A Novel CPW DC-Blocking Topology with Improved Matching at W-Band

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***Abstract*—This letter presents the design of a new wide-band DC-blocking filter in coplanar waveguide (CPW) technology. Compared with the well-known coplanar open-end series stub, the matching of the new filter is theoretically shown to be improved from −10 to −20 dB over the whole W-band. The calculation is made by combining coupled line theory with a simple variational calculation of modal parameters. Measurements carried out in the W-band show very good agreement with the theory.**

I. Introduction

INTEREST in coplanar waveguide (CPW) passive elements at millimeter waves arose from the necessity to realize low-cost millimeter-wave integrated circuits. No ground plane indeed is needed in the case of CPW topology. For passive elements, efficient design rules do exist in microstrip technology but are still missing for the CPW one. Open- and short-end series stubs have been introduced by Houdart [1]. In the 1990's a full-wave space domain model has been developed for those structures [2], and much work has been devoted to efficiently modeling CPW discontinuities as lumped L and C elements for filter design. Several bandpass and bandstop filter applications have been investigated recently [3]. The present letter is devoted to another important element widely used in microstrip technology: the DC-blocking filter. Basically, it can be made out of a single cell of a bandpass filter, consisting of an open-end series (OES) stub [3]. For wide-band applications, however, the device must have a large bandwidth and low return loss in the passband. Such specifications are difficult to meet with the conventional OES CPW stub.

In this letter, we present a novel CPW topology for DC-blocking filters. Its design is based on a new model we developed for a section of coupled CPW lines. We associate to this structure a coupled line (CL) formalism, and we compute the CL modal parameters in a new way: coupling coefficients and propagation constants are expressed as explicit functions of the geometry and electrical parameters of the structure, so that no iterations are required. Hence, the mathematical and numerical complexity are reduced, and the synthesis is fast. Also, the partial modal characteristic impedances are computed in such a way that reciprocity is ensured. This efficient model is first validated on the conventional OES DC-blocking CPW stub. Next, we design a tapered CPW bandpass cell, by linearly varying the center spacing between the inner slots of the OES CPW stub. Due to the numerical efficiency of the model, subsections of parallel coupled lines can be cascaded at will to approximate the linear variation of the inner spacing. The tapered prototype exhibits a matching level greater than 20 dB in the W-band, which is confirmed by measurement of devices on GaAs.

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II. Description of the Model

We consider the transverse section in the plane AA' of the interdigitated straight capacitive cell in Fig. 1(a). Because of the symmetry of the structure, the problem reduces to the description of coupling between two slots of respective widths S_1 and S_2. Associating equivalent currents and voltages to an infinitesimal section of this coupled line area, they are described by the coupled mode solution for two lines [4]

$$V_1(z) = \sum_{k=c,\pi} (A_k e^{-\gamma_k z} + B_k e^{\gamma_k z}) \tag{1a}$$

$$V_2(z) = \sum_{k=c,\pi} K_k (A_k e^{-\gamma_k z} + B_k e^{\gamma_k z}) \tag{1b}$$

$$I_1(z) = \sum_{k=c,\pi} Y_{1,k} (A_k e^{-\gamma_k z} - B_k e^{\gamma_k z}) \tag{1c}$$

$$I_2(z) = \sum_{k=c,\pi} K_k Y_{2,k} (A_k e^{-\gamma_k z} - B_k e^{\gamma_k z}). \tag{1d}$$

Hence, the problem reduces to the computation of transmission line parameters for electric field configurations associated with each of the first two (dominant) electromagnetic modes supported by two slots of unequal widths in a ground plane lying on a substrate. Since the structure is of slot-type, the voltages V_1 and V_2 of the equivalent lines 1 and 2 are easily related to the x-component of the electric field existing across the two slotlines. For each mode c or π, the ratio of electric field magnitudes yields the coupling factor $K_{c \text{ or } \pi}$, while their z-dependence yields the propagation constant $\gamma_{c \text{ or } \pi}$. It has been shown [5] that the propagation constant of a multilayered planar line can be obtained by solving a second-order equation whose coefficients are proportional to square magnitudes of spectral expressions holding for the x-dependence of the trial electric field across the slots. The x-component of electric field associated to each line exists only in complementary slot areas. The trial spectral fields are thus a linear combination of known spectral basis functions, which hold for trial x-dependencies of electric field across each slot and are normalized such that their integral over the slot is equal to unity

$$\tilde{e}_x(k_x) = \tilde{e}_{x,\text{slot1}}(k_x) + K\tilde{e}_{x,\text{slot2}}(k_x). \tag{2}$$

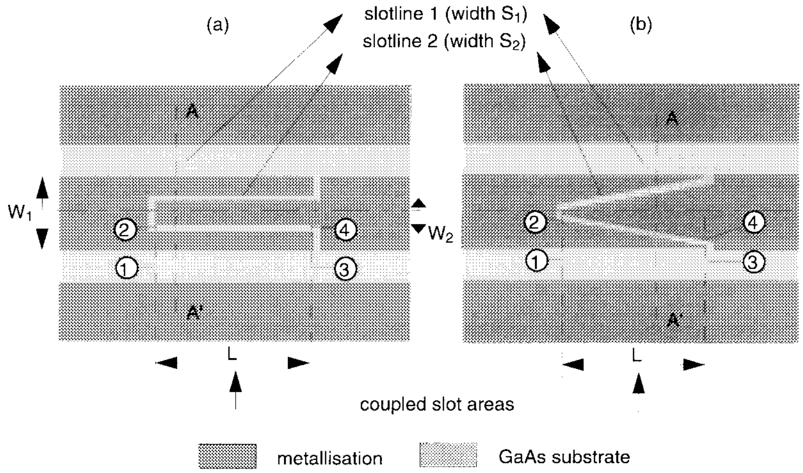

Fig. 1. Topologies of OES CPW stubs. (a) Conventional cell as described in [2] and [3]. (b) New tapered DC-blocking filter prototype. Circled numbers refer to port numbering in coupled slotline area.

Hence, it can easily be shown that the solution of [5, eq. (6)] is a ratio of square magnitudes of spectral expressions and may be rewritten as

$$\gamma = \frac{\tilde{N}}{\tilde{D}} = \frac{\tilde{N}_{11} + 2K\tilde{N}_{12} + K^2\tilde{N}_{22}}{\tilde{D}_{11} + 2K\tilde{D}_{12} + K^2\tilde{D}_{22}} \tag{3}$$

where $\tilde{N}, \tilde{D}$ are linear combinations of coefficients of [5, eq. (6)] which are proportional to the product $[\tilde{e}_x(k_x) \cdot \tilde{e}_x(k_x)^*]$ while $\tilde{N}_{mn}, \tilde{D}_{mn}$ are the fractions of $\tilde{N}, \tilde{D}$ which are proportional to $[\tilde{e}_{x,\mathrm{slot}m}(k_x) \cdot \tilde{e}_{x,\mathrm{slot}n}(k_x)^*]$. Numerators and denominators in (3) involve only known analytical expressions, exception made of the coupling factor K.

Taking advantage of the variational nature of (2), the Rayleigh–Ritz procedure imposes that the solution is extremum. Taking the first derivative of the right-hand side of (2) with respect to K, a second-order equation in K is obtained. Its two solutions yield the coupling coefficients K_c and K_π associated with each mode. Hence, reintroducing each solution in the explicit solution (2) yields the corresponding complex propagation constants γ_c and γ_π. This also completes the description of the spectral fields (2), and of the spectral coefficients of [5, eq. (6)].

We have also shown in [5] that those coefficients combined with the propagation constant provide a straightforward expression for the total complex power flowing into an semi-infinite section of the structure. Using it for each mode yields the total power transported by each mode, noted P_c and P_π. At plane $z = 0$, they are related to the circuit expressions (1) by

$$\begin{aligned} P_{c,\pi} &= \tfrac{1}{2}[V_1(0)I_1(0)^*]_{c,\pi} + \tfrac{1}{2}[V_2(0)I_2(0)^*]_{c,\pi} \\ &= (Y_{1,c,\pi} + |K_{c,\pi}|^2 Y_{2,c\pi})|A_{c,\pi}|^2. \end{aligned} \tag{4}$$

At this stage, we do not know the exact repartition of the power associated to a given mode between the two lines. In [3], partial modal impedances are deduced from a full-wave formalism associated to the spectral domain method. However, the physical relevance of such approach is not guaranteed in case of coplanar lines because we observed that using it leads to spurious nonreciprocities, as already obtained for coupled microstrips [6]. Hence we prefer to use an approach compatible with the reciprocal coupled-mode formalism (1). It can be indeed shown that the ratio of partial modal characteristic impedances is related to the product of the two coupling factors [4]

$$\frac{Y_{1,c}}{Y_{2,c}} = \frac{Y_{1,\pi}}{Y_{2,\pi}} = K_c \cdot K_\pi. \tag{5}$$

From (4) and (5) the four characteristic impedances of each mode on each line are obtained as a function of the modal powers P_c and P_π, computed according to [5, eq. (10)] for an unitary voltage on slot 1 $(A_c = A_\pi = 1)$. Our approach has the advantage to be fully analytical, hence very fast. The fields are described in the spectral domain, but the coupling factor and the propagation constants are obtained from simple analytical functions of those known fields. Also, the partial modal impedances are obtained in accordance with the coupled mode formalism (1), (5), which guarantees the reciprocity of the obtained formulation.

III. Design of New Prototype

For the new tapered version, the coupled-slot area is divided into N subsections of very small length, wherein the spacings between inner slots 2 are assumed to be constant. Hence, the modal parameters Y, γ, K are kept constant on the subsection. The 4-port chain matrix of each subsection is computed from (1) and the global 4-port is obtained by cascading the chain matrices of the subsections. Because of the explicit formalism (2), (3) the subdivision of the coupling area can be refined at will while keeping the computation time low. Ports of the global 4-ports are located in the planes $z = 0$ and $z = L$, at the ends of the coupling area (Fig. 1). Port 2 is loaded by an open-ended slotline. Port 4 is connected serially with port 3, and the resulting 2-port is obtained between accesses 1 and 4. The 2-port modeling the conventional OES stub is directly found by considering only one subsection.

IV. Validation of the Design

Fig. 2 compares the performances obtained at W-band for the two configurations depicted in Fig. 1. The dashed curves depict simulated S-parameters. Curves labeled "p" are

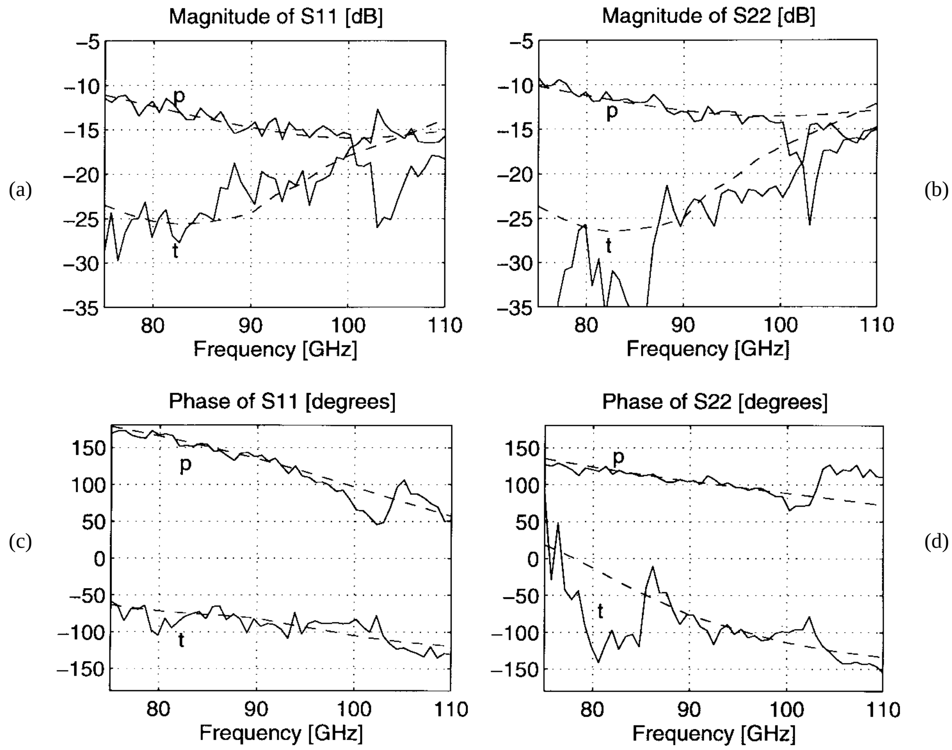

Fig. 2. Compared performances of parallel OES CPW stub in Fig. 1(a) (curves labeled p) and of new tapered topology in Fig. 1(b) (curves labeled t). (a) Magnitude of input reflection coefficient. (b) Magnitude of output reflection coefficient. (c) Phase of input reflection coefficient. (d) Phase of output reflection coefficient. Dashed lines: modeling; solid lines: measurement. Geometrical dimensions are $S_1 = 50\ \mu$m, $S_2 = 5\ \mu$m, $W_1 = 75\ \mu$m, $W_2 = 30\ \mu$m, $L = 300\ \mu$m, and thickness of GaAs wafer is equal to $350\ \mu$m.

obtained by applying (1), (5) to a single section of parallel coupled slots to simulate the behavior of the OES stub of Fig. 1(a). It is observed that the input and output reflection coefficients [Fig. 2(a), (b)] are both comprised between −15 and −10 dB in most part of the band. This prediction is validated by measurement (solid). Also, the model predicts the asymmetry of the device, in magnitude and in phase [Fig. 2(c), (d)]. Curves labeled "t" shows the corresponding results for the tapered prototype in Fig. 1(b). Ten sections of parallel coupled slots are cascaded to approximate the linear variation of the spacing between the inner slots. The model (dashed) predicts a significant improvement on the input and output reflection levels [Fig. 2(a), (b)]. Both reflection coefficients are better than −20 dB over a 25-GHz bandwidth. The predicted improvement of the matching is successfully confirmed by measurement (solid lines labeled "t"). It has to be mentioned that, for the two topologies, modeled and measured transmission coefficients (not shown here) agree very well.

The value of W_2 influences the bandwidth of the OES. At finite S_2 a small spacing W_2 induces a high characteristic impedance for the inner CPW formed by slots 2, hence a rapid variation of the series stub phase around resonance. Taking a moderate to large value for W_2 enlarges the bandwidth, but also the coupling between slots 2 and 1: the modal impedances of the CL section differ from that of outer CPW at access ports, and the reflection levels increase. Linearly varying W_2 confines the coupling effects into a small fraction of the device length only, while minimizing the characteristic impedance of inner CPW stub at its input [port 4 in Fig. 1(b)]. Also, the center frequency of parallel OES subsections varies with W_2 so that minimum reflection occurs at a different frequency in each subsection. Hence, the tapered CPW DC-block has improved input and output reflection levels over a wide band.

V. Conclusion

In this letter, we propose a new tapered CPW DC-blocking filter with improved matching at W-band (better than −20 dB). Its design is facilitated by using a simple explicit model we developed for coupled transmission lines parameters. The theoretical performances of the new tapered designed prototype are in good accordance with the measurements.

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