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THE DESIGN OF A NOVEL COMPACT CRLH BANDPASS FILTER WITH HARMONICS SUPPRESSION

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Abstract—This paper presents the design procedure and implementation of an innovative compact bandpass filter in UHF band based on the Composite Right/Left Handed Transmission Line (CRLH-TL) by using parasitic effects of these structures. The CRLH-TL is utilized in order to minimize the overall size of component and reject higher ordered harmonics. The metamaterial transmission line is about one-eighth of a wavelength long and acts as both resonator and inverter which provide capacitive coupling between neighboring sections. To show the validity of design procedure, two prototype bandpass filters were fabricated and tested. We have shown that very good agreement exists between simulation results and those obtained by measurement.

1. INTRODUCTION

RF circuits and components play an essential role in mobile communication and broadcasting, thus miniaturization and integration of such components are considered as important requirement that can be accomplished by using of evanescent mode waveguide filters or meander lines microstripe filters. In the design of such filters we usually deal with half wavelength resonators. Recently metamaterial structures have found vast applications for improving the component performance and also minimizing their overall size [1, 2, 4].

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CRLH transmission lines have been used for constructing various filters in several papers. These CRLH structures are artificial TLs constructed of cascaded unit cells, composed of capacitors and inductors such as metal-insulator-metal (MIM) capacitors, interdigital capacitors, meander-line inductor, stub inductors, spiral inductors, etc. In [5], a UWB filter is proposed with the use of Metal-insulator-metal (MIM) capacitors and stub inductors. Two direct coupled bandpass filters are designed and attractive features of a class of filters using series-connected inductance line interconnected by two asymmetrical interdigital capacitors structures are observed theoretically and experimentally the same as size reduction and suppression of harmonic resonances in [6]. A miniaturized dual-band bandpass filter suggested in [7] is based on the subwavelength resonators structures that are realized through the zeroth order resonance and inverter structures which control the coupling between neighboring zeroth order resonator with use of symmetrical type Interdigital line and short-circuited stub.

This paper presents a novel bandpass filter based on the Composite right/left-handed metamaterial, where CRLH transmission line structure is implemented in the form of interdigital capacitors and stub short-circuited. However, this unit cell is used to design CRLH TLs in [3], but, we use parasitic effects of these distributed elements to design a novel compact bandpass. The proposed filter provides high level of miniaturization and efficient harmonic rejection ability in UHF band.

The principle of the CRLH TLs is described in Section 2. Proposed topology and design procedure of this novel bandpass filters are presented in Section 3 and Section 4, respectively. Finally, to validate the design formula, two sample filters at 610 MHz and 2300 MHz will be designed as examples. In Section 5, it will be shown that the simulated and measured performances are good in agreement.

2. COMPOSITE RIGHT/LEFT-HANDED MTMS

In the last decade, planar filters based on LH-TL, ring resonator structure and interdigital capacitor received an increasing attention because of their unique properties and excellent performances [8]. However, the LH structures presented originally were impractical for microwave application because of their very lossy and narrow bandwidth characteristics [3]. This was the motivation for introducing the term “composite right/left-handed” (CRLH). The equivalent circuit model for a loss-less CRLH TL is shown in Fig. 1(a) and The CRLH dispersion diagram is plotted in Fig. 1(b). These figures’

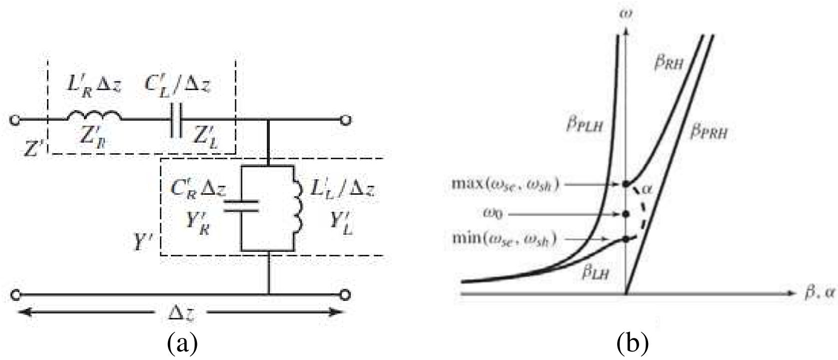

Figure 1. (a) the equivalent circuit model for a loss — less CRLH TL. (b) The CRLH dispersion diagram.

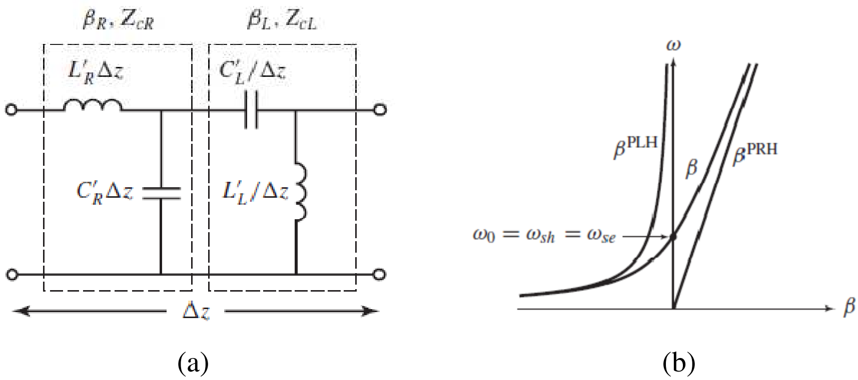

Figure 2. (a) The equivalent circuit model for a balanced loss — less CRLH TL. (b) The balanced CRLH dispersion diagram.

parameters are described in [3].

The CRLH TL exhibits interesting properties in the particular case where the series and shunt resonant frequencies are equal. This case is called the balanced case. The Simplified equivalent circuit model and dispersion diagram for the unit cell of an ideal CRLH TL under the condition of balanced resonances are shown in Fig. 2(a) and Fig. 2(b), respectively. In the transition frequency ω_0, the propagation constant and phase shift along a line are zero. Below the transition frequency ω_0, a CRLH transmission line behaves as a left-handed transmission line and above this frequency that is basically a conventional right-handed line. As a consequence of this combined behavior, the phase response is not linear with respect to the frequency. In addition, resonance is independent of the length of the structure but it is only dependent on the reactive loadings. Because of particular property of CRLH TLs, microwave filters using these structures have a potential for effective volume minimization [3].

3. PROPOSED TOPOLOGY

The topology of the proposed band-pass filter is shown in Fig. 3 [9, 10]. As shown, the band-pass filter uses admittance inverters and shunt-type resonator. The admittance inverters and shunt-type resonators are represented in the layout by interdigital capacitors and stub short-circuited, respectively.

Generalized model of interdigital capacitor and its equivalent circuit are shown in Fig. 4(a) and Fig. 4(b), respectively. For the

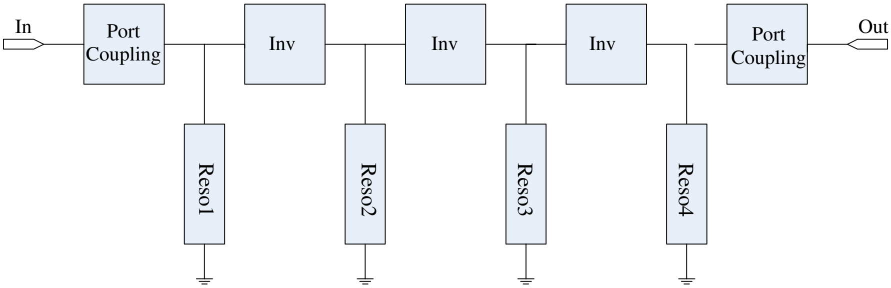

Figure 3. Circuit expression of the proposed band-pass filter.

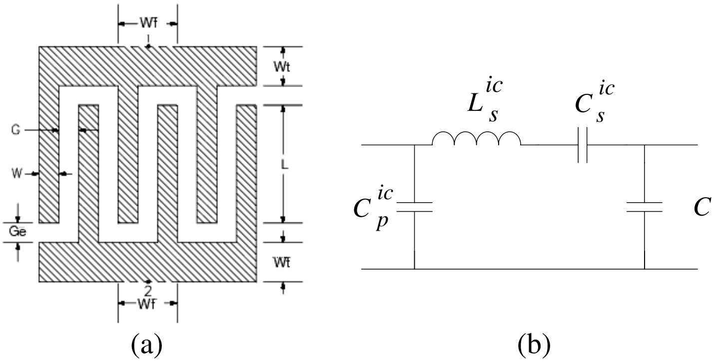

Figure 4. (a) Generalized model of interdigital capacitor. (b) The equivalent circuit model of interdigital capacitor.

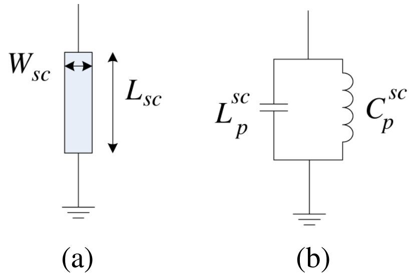

Figure 5. (a) Generalized model of stub short-circuited. (b) The equivalent circuit model of stub short-circuited.

purpose of calculation of the circuit model's parameters L_s^{ic}, C_s^{ic} and C_p^{ic}, firstly, the scattering parameters of the interdigital capacitor are determined by either full-wave simulation or measurement [5]. The S parameters of the interdigital capacitor are then converted into admittance or Y parameters and impedance or Z parameters, respectively, using standard conversion formulas. Finally, the parameters of Interdigital capacitor equivalent circuit are obtained [3]. The inductance L_s^{ic} is very small and can be neglected. In this paper, the scattering parameters of the interdigital capacitor are determined by model simulation software ADS2008.

Generalized model of stub short-circuited and equivalent circuit for Stub short-circuited are depicted in Fig. 5(a) and Fig. 5(b), respectively. The admittance of stub short-circuited and its derivative in relation to ω must be equal to the admittance of equivalent circuit and its derivative to ω, yields a linear system of two equations with the two unknowns L_r^{sc} and C_r^{sc}, that are easily which found to be:

$$L_r^{sc} = \frac{2}{\omega_0 Y_0[\cot\theta + \theta(1+\cot^2\theta)]} \tag{1}$$

$$C_r^{sc} = \frac{Y_0}{2\omega_0}\left[\theta\left(1+\cot^2\theta\right) - \cot\theta\right] \tag{2}$$

where θ and Y_0 represent electric length and admittance of stub, respectively. Using interdigital and stub short-circuited, a bandpass miniaturized filter can be established. In next chapter design procedure is presented.

4. DESIGN PROCEDURE

Figure 6 shows a layout of the new metamaterial bandpass filter. The equivalent circuit of this filter with replacing equivalent circuit of interdigital capacitors and stub short-circuited is shown as Fig. 7. An admittance inverting circuit is depicted in Fig. 8. The design procedure can be formulated by using the admittance inverting circuit in Fig. 7 leading to the simplified equivalent circuit as Fig. 9.

In Fig. 9, as shown the negative shunt capacitance of the J-inverters must be subtracted from the positive resonator capacitance to give the net shunt capacitance actually inserted in the circuit. Moreover, the capacitances C_{rj} are the effective capacitances for determining the resonant frequency and susceptance slope parameters of the resonators [11].

Inverting admittance is not satisfied in the interior of the filter. Let us suppose that $C_{01} = C_{p1}$, then The imaginary part of the admittance is looking from Resonator 1 in Fig. 7 out toward C_{01} and $G_A \| C_{p1}$

Figure 6. Layout of the new metamaterial bandpass filter.

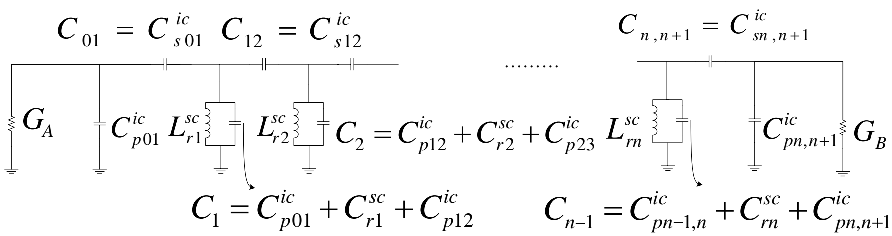

Figure 7. Equivalent circuit of proposed filter with replacing equivalent circuit of interdigital capacitors and stub short-circuited.

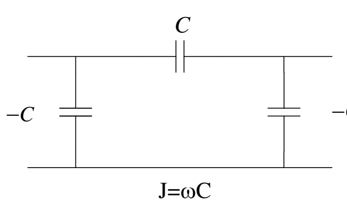

Figure 8. Schematic of the admittance inverter to be used with shunt-type resonators

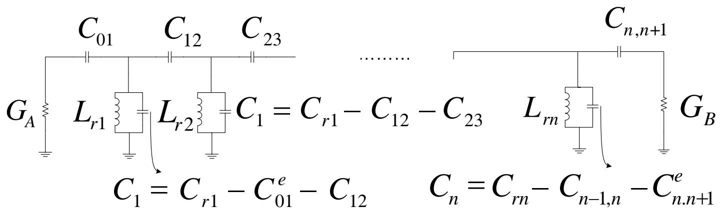

Figure 9. Equivalent circuit of proposed filter by using admittance inverting circuits.

in series, can be dealt with satisfactorily by replacing it by a shunt capacitive of the same size which leads to Eq. (3). Since C^e_{01} effectively increases the shunt capacitance of Resonator 1, this amount should be subtracted from C_{r1} as indicated in Fig. 9 when computing the net shunt capacitance to be used in constructing resonator 1. Of course, the same reasoning applies for design of the $C^e_{n,n+1}$ coupling at the other end of the filter which is given by Eq. (4) [11].

$$C^e_{01} = \frac{B_{01}G_A^2 + 2B_{01}^3}{\omega_0\left(G_A^2 + 4B_{01}^2\right)} \tag{3}$$

$$C^e_{n,n+1} = \frac{B_{n,n+1}G_B^2 + 2B_{n,n+1}^2}{\omega_0(G_B^2 + 4B_{n,n+1}^2)} \tag{4}$$

where B_{01} and $B_{n,n+1}$ are defined as $B_{01} = \omega_0 C_{01}$ and $B_{n,n+1} = \omega_0 C_{n,n+1}$. Design formulate for the design of a miniaturized band pass filter may be the following:

1. First choose Y_0 and the physical length of the shorted stub ($\lambda/8$) and then parameters L_{rj} and C_{rj} are determined by the following relations:

$$L_{rj} = \frac{2}{\omega_0 Y_0[1 + \frac{\pi}{2}]} \quad (\text{for } j = 1 \text{ to } n) \tag{5}$$

$$C_{rj} = \frac{1}{L_{rj}\omega_0^2} \quad (\text{for } j = 1 \text{ to } n) \tag{6}$$

2. We can now determine $J_{i,j+1}$ from C_{rj} by the following relations:

$$J_{01} = \sqrt{\frac{G_A\omega_0 C_{r1} w}{g_0 g_1 \omega_1'}} \tag{7}$$

$$J_{j,j+1} = \frac{w\omega_0}{\omega_1'}\sqrt{\frac{C_{rj}C_{rj+1}}{g_j g_{j+1}}} \quad \text{for } j = 1 \text{ to } n-1 \tag{8}$$

$$J_{n,n+1} = \sqrt{\frac{G_B \omega_0 C_{rn} w}{g_n g_{n+1}\omega'}} \tag{9}$$

where in these formulas, the coefficients g_j are the Chebyshev coefficients of the equivalent low-pass filter prototype and define the bandwidth ripple. The parameter ω_1' is the cut-off frequency of the low-pass prototype and G_A and G_B are the terminating conductance's of the circuit. Finally, w is defined as the fractional bandwidth.

3. In the next step, series (interdigital) capacitors are calculated by:

$$C_{01} = \frac{J_{01}}{\omega_0\sqrt{1-\frac{4J_{01}^2}{G_A^2}}} \tag{10}$$

$$C_{j,j+1} = \frac{J_{j,j+1}}{\omega_0} \quad j = 1 \text{ to } n-1 \tag{11}$$

$$C_{n,n+1} = \frac{J_{n,n+1}}{1-\frac{4J_{n,n+1}^2}{G_B^2}} \tag{12}$$

4. Then C_{01}^e and $C_{n,n+1}^e$ can be calculated by Eq. (3) and Eq. (4).
5. Shunt capacitors, C_j ($j = 1$ to n) can be determined.
6. When C_j ($j = 1$ to n) are calculated, the parameters C_s^{ic} and C_p^{ic} of any interdigital capacitor can easily be determined. Finally, after this parameter extraction procedure, we can obtain physical dimension of interdigital capacitor by simulation software or other ways.

The above formulae is validate for even order filter. The difference between even order and odd order filters is located only in asymmetric shunt capacitance of interdigital equivalent circuit near the central stub short-circuited, which is presented in Fig. 10.

5. SIMULATION AND EXCREMENTAL RESULTS

In order to validate the design principles stated previously, two new type bandpass filter were designed. The new type bandpass filter with a bandwidth of 30 MHz centered at 610 MHz was designed on the RO4003 substrate ($\varepsilon_r = 3.55$, $h = 0.5$ mm, $\tan\delta = 0.003$). Fig. 11 gives the photograph of this filter. Its overall area is 60×40 mm which amount to the reduction size by over 80% compared to conventional coupled line filter.

The filter was simulated using circuit simulator ADS2008. The measurement and simulated result presented in Fig. 12. It is shown

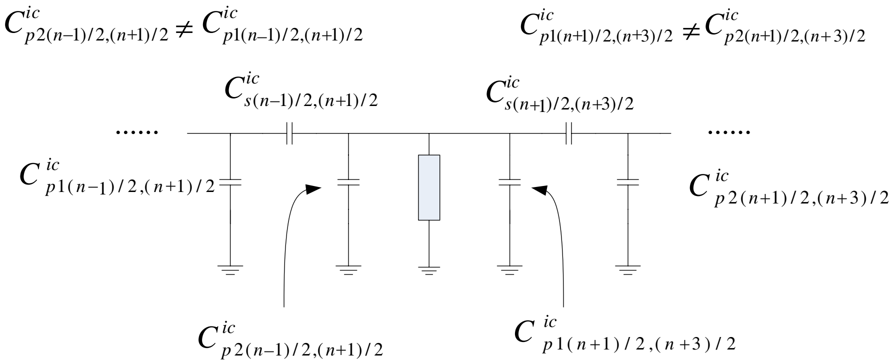

Figure 10. Interdigital equivalent circuit near the central stub short-circuited for odd order filter.

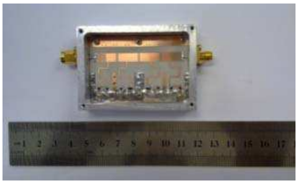

Figure 11. The photograph of the designed new type bandpass filter centered at 610 MHz.

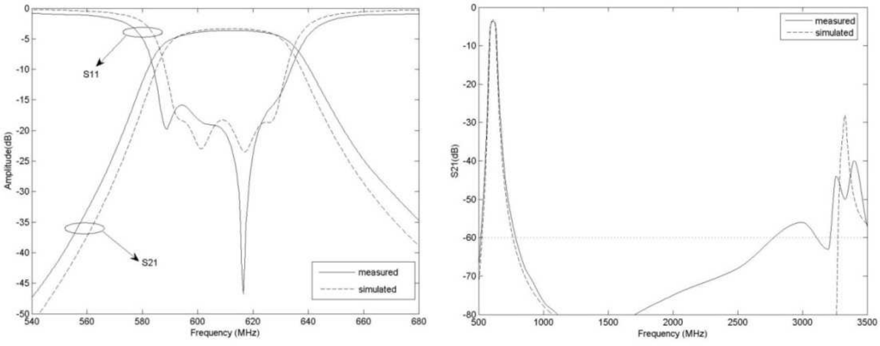

Figure 12. The measurement and simulated result for new type bandpass filter centered at 610 MHz.

that the experimental result is in good agreement with simulation and this filter works excellently in suppressing and harmonics almost up to the 4th order.

As seen from Fig. 11, the stub shorted to the ground plane by a $0\,\Omega$ resistance. An important benefit of this bandpass filter is the possibility of a little center frequency tuning by adjusting position of $0\,\Omega$ resistance. Moreover, Fig. 13 shows Measured group delay of our proposed bandpass filter which is almost flat over the whole pass band.

For the purpose of comparison, a photograph of the fabricated prototype and measured results of the another metamaterial bandpass filter type of order 4 with a bandwidth of 100 MHz centered at 2300 MHz together with the conventional one are shown in Fig. 14 and Fig. 15, respectively. The layout board of dimension 40 mm × 18.5 mm is fabricated on the substrate of the RO4003 with substrate thickness of 0.5 mm along with those for a conventional band pass filter. Fig. 16 shows the measured and simulated results of this new metamaterial

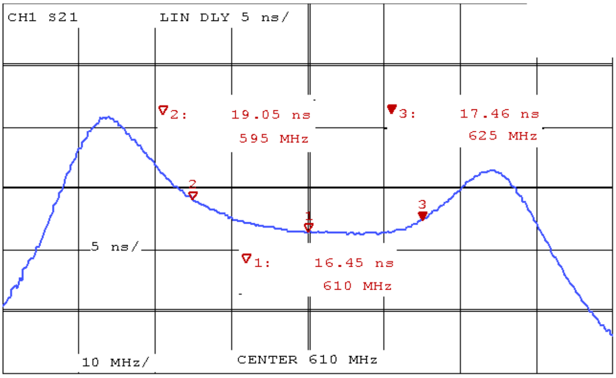

Figure 13. Measured group delay of our proposed bandpass filter.

Figure 14. The photograph of the designed new type bandpass filter centered at 2300 MHz (down) compared with the conventional one (up).

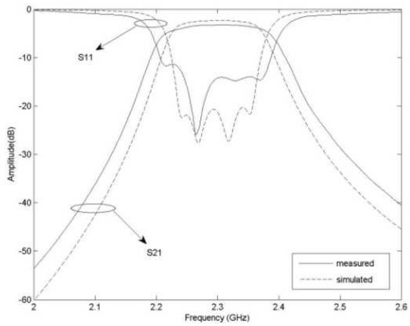

Figure 15. The measurement and simulated result for new type bandpass filter centered at 2300 MHz.

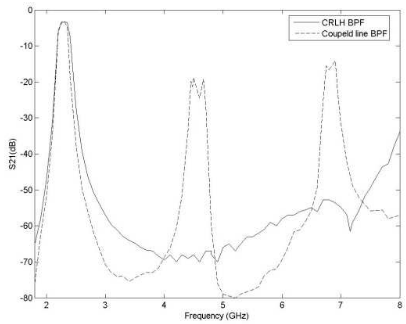

Figure 16. The measurement of the designed new type bandpass filter with excellent harmonic suppression compared with conventional one.

bandpass filter. The measured results are obtained with Rohde & Schwartz ZVA vector network analyzer. ADS2008 is used to simulate the circuit shown in Fig. 16 and a good agreement is obtained between them. It is noted that the metamaterial based bandpass filter is much smaller in size as compared to a conventional filter of same frequency and performance. Also, Fig. 15 shows that the proposed filter works excellently in harmonic suppression compared with the conventional one.

6. CONCLUSION

This paper presents the design formula for an innovative compact bandpass filter working in the UHF band by using a Composite Right/Left Handed Transmission Line. To validate the design formula, a sample bandpass filter was designed with a miniaturization 80% at 610 MHz, and the layout board of bandpass filter was fabricated and measured. The proposed metamaterial bandpass filter works excellently in suppressing the harmonics almost up to the 4th order. It was shown that the simulated and measured performances were in good agreement.

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