行政院國家科學委員會補助專題研究計畫成果報告

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※　　　　　　　線射頻濾波器　　　　　　　　　　　※
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計畫類別：□個別型計畫　　□整合型計畫

計畫編號：NSC　89－2213－E－009－057

執行期間：　88 年　8 月　1 日至　89 年 7 月　31 日

計畫主持人：郭仁財

共同主持人：

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執行單位：國立交通大學電信工程學系

中　華　民　國　89 年　7 月 31 日

行政院國家科學委員會專題研究計畫成果報告

利用網路合成法以密集微小化平面振子
設計高品質無線射頻濾波器

Planar Microstrip Filters with Transmission Zeros at Real or Imaginary Frequencies Using Compact Miniaturized Hairpin Resonators

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一、中文摘要

本計畫利用髮夾型及開迴路方形共振子，設計單一傳輸零點的橢圓函數響應的平面微帶濾波器。文中用網路合成技術，由給定的頻率響應函數計算出所需的耦合係數，再利用電磁模擬軟體，決定耦合共振子之間的適當距離，以完成整個電路的佈局。文中設計並製作4%及10%頻寬的帶通濾波器，並與實際量測結果比較。

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關鍵詞：微小化電路、微波濾波器、耦合係數、髮夾形共振子、開迴路方形共振子。

Abstract

Four–pole cross-coupled planar microwave filters are designed using hairpin resonators and square open-loop resonators. The synthesis technique for the design of filter is described. The coupling coefficients for the three basic coupling structures are then determined by the filter specification. The full-wave simulator IE3D is used to design the resonators and the three basic coupling structures. Experimental results are compared with computed responses.

Keywords: miniaturized circuit, microstrip filters, coupling coefficient, hairpin resonator, square open-loop resonator.

二、緣由與目的

Recently, expanding wireless and mobile communication systems have presented new challenges to the design of high quality miniature RF filters. Planar filters would be preferred since they can be fabricated using printed circuit technology with low cost. Obviously, the size of planar filters with parallel-coupled, half-wavelength microstrips [1] is too large to be used in modern systems. Thus size reduction has been an important issue in developing RF filters.

The hairpin filters [2]-[4] make progress in size reduction from the parallel-coupled lines structure. The whole filter consists of a cascade of U-shaped resonators [2, 3]. If the orientations of the hairpin resonators alternate, the structure is capable of considerable bandwidth; if they do not, the structure has attractive properties for design of compact, narrow-band filters [3]. In [4], quasi-elliptic function filters are realized by the U-shaped microstrips with different orientations and offsets to establish an adequate coupling combination among the resonators.

Further progress in size reduction is made by the compact miniaturized hairpin resonator filters [5], where the two arms of the U-shaped microstrip are further folded to form a pair of closely coupled lines to enhance the capacitive nature of open-end arms. The area of such a resonator is no more than half of that of a U-shaped resonator. Also of interest are filters with microstrip square open-loop resonators [6], of which the lateral size is only one eighth guided wavelength at the midband frequency.

We have made several bandpass filters with compact miniaturized hairpin resonator for relatively narrow bandwidth, and with open square open-loop resonators for relatively wide bandwidth design. For a design with center frequency at f_0 = 2.46 GHz for comparison with existing literature, it shows that the area occupied by a compact miniaturized hairpin resonator is only 75% of that of a square open-loop resonator. For the coupling between two compact miniaturized hairpin resonators, it can be shown that the coupling coefficient will be difficult to be made large enough for designing filters with fractional bandwidth greater than 10%. For this reason, we turn to use square open-loop resonators for filters with 10% bandwidth. For the design with f_0= 2 GHz and fractional bandwidth (*FBW*) $\Delta f / f_0$ = 10%, the area used by a compact miniaturized hairpin resonator is only 55% of that of a square open-loop resonator. Thus, from the circuit area point of view, it is beneficial designing filters using compact miniaturized hairpin resonators.

三、研究報告

The compact miniaturized hairpin resonators show in Fig.1 are evolved from sections of half-wavelength open microstrips, thus the fundamental resonance of each resonator occurs in odd-mode. This means that at the resonance the voltage at the central valley of the folded microstrip is a minimum, and the voltages at both ends of the coupled lines have maximal values with opposite signs. The condition for this fundamental resonance can be derived from the concept of stepped impedance resonators [5].

In designing a single resonator for the first resonator, the width of the microstrip is determined by setting its characteristic impedance to 50Ω and in making the whole structure compact: 1) the peripheral of each resonator is made a square; 2) the length of the central open-circuited coupled lines is extended to its extremes; and 3) the gap and width of the coupled lines are made as small and large, respectively, as possible. As a result, the lateral size of the resonator is 6.1 mm, which is 13% less than that of a square open-loop resonator reported in [6], for designing a filter with the same center frequency.

The filters in Fig.1 and Fig.2 have a 2ξ2 configuration. Significant couplings exist between any two non-diagonally neighboring resonators. In our design, the coupling between resonators 1 and 2 is identical to that between resonators 3 and 4. Thus there are three basic coupling struc-tures to be investigated. The coupling in each structure can be specified by the two dominant resonant frequencies, which are split off from the resonance condition due to the electromagnetic coupling [5]. Let f_a be the lower of the two resonant frequencies, and f_b be the higher one. The coupling coefficient M_{ij} for resonators i and j can be calculated as [6]

$$M_{ij} = \pm \frac{f_b^2 - f_a^2}{f_b^2 + f_a^2} \qquad (1)$$

where the upper sign (+) applies to M_{12}, M_{23}, and M_{34}, while the lower sign (-) to M_{14}. Noted that the meaning of positive or negative coupling is rather relative. This means that if we refer to one particular coupling as the positive coupling, then the negative coupling would imply that its phase response is opposite to that of the positive coupling. It is to be noted that (1) is applied by separating off two resonators at a time, and that M_{12} and M_{34} are also subject to an offset d. We use the 3-D full-wave simulator IE3D [8] to determine the resonant frequencies of each coupling structure. The way to obtain M_{ij} is to model the pair of resonators and compute their transmission response. Each M_{ij} decreases as the corresponding distance is increased, as expected. For sufficiently accurate simulation results, it is found that a discretization of 80 cells per wavelength is required for our particular circuit configuration.

The design and synthesis of cross-coupled filters are distinguished by the location of their transmission zeros (attenuation poles) [9]. The transmission zeros improve the skirt attenuation performance of the response. The low-pass prototype filters may be synthesized as the admittance-inverters equivalent circuit. The extra cross-coupling of admittance inverter J_{m-1} is negative for the filter having attenuation poles [9]. An approximate synthesis technique will be presented which may be achieved by introducing cross coupling between one pair of nonadjacent elements of the standard Chebyshev filter while leaving all but one of the elements unchanged.

The four-pole cross-coupled bandpass filters could be synthesized with the aid of a low-pass prototype filter. Once the element values are found, the external quality factor Q_e and the coupling coefficients of the filter can be calculated by

$$Q_e = \frac{g_0 g_1}{FBW} \qquad (2)$$

$$M_{12} = M_{34} = \frac{FBW}{\sqrt{g_1 g_2}} \qquad (3)$$

$$M_{23} = \frac{FBW \cdot J_2}{g_2} \quad (4)$$

$$M_{14} = \frac{FBW \cdot J_1}{g_1} \quad (5)$$

Results

A four-pole elliptic function filter response can be realized using proper cross couplings. For the particular 2×2 configuration in Fig.1 and Fig.2, the cross couplings give the input signal two paths from the input port to the output. Through the two paths, the signal magnitude and phase are altered differently. Thus, at the output port, the multipath effect may cause attenuation poles at finite frequencies if the couplings are properly designed [4]. Fig.3, Fig.4, and Fig.5 compare the measured and the calculated responses of the three four-pole elliptic function bandpass filters which were designed and fabricated on a RT/Duroid 6010 substrate with a relative dielectric constant of 10.2, thickness 1.27 mm, and a loss tangent $\tan\delta = 0.002$. The measurements are performed with the HP8720C network analyzer. The passband insertion loss is less than 3.5 dB, which are mainly resulted from the conductor loss and radiation. In Fig.3, the center frequency f_0 is 2.46 GHz and the FBW is 4%. Fig.4 and Fig.5 show the responses of the bandpass filters with the compact miniaturized hairpin and square open-loop resonators. The center frequency f_0 = 2 GHz and the FBW = 10%. In Fig.4, the measured and calculated responses have also a good agreement. It is to be noted that in Fig.4, the upper stopband dose not show the attenuation pole, and in Fig.5, the location of transmission zeros are not symmetric. This could be from the fact that the coupling coefficient M_{14} is frequency-dependent. As the cross coupling is resulted from the electric coupling between resonators 1 and 4, the stronger the cross coupling, the closer is the transmission zero approaching to the passband edge [7]. Note that the second filter exhibited a wide upper stopband with a rejection better than 30 dB up to about 4.2 GHz and the third filter exhibited a upper stopband with a better than 20 dB up to about 3.4 GHz. The harmonic dispute of the second filter is better than the third filter.

四、結論

We have presented the design of four-pole elliptical function bandpass filters. The design is based on the knowledge of the coupling coefficients of the three basic coupling structures. The measured responses have good agreement with the theoretical predictions. The compactness in circuit size makes the design of cross-coupled filters using the miniaturized hairpin resonators attractive for further developments and applications in modern mobile radio systems.

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六、圖表

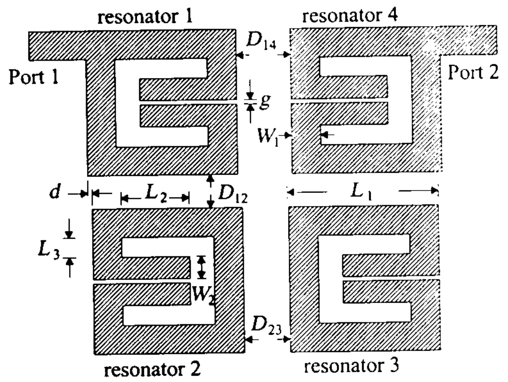

Fig.1 The layout of a compact miniaturized hairpin resonator filter with elliptical function response.

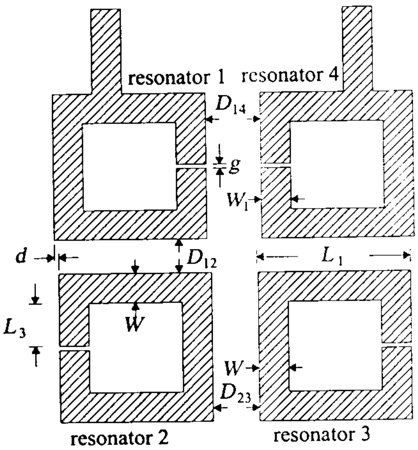

Fig.2 The layout of a square open-loop resonator filter with elliptical function response.

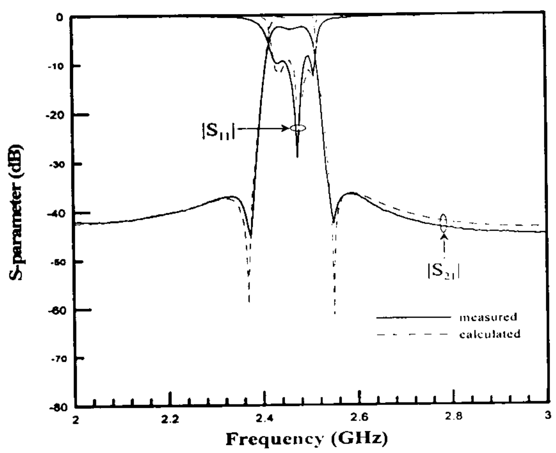

Fig.3 The measured and simulated S-parameters of a compact miniaturized hairpin resonator filter (f_0 = 2.46 GHz, *FBW*=4%, W_1 = 1.187 mm, W_2 = 0.91 mm, L_1 = 6.124 mm, L_2 = 2.56 mm, L_3 = 0.8 mm, and g = 0.173 mm, D_{12} = 1.44 mm, D_{23} = 1.82 mm, D_{14} = 2.22 mm, d = 0.20 mm).

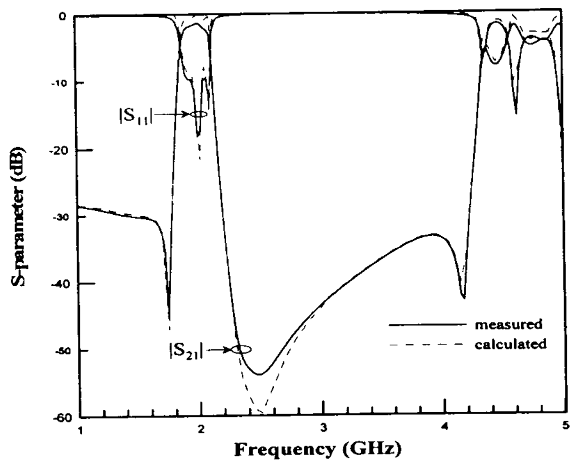

Fig.4 The measured and simulated S-parameters of a compact miniaturized hairpin resonator filter (f_0 = 2 GHz, *FBW*= 10%, W_1 = 0.6 mm, W_2 = 0.3 mm, L_1 = 6.1 mm, L_2 = 4.2 mm, L_3 = 1.75 mm, and g = 0.17 mm, D_{12} = 0.35 mm, D_{23} = 0.5 mm, D_{14} = 1.2 mm, d = 0.35 mm).

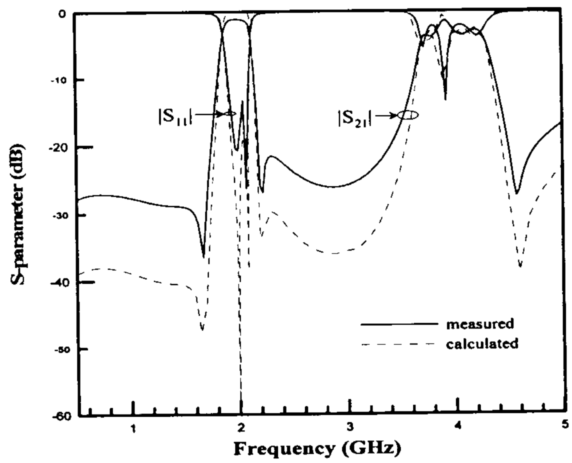

Fig.5 The measured and simulated S-parameters of a square open-loop resonator filter (f_0 = 2 GHz, *FBW*= 10%, W = 0.6 mm, L_1 = 8.85 mm, L_2 = 2.9 mm, and g = 0.2 mm, D_{12} = 0.7 mm, D_{23} = 0.8 mm, D_{14} = 2 mm, d = 0.6mm).