A Compact Filtering Microstrip Antenna With Quasi-Elliptic Broadside Antenna Gain Response

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***Abstract*—Design, fabrication, and measurement of a compact filtering microstrip antenna with second-order quasi-elliptic broadside antenna gain response are presented. A U-shape radiating patch is excited by a T-shape resonator through an inset coupling structure. The U-shape patch acts as a radiator as well as the last stage of the filter, and the inset coupling structure can be treated as the admittance inverter in filter design. The design procedure follows the circuit approach—synthesis of bandpass filters. The broadside gain of the filtering antenna has two poles in passband and two broadside radiation nulls (zeros) at the band edges for improving selectivity. Compared to the conventional inset-fed microstrip antenna, with a little extra circuit area, the proposed filtering antenna has a flatter passband response, better frequency skirt selectivity, and almost twice wider bandwidth. The measurement result shows a good agreement with the simulations.**

***Index Terms*—Band-edge selectivity, filter synthesis, filtering antennas, microstrip antennas, quasi-elliptic function bandpass filter.**

I. Introduction

In RECENT years, there has been prosperous development on microwave communication systems. One of the focusing issues nowadays is multifunctional component design, which miniaturizes the circuit size and improves overall performance.

The antenna and the bandpass filter are two indispensable components in a wireless communication system. In most of the RF front-end systems, the bandpass filter is cascaded right after the antenna for filtering the spurious signals. Conventionally, these two components are designed separately and connected by a section of 50-Ω transmission line. This line may not be matched perfectly within the whole frequency range in interest, so the transmission-line section forms a resonator at some frequency and causes interference. The extra line not only degrades the performance of the system, but also occupies the additional circuit area. In addition, polarization purity is one of the important issues for antenna design. The unwanted cross-polarized radiation cannot be filtered by any kind of electrical circuit, except by the antenna itself. A well-designed antenna should own the ability to select the copolarized signals and reject the cross-polarized ones.

Manuscript received February 15, 2011; revised March 24, 2011; accepted April 06, 2011. Date of publication April 29, 2011; date of current version May 16, 2011. This work was supported in part by the National Science Council, Taiwan, under Contract NSC97-2221-E-009-041-MY3.

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Digital Object Identifier 10.1109/LAWP.2011.2147750

Recently, the concept of *filtering antenna* was evolving. The design purpose and process of the filtering antenna are fairly different from those of the conventional antenna. The filtering antenna is not merely another impedance matching approach, but also a shaping for a filter-like frequency response both for antenna gain (corresponding to insertion loss in filter terminology) and input return loss. Several research works were on the filtering antenna design through the filter synthesis that the antenna acts as one of the resonators. The filtering antennas were achieved by different forms, such as circular patch antennas [1], slot dipole antennas [2], monopolar antennas [3], and rectangular patch antennas [4], [5].

In the authors' previous work [5], a filtering microstrip antenna with the inset-coupling structure was proposed. Based on [5], this letter proposes a performance-enhanced filtering microstrip antenna with second-order quasi-elliptic function antenna gain response. The cross-polarized radiation in broadside direction, which was not considered in [5], is now completely eliminated because of geometrical symmetry, so the band-edge selectivity and polarization purity are both greatly improved.

II. Design of the Filtering Antenna

The geometry of the proposed filtering microstrip antenna is depicted in Fig. 1(a). The structure is printed on one side of a Rogers RO4003 substrate with dielectric constant 3.38 and thickness 0.508 mm. The antenna has an all-planar structure with geometrical symmetry to the *yz* plane.

The filtering antenna consists of three parts—the U-shape radiating patch, the T-shape resonator, and the feeding microstrip line. The U-shape patch has a dimension of $w_\mathrm{p} \times l_\mathrm{p}$, with l_p about half a guided wavelength at the operating frequency. The T-shape resonator can be divided into three sections. Two of the arms are extended in the x-direction with length l_1 and width w_0. The other one is inset into the U-shape patch in the y-direction with length l_2 and width w_0. (l_1+l_2) is about a half-wavelength long. The resonator and the patch have gaps g_1, g_2, and g_3 apart. The feeding 50-Ω microstrip line with width w_0 (=1.17 mm) is inserted at the bottom.

Fig. 1(b) shows the corresponding equivalent circuit model of the filtering antenna. In an antenna designer point of view, the T-shape resonator and the U-shape patch are modeled by the parallel L_1C_1 [6] and $R_LL_2C_2$ [7] circuits, respectively. In the filter designer point of view, at the bottom of Fig. 1(b), the T-shape resonator corresponds to the first-stage resonator L_1C_1, and the U-shape patch corresponds to the second-stage resonator L_2C_2 with the load R_L. The coupling between these two resonators mainly comes from the quarter-wavelength parallel section (l_2), which is similar to the parallel coupled

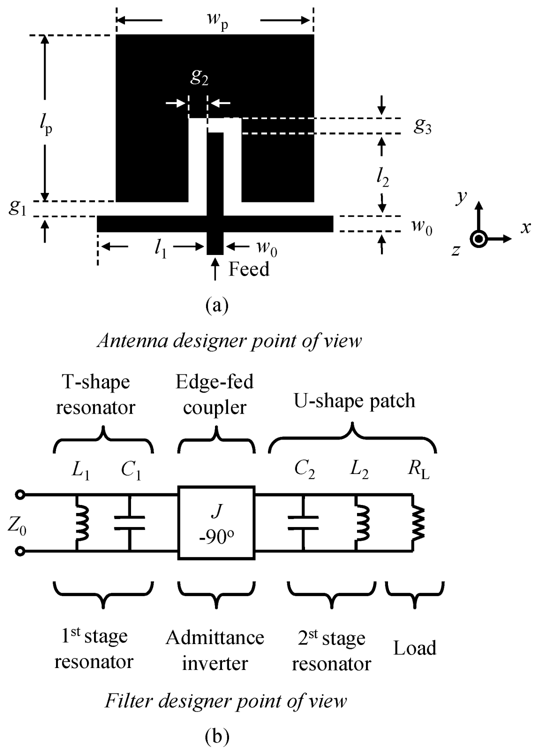

Fig. 1. (a) Topology and (b) equivalent circuit of the proposed filtering microstrip antenna.

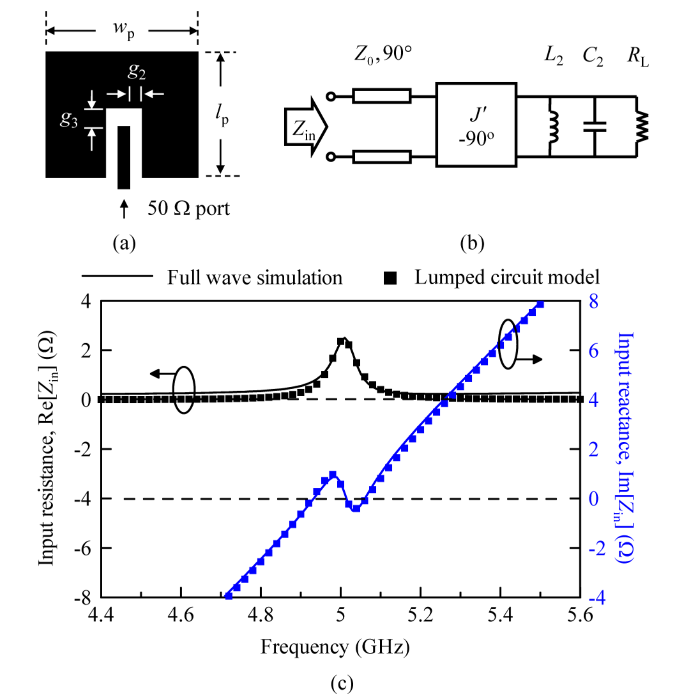

Fig. 2. (a) Test structure for extracting parameters of the U-shape patch with $w_p = 18.0$ mm, $l_p = 15.3$ mm, $w_0 = 1.17$ mm, $g_2 = 1.55$ mm, and $g_3 = 1.17$ mm. (b) Equivalent circuit with $C_2 = 37.4$ pF, $L_2 = 27.0$ pH, $J' = 4.8$ mS. (c) Input impedances of the lumped circuit model and full-wave simulation.

lines, and thus can be modeled by a J (admittance) inverter [6]. Since the equivalent circuit is exactly the same as the bandpass filter prototype, by employing the synthesis method of filters, the filtering microstrip antenna can be designed with the filter response. The antenna gain is analogous to the filter insertion loss, and the antenna radiation nulls at certain frequencies correspond to the filter transmission zeros.

In the filtering antenna design, we do not have the "port 2" for testing the external quality factor and coupling coefficients. Therefore, the conventional design process is greatly modified.

The design procedure starts from the second-order Chebyshev bandpass filter prototype. The first step is to determine the filter parameters, such as the operation frequency, the bandwidth, and the ripple level. Then, the lumped element values in Fig. 1(b) are available from literature [6]. The next step is to design the first- and second-stage resonators via the T-shape half-wavelength resonator and the U-shape patch, respectively. Finally, put them all together and perform some fine-tuning. The full-wave simulation throughout this letter is done by Ansoft High Frequency Structure Simulator (HFSS) [8]. The detailed design process is given as follows.

A. Second-Order Chebyshev Bandpass Filter Prototype

The bandpass filter prototype is chosen to be with a second-order Chebyshev equal-ripple response, with the ripple level $L_a(\mathrm{dB}) = 0.3$, center frequency 5 GHz, and port impedance $Z_0 = 50\ \Omega$. For an ideal Chebyshev filter, the minimum return loss in passband is

$$\mathrm{RL(dB)} = -10\log\left(1 - 10^{-L_{\mathrm{a}}(\mathrm{dB})/10}\right). \quad (1)$$

In this case, $\mathrm{RL} = 11.75$ dB. The fractional bandwidth Δ in this letter is set as 2%, which is about twice that of a well-matched ordinary inset-fed microstrip antenna. Once we get these filter parameters, the element values in Fig. 1(b) can be easily obtained [6]: $C_1 = C_2 = 37.4$ pF, $L_1 = L_2 = 27.0$ pH, $J = 26.0$ mS, and $R_L = Z_0 = 50.0\ \Omega$.

B. U-Shape Patch Design

Fig. 2(a) depicts the structure for extracting the equivalent lumped-circuit element parameters of the U-shape patch, and Fig. 2(b) is its equivalent circuit. The coupling mechanism affects the response of the resonator a lot. Thus, while extracting the parameters, the J-inverter is inseparable. Note that, in Fig. 1, the coupling microstrip section is a resonator, but in Fig. 2(a), it is not. Therefore, the value of J' in Fig. 2(b) is different from the J in Fig. 1(b) and is only for extracting the parameters of the U-shape patch.

Fig. 2(c) shows input impedances of both the full-wave and lumped-circuit model simulated result. The values of the parameters are listed in the figure caption. There are $R_L L_2 C_2$—three parameters to be determined. The patch width w_p mainly affects the radiation resistance R_L, and the patch length l_p dominates the resonant frequency, i.e., the product of L_2 and C_2. The ratios of L_2 and C_2 are determined by the substrate thickness. According to Fig. 2(c), the input impedance fits well within the whole band in which we are interested.

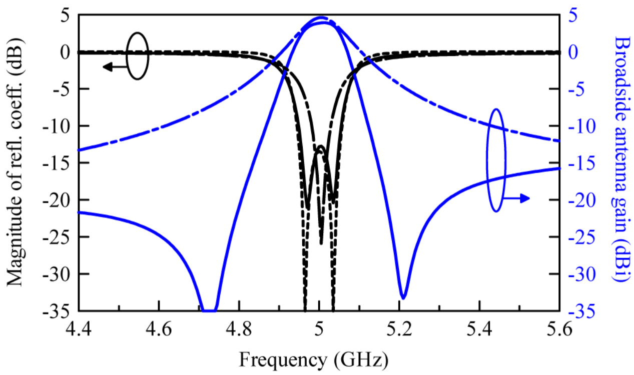

Fig. 3. Magnitude of reflections and broadside $(+z)$ antenna gains for the filtering antenna (solid lines), ordinary inset-fed microstrip antenna (dash-dot-dash lines), and the equivalent circuit model (dashed line). The dimensions of the microstrip filtering antenna are $w_{\rm p} = 18.0$ mm, $l_{\rm p} = 15.3$ mm, $l_1 = 10.91$ mm, $l_2 = 7.43$ mm, $g_2 = 1.55$ mm, and $g_1 = g_3 = w_0 = 1.17$ mm.

C. T-Shape Resonator Design

The product and ratio of C_1 and L_1 determine the resonant frequency and external quality factor, respectively. The required external quality factor of the first stage can be calculated by [6]

$$Q_{\rm e1} = Z_0\sqrt{\frac{C_1}{L_1}}. \qquad (2)$$

Inserting the previously calculated values, the required first-stage external quality factor is $Q_{\rm e1} = 59$.

The L_1C_1 resonator in Fig. 1(a) is realized by the T-shape resonator. $(l_1 + l_2)$ is almost half a wavelength at 5 GHz. By adjusting the lengths l_1 and l_2, we can reach different values of $Q_{\rm e1}$'s and keep the resonant frequency at 5 GHz simultaneously. When $l_1 = 10.91$ mm and $l_2 = 7.43$ mm, this results in $Q_{\rm e1} = 59$, which satisfies the requirement.

D. Synthesis of the Filtering Antenna

Now, we have these two resonators in hand. What we have to do next is put them all together. The value J of the admittance inverter is determined by the coupling strength. With narrower gap size g_2 for the structure in Fig. 1, we get tighter coupling between resonators so that the two poles repel one another, resulting in a wider bandwidth. By this principle, the appropriate bandwidth is obtained while $g_2 = 1.55$ mm.

Fig. 3 shows the simulated results. The solid lines are the reflection coefficient and broadside $(+z)$ direction antenna gain of the filtering antenna, which are obtained by full-wave simulation. The dash-dot-dash lines are those of an ordinary inset-fed microstrip antenna for comparison. The dashed line is the calculated reflection coefficient of the equivalent circuit [Fig. 1(b)].

The filtering antenna and the equivalent circuit have almost the identical reflection coefficient response—two transmission poles in passband, centered at 5 GHz, and 2 % fractional bandwidth. Compared to the response of the ordinary inset-fed microstrip antenna (dash-dot-dash line), the filtering antenna has twice wider bandwidth.

The radiation pattern of the filtering antenna is almost the same as that of the inset-fed microstrip antenna. The cross polarized radiation level is lower than -15 dBi in all directions throughout the band in interest. Referring to the antenna gain response in Fig. 3, the solid line is the broadside copolarized gain of the filtering antenna. The cross-polarized gain is negligible because there is a cross-polarized null in broadside $(+z)$ direction due to geometrical symmetry. There are two radiation nulls at 4.7 and 5.2 GHz, respectively, so it owns a sharp band skirt and good selectivity. The filtering antenna has a flat passband response because of the equal-ripple design. Moreover, the out-of-band antenna gain response is lower than -15 dBi, that is, a good out-of-band rejection is obtained. Compared to the gain response of the inset-fed microstrip antenna (dash-dot-dash line), the filtering antenna (solid line) sacrifices a little gain at the center frequency, but the passband response is flatter and the band-edge roll-off is much steeper.

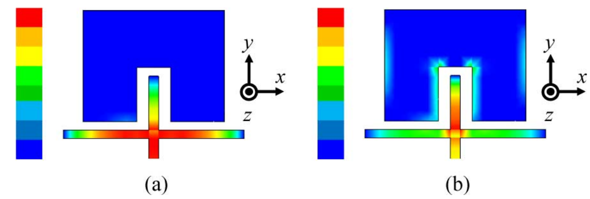

Fig. 4. Average current distributions at (a) 4.7 and (b) 5.2 GHz.

III. Discussions

In this section, discussions are on the broadside antenna nulls and the effect of geometrical symmetry.

A. Broadside Radiation Nulls

The open end of a microstrip line radiates. At 4.7 GHz (the resonant frequency of two quarter-wavelength microstrip arms extending in x-direction), referring to the current distribution in Fig. 4(a), the strongest currents concentrate on these two arms, but flow in opposite direction. Thus, the cross-polarized radiation cancels in broadside $(+z)$ direction within the whole frequency range. On the other hand, the copolarized radiation is contributed by the nonresonant microstrip section in y-direction. Since there is less current on this section, the radiation level would be very low at this frequency. In summary, there would be no radiation in broadside direction at the very resonant frequency, and it forms the radiation null.

The radiation null at 5.2 GHz is induced by the interaction between the U-shape patch and the inset section of the microstrip. Observing the current distribution in Fig. 4(b), there are two possible sources for copolarized radiation: One is the open-end of the feed microstrip, and the other one is upper and lower edges of the U-shape patch. These two radiation sources are 180° out of phase and tend to cancel each other's contribution. At a certain frequency, these two sources have the same radiation level and thus form the radiation null.

B. Symmetry

Here, discussion is on the geometrical symmetry and its effects on the broadside $(+z)$ direction gain. An asymmetric circuit is established as shown in Fig. 5. Compared to the symmetric structure in Fig. 1, the T-shape resonator is replaced by an L-shape one. The dimensions of the asymmetric microstrip antenna are given as follows: $l_1' = 9.6$ mm, $l_2 = 7.43$ mm, $w_0 = g_2 = g_3 = 1.17$ mm, $l_{\rm p} = 15.3$ mm, and $w_{\rm p} = 18$ mm.

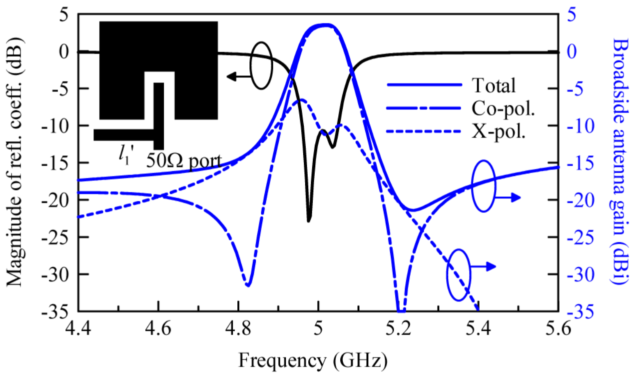

Fig. 5. Reflection coefficient and antenna gains in broadside $(+z)$ direction. (Solid line: total gain. Dash-dot-dash line: copol. Dashed line: x-pol.)

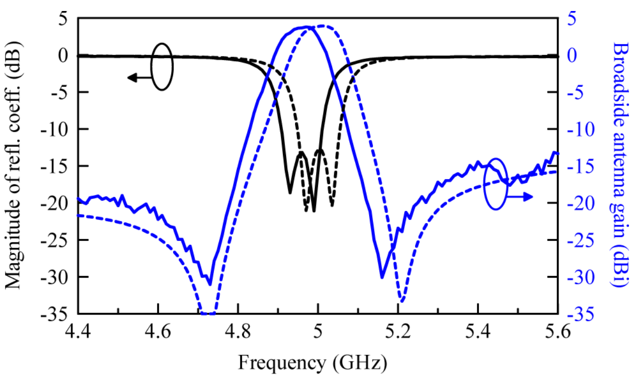

Fig. 6. Magnitude of reflection coefficients and antenna gains in broadside $(+z)$ direction. (Solid lines: measurement. Dashed lines: full-wave simulation.)

Only the section l_1 is shortened for keeping the same resonant frequency.

Fig. 5 also shows the full-wave simulated reflection coefficient and antenna gain in broadside $(+z)$ direction. The reflection coefficient and copolarized gain (dash-dot-dash line) has almost the same response as those in Fig. 3, only the radiation null at lower frequency is shifted from 4.7 to 4.8 GHz. The lower frequency null is caused by the resonance of the l_1 section, as discussed in Section III-A, so the frequency shifts as l_1 varies. The cross-polarized gain (dashed line) in Fig. 5 has a considerable level. The cross-polarized radiation mainly originates from the current flowing in x-direction on the L-shape resonator. Since one of the arms is removed, no cancellation occurs. The radiation nulls are buried because of the cross-polarized gain.

The symmetric filtering microstrip antenna in Fig. 1 solves the problem. Symmetry eliminates the cross-polarized gain, so the symmetric filtering microstrip antenna has better selectivity and purer polarization. In another point of view, for any antenna that is symmetric to the yz plane, the symmetric plane can be considered as an H-wall. On this wall, the boundary condition forbids the existence of the E-theta component. That is, there must be a cross-polarized radiation null in z-direction.

IV. Measurement Result

Fig. 6 shows the comparison between simulated and measured reflection coefficients and the total antenna gains in broadside $(+z)$ direction. The measured reflection coefficient (solid line) in Fig. 5 stays lower than -12 dB and has two transmission poles in passband as expected. The measured passband endures a slight shift about 30 MHz toward the lower frequency (that is, only 0.6% relative to the center frequency), which may stem from the variation of the dielectric constant of the substrate. Also shown in Fig. 6 are the measured copolarized broadside $(+z)$ direction antenna gain responses, which agree with the simulated one—flat passband response, two nulls, and low out-of-band radiation level. The measured cross-polarized component at broadside $(+z)$ direction is neglected because its level is much lower (20 dB and more).

V. Conclusion

A second-order filtering microstrip antenna with quasi-elliptic function antenna gain response has been presented. The design process follows the synthesis method of a second-order Chebyshev bandpass filter. By introducing two radiation nulls (or transmission zeros), we obtain a pair of steep skirt at band edges. The mechanisms of the broadside radiation nulls are introduced, by which the frequencies of the nulls can be designed by easily adjusting the parameters. The effect on symmetry for improving selectivity is also discussed. The full-wave simulation shows good agreement with experiments in reflection coefficient, antenna gain response, and the radiation patterns.

Compared to the ordinary inset-fed microstrip antenna, with little extra circuit areas, the proposed filtering antenna has almost twice wider bandwidth as well as good filtering abilities—that is, a flatter passband response and better frequency skirt selectivity. With the filtering microstrip antenna, most of the function for filtering the spurious signals can be achieved. The specification of the following filter, if any, can thus be greatly loosened. Because of the all-planar structure, it can be easily integrated in modern microwave circuit design, such as the prefiltering of the printed antenna array.

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