Wideband Bandpass Filter Design with Three-Line Microstrip Structures

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Abstract – **Systematic procedure is described for designing bandpass filters with wide bandwidths based on parallel coupled three-line microstrip structures. It is found that the tight gap sizes between the resonators of end stages and feed lines, required for wideband filters based on traditional coupled line design, can be greatly released. The relation between the circuit parameters of a three-line coupling section and an admittance inverter circuit is derived. A design graph for substrate with ε_r = 10.2 is provided. Two filters of orders 3 and 5 with fractional bandwidths 40% and 50%, respectively, are fabricated and measured. Good agreement between prediction and measurement is obtained.**

I. Introduction

Microstrip realization of parallel coupled-line filters has been found to be one of the most commonly used filters in many practical wireless systems for over 30 years [1]-[2]. Major advantages of this kind of filters include its planar structure, relatively wide bandwidth, and a simple design procedure. Based on the insertion loss method [1], a given bandpass filter function of Butterworth or Chebyshev type can be easily synthesized, and the filter performance can be improved in a straightforward manner at the expense of increasing the order of filter.

It is because broadband wireless access (BWA) is an important issue in current developments of modern wireless communication systems, bandpass filters with relatively wide bandwidths are frequently required in the RF front ends to meet this trend. When these filters with wide bandwidth are to be realized by parallel coupled microstrip lines, the main limitation will be the small gap sizes of the first and the last coupling stages. The more fractional bandwidth, the smaller the gap size is required to enhance the coupling. Unfortunately, the smallest gap size that can be made with good repetition in a general laboratory is around 5mil or 0.13mm using wet chemical etching process.

Shrinking the gap size is not the only way to increase coupling of coupled lines. It is the idea of this work to add a third line, which is identical to the coupled ones, on the other side of the input line, to increase the coupling so that the gap size can be released for a specific coupling. Thus, the basic building block or coupling section for such a filter becomes a symmetric three-line microstrip structure as shown in Fig.1.

During the past, typical publications related to this work are [2]–[4]. In [2], a modified parallel-coupled filter structure is proposed. The tanks are aligned alternatively along two parallel traces so that less space than the traditional arrangement is used. The coplanar waveguide filters in [3] are also of a three-line approach. Lumped circuit models of the discontinuities formed between adjacent sections have to be incorporated into the circuit simulation to get better prediction for the filter performance. In [4], the presented three-line filter consists of only one coupling section.

The presentation is organized as follows. In Section II, starting with treating a coupled three-line section as a six-port network, we demonstrate how it is equivalent to an admittance inverter for use in the filter design. In Section III, a design graph with circuit parameters of the three-line system is provided, and the gap sizes required for the first stage of a bandpass filter with two-line sections are compared with those for our three-line cases. Two filters are fabricated and measured, and their results are presented in Section IV.

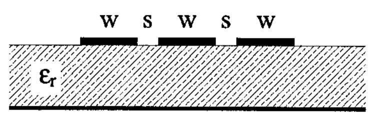

Fig.1. Cross-section view of a three-line microstrip structure.

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II. Equivalent Circuit of a Coupled Section

For the three-line microstrip structure shown in Fig.1, there are three quasi-TEM or dominant modes. The inductance matrix [L] and capacitance matrix [C] per unit length for the structure can be obtained by invoking the spectral domain approach (SDA) [4]-[5]. Due to the symmetry of the structure, the eigenvoltage matrix for these dominant modes can be written as

$$[M_V]=\begin{bmatrix} 1 & 1 & 1 \\ m_1 & 0 & -m_3 \\ 1 & -1 & 1 \end{bmatrix} \tag{1}$$

Each vector of [M_V] is the eigenvector of the matrix product [L][C]. The eigenvoltage matrix can be used to derive the relation between the port voltages and port currents defined in Fig.2(a) as

$$\begin{bmatrix} V_a \\ V_b \end{bmatrix}=\begin{bmatrix} Z_a & Z_b \\ Z_b & Z_a \end{bmatrix}\begin{bmatrix} I_a \\ I_b \end{bmatrix} \tag{2}$$

where $[V_a] = [V_1, V_2, V_3]^T$, $[V_b] = [V_4, V_5, V_6]^T$, $[I_a] = [I_1, I_2, I_3]^T$, and $[I_b] = [I_4, I_5, I_6]^T$, and the impedance matrices [Z_a] and [Z_b] are found to be

$$[Z_a] = [M_V]\ \mathrm{diag}[-jZ_{mi}\cot\theta_i]\ [M_V]^{\mathrm{T}} \tag{3}$$

$$[Z_b] = [M_V]\ \mathrm{diag}[-jZ_{mi}\csc\theta_i]\ [M_V]^{\mathrm{T}} \tag{4}$$

where $\theta_i = \beta_i l$ with β_i being the phase constant of the i^{th} mode and l the length of the coupled section, and Z_{mi} is given as

$$Z_{m1}=\frac{Z_{o1}}{m_1^2+2} \tag{5a}$$

$$Z_{m3}=\frac{Z_{o3}}{m_3^2+2} \tag{5b}$$

with Z_{oi} the characteristic impedance of mode i [6].

It is to be noted that when the input line is connected to the center line of the symmetric three-line structure in Fig.1, only modes 1 and 3 will be excited, since mode 2 is an odd-mode. For the six-port network shown in Fig.2(a), let port 2 be the input port and ports 4 and 6, which are connected together, be the output port. The corresponding terminal conditions are

$$I_2 = I_i \tag{5a}$$

$$V_2 = V_i \tag{5b}$$

$$I_1 = I_3 = I_5 = 0 \tag{5c}$$

$$I_4 + I_6 = I_o \tag{5d}$$

$$V_4 = V_6 = V_o \tag{5e}$$

The network becomes a two-port. We follow the approximations used in [3] to establish the equivalence between the coupled line section in Fig.2(a) and the inverters in Fig.2(b) and Fig.2(c). Firstly, assume that the three modal phase constants are approximately the same, and let $\beta_i l = \pi/2$ at center frequency. Secondly, compare the two-port Z-parameters of the circuits in Fig.2(a) with those of Fig.2(b), then we can obtain

$$m_1Z_{m1} - m_3Z_{m3} = JZ_AZ_B \tag{6a}$$

$$m_1^2Z_{m1} + m_3^2Z_{m3} = Z_A\,(J^2Z_AZ_B + 1) \tag{6b}$$

$$Z_{m1} + Z_{m3} = Z_B\,(J^2Z_AZ_B + 1) \tag{6c}$$

Thirdly, approximate the admittance inverter in Fig.2(b) by the circuit in Fig.2(c) with $Z_o^2 = Z_AZ_B$. The next step of approximation in our design is to reduce the products of (6b) and (6c). Taking $m_1^2 + m_3^2 \approx 2m_1m_2$, we have

$$m_1Z_{m1} + m_3Z_{m3} \approx Z_o\,(J^2Z_o^2 + 1) \tag{7}$$

This is a key step in our design procedure which can greatly simplify the determination of line width and line spacing of each coupled line section in a general N^{th} order three-line bandpass filter. It is to be noted that from the design equations for a bandpass filter with N + 1 coupled line sections in [1], the value of JZ_o for each admittance inverter can be simply determined from the values of lumped circuit elements of the low-pass filter prototype. Once JZ_o is known, (6a) and (7) can be solved simultaneously to determine the values of m_1Z_{m1} and m_3Z_{m3} for each coupled section.

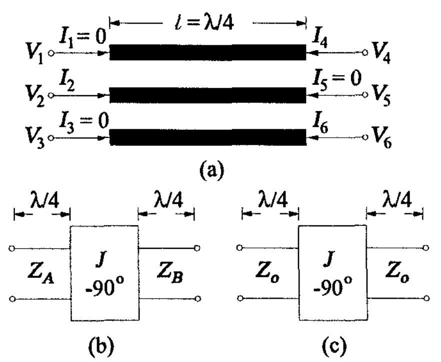

Fig.2. (a) Voltages and currents for a coupled three-line section as a six-port network. (b) Equivalent admittance inverter. (c) Further approximated admittance inverter.

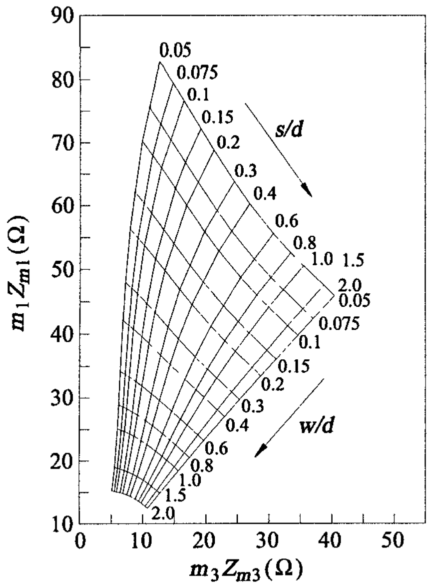

Fig.3. The bandpass filter design graph for a symmetric three-line microstrip structure. The substrate $\varepsilon_r = 10.2$.

III. The Design Graph and a Comparison with Conventional Coupled Line Filters

The line width and spacing for each coupled three-line section can be determined with the aid of the design graph as shown in Fig.3, where the substrate $\varepsilon_r = 10.2$. It is to be noted that the results do not scale with dielectric constant, so new design graph must be made if the value of dielectric constant is changed [1]. Note also that the vertical and horizontal axes for the plots in Fig.3 are m_1Z_{m1} and m_3Z_{m3}, respectively. In fact, to obtain sizes of circuit layout with more accuracy, we determine the values of w/d and s/d by using a root-searching program based on the Newton's method in two dimensions and a database in readiness.

Next, we compare the gap sizes for the first coupling stages of parallel coupled line filters built by two-line sections with those built by the three-line sections. From the design equation provided in [1], the inverter value for the first stage can be calculated as

$$J_1 = \frac{1}{Z_0}\sqrt{\frac{\pi\Delta}{2g_1}} \quad (8)$$

where Δ is the fractional bandwidth, and g_1 the first element value for the low-pass filter prototype. Thus, the corresponding Z_{oo} and Z_{oe} for a two-line section, and m_1Z_{m1} and m_3Z_{m3} for a three-line section can be known. The gap sizes are then determined by invoking the design graph as in Fig. 3. Table I compares the gap size to substrate thickness ratios s/d, for the first coupled section of a third order Chebyshev filter, on substrate with $\varepsilon_r = 10.2$, with 0.5dB passband ripple and Δ from 30% to 60%, based on two-line and three-line microstrip approaches. All the s/d values for the three-line sections are more than two times that for the two-line sections. Thus, when the gap size becomes a limitation in fabricating a wideband parallel coupled two-line filter, the three-line sections can provide an effective solution.

TABLE I

Values of s/d of the First Coupling Stage for Designing a Parallel Coupled Line Filter

Δ	30%	40%	50%	60%
Two-line	0.177	0.151	0.137	0.129
Three-line	0.415	0.354	0.312	0.281

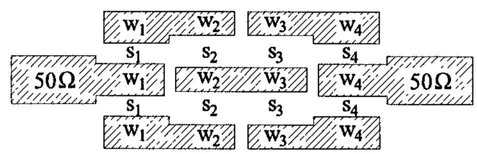

Fig.4. The layout of the three-line filter for $N = 3$. The length of each coupled line section is $\lambda_g/4$.

IV. The Fabricated Filters and Measurements

Two filters are designed and fabricated as illustrative examples. The specifications for the first filter are $N = 3$, 0.5 dB passband ripple, center frequency $f_o = 2.45$ GHz, and the fractional bandwidth $\Delta = 40\%$. The pattern of the circuit layout is shown as in Fig.4. The length of each coupled line section is approximately $\lambda_g/4$.

Before fabricating the circuit, we still have to compensate the parasitic capacitances resulted from the open ends of each straight resonator and the series gap coupling. These two parasitics will lead to series gap and end effect line extensions. Based on the capacitance equations in the equivalent lumped model given in [7], the former can be eliminated by cutting a short length of each line section at both ends, then a further shortening is required for compensating the series gap capacitance.

We use the full-wave simulator IE3D [8] to validate our design before the circuit is fabricated. Fig.5 shows the simulated and the measured S-parameters results. Their $|S_{11}|$ and $|S_{21}|$ responses have good agreement.

The second filter has $N = 5$ and $\Delta = 50\%$, and all the other specifications are identical to those of the first one. The results are plotted in Fig.6. The $|S_{21}|$ has better stopband rejection than that in Fig. 5 as it has a higher order. Again, the measured responses agree quite well with the simulated results.

V. Conclusion

It is believed that for the first time a systematic design procedure is given for synthesizing parallel coupled line filters based on a three-line microstrip system. One of the key steps is to establish the equivalence between a coupled three-line section and an admittance inverter circuit.

The whole structure is more compact than the traditional parallel coupled two-line filters. The most important advantage of such a design is that the lower limit of the tight gap size for designing filters of wide bandwidth can be greatly released as compared with the traditional two-line design.

With each line section being compensated by the end effect and series gap line extensions, a good agreement between the predicted and measured results is obtained.

Acknowledgement

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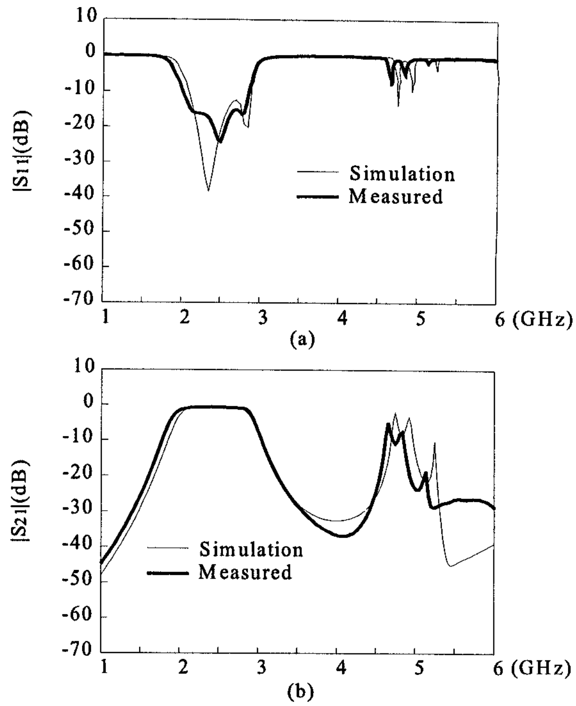

Fig.5. Measured and simulated results for a Chebyshev filter. $N = 3$, ripple level = 0.5 dB, $\Delta = 40\%$. $w_1/d = w_4/d = 0.133$, $w_2/d = w_3/d = 0.208$, $s_1/d = s_4/d = 0.354$, $s_2/d = s_3/d = 0.478$, $d = 1.27$ mm, $\varepsilon_r = 10.2$. (a) $|S_{11}|$. (b) $|S_{21}|$.

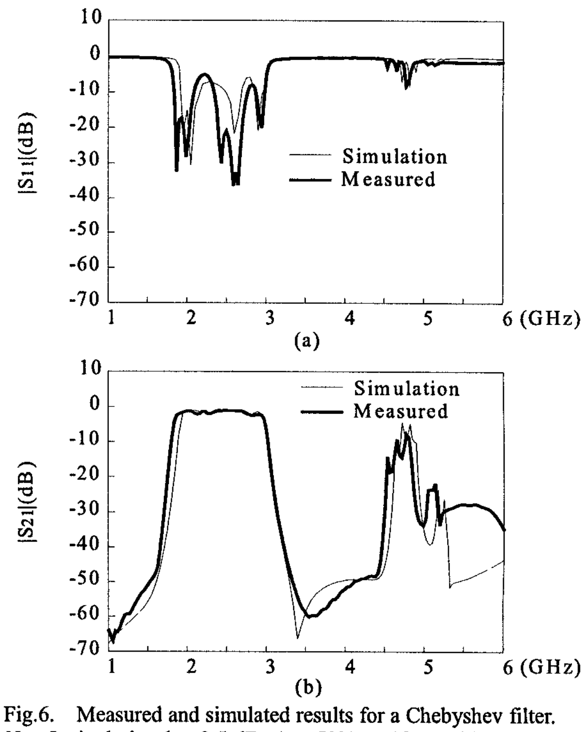

Fig.6. Measured and simulated results for a Chebyshev filter. $N = 5$, ripple level = 0.5 dB, $\Delta = 50\%$. $w_1/d = w_6/d = 0.112$, $w_2/d = w_5/d = 0.173$, $w_3/d = w_4/d = 0.224$, $s_1/d = s_6/d = 0.324$, $s_2/d = s_5/d = 0.415$, $s_3/d = s_4/d = 0.513$, $d = 1.27$ mm, $\varepsilon_r = 10.2$. (a) $|S_{11}|$. (b) $|S_{21}|$.