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A Compact Microstrip Crossover Using NRI-TL Metamaterial Lines

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Abstract— A practical implementation of a miniaturized crossover is presented using negative-refractive-index transmission-line (NRI-TL) metamaterial lines. NRI-TL metamaterial theory has demonstrated that by loading a conventional transmission line with lumped LC elements, a large phase shift per-unit-length can be achieved. Employing this technique, the conventional microstrip transmission lines in a three-section branch-line coupler crossover are replaced with 90° NRI-TL lines, and thus a compact NRI-TL crossover with an area 47 times smaller than the area of its conventional counterpart is demonstrated. Measured results demonstrate that the proposed NRI-TL metamaterial crossover exhibits a -10 dB $|S_{11}|$ bandwidth of 33% and a |3 dB| $|S_{31}|$ bandwidth of 26% around 1 GHz.

Index Terms— Negative-refractive-index transmission-line (NRI-TL), metamaterial, crossover.

I. INTRODUCTION

Planar microwave circuits frequently use microwave transmission lines that intersect with each other. A crossover has been a well-known device suited for such intersection points after its first successful implementation by Wight [1]. Since then, a number of crossover configurations have been presented to primarily increase the bandwidth of the device and to decrease the overall size [2-4]. A broadband crossover has been presented using single and multi-section branch-line coupler configurations [5, 6]. Referring to Fig. 1, each branch in this configuration is set to incur a 90° phase shift for the successful transmission of a signal from Port 1 (P1) to Port 3 (P3), and from Port 2 (P2) to Port 4 (P4). A $\lambda/4$ section of conventional transmission line is required in each branch of the crossover, which results in a physically large device. For example, at 1 GHz, the size of a three-section crossover [6] is $55 \times 165 \text{ mm}^2$, which is too large for most applications in modern microwave

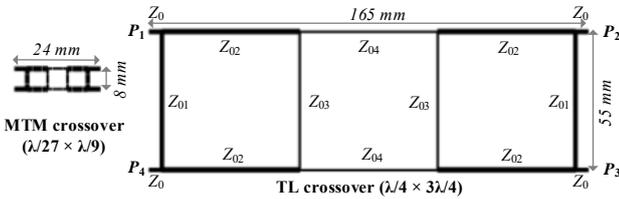


Fig. 1 Footprint comparison between a microwave crossover using conventional transmission lines (right) and a compact NRI-TL metamaterial (MTM) microwave crossover (left).

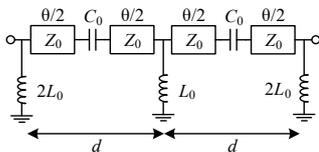


Fig. 2 Two-stage NRI-TL metamaterial line formed by cascading two Π unit cells for the implementation of each branch of the crossover.

electronics.

II. NRI-TL METAMATERIAL CROSSOVER

In this work, a practical implementation of a miniaturized three-section branch-line crossover using NRI-TL metamaterial lines is proposed, which is implemented in a simple and realizable fashion. Each branch of the crossover is realized using a two-stage NRI-TL metamaterial line of Π unit cells [7, 8]. This approach also avoids the unrealizable overlapping of three capacitors at the junctions of each branch while using a T-unit cell (for example in [9]). A single NRI-TL metamaterial Π unit cell consists of a host transmission line with characteristic impedance Z_0 , series inductance per unit length L , shunt capacitance per unit length C , and length d , which is loaded with a lumped-element series capacitor C_0 and two shunt inductors $2L_0$. A cascade of two metamaterial Π unit cells to form a two-stage NRI-TL metamaterial line is presented in Fig. 2. The conventional host transmission line realizes a negative phase response, while the series capacitive and shunt inductive loading forms a backward-wave (BW) line with a positive phase response. Under effective medium conditions, the phase shift for N NRI-TL metamaterial unit

Table 1: Host transmission line widths and loading element values

Line impedance	Line width	Ideal		Available	
		L_0	C_0	L_0	C_0
$Z_0 = Z_{01} = 50 \Omega$	1.54 mm	9.32 nH	3.73 pF	9.42 nH	3.90 pF
$Z_{02} = 45 \Omega$	1.79 mm	8.39 nH	4.14 pF	8.50 nH	4.30 pF
$Z_{03} = Z_{04} = 81 \Omega$	0.69 mm	15.11 nH	2.30 pF	14.6 nH	2.40 pF

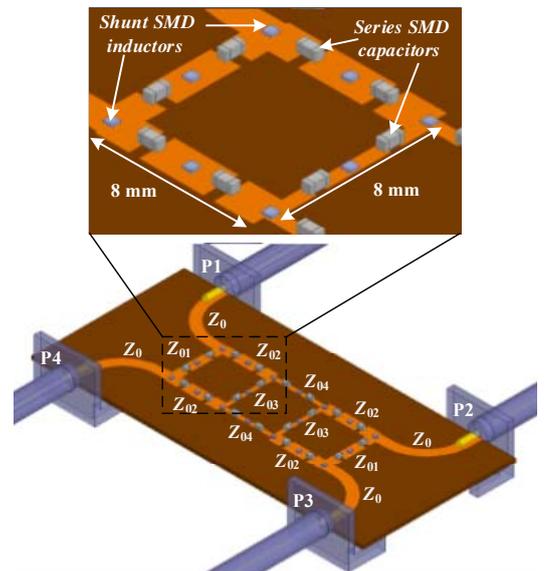


Fig. 3 Perspective and zoomed-in view of the full-wave EM simulation prototype of the three-section NRI-TL metamaterial crossover.

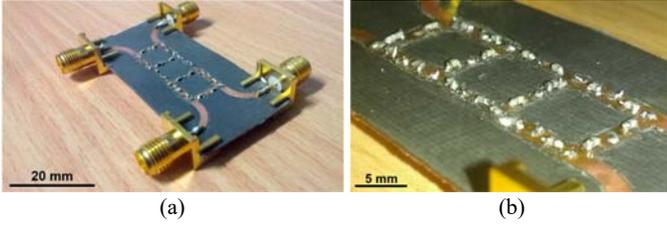


Fig. 4 Fabricated prototype of the compact NRI-TL metamaterial microwave crossover (a) the device (b) zoomed-in view of the soldered lumped-element components.

cells is given by [7]:

$$\Phi_{MTM} \approx N(\phi_{MTM}) = N\left(-\omega\sqrt{LC}d + \frac{1}{\omega\sqrt{L_0C_0}}\right) \quad (1)$$

This is valid under the impedance matching condition:

$$Z_0 = \sqrt{\frac{L}{C}} = \sqrt{\frac{L_0}{C_0}} \quad (2)$$

At the initial design phase, the impedance of each branch was calculated based on the three stage branch line coupler presented in [6]. The calculated characteristic impedance values of the transmission lines were $Z_0 = Z_{01} = 50 \Omega$, $Z_{02} = 45 \Omega$, and $Z_{03} = Z_{04} = 81 \Omega$. Each branch of the original crossover consisted of a 90° microstrip transmission line, which was replaced with a compact two-stage NRI-TL metamaterial line, where each unit cell had an electrical length of $\theta = 5^\circ$. A total of 10 such branches formed the three-stage metamaterial crossover. The ideal lumped-element component values for each unit cell were calculated using the guidelines presented in [7] and are listed in Table 1. Note that the calculated L_0 and C_0 values simultaneously need to satisfy the conditions of equations (1) and (2). In total, 20 surface-mount device (SMD) capacitors and 18 SMD inductors were used to implement the crossover, as can be seen in Fig. 3.

The initial crossover was designed, simulated and evaluated in the Keysight – Advanced Design System (ADS) simulator to operate at 1 GHz on a Rogers RT/duroid 5880 substrate ($\epsilon_r = 2.2$, $\tan\delta = 0.0009$, $h = 0.787$ mm). At first, ideal lumped-element components (see Table 1) were used in each branch line to ensure the theoretical device operation. In the next stage, the ideal components were replaced with the closest available surface-mount component values from the Murata chip capacitor (GJM03 series) and Coilcraft chip inductor (0402CS series) component libraries, and the realistic widths of the transmission lines on the Rogers 5880 substrate were calculated (see Table 1). In the next and final stage, a full-wave EM simulator was used to evaluate the device performance in a realistic environment. Fig. 3 shows the geometrical configuration and component loading of the proposed device. It can be observed that due to the small size of the device, SMA connectors cannot be mounted directly to the ports of the crossover, therefore 50Ω curved feed lines were added, and the SMA connectors were attached to these. The SMD capacitors were mounted in series along with the microstrip line trace, whereas the SMD inductors were placed vertically through the substrate. When assembling each of the individually-designed metamaterial branches to form the crossover, a shunt inductor is present at both ends of each branch, however, at each junction it is not practically feasible to place adjacently three separate inductors corresponding to separate branches in a very small area. Therefore, an equivalent parallel inductance was evaluated ($L_{01} \parallel L_{02}$ for a junction of two branches and $L_{01} \parallel L_{02} \parallel L_{03}$ for a junction of three branches). The closest possible component value in the Coilcraft chip inductor (0402CS series) component library was used to implement this new

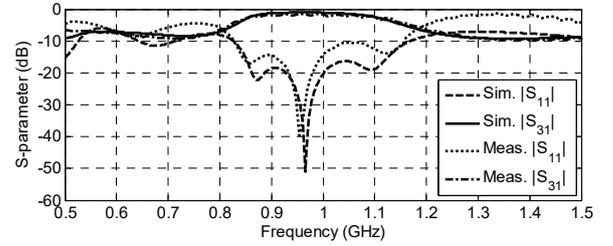


Fig. 5 Simulated and measured S-parameter responses for the compact NRI-TL metamaterial microwave crossover.

equivalent inductance value. Although the crossover was initially designed at a design frequency of 1 GHz, due to the limited and discrete range of component values available in the Murata and Coilcraft SMD component libraries, the final design was optimized at a slightly shifted design frequency of $f_0 = 0.965$ GHz.

III. FABRICATION AND MEASUREMENTS

The microstrip conductor and gap patterns shown of Fig. 3 were implemented on the Rogers 5880 substrate using an LPKF ProtoMat H100 milling machine. Holes of 0.4 mm diameter were drilled for the vertical placement of the SMD inductors. Then, the lumped components were soldered onto the metallic traces. Fig. 4a shows the fabricated prototype of the compact metamaterial crossover while Fig. 4b shows the soldered SMD components. To measure its performance, the two ports of an Agilent E8363B vector network analyser were connected to Port 1 and Port 3, while two 50Ω broadband loads were connected to the remaining two ports (Ports 2 and 4) of the device. The reflection coefficient at Port 1 remains below -10 dB from 0.823 GHz to 1.155 GHz, as can be observed in Fig. 5, which amounts to a 33.57% fractional bandwidth. At the design frequency of 0.965 GHz, the simulated and measured $|S_{11}|$ of the device is -50.93 dB and -27.34 dB, respectively, indicating good impedance matching. The simulated and measured transmission coefficients $|S_{31}|$ at the same frequency are -1.026 dB and -1.55 dB, respectively. A 3 dB transmission bandwidth can be defined as the bandwidth over which $|S_{31}|$ does not fluctuate beyond 3 dB from the value at $f_0 = 0.965$ GHz. For this device, the simulated and measured 3 dB bandwidths were 27% (from 0.85 GHz to 1.11 GHz), and 26% (from 0.85 GHz to 1.10 GHz), respectively. Similar results were recorded when the crossover was connected in the opposite direction, i.e. for S_{33} and S_{13} , indicating that the device is reciprocal. Also, analogous results were observed in the other crossover path, i.e. for transmission from P2 to P4 and in the opposite direction from P4 to P2. Since these results are almost identical to the ones presented in Fig. 5, they are omitted for brevity. The small discrepancies between the simulated and measured S-parameters can be attributed to the $\pm 5\%$ component tolerances and the soldering and component placement imperfections that can be observed in Fig. 4b. These discrepancies can be reduced by efficient component placement and accurate soldering while in mass production. Nevertheless, the good reflection and transmission responses of the crossover at the design frequency indicate that the proposed device is well suited for use in next-generation microwave systems.

IV. CONCLUSION AND FUTURE WORK

A simple and practical implementation of a compact NRI-TL metamaterial microwave crossover is presented in this work. The overall footprint has been reduced significantly by a factor of 47, while the signal reflection and transmission characteristics of the device are maintained at a good level. The design frequency of the proposed device around 1 GHz has been selected to enable rapid

prototyping using commercially-available SMD components, however, the same design procedure can be used to design NRI-TL metamaterial crossovers at higher frequencies. Future work would involve the replacement of SMD components with inter-digital capacitors and printed inductors, in order to further decrease the cost and losses of the device. RF front end microwave and mm-wave circuit boards hosting the feeding network of massive- multi-input multi-output (MIMO) transceiver antenna arrays (128 or 256 elements) can be considered as the most important application of this work.

IV. ACKNOWLEDGMENTS

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