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Compact Dual-Band Differential Power Splitter with Common-Mode Suppression and Filtering Capability based on Differential-Mode Composite Right/Left Handed (DM-CRLH) Transmission Lines

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***Abstract*— This paper presents a compact differential-mode (balanced) power splitter with filtering capability and dual-band functionality. This multifunctional device is based on differential-mode composite right/left handed (DM-CRLH) transmission lines specifically designed to achieve the dual-band behavior and to suppress the common-mode noise over a broad frequency band. The DM-CRLH lines are implemented by means of pairs of CRLH lines with series connected interdigital capacitors and strip inductors in the series branches, and mirrored step impedance resonators (SIRs) in the shunt branches. For the differential-mode, the SIRs are described by parallel resonant tanks, and the resulting equivalent circuit model is the canonical circuit of a CRLH line, which can be used for designing the dual-band balanced splitter/filter. However, the SIRs are opened under common-mode operation, introducing transmission zeros for that mode. By tailoring the position of these transmission zeros, the common-mode can be effectively rejected over a frequency band that covers the splitter operating frequencies. The fabricated dual-band balanced power splitter exhibits a measured matching level better than –11dB, and measured insertion losses (power splitting) lower than 4.1dB at the design frequencies (f_1=1.8GHz and f_2=2.4GHz). The common-mode suppression at f_1 and f_2 is better than 30dB.**

***Index Terms*– Power splitter, differential lines, composite right/left handed (CRLH) lines, dual-band.**

I. Introduction

In recent years, there has been an increasing interest for the design and fabrication of dual-band microwave components, including power splitters and dividers [1]-[9]. Among the considered approaches, compact designs have been achieved by using composite right/left handed (CRLH) transmission lines [3],[9]. By means of these artificial lines, quarter-wavelength transformers (typically required for the implementation of power splitters and dividers) functional at two arbitrary frequencies can be implemented [3],[10][11].

In the previous designs [1]-[11], the multiband splitters/dividers are single-ended. The purpose of this paper is to demonstrate the potential of differential-mode CRLH (DM-CRLH) lines for the design of differential (or balanced) power splitters functional at two frequencies, with intrinsic common-mode noise rejection, and filtering capability. This work is motivated by the increasing demand of balanced circuits, caused by their high immunity to environmental noise and electromagnetic interference (EMI). In such balanced circuits, a figure of merit is the common-mode rejection ratio (CMRR), defined as the ratio between the transmission coefficients for the differential and common modes at the frequencies of interest. Therefore, the design of a compact dual-band balanced power splitter able to efficiently suppress the common mode without the need of cascading additional common-mode filters is of the highest interest. As will be shown, the considered DM-CRLH lines are key components to satisfy all these requirements. To the author’s knowledge there are few designs of balanced power splitters. In [12], Xia *et al*. reported a single-band balanced power divider, whereas a dual-band design was proposed in [13] by the same group. This latter design exhibits good performance, including high common-mode rejection at the two operating frequencies. However, the device is based on dual-band T-shaped 90º and 180º phase shifting lines, being thus relatively large.

The main contribution/novelty of the paper is the use of mirrored stepped impedance resonators (SIRs) for the implementation of DM-CRLH transmission lines with inherent common-mode noise suppression, and their application to a dual-band differential-mode power splitter. For differential-mode excitation, the reported lines are described by the canonical circuit model of a CRLH line, avoiding parasitic elements and easing the design procedure.

II. DM-CRLH lines with intrinsic common-mode rejection based on stepped impedance resonators (SIR)

A typical topology of the proposed order-2 DM-CRLH transmission lines is depicted in Fig. 1. In a first order approximation, the interdigital capacitors are modeled by a capacitance C_s, and the narrow strips cascaded to them by an inductance L_s (notice that the ground plane has been windowed –etched– to enhance the per-unit length inductance of the strip and to minimize the effects of the parasitic capacitance, C_{par}, of the series branch). The shunt branches consist of a symmetric stepped impedance topology, that can

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be seen as a pair of mirrored stepped impedance resonators (SIRs) and can be described by a cascade of grounded capacitors and series inductors (like the circuit model of a stepped impedance low pass filter). Thus, the lumped element equivalent circuit model of the proposed DM-CRLH transmission line, as well as the models for differential and common mode excitation are depicted in Fig. 2. For the differential mode, the symmetry plane is virtually grounded (electric wall), and the shunt branch is simply described by an LC parallel resonant tank. Therefore, the circuit model for this mode is identical to the canonical circuit model (π-model) of an order-2 single-ended CRLH transmission line (the order of a single-ended CRLH line indicates the number of reactive elements of the series and shunt branches, but notice that this model is also the canonical circuit model of an order-3 bandpass filter). Conversely, the symmetry plane is an open-circuit (magnetic wall) for the common mode, and the shunt branch is modeled by the parallel combination of the capacitance $C = C_p + C_{\mathrm{par}}$ and the series resonator L_p-C_{zi} (with i = 1,2). Therefore, transmission zeros at frequencies given by

$$f_{Zi}^{cc} = \frac{1}{2\pi} \frac{1}{\sqrt{C_{zi} L_p}} \tag{1}$$

are expected for the common mode. Notice that by choosing $C_{z1} \neq C_{z2}$, the transmission zeros are located at different frequencies, and this enhances the rejection level and bandwidth for the common mode. Therefore, these DM-CRLH lines exhibit inherent common mode suppression.

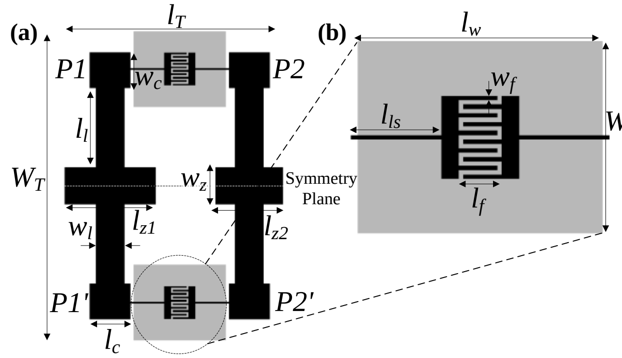

Fig. 1. Typical topology of the proposed order-2 DM-CRLH transmission line (a) and detail of the series branch (b). The upper metal level is indicated in black, whereas the ground plane windows are depicted in grey. Dimensions are: l_T = 18.5mm, W_T = 26.4mm, l_c = 3.4mm, w_c = 3mm, l_l = 6.7mm, w_l = 2.4mm, l_{z1} = 7.7mm, l_{z2} = 5.7mm, w_z = 3.1mm, l_{ls} = 2.8mm, w_f = 0.15mm, l_f = 1.3mm, l_w = 7.8mm and W_w = 6.5mm.

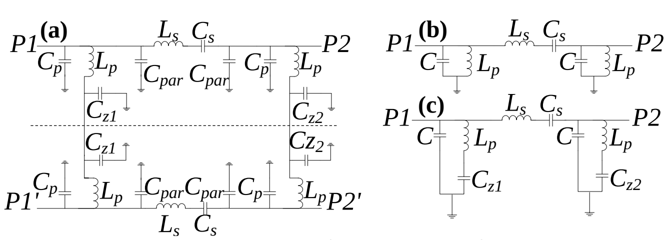

Fig. 2. Equivalent circuit model of the structure of Fig.1. (a) complete model; (b) differential-mode model; (c) common-mode model.

III. Device Design and Results

The differential dual-band power splitter proposed in this work is based on a dual-band 35.35 Ω impedance inverter. If the electrical length and the characteristic impedance for the odd mode, given by [3]

$$\cos\phi = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \tag{2}$$

$$Z_{oo}(\omega) = \sqrt{\frac{Z_s(\omega) Z_p(\omega)/2}{1 + \frac{Z_s(\omega)}{2Z_p(\omega)}}} \tag{3}$$

respectively, are set to the required values at the operating frequencies, that is, $\phi(\omega_1) = -90^{\circ}$, $\phi(\omega_2) = +90^{\circ}$, $Z_{oo}(\omega_1) = Z_{oo}(\omega_2) = 35.35\Omega$, the resulting four equations univocally determine the element values of the circuit of Fig. 2(b). In (2) and (3), Z_s and Z_p are the impedances of the series and shunt branches of the differential-mode circuit model. By considering the lower and upper operating frequencies to be given by $f_1 = \omega_1/2\pi$ = 1.8GHz and $f_2 = \omega_2/2\pi$ = 2.4GHz, the element values are found to be: C_s = 0.625pF, L_s = 9.376nH, C = 7.503pF, and L_p = 0.781nH. Notice that L_s is relatively large; therefore, slots are etched in the ground plane (see Fig. 1) to enhance the strip inductance. The common mode transmission zero frequencies are set to f_{z1}^{cc} = 1.8GHz and to f_{z2}^{cc} = 2.4GHz, in order to obtain a significant common-mode noise rejection within a wide band covering the operating frequencies. From expression (1), the patch capacitances of the central metallic regions of the device are found to be C_{z1} = 10pF and C_{z2} = 5.5pF.

For layout synthesis, the dimensions of the different semi-lumped elements (interdigital capacitors, straight line inductors and patch capacitors) have been first estimated from available expressions [14]. Then, the reactances of these elements have been inferred from electromagnetic simulation and the final dimensions have been adjusted by curve fitting the reactances to those given by the elements values. The fact that the parasitic capacitances of the series branch, C_{par}, are absorbed by the patch capacitances C_p is one of the reasons for choosing the π-circuit topology. The other important reason is that with this topology two common-mode transmission zeros can be independently adjusted by tailoring the dimensions of the patches present at the central positions. The synthesized differential dual-band common-mode suppressed impedance inverter is the one depicted in Fig. 1 (the *Rogers RO3010* substrate with dielectric constant ε_r=10.2 and thickness h=254μm has been considered). The simulated differential and common-mode responses, considering 35.35Ω reference impedances for both modes are depicted in Fig. 3. Good matching level and the required phase shift at the design frequencies is observed. Moreover, strong attenuation for the common-mode results by virtue of the transmission zeros for that mode.

After cascading a 50Ω differential access line at the input port and two 50Ω differential access lines at the output port of the inverter (like in a single-ended power splitter implemented from a 35.35Ω impedance inverter), the differential dual-band power splitter has been fabricated by means of a *LPKF-H100* drilling machine. The photograph of the device is depicted in Fig. 4 (the differential ports are indicated) and the simulated and measured frequency responses are shown in Fig. 5. The agreement between circuit, electromagnetic simulation and measurement is reasonable (discrepancies in measurement are attributed to inaccuracies of the drilling process, which may be significant when narrow and malleable substrates are used). The measured insertion and return losses at the design frequencies are $IL(f_1)$=3.2dB, $RL(f_1)$=11.3 dB, $IL(f_2)$=4.1dB, and $RL(f_2)$=11dB (notice that the ideal insertion loss level for a power splitter is IL=3dB), the device exhibits a common-mode rejection level better than 26dB up to 3GHz, with measured common mode rejection ratios at the operating frequencies as high as CMRR(f_1)=31.6dB, and CMRR(f_2) =34.3dB. Finally, the filtering action of the splitter, due to the intrinsic bandpass filter functionality of CRLH transmission lines, is clearly visible in Fig. 5. Notice that the differential mode insertion loss is close to the ideal value (3dB) between f_1 and f_2, and the input matching between these frequency is also good. However, rather than a broadband power splitter, the proposed device must be considered a dual-band structure since it has been designed by forcing its functionality at the design frequencies. The fact that these frequencies are not so far leads to reasonable power splitting and input matching between these frequencies as well.

Device dimensions are $0.34\lambda_g \times 0.45\lambda_g$, where λ_g is the guided wavelength at f_1. By narrowing the inductive strips of the shunt branches, the lateral dimensions of the splitter can be further reduced. In the designed device, the limit has been dictated by the minimum distance between access lines for connectors soldering. Nevertheless, the proposed device is by far smaller than the one reported in [13].

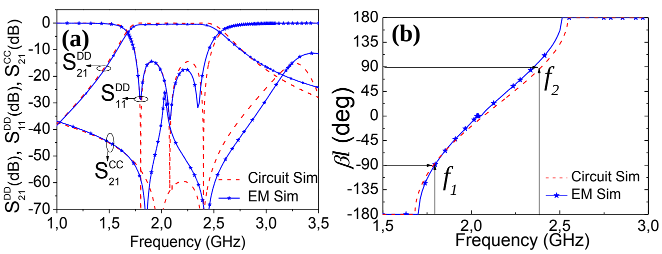

Fig. 3. Magnitude (a) and phase response (b) of the designed differential-mode dual-band impedance inverter depicted in Fig. 1. For the common mode, only S_{21} is shown.

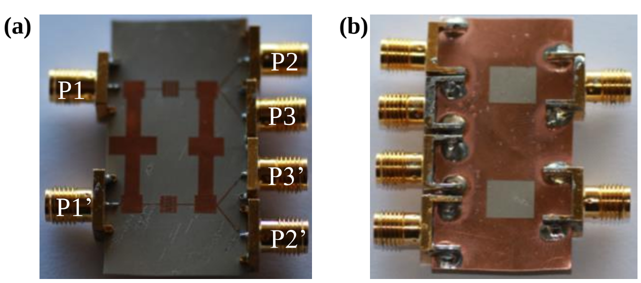

Fig. 4. Photograph of the fabricated differential-mode dual-band power splitter.(a) Top view; (b) bottom view. The differential ports are indicated.

IV. Conclusion

It has been demonstrated that compact differential-mode power splitters with dual-band functionality and intrinsic common-mode rejection can be implemented by means of differential-mode CRLH transmission lines based on stepped impedance resonators. The proposed semi-lumped based solution gives small device size, is fully compatible with planar technology, and provides high common-mode rejection at the frequencies of interest. Moreover, the proposed power splitter acts also as a filter for the differential mode. This multi-functionality represents an added value of the approach, of interest in applications where size is a critical aspect.

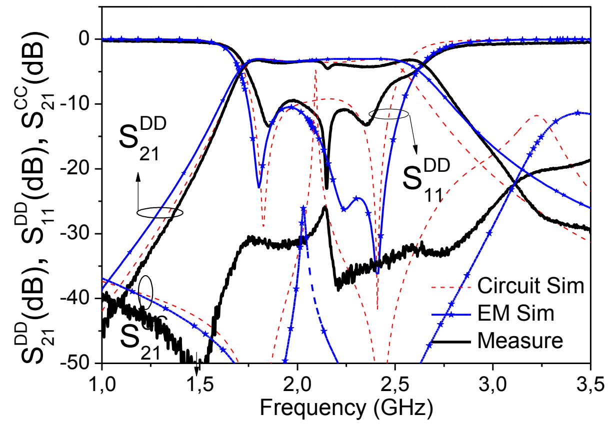

Fig. 5. Frequency response of the power splitter depicted in Fig. 4.The S-Parameters at port 3 (S_{31}^{DD} and S_{31}^{CC}) are not shown since they are roughly the same than those obtained at port 2.

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