An Elliptic Low-pass Filter With Shorted Cross-over and Broadside-coupled Microstrip Lines

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Abstract — **An S-band two-layer elliptic type microstrip low-pass filter (LPF) is presented. This filter employs an open-ended cross-over and broadside-coupled microstrip lines with shorting vias to obtain the transmission zeroes near the passband. An LPF has been fabricated and measured; simulation and measurement results prove validity of the proposed structure.**

I. INTRODUCTION

Conventional microstrip low-pass filters, such as LC-ladder type filter using stepped-impedance transmission line or open-circuited stubs, have been widely used in microwave systems. To obtain an even sharper rate of cutoff for a given number of reactive elements, filters with elliptic function response are often desired, which can give infinite attenuation poles at finite frequency. Figure.1 illustrates the commonly used network structure for elliptic function lowpass prototype filters. The shunt branches of series-resonant circuits are introduced for realizing the finite-frequency transmission zeros, since they short out transmission at resonance. Usually the series resonant is difficult to be realized in microstrip application. There are few ways to obtain the elliptic LPF for microstrip applications. One of them is to make use of stepped-impedance lines shunted to the main transmission line to approximate the L-C elements shunted along a transmission line. However, the filter using stub resonator always requires characteristic impedances that are difficult to realize in practice [1]-[2]. Recently, with the development of integration technology and rapid growth of LTCC technique, multilayered structures are becoming popular in filter design.

In this paper, we present newly developed S-band LPF using two-layer microstrip structure. The cross-over and broadside-coupled microstrip structures used in this filter are analyzed with network theory and quasi-static approach. Equivalent circuits are extracted and employed for optimization of the design of the low pass filter. The design procedure is verified by comparing with analysis results.

II. ANALYSIS AND DESIGN

Broadside coupling has been used for obtaining tight coupling between microstrips. Fig. 2 (a) is the cross-section of the broadside-coupled microstrip line. Usually, a quasi-static equivalent circuit is applied to unit length of this kind of transmission line [3]. The per-unit-length capacitances are determined by utilizing the Zeland IE3D™. Fig. 2(b) gives the unit length equivalent circuit. In this design, the bottom quarter wave length line (at resonant frequency) is kept open at the two ends, and a via to the ground is put under the strip (Fig.2(c)). Malherbe gave the network model of TEM coupled lines in homogeneous medium [4]. For the coupling structure in Fig. 2(c), suppose the under-crossing line are ¼ wavelength, an equivalent circuit in Fig. 2(d)(f) could be derived. However, it is not accurately suitable for coupled line in inhomogeneous medium where the phase velocities of the two modes are not equal. For the broadside coupled microstrip line, there exists c and π mode related to in-phase and out-of-phase excitation. In this design, The coupling structure was modeled into several cells (each cell is around 0.025 wavelength) in Fig. 2(b) in series and inserted with a shorting via (or a lumped inductance) among them (Fig. 2(e)). Unit length capacitances and inductances were multiplied by the cell length. This network is analyzed and optimized to give the estimated coupled structure parameters which could match the L-C parameters in the prototype (Fig. 2(f)) in the required frequency. However, the structure parameters (the widths of the coupled lines and the length of the under-crossing strip) have to be optimized and adjusted with full-wave simulations to obtain the ultimate system requirements.

Fig. 3 illustrates the side and top view of the cross-over microstrip line structure. The under-crossing strip line with two open ends has a characteristic admittance of Y_1; it is separated by the crossing region of length l_a and l_b. The main transmission line on the top layer is with characteristic admittance of Y_0.

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The equivalent circuit of the structure is shown in Fig. 4. The overlapping region is modeled as a capacitive lumped π-network which includes the self capacitances C_{s1}, C_{s2} and mutual coupling capacitance C_{12}. The lumped capacitances are computed with quasi-static technique, which solve excess charge densities on the strips using electro-static 3-D Green function and the Galerkin method in spectrum domain [5-6]. This approach has been proved valid for various microstrip discontinuity problems. L_t is the effective inductance introduced in the longitudinal direction when the width of under-crossing strip is not neglectable: $L_t = \frac{Z_\pi \tan(\beta w_1/2)}{\omega}$, Z_π corresponds to the characteristic impedance of the coupled structure in π-mode (out of phase excitation mode), and it is determined by per-unit-length capacitance of broadside-coupled microstrip line analysis.

In the equivalent circuit, the admittance at point B is

$$Y_B = jB_B = j[\frac{B_A \cdot B_{12}}{B_A + B_{12}} + B_{s2}],$$

in which, $B_{12} = \omega C_{12}, B_{s2} = \omega C_{s2}$, B_A is the susceptance at point A, and it is defined by:

$$B_A = Y_1 \tan(\beta l_{e1}) + Y_1 \tan(\beta l_{e2}) + B_{s1}$$
$$= Y_1 \frac{\sin(\theta_{e1} + \theta_{e2})}{\cos\theta_{e1} \cdot \cos\theta_{e2}} + B_{s1},$$

($\theta_{ei} = \beta l_{ei}\ (i=1,2)$, $l_{e1} = l_a + l_{ea}$, $l_{e2} = l_b + l_{eb}$, $l_b = l_{b1} + l_{b2}$, $B_{s1} = \omega C_{s1}$). l_{e1} and l_{e2} is the equivalent length of transmission line considering fringing fields at the open end discontinuities of under-crossing strip line.

To achieve serial resonance at point B, it must be satisfied that $B_A + B_{12} = 0$, i.e. $\pi/2 < \theta_{ei} < \pi$, and the resonant frequency is

$$\omega_0 = Y_1 \frac{-\tan(\beta l_{e1}) - \tan(\beta l_{e2})}{C_{s1} + C_{12}}.$$

If the upper substrate thickness h_1 is small compared with microstrip line width and h_2, C_{12} is large in comparison with C_{s1} and C_{s2}. For a relatively narrow bandwidth ($\Delta\omega << \omega_0$), this structure can be approximated by a series LC block at the resonant frequency and it is obtained that

$$L \cong \frac{C_{s1} + C_{12}}{-\omega_0 C_{12}(Y_1 \tan(\beta_0 l_{e1}) + Y_1 \tan(\beta_0 l_{e2}) + \omega_0 C_{s1})}, C = \frac{1}{\omega_0^2 L}.$$

From the equivalent circuits, it can be easily found that the size of the structure could be further reduced if use only half of the crossing line as in Fig 1. (c). Supposing the value of C_{12} doesn't change, the equivalent L value will increases by twice, and the resonant frequency decreases, it satisfies $\omega_0 = Y_1 \frac{-\tan(\beta l_{e1})}{C_{s1} + C_{12}}$.

Now we add a via from the under-crossing strip to the ground. The via is modeled by a lumped inductance L_v. For a straight conductor of length l and radius R, neglecting the effect of nearby conductors, its self inductance value is given by $L_v = \frac{\mu_0 l}{2\pi}[\ln(\frac{2l}{R}) - 0.75]$. Suppose the via is positioned where the under-crossing strip l_b is separated by l_{b1} and l_{b2},

$$B_A = Y_1 \frac{Y_1 \tan(\beta l_{e2}') + B_v + Y_1 \tan(\beta l_{b1})}{Y_1 - (Y_1 \tan(\beta l_{e2}') + B_v) \cdot \tan(\beta l_{b1})}$$
$$+ Y_1 \tan(\beta l_{e1}) + B_{s1}$$

in which, $B_v = \frac{-1}{\omega L_v}$.

The parameters in the structure are optimized with Ansoft Serenade™ until the equivalent L, C values match to the filter prototype at the resonant frequency.

III. SIMULATION AND MEASUREMENT

In this design, the LPF with two transmission zeroes takes the prototype in Fig. 1. This prototype has a five-order elliptic function response, and it has a passband ripple L_{Ar}=0.08dB, and cutoff frequency at f_c=2.15GHz, stop band ratio of 1.1. The two finite-frequency attenuation poles occurs at f_1=2.45GHz, f_2=3.36GHz. The circuit parameters are: L_1=0.815nH, L_2=6.94nH, C_1=0.61pF, L_3=4.544nH, L_4=1.6nH, C_2=1.4pF, L_5=2.67nH.

Fig. 5 shows the ultimate filter structure on a two-layer Roger4003 substrate. The size of the whole filter is 30mm×14mm×2.35mm. The upper and bottom layer have the same dielectric constant of 3.38, and with thickness of 31mil, and 62mil respectively. The width of the upper strip is 3.6mm, and the strip is connected to SMA by two tapered matching lines. Other parameters are as followings: w_1=10mm, l_a=l_b=4.2mm, l_{b1}=0.3mm, d=2mm, l_2=14mm, the width of bottom coupling strip w_2=3.6mm, the radii of the two vias are 0.3mm. The whole filter structure is simulated with IE3D based on MOM. The simulated and measured S-parameters are shown in Fig. 6.

IV. CONCLUSION

In this paper, a five-order elliptical microstrip LPF employing a two-layer orthogonally cross-over microstrip line and a broadside-coupled microstrip line combined with shorting vias has been proposed. The equivalent circuit for each two-layer element was derived with efficient quasi-static analysis method. The whole system was designed and optimized with commercial software from the extracted equivalent circuits. The numerical simulation and measurement of the fabricated filter prove the validity of the analysis and design. The proposed filter structure is appropriate in LTCC multi-layered modules.

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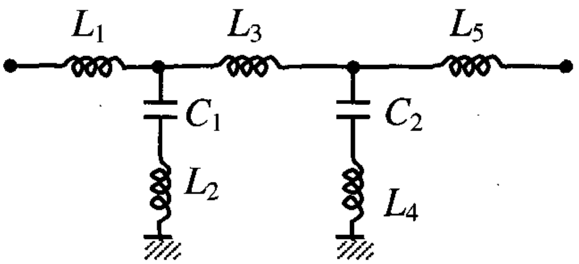

Fig.1. A commonly used lumped elliptic function LPF prototype

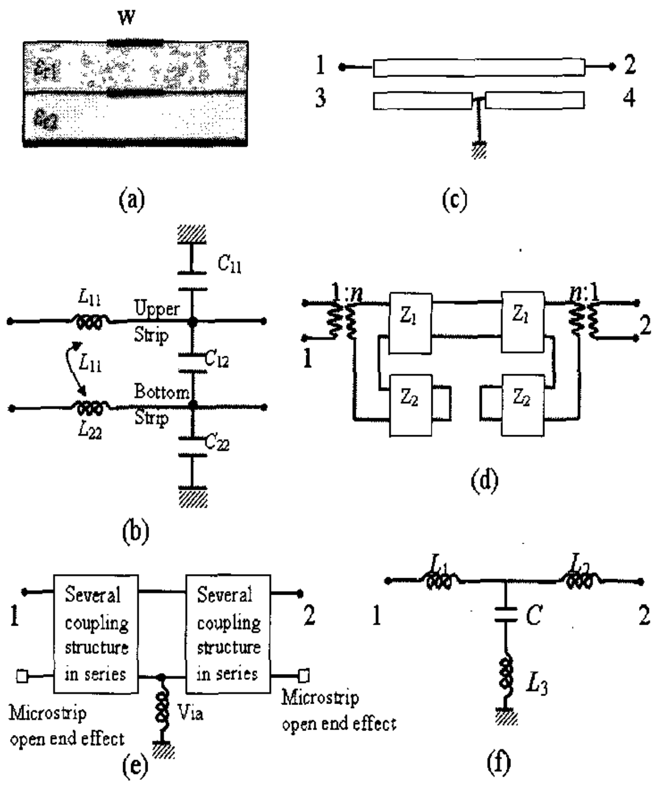

Fig.2. Broadside coupled microstrip structures and equivalent circuits

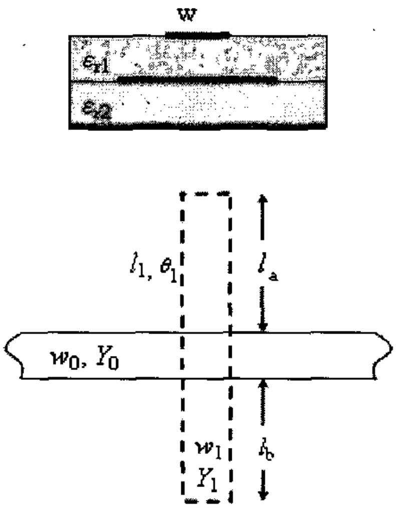

Fig.3. The structure of cross-over miscrostrip lines

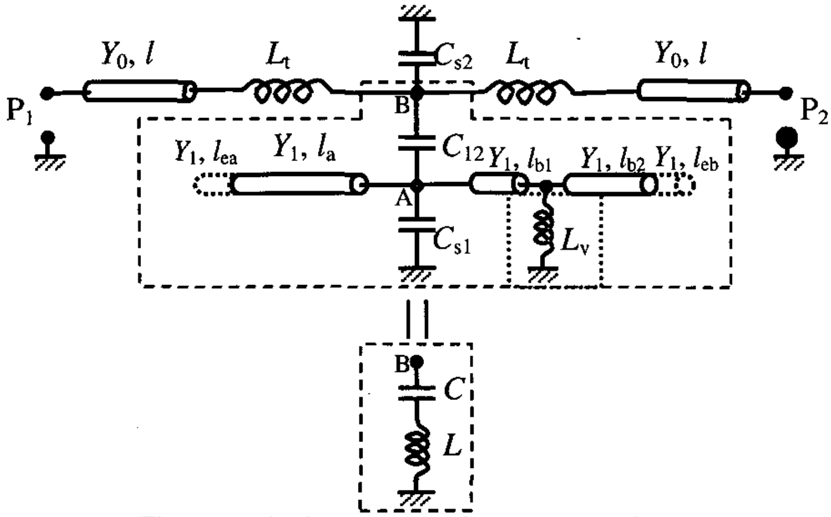

Fig.4. Equivalent circuit of cross-over microstrip lines

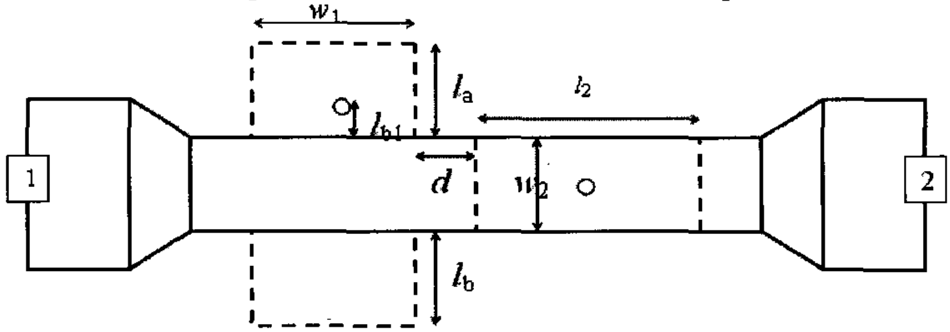

Fig.5. Structure of the LPF

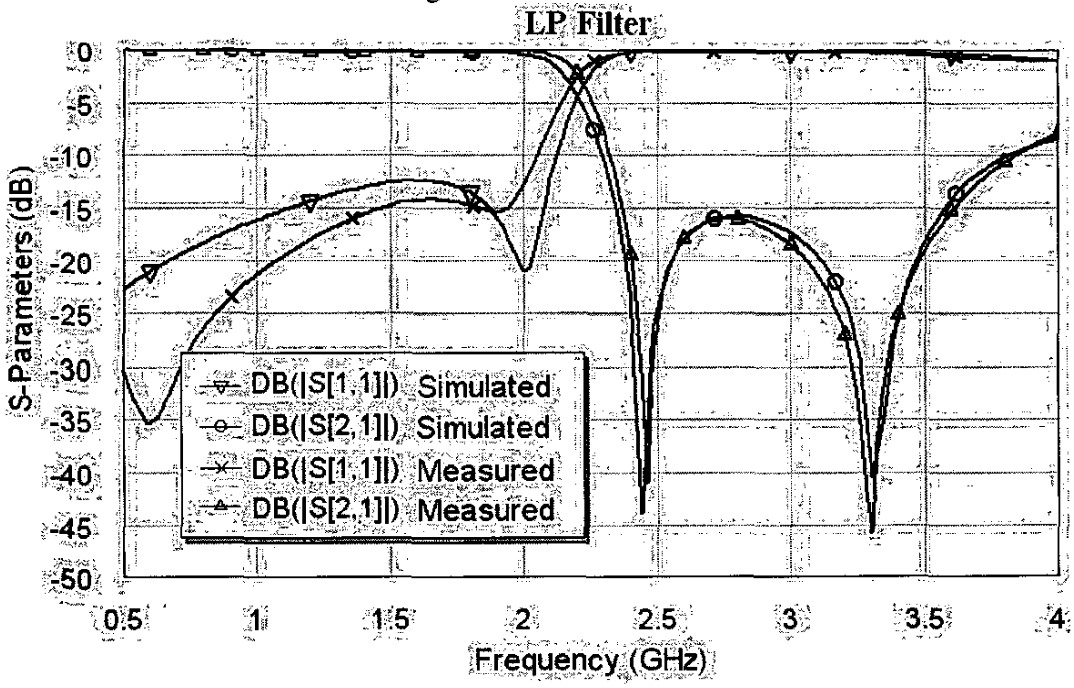

Fig.6. Simulated and Measured results of the LPF