Compact CPW-Fed Comline Filter in Substrate Integrated Waveguide Technology

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Abstract—In this letter, the design and experimental results of compact low loss combline filters, based on the extension of the classical coaxial waveguide resonator to Substrate Integrated Waveguide (SIW) technology, are successfully demonstrated. A 3-pole 5% FBW Chebyshev filter has been designed, fabricated, and measured. The fabricated device shows an excellent agreement with simulated results. These structures keep the low-cost fabrication scheme of single-layer PCB processing, while requiring less than half the area compared to a conventional SIW design.

Index Terms—Microwave filter, compact filter design, combline resonator, substrate integrated waveguide.

I. INTRODUCTION

COMBLINE filters have been extensively used in metallic waveguide technology for implementing compact size filters. Although the insertion of metallic posts as perturbing elements of the TE_{101} mode in SIW circuits had been already explored in [1] for reconfiguration purposes, the authors have recently proposed that a combline resonator in SIW technology can be built by inserting an opened-shorted plated via hole inside the cavity [2].

In this letter, an X-band combline filter is successively designed, fabricated and measured in SIW technology. A model for the proposed resonator is considered, and a systematic procedure for the design of these miniaturized coupled resonator bandpass filters is presented. Measurements and simulation results are in good agreement, showing the potential advantages of this structure in terms of size, design flexibility and spurious rejection, without increasing insertion losses compared to its conventional SIW counterpart.

II. ANALYSIS AND DESIGN

A. TEM model and resonant frequency

The structure of the proposed combline SIW resonator is shown in Fig. 1. The inner via is short-circuited at the bottom metallization and open ended at the top. A capacitance is therefore established between a metallic disk, connected to the inner via, and the top metallization of the SIW cavity through the fringing fields across an annular gap etched on the metal. Narrow bandpass filters can be implemented by inserting inductive windows coupling adjacent resonators.

The susceptance $B(\omega)$ of the resonator can be expressed as

$$B(\omega) = \omega C_d - Y_0 \cot \beta h$$

being $C_d$ the capacitance between the disk and the top plane of the SIW cavity, $Y_0$ the characteristic admittance of the TEM mode short-circuited transmission line formed by the inner (circular) and outer (rectangular) conductors, $\beta = c/\sqrt{\epsilon_r}$ the phase constant of the coaxial line at frequency $\omega$, and $h$ the substrate thickness.

The inductive contribution comes from the TEM mode short-circuited resonator, that can be seen as a short piece of coaxial transmission line of length $h$ and characteristic admittance $Y_0$ embedded into the dielectric of permittivity $\epsilon_r$.

For a circular inner conductor of diameter $d$ and a square contour of side $w$, the characteristic admittance can be well approximated when $w >> d$ by [3]

$$Y_0 = \left[ \frac{60}{\sqrt{\epsilon_r}} \ln \left( \frac{1.079w}{d} \right) \right]^{-1}$$

Neglecting the thickness of the conductors, the capacitance established between the metal disk of radius $r$ and the SIW top metal can be computed from [4]

$$C_d = \frac{2\pi r \epsilon_0 (1 + \epsilon_r)}{\ln \left( 1 + \frac{r}{w} \right)} \int_0^\infty |J_0(\zeta r) - J_0(\zeta (r + s))| \frac{J_1(\zeta r)}{\zeta} d\zeta$$

where $s$ is the annular gap and $J_n$ the $n$th-order Bessel function of the first kind.

The resonant frequency can be obtained from the condition $B(\omega_0) = 0$. The field pattern of the first resonant modes of the
The first resonant modes of a combline cavity in SIW technology, and resonant frequency given by full-wave EM simulations (solid) and the TEM resonator model (dashed) as a function of the disk radius for different values of $\frac{d}{w}$ (right). Fixed resonator dimensions are $w = l = 8 \text{ mm}$, $h = 1.524 \text{ mm}$, $\epsilon_r = 3.55$, $s = 150 \mu\text{m}$. A better accuracy of the TEM model is obtained for higher values of $r$ (and therefore of $C_d$), where the contribution of the via post is minimized.

Combline resonators in SIW technology are shown in Fig. 2. As can be seen, the first spurious responses are those due to the perturbed TE$_{n0m}$ modes of the SIW cavity [5]. A comparison between the resonant frequency given by the TEM model of the resonator and the obtained using 3D full-wave simulations is also presented in Fig. 2, showing good agreement.

The input/output coupling to the resonator is realized through a modified coplanar waveguide (CPW) probe. The probe excites the TE$_{101}$ mode in the SIW cavity that couples to the TEM mode of the combline resonator. The probe is composed of a CPW to microstrip transition short-circuited to the SIW top plane. Two slots are added at the end of the microstrip to increase the coupling.

B. Filter design

The prototype of the combline bandpass filter using shunt resonators and frequency-invariant admittance inverters is shown in Fig. 3. Given a filter response with centre frequency $w_0 = \frac{1}{\omega_H \omega_L}$ and bandwidth $\Delta \omega$, being $\omega_H$ and $\omega_L$ the upper and lower cutoff frequencies, the loading capacitance $C_d$ and the admittance of the coaxial resonator $Y_0$ can be obtained by mapping the resonator susceptance and the admittance inverters of the bandpass filter with those of the low-pass prototype.

The former condition can be expressed as

\[ B(\omega) = \omega' \left( \frac{b \Delta \omega}{\Omega_c} \right) \]  

where $b$ is the resonator slope parameter, $\omega'$ and $\Omega_c = 1 \text{ rad/s}$ the angular frequency and cut-off frequency of the low-pass prototype respectively. The level of $b$ will impact the unloaded quality factor of the resonator, and must be a trade-off between low losses, compactness and feasibility of the synthesized values of $C_d$ and $Y_0$.

Now, substituting (1) into (4) for the corresponding passband edge frequencies ($\omega' = \pm 1 \rightarrow \omega = \{\omega_H, \omega_L\}$), we can obtain

\[ C_d = \left( \frac{b \Delta \omega}{\omega_0} \right) \frac{\cot \theta_H + \cot \theta_L}{\omega_H \cot \theta_L - \omega_L \cot \theta_H} \]  

\[ Y_0 = \left( \frac{b \Delta \omega}{\omega_0} \right) \frac{\omega_H + \omega_L}{\omega_H \cot \theta_L - \omega_L \cot \theta_H} \]

Finally, the admittance inverters can be computed from the low-pass prototype coefficients $g_0, g_1, \ldots, g_{n+1}$ using the well-known expressions [6]

\[ J_{0,1} = \sqrt{\frac{Y_A \left( b \Delta \omega \right)}{g_0 g_1}} \quad J_{n,n+1} = \sqrt{\frac{Y_A \left( b \Delta \omega \right)}{g_n g_{n+1}}} \]  

\[ J_{i,i+1} = \left( \frac{b \Delta \omega}{\omega_0} \right) \frac{1}{g_i g_{i+1}} \]

for $i = 1$ to $(n - 1)$, where $Y_A$ is the admittance of the input/output access ports.

Following this approach, a 3-pole Chebyshev filter centered at 9.8 GHz with 5% in-band ripple has been synthesized. A substrate thickness $h = 1.524 \text{ mm}$ and permittivity $\epsilon_r = 3.55$ have been considered. Thus, a loading capacitance $C_d = 0.2 \text{ pF}$ and resonator admittance $Y_0 = 8.22 \text{ mS}$ have been obtained for a susceptance slope parameter $b = 0.014 \text{ S}$. The filter values were $J_{01} = J_{24} = 0.0036$ and $J_{12} = J_{23} = 0.0007$ for an input/output impedance $Z_A = 65 \text{ Q}$. The synthesized response can be seen in Fig. 5.

C. Layout

Firstly, the resonators are dimensioned. The diameter of the inner via $d$ is chosen sufficiently small in order to reduce the size of the cavity for a given admittance level. However, this will be ultimately limited by the minimum via size of the fabrication process and its tolerances. The dimensions of a square SIW cavity with the synthesized admittance $Y_0$ can be obtained from (2). Next, given the required loading capacitance, the disk radius and annular gap can be obtained numerically from (3). Again, the minimum annular gap will also be limited by the fabrication process and the tolerances.

Once the resonators are designed, the input/output and inter-resonators couplings are dimensioned. The $Q_{ex}$ and $k_{i,j}$ coefficients are obtained from the synthesized $J_{i,i+1}$ as

\[ Q_{e,i} = \frac{b Y_A}{J_{0,1}^2}, \quad Q_{e,n} = \frac{b Y_A}{J_{n,n+1}^2}, \quad k_{i,i+1} = \frac{J_{i,i+1}}{b} \]

for $i = 1$ to $(n - 1)$.

Then, using full-wave EM simulations, the external quality factor can be estimated from the group delay of the reflected response of a singly-loaded cavity [7]. For a given CPW-to-microstrip transition, the length of the slots $l_{m,n}$ is modified in order to increase the coupling. The coupling coefficient can also be obtained from the reflected response of two coupled.

\[ Y_{01} \quad J_{01} \quad J_{24} \quad J_{12} \quad J_{23} \quad Y_{n+1} \]

\[ J_{n,n+1} = \sqrt{\frac{Y_A \left( b \Delta \omega \right)}{g_n g_{n+1}}} \]  

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TABLE I

| l₁, l₂ | 5.85 mm | d₁, d₂, d₃ | 250 μm | w₁, w₂, w₃ | 8 mm | r₁, r₂ | 0.8 mm | r₁, r₂ | 0.85 mm | s₁, s₂, s₃ | 150 μm | w₁, w₂, w₃ | 200 μm | d₃|SIW | 350 μm |
|--------|---------|------------|--------|-------------|-------|--------|--------|--------|--------|------------|--------|-------------|--------|----|

singly-loaded cavities. The coupling between both resonators is obtained as a function of the iris width. The $Q_e$ and $k_{i,j}$ curves for the designed filter can be seen in Fig. 4.

Finally, a fine-tuning procedure is performed using full-wave EM simulations. During this process, the characteristic admittance and loading capacitance of the resonators must be slightly modified due to the loading effect of the inverters. Both distributed circuit parameters are finely controlled through the modification of the disk radius and the SIW cavity lengths. The layout dimensions of the previously synthesized filter are shown in Table I. The EM simulated response of the filter compared to the synthesized model can be seen in Fig. 5.

III. EXPERIMENTAL RESULTS

The filter was fabricated in a 1.524 mm thick Rogers RO4003C substrate ($\varepsilon_r = 3.55, \tan \delta = 2.7 \cdot 10^{-3}$) using standard single-layer PCB processing technology. Filter size is $17.7 \times 8$ mm$^2$, while the equivalent conventional SIW filter would require about $36 \times 12$ mm$^2$ for square cavities using the dominant TE$_{101}$ mode. The device was measured using a network analyzer on a probe station with GSG pitch probes, and S-parameters were re-normalized to the port impedance of 50 Ω.

Excellent agreement with the measurements as can be seen in Fig. 6. Wideband response of the filter shows a spurious-free band of almost one octave with a rejection level better than 30 dB up to 17 GHz.

IV. CONCLUSIONS

The design, performance and manufacturability of novel combine filters in SIW technology have been demonstrated. The measured results show excellent agreement with the EM simulations when all involved physical effects are considered. The proposed topology presents important advantages in terms of size compactness, spurious rejection and design flexibility.

REFERENCES