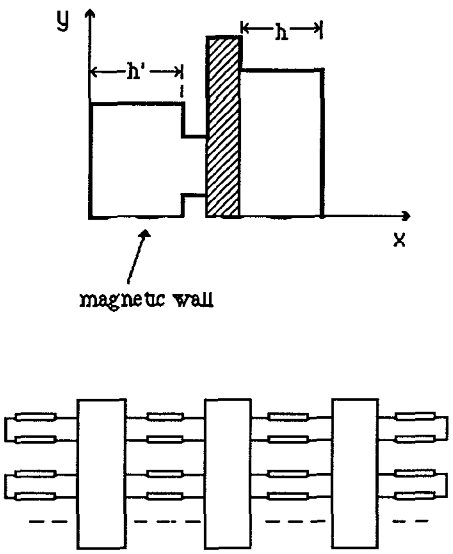

Fig. 2 Reduced cross section of NSCPW for analysis (a) and generalized equivalent transverse circuit (b)

Fig. 2b shows the corresponding generalized equivalent transverse circuit. Each waveguide section is represented by a (theoretically infinite) number of transmission lines, each corresponding to a different mode. Contrary to [5] and [6], however, different number of modes are chosen in the various section in order to properly account for the relative convergence phenomenon. The mode order in the z direction correspond to the number of half-wavelengths contained within the length L of the cavity. In fact, the z dependence of the electromagnetic field is of the form of a standing wave with spatial wavenumber

$$\beta = n\pi/L \qquad (1)$$

n being the number of half wavelengths. The step discontinuities between the waveguide sections are represented by generalized multiport networks. They produce the coupling among the transverse waveguide modes (or LSE, LSM) [7].

The resonant condition of the generalized equivalent transverse circuit of Fig. 2b constitutes the dispersion relation for the structure which has the form

$$f(\omega,\beta)=0 \qquad (2)$$

For any given distance L and resonant order n, (1) and (2) determine the resonant frequency f_{in} of the cavity for the ith mode, thus the frequency associated with the value of β given by (1). The complete CPW spectrum can be determined by numerically simulating different resonant experiments either with different cavity lengths L, or, if L is chosen large enough, increasing the resonant order n. The latter solution corresponds to the experimental procedure described next.

3. EXPERIMENTAL CHARACTERIZATION

To validate the theoretical results an experimental method has been developed to determine the propagation constants of all lower-order modes across a very large (15:1) bandwidth. They can be obtained from only two broadband transmission measurements of a properly designed two-port resonator formed by the waveguide structure under test. The two measurements correspond to the even-symmetrical and odd-symmetrical modes, respectively.

A brief initial investigation revealed that none of the well known methods:
- open end resonance line (with capacitive end-coupling);
- ring resonance technique;
- nodal shift technique
- four short/four open technique

is applicable for this purpose. These techiques are only useful for the fundamental mode and require at least two measurements on different line lengths to eliminate detuning effects introduced by the field coupling probes.

Instead, a method similar to [8] has been used, where a small probe-current was introduced into one of the two short-circuits terminating a resonant microstrip line. For our measurements we used a complete closed cavity of about 5 wavelengths at center frequency with massive short circuit plates at both ends.

A two-port resonator is realized by introducing one coaxial launcher above and another below the substrate at one end of the resonator. By shorting the center conductors to ground and leaving part of the outer conductor open to the waveguide, weak magnetic coupling with negligible detuning effect is obtained. Selective excitation of even-symmetrical modes is possible by orienting the coaxial launchers in the axial direction or perpendicular to the substrate. Odd-symmetrical modes are excited with the center conductor oriented transversely to the longitudinal axis and parallel to the substrate.

With a proper choice of the coupling aperture sizes, broadband coupling to all modes of interest with small frequency variation is achieved. A local maximum of the transmission between the two coaxial ports occurs at every frequency where the cavity length is an integer multiple of $\lambda/2$ for one of the propagating modes excited. This is due to a maximum transverse magnetic field of the respective mode when a virtual short is transformed to the excited end of the resonator.

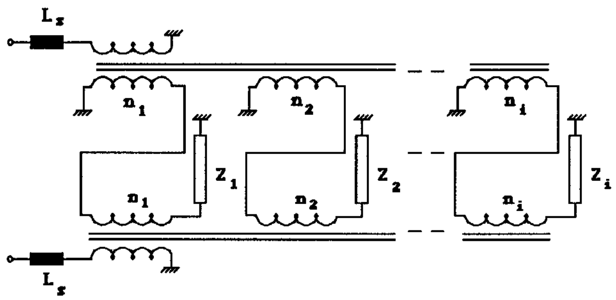

Fig. 3 Simplified equivalent circuit of the experimental set up

The contribution of the different modes to the resultant transmission can be explained by the approximate equivalent circuit of Fig. 3. Here n_i denotes the coupling coefficient (both mode- and frequency-dependent), Z_i is the multiple resonant input impedance of *i*th mode and L_s denotes the lumped series inductance of the launchers. Due to the high Q (typically 500) of the individual resonances, their interference is very small. The mode number i and longitudinal order n can be determined directly from one broadband swept frequency measurement. A typical measurement plot is shown in Fig. 4. Due to the absence of detuning effects, the propagation constant β_i of mode *i* at frequency f_{ni} can be determined using (1), or, in normalized form

$$\frac{\beta_i}{k_o} = \frac{n\,c_o}{2Lf_{ni}}$$

Typical transmission values are -30 to -5 db, depending on mode and frequency.

4. RESULTS

Two housings have been fabricated for experimental verifications, one for a NSCPW and one for a NSSS. With reference to Fig.1, dimensions are b=22.86mm, b'=15.8, h=b/2 for both structures; h'= h for the NSCPW and h'=0.1 b for the NSSS. The geometries of the printed circuits were determined by the theory of section 2 to have 100 and 50 Ω characteristic impedance respectively. Transmission measurement were performed in the frequency range 0.5 to 15 GHz on cavities 15 cm long. This allowed the evaluation of the dispersion characteristics β vs. f of the first three or two modes of the guiding structures. The orientation of the coaxial probes was such that only the even-symmetrical characteristics were evaluated. To re-utilize the circuits for further experiments it was decided not to solder the circuits to the housing. To properly hold the circuits and to ensure good electrical contact between the metal fins and the housing, grooves were milled in the lower housing. These were included in the theoretical simulations.

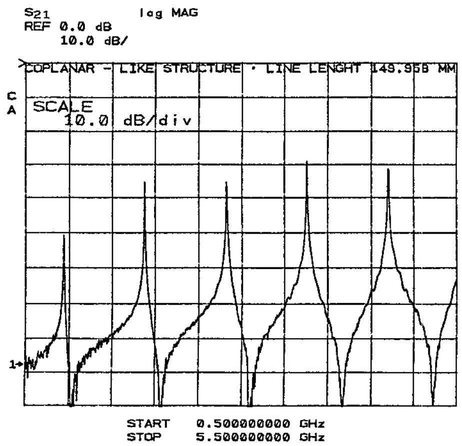

Fig. 4 Experimental transmission measurement

The theoretical and experimental results for NSCPW and NSSS are compared in Fig. 5. An excellent agreement is observed. Error is always less than 2%, generally around 1%. Both structures exhibit very low, almost negligible, dispersion of the dominant quasi-TEM mode. In the frequency range examined the NSSS has only one higher order mode instead of two as for the NSCPW. This is due to the reduced housing height h'. Additional measurements were performed on structures obtained by interchanging the substrates with the housings. The theoretical and experimental results for these cases are again in very close agreement.

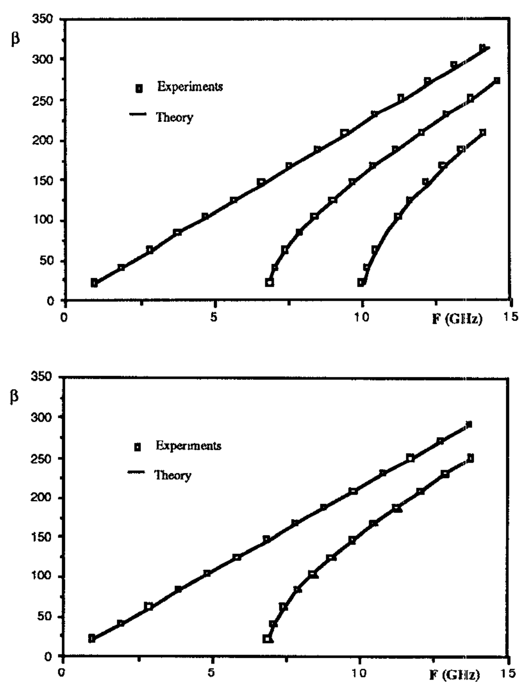

Fig. 5 Theoretical and experimental even mode spectra of NSCPW (a) and NSSS (b).

A theoretical investigation has then been made to characterize the structure in terms of bandwidth and characteristic impedance for various structural parameters. The characteristic impedance has been evaluated according to the voltage/power definition [4]

Figs. 6 and 7 show the characteristic impedance of the NSCPW as a function of the strip width w (normalized to w+2s) for different values of w+2s (normalized to b'), Fig. 6, and for different h' (normalized to b), Fig. 7. The case (w+2s)/b' =1 in Fig. 6 corresponds to the NSSS case. These figures show that, for the structures analyzed, the range of realizable characteristic impedances range from about 30 to 200 ohms.

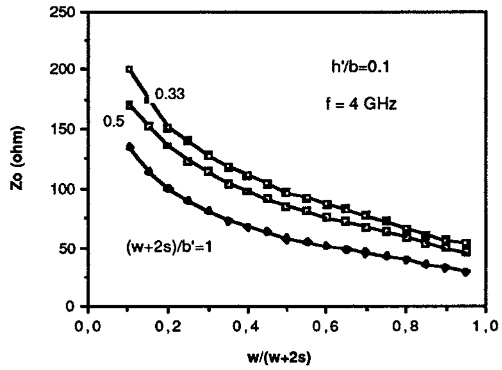

Fig. 6. Characteristic impedance of the NSCPW as a function of normalized strip width, for different (w+2s)/b'.

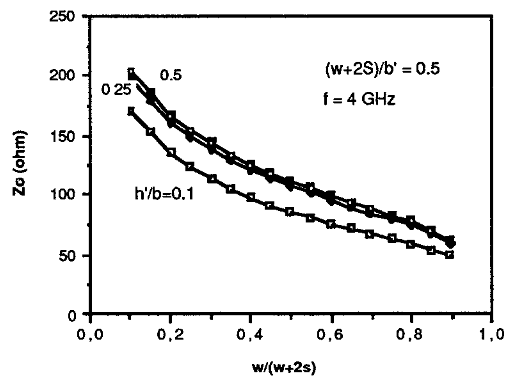

Fig. 7. Characteristic impedance of the NSCPW as a function of normalized strip width, for different h'/b.

Fig. 8 finally illustrates the behaviors of the cutoff frequencies of the first higher order modes of the NSCPW as functions of the shape ratio and for two values of the normalized height h'/b. The cutoff frequencies are normalized to that of the dominant mode of a rectangular waveguide with same widht b. f_{co} is the cutoff frequency of the lowest odd mode, while f_{ce} is the cutoff frequency of the lowest even mode (apart from the quasi TEM which is also even).

Contrary to the odd mode which is strongly affected by both strip width w and height h' of the lower half housing, the cutoff of the even mode is almost insensitive to both parameters. The cutoff frequency of the odd mode can be made higher than that of the even mode by reducing either h'/b or w/(w+2s), so maximizing the useful bandwidth of the line.

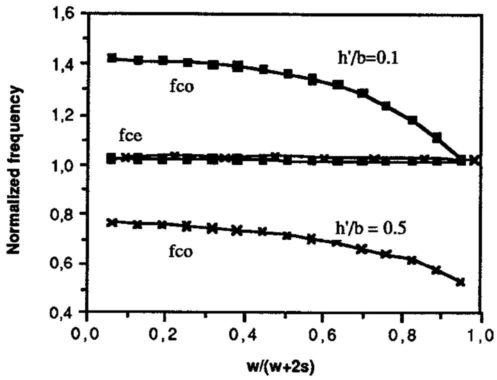

Fig. 8 Normalized cutoff frequencies of the first two higher order modes of NSCPW vs. normalized strip width

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