Novel Planar, Square-Shaped, Dielectric-Waveguide, Single-, and Dual-Mode Filters

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***Abstract*—This paper presents a novel concept of a square-shaped dielectric-waveguide resonator, which can easily realize a dielectric-waveguide cross-coupled filter or a dielectric-waveguide dual-mode filter, using the conventional printed circuit board (PCB) process. This planar dielectric-waveguide resonator has a higher Q value than a microstrip resonator. The physical mechanisms of both single- and dual-mode filters are elucidated. Some new coupling structures are developed for undergoing the PCB process. Filter design procedures are detailed. Design curves for the coupling coefficients of internal- and external-coupling structures are determined by full-wave finite-element-method electromagnetic calculations. Measurement results are highly consistent with theory for all trial filters. This study offers an effective means of producing low-cost high-performance planar circuit filters.**

***Index Terms*—Cross-coupled filter, dielectric-waveguide filter, printed circuit board (PCB) filter, square-shaped waveguide resonator, waveguide single- and dual-mode filter.**

I. Introduction

HIGH-PERFORMANCE microwave filters are important circuit components in wired or wireless communication, the military, science, and instrumentation applications. The performance of a waveguide filter is well known to be superior to that of a microstrip-, a strip-, or a coaxial-line filter, due to its higher resonator Q value and higher power-handling capability. Recently, two waveguide filters, namely, the dual-mode and the cross-coupled filters [1]–[3] or a combination of the two [4]–[6] has become increasingly important. A dual-mode filter consists of resonators with two degenerated modes that offer the advantage of reduced size and, sometimes, high-Q value [1], [7]. A conventional waveguide dual-mode filter [4], [5] employs the mode degeneration of a square or a circular waveguide. The polarizations of the E-fields of two degenerated waveguide modes are mutually perpendicular, hence, the term field-polarization degeneration [8]. The other important type of filter, namely, the cross-coupled filter, has also attracted much attention due to its flat group delay or quasi-elliptic response [2]–[6]. However, a conventional metallic waveguide dual-mode or cross-coupled filter is too large and too difficult to mass produce.

A dielectric waveguide can significantly reduce the size of a waveguide filter without greatly diminishing its performance. Konishi [9] described a series of dielectric waveguide filters of high dielectric-constant material. The filters in [9] are primarily from empty metallic waveguide filters, except in that an approximation to a magnetic wall can be formed at the dielectric–air interface if the dielectric constant is sufficiently high. This magnetic wall approximately halves the size of the filter. Sano and Miyashita [10] suggested a dielectric-waveguide filter made of high dielectric-constant material and inductive strip coupling. The filters in [9] and [10] both require firing at a high temperature. Tzuang *et al.* [11] described a dielectric waveguide filter with inductive iris coupling, suitable for the printed circuit board (PCB) process. However, the filters in [9]–[11] are neither dual mode, nor cross-coupled.

Awai *et al.* [12] proposed a dielectric-waveguide dual-mode filter that uses high-temperature-fired high-dielectric-constant material and an extra I/O board, but it is suited only to a two-pole filter and is difficult to match with the PCB process. Sano and Miyashita [13] posited another type of dielectric-waveguide dual-mode filter that employs high-temperature-fired dielectric material. The I/O circuit in [13] should be a coaxial line and is not a planar circuit. The filters in both [12] and [13] are three-dimensional structures and difficult to match with a planar circuit.

This investigation describes a dielectric-waveguide filter, which uses square-shaped resonators and is easily realized by the conventional PCB process. The proposed basic resonant cavity is a square PCB patch with continuously drilled plated through-holes, i.e., the side metal, as depicted in Fig. 1(a). The plated through-hole is commonly used in the PCB or low-temperature ceramic co-fired (LTCC) fabricating processes. For filter applications, this hole can be used to generate a waveguide channel for a stripline or a microstrip filter to enhance the performance of the filter, as proposed in [14], or it can be used in the sidewalls of a dielectric waveguide to realize a waveguide filter, as proposed in [15]. Here, the top view (but not the cross-sectional view) of the proposed resonator is a square. The square shape is such that the proposed resonator can be either a single- or dual-mode resonator.

The proposed square-shaped dielectric-waveguide resonator offers the following advantages. First, a square-shaped dielectric-waveguide resonator has a degenerated resonance of modes TE_{201} and TE_{102}. This mode degeneration can be used to realize a dual-mode filter. This mode degeneration, called field-distribution degeneration, is due to a field distribution different from that of a dual-mode waveguide resonator with so-called field-polarization degeneration. Therefore, new coupling structures must be specially developed. This proposed square-shaped dielectric-waveguide resonator shows its degenerated resonant modes in the first higher order mode rather

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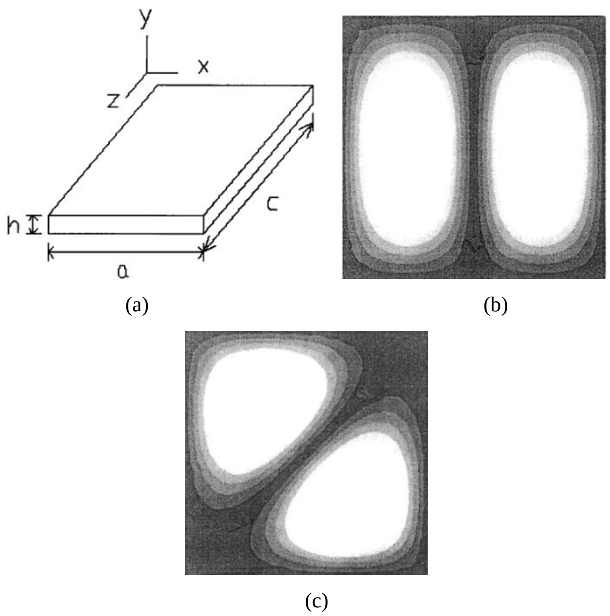


Fig. 1. Physical structure and field distributions of the dielectric-waveguide resonator. (a) Physical structure. (b) Normal mode of TE_{102}. (c) Diagonal mode of TE_{102}.

than the fundamental mode, as in the conventional waveguide dual-mode resonator. Measurement results indicate that the unloaded Q value of the first higher order mode exceeds that of the fundamental mode. The second advantage is that the fundamental mode (TE_{101}) of a square-shaped dielectric-waveguide resonator can easily support a cross-coupled filter due to its symmetrical field distribution. A filter with either flat group delay or quasi-elliptic response can be realized, using a newly proposed coupling structure, which will be discussed in Section II. The third advantage is that, although the shape of the proposed square-shaped dielectric-waveguide resonator is similar to that of a square microstrip-patch dual-mode resonator [16], the current distribution differs completely. The square microstrip patch has a very high current density at the edge of the patch that induces high conductor loss and low-Q value. The current density of a square-shaped dielectric-waveguide resonator is much more evenly distributed. Therefore, the conductor loss should be lower than that in a microstrip-patch dual-mode resonator. Furthermore, the radiation loss of a square-shaped dielectric-waveguide resonator is also markedly reduced by its closed structure.

More importantly, this novel planar single- and dual-mode filter can be manufactured by most popular PCB processes and is easily mass-produced. The metal sidewall of the resonator is fabricated by a sequence of continually drilled plated through-holes. The proposed square-shaped dielectric-waveguide filters have the advantages of a higher unloaded Q value, smaller circuit size, easier interfacing with other circuits on a PCB, and lower fabricating cost.

This study addresses the behaviors of the first three resonant modes TE_{101}, TE_{102}, and TE_{201} in depth. The fundamental mode (TE_{101}) is used in a single-mode filter. The TE_{102} and TE_{201} modes are two degenerated modes of a squared-shaped waveguide resonator that are suited to a dual-mode filter. The filters with Chebyshev responses are designed using either single- or dual-mode resonators. Some cross-coupling structures for the single-mode filter are also elucidated to provide appropriate magnitude and phase for the cross-coupling. The filters with a flat group delay or quasi-elliptic response with a cross-coupled structure and proper coupling magnitude and phase are investigated.

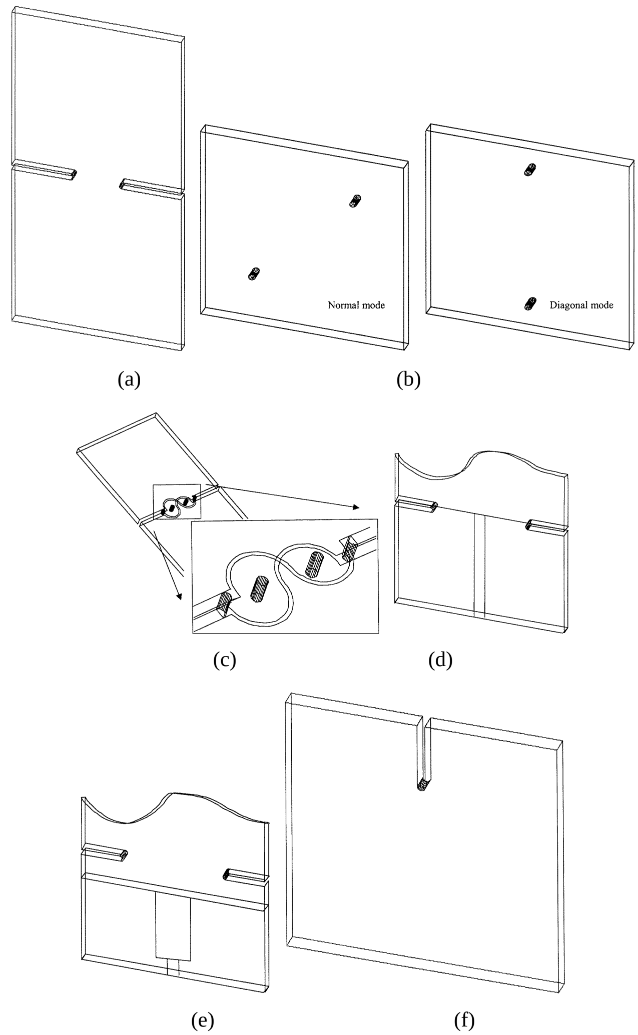


Fig. 2. Coupling structures and the frequency-tuning mechanism. (a) Inductive iris for inter-coupling and positive cross-coupling. (b) Two plated through-holes for intra-coupling. (c) Structure for negative cross coupling. (d) Microstrip direct-coupled-type external coupling. (e) Quarter-wave transformer-type external coupling. (f) Slot for frequency tuning.

II. Resonator, Coupling Structures, and Frequency-Tuning Mechanisms

A. Square-Shaped Dielectric-Waveguide Resonator

Equation (1) defines the resonant frequency of the $\mathrm{TE}_{\mathrm{nml}}$ mode of a dielectric-waveguide resonator as follows and is as shown in Fig. 1(a):

$$f_{r_{\mathrm{nml}}} = \frac{v_c}{2\sqrt{\varepsilon_r}} \sqrt{\left(\frac{n}{a}\right)^2 + \left(\frac{m}{h}\right)^2 + \left(\frac{l}{c}\right)^2} \tag{1}$$

where n, m, and l are the mode indexes of the $\mathrm{TE}_{\mathrm{nml}}$ mode, a, h, and c are physical dimensions of the resonator along corre-

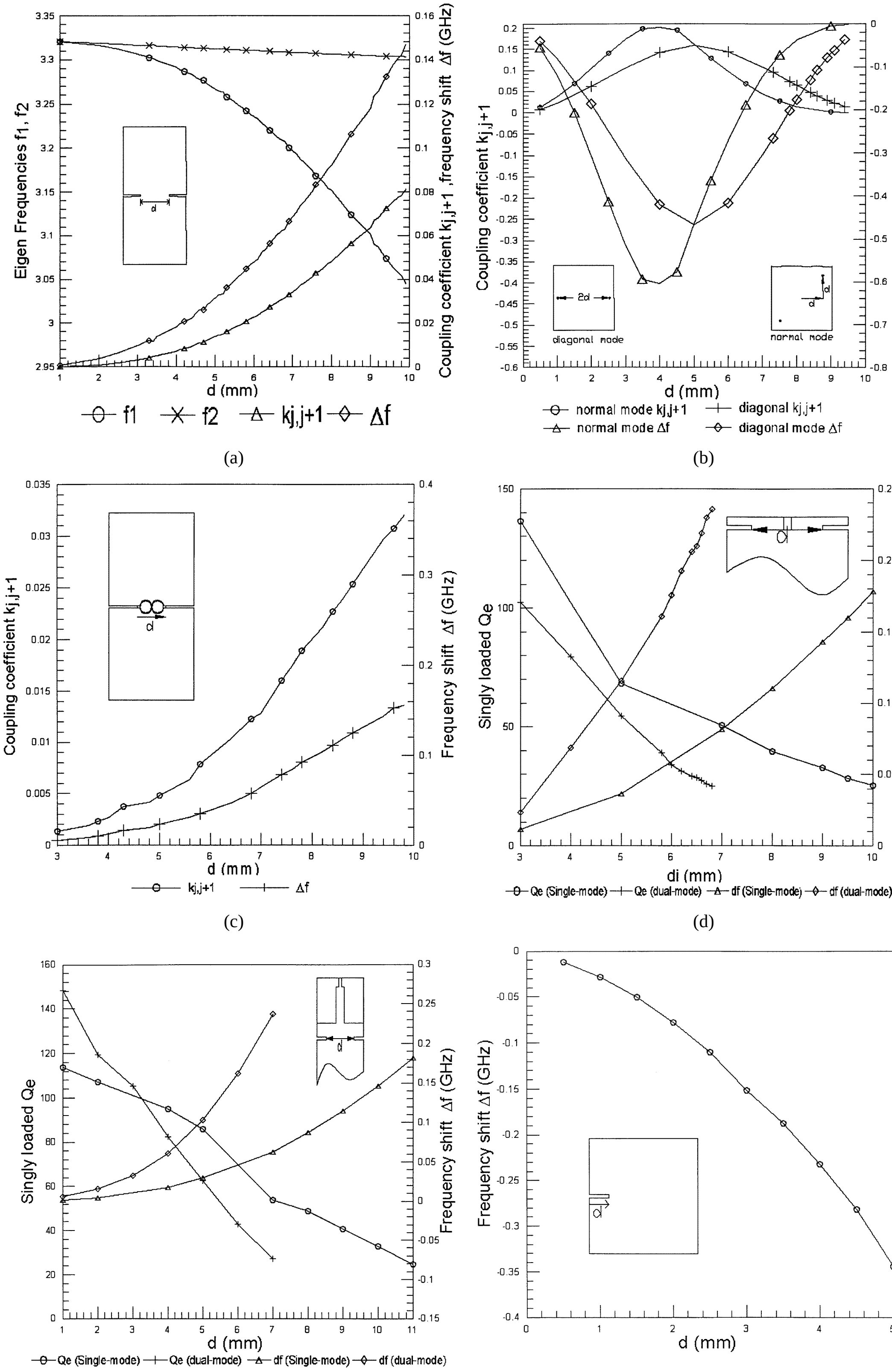


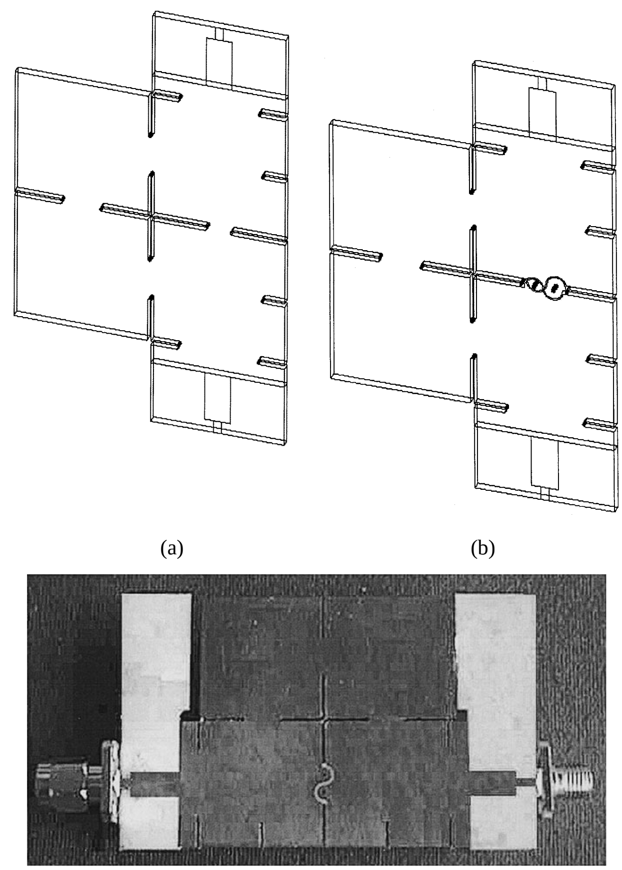
Fig. 3. Design curves. (a) Inter-coupling between two different cavities. (b) Intra-coupling between two degenerated modes. (c) Cross-coupling with inverse phase. (d) External coupling of the microstrip direct-coupled type. (e) External coupling of the quarter-wave transformer type. (f) Frequency tuning versus depth of the slot.

sponding axes in Fig. 1(a), ε_r is the relative dielectric constant of the dielectric substrate, and v_c is the velocity of light in free space.

The physical parameters of the proposed square-shaped dielectric-waveguide cavity are chosen as $a = c = 20$ mm, $h = 1.27$ mm, and $\varepsilon_r = 10.2$ to confirm the theory. Ac-

TABLE I
THREE FOUR-POLE FILTERS (FBW: 0.03, PASSBAND RETURN LOSS: 20 dB)

Type	Pole position	Q_e	K_{12}	K_{23}	K_{34}	K_{14}
Chebyshev	N.A.	23.76	3.24×10^{-2}	2.38×10^{-2}	3.24×10^{-2}	N.A.
Flat group delay	(1.05,0)	28.0	2.9×10^{-2}	1.68×10^{-2}	2.9×10^{-2}	1.11×10^{-2}
Quasi-elliptic	(0,1.5)	31.25	2.46×10^{-2}	2.55×10^{-2}	2.46×10^{-2}	-1.12×10^{-2}



(a) (b) (c)

Fig. 4. Circuit layouts and photograph of the four-pole single-mode filters. (a) Circuit layout of the flat group-delay response filter. (b) Circuit layout of the quasi-elliptic response filter. (c) Photograph of the quasi-elliptic response filter.

cording to (1), this resonator resonates at $\mathrm{fr}_{101} = 3.32$ GHz and $\mathrm{fr}_{102} = \mathrm{fr}_{201} = 5.25$ GHz.

The distribution of the electric-field strength of the TE_{101} mode is that of a drum that symmetrically moves up and down in the y-direction.

For mode TE_{102} (or TE_{201}), the field distribution is divided into two halves, of which the electric fields are of the same magnitude, but antiphase. In general, two basis functions describe the distribution of the electric-field magnitude of mode TE_{102} (or TE_{201}); the first is divided along the center line and the second is divided along the diagonal line, as depicted in Fig. 1(b) and (c), respectively. The fields in Fig. 1(c) can be represented as the sum and difference of the fields in Fig. 1(b) and vice versa [7]. The field distribution in Fig. 1(b) is called the normal mode and that in Fig. 1(c) is the diagonal mode.

The measured unloaded Q values are 270 for mode TE_{101} and 360 for mode TE_{102} (or TE_{201}). The substrate used here is a Rogers RT/Duroid 6010 with a dielectric constant of 10.2, dielectric thickness of 1.27 mm, and dielectric loss tangent of 0.0023. The Q values are not high primarily because of the relatively high loss tangent of the substrate used. According to the data, the simulated Q's are 295 and 305 for TE_{101} and TE_{102} modes, respectively. Setting the loss tangent one order of magnitude lower (i.e., at 0.00023) could yield simulated unloaded Q up to 773 for mode TE_{101} and 832 for mode TE_{102}.

B. Coupling Structures

Coupling involves both internal and external coupling to form a bandpass filter internal and external coupling, which are used in the proposed filters and are described as follows.

1) Internal-Coupling Structures: Internal coupling can be classified into three types, namely: 1) inter-coupling; 2) intra-coupling; and 3) cross coupling. Inter-coupling and intra-coupling are couplings between adjacent cavities in the main coupling path. Inter-coupling is the coupling between two physically different cavities, and intra-coupling is that between degenerated modes [17]. Cross-coupling is between nonadjacent cavities.

Fig. 2(a) displays an inductive coupling iris. This inductive iris is used as the inter-coupling structure for the proposed single-mode filter and a dual-mode filter of order above two.

Intra-coupling in a single cavity is achieved by introducing a perturbation element such as tuning screws, corner cuts [18], slots, etc. in the plane orthogonal to the resonator. Here, two plated through-holes, shown in Fig. 2(b), are proposed as the intra-coupling structure because they undergo perfectly with the PCB process. The proposed plated through-hole intra-coupling is similar to the concept outlined in [18]. These plated through-holes constitute an inductive coupling.

The two types of cross-couplings correspond to positive and negative polarity of the coupling phase. An iris-type cross-coupled structure is a positive cross-coupling and should induce real axis poles for a flat group-delay response because the main couplings of a single-mode filter are all of the inductive iris type. A negative cross-coupling structure should be used if a quasi-elliptic response is required. The negative cross-coupling of the proposed resonator is more difficult to achieve than that of a conventional waveguide dual-mode filter with field-polarization degeneration. Walker and Hunter described a dielectric-loaded waveguide filter [19] for which an inverted coupling loop is proposed to realize a negative coupling. However, the structure proposed in [19] is not applicable to our case due to a different process of fabrication. Here, a structure including iris coupling and a balanced microstrip line with a pair of plated through-holes is presented to twist the phase of the signal, as shown in Fig. 2(c). This structure perfectly fits the PCB process and provides accurate antiphase coupling with an inductive iris.

2) External-Coupling Structures: The coupling structure between the feeding microstrip line and proposed square-shaped waveguide cavity is extremely important because it allows the filter to be fabricated with other microwave circuits on the same PCB using a single process. In [11], an interface is achieved by tapering the microstrip line to a dielectric waveguide, but the transition area is too large. Here, two types of microstrip-line interfaces are proposed, including the direct-coupled and quarter-wave transformer types. The direct-coupled type, illustrated in Fig. 2(d), directly

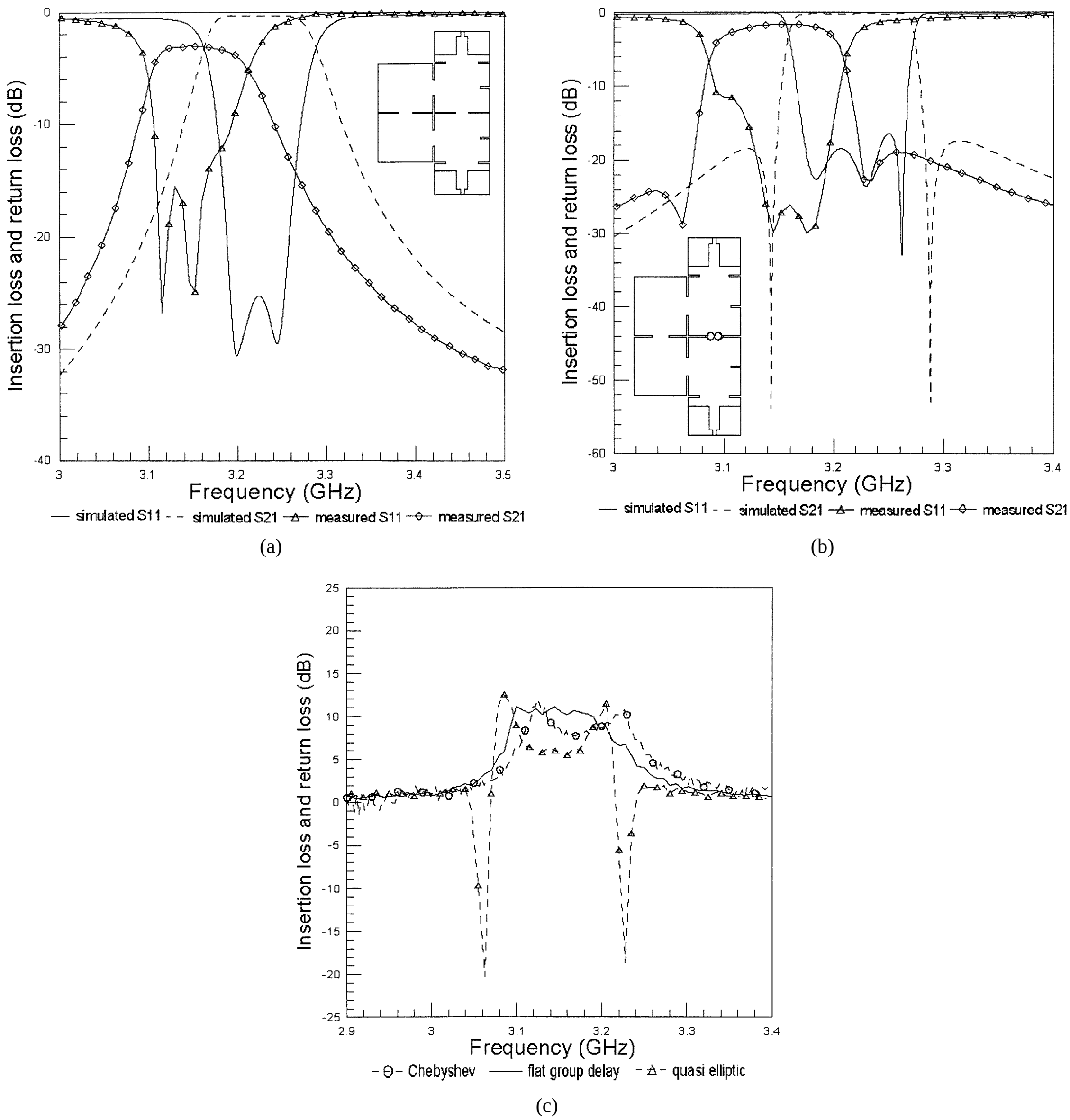


Fig. 5. Simulation and measurement performances of the single-mode filters. (a) Flat group-delay response. (b) Quasi-elliptic response. (c) Comparisons of measurement group-delay performances.

connects the 50-Ω microstrip line to the dielectric-waveguide resonator through an inductive window. Varying the width of the inductive window changes the external-coupling coefficient accordingly. The quarter-wave transformer type, shown in Fig. 2(e), is a microstrip-to-dielectric-waveguide transition that uses a quarter-wave microstrip-line impedance transformer to transform the 50-Ω microstrip-line impedance to dielectric-waveguide impedance. These two interface circuits can be used in all proposed filters.

As mentioned earlier, the field distribution of the degenerated modes of the proposed cavity can be either normal or diagonal. If the external-coupling position is placed at the center of the edge of the square, then the excited field will be in the normal mode, and the plated through-holes for intra-coupling should be placed along the diagonal line of the square. If, however, the external-coupling position is at the corner of the square, then the excited field will be in the diagonal mode, and the plated through-holes for intra-coupling should be placed on the center line of the square. Here, both external-coupling positions will be considered for the two-pole dual-mode filter case.

C. Frequency-Tuning Mechanism

Coupling between two resonators causes the resonant frequency to split into f_1 and f_2, where f_1 and f_2 are the lower and higher resonant frequencies, respectively. The mean frequency of these coupled-cavities, which equals $(f_1+f_2)/2$, differs from the original single-cavity resonant frequency, which is determined by (1). Similarly, external coupling will also cause the resonant frequency to shift. The shifting of the central frequency can be either upward or downward, and is governed by the coupling structures. This resonant frequency shift must be tuned out.

Two methods are proposed here to compensate for the shifting of the central frequency. The first method is to shrink or expand the size of the resonator by a factor equal to the frequency shift divided by the central frequency. Section IV will apply this method to realize the duplexer, which includes two connected dual-mode filters. The second method employs a slot, as shown in Fig. 2(f), to reduce the effective size of the resonator. All of the single- and dual-mode filters proposed here use this frequency-tuning mechanism to tune out the frequency shift.

III. Filter Design

This section considers filter design by synthesis based on three-dimensional (3-D) finite-element method (FEM) electromagnetic (EM) calculation. The HFSS EM solver of Ansoft Inc., Pittsburgh, PA, is used to perform the calculations. The filter is designed as follows.

The first step is to obtain the design curves by FEM calculation.

1) For a specific dielectric substrate (with known dielectric constant and thickness), determine the dimensions of the dielectric-waveguide resonator by (1) for the desired resonant frequency f_{c0}.
2) Solve the eigenvalue problem for each of the internal-coupling structures as described in Section II. Two eigenfrequencies f_1 and f_2 should be obtained near the central frequency f_{c0}, which correspond to the two eigenmodes of the coupled resonators. The coupling coefficient $k_{j,j+1}$ between the jth and $j+1$th resonators can be determined by the following equation:

$$k_{j,j+1} = \frac{f_1^2 - f_2^2}{f_1^2 + f_2^2}. \tag{2}$$

The coupling dictates that the central frequency $f_{cj,j+1}$ becomes $(f_1+f_2)/2$. The frequency shift $\Delta f_{j,j+1}$, which equals $f_{c0} - f_{cj,j+1}$, should be tuned out by the proposed frequency-tuning slot. Fig. 3(a)–(c) plots the design curves for $k_{j,j+1}$ and $\Delta f_{j,j+1}$.
3) The strength of external coupling can be represented by an external quality factor Q_e, which is determined by the loaded resonator method. Again, the frequency shift Δf_{ext} occurs due to the external coupling. Δf_{ext} must also be tuned out.

Awai *et al.* claimed in [12] that external coupling is typically impacted by the intra-coupling coefficient in a dual-mode ring resonator. This claim also holds for the case considered here; therefore, a method to adjust the external coupling is required. An accurate Q_e is obtained by first drilling two intra-coupling plated through-holes into a dual-mode cavity. The positions of the intra-coupling plated through-holes are determined in advance by solving the eigenvalue problem. The external quality factor Q_e can then be determined using (3) [20] as follows:

$$\left(\frac{1}{Q_e}\right)^2 = k_{1,2e}^2 - k_{1,2d}^2 \tag{3}$$

where the coupling coefficient $k_{1,2e}$ is that of the coupled cavities without external coupling and the coupling coefficient $k_{1,2d}$ is that of the coupled cavities with external coupling.

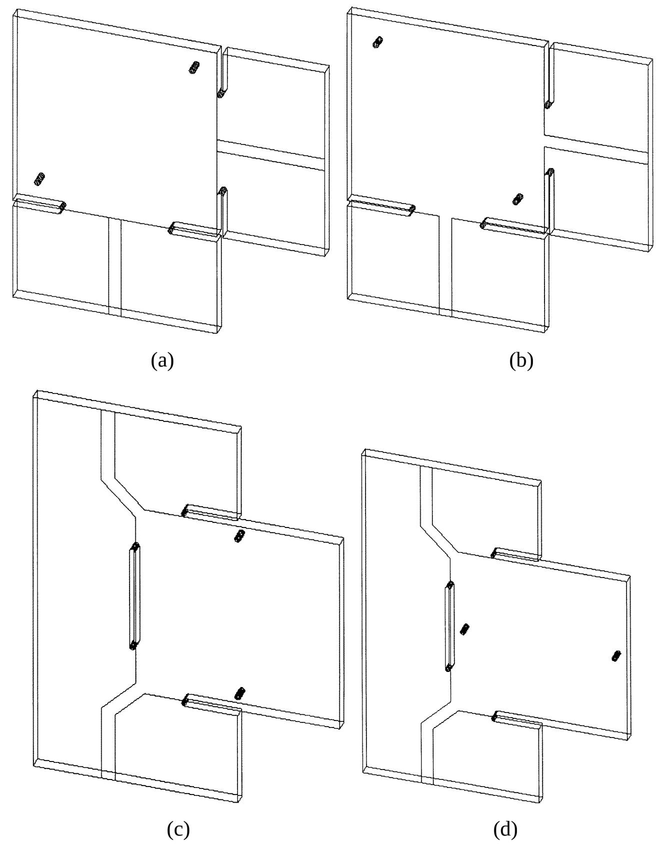


Fig. 6. Circuit layouts of the two-pole dual-mode filter. (a) Filter with normal-mode field distribution and the first kind of relative intra-coupling to the I/O position. (b) Filter with normal-mode field distribution and the second kind of relative intra-coupling to the I/O position. (c) Filter with diagonal-mode field distribution and the first kind of relative intra-coupling to the I/O position. (d) Filter with diagonal-mode field distribution and the second kind of relative intra-coupling to the I/O position.

Fig. 3(d) and (e) displays the design curves for Q_e and Δf_{ext} for the direct coupled and transformer types, respectively.
4) Frequency shifts due to the internal and external coupling should be tuned out. The frequency shift can be either positive (upward) or negative (downward). Summing all the frequency shift terms for each cavity, and using the frequency-tuning mechanism elucidated above, the frequency shift can be tuned out by the appropriate tuning depth. Fig. 3(f) shows the curve of frequency shift against tuning depth.

The second step is to obtain the coupling values from the low-pass prototype filter.

Two kinds of low-pass prototype filters are used in this study. The first kind is the conventional Chebyshev or Butterworth low-pass prototype filter, whose coupling coefficients can be determined from the prototype values following the steps given in [21]. The second kind is the generalized Chebyshev low-pass prototype filter. In the cases considered here, four resonators are used to realize all generalized Chebyshev bandpass filters. The coupling coefficients of these generalized Chebyshev bandpass filters can be established by transforming the prototype values according to the steps described in [22]. The coupling coefficients determined by the method in [22] include cross-coupling between resonators 1 and 4; the polarity of the cross-coupling

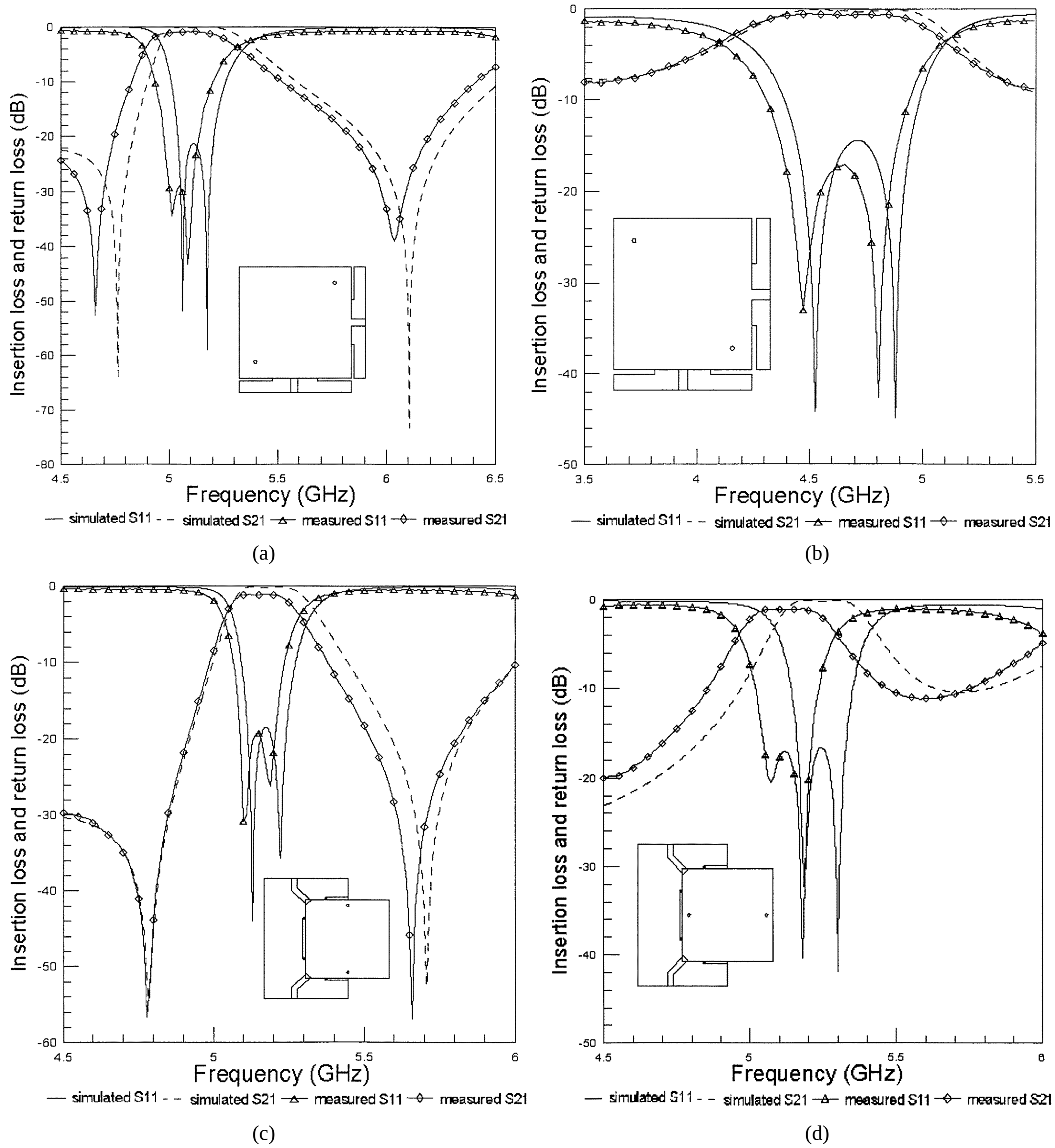


Fig. 7. Simulation and measurement performances of the filters in Fig. 6. (a) Performances of the filter in Fig. 6(a). (b) Performances of the filter in Fig. 6(b). (c) Performance of the filter in Fig. 6(c). (d) Performances of the filter in Fig. 6(d).

is also important because it governs whether the response is quasi-elliptic or a flat group delay.

IV. Simulated and Measured Results Pertaining to the Example Filters

Some example filters are designed and fabricated to prove the validity of the theory following the design described in Section III. They are described as follows. Three of the four-pole single-mode filters are fabricated to compare various types of responses, namely, the conventional Chebyshev response, quasi-elliptic response, and flat group-delay response. Four of the two-pole dual-mode filters are realized corresponding to two kinds of degenerated basis functions and two kinds of related intra-coupling to feeding structure positions. A four-pole dual-mode filter is realized to establish the feasibility of the proposed dual-mode filter with large numbers of poles. Finally, a diplexer is realized by connecting two of the two-pole dual-mode filters using a microstrip T-junction.

A. Four-Pole Single-Mode Filters

Three four-pole single-mode filters with the design parameters listed in Table I are fabricated. Fig. 4 presents the circuit layouts for positively and negatively cross-coupled filters, as well as a photograph of the latter. Fig. 5 presents the simulated and measured performances of the filters corresponding to Fig. 4. The results are very consistent. The filter with the quasi-elliptic response has a newly proposed negative-phase cross-coupling structure, as depicted in Fig. 4(b). This coupling structure can be very easily realized by the conventional PCB process.

B. Two-Pole Dual-Mode Filters

The two-pole dual-mode filter is the most basic dual-mode filters. Nevertheless, they exhibit much interesting physical behavior.

As stated in Section II, the proposed dual-mode resonator has two types of basis functions, namely, the normal and diagonal modes, both of which can realize a dual-mode filter according to the position of the feeding point. Fig. 6 depicts the layouts of the two types of two-pole dual-mode filters.

Another interesting phenomenon is observed. According to the theory of [1] and [23], the two-pole dual-mode filter responds in two different ways. The first response is similar to a quasi-elliptic response with transmission zeros near the passband due to the parasitic source to load coupling. The second response is similar to the flat group-delay response, which is caused in the same way. Whether the filter exhibits the first or second kind of response depends on the relative position of the two intra-coupling holes. For filters with the first kind of response, the input and output feedings are on the same side of the two intra-coupling holes, as depicted in Fig. 6(a) and (c). However, for filters with the second kind of response, the feedings are on the opposite side of the two intra-coupling holes, as depicted in Fig. 6(b) and (d). The response of the filters (positions of transmission zeros or flatness of the group delay) is not controllable because the parasitic coupling described here cannot be easily controlled. Fig. 7 displays the calculated and measured results for the filters in Fig. 6. They match very closely.

C. Four-Pole Dual-Mode Filter

A four-pole dual-mode filter is realized to demonstrate the feasibility of realizing a dual-mode filter with a large number of poles. The central frequency is 5.25 GHz, the fractional bandwidth is 0.03, and the passband ripple is 0.01 dB.

Fig. 8 presents the circuit layout and a photograph of the four-pole dual-mode filter. Fig. 9 presents the simulated and measured results.

D. Diplexer

The microstrip direct feed-type external-coupling structure is very compact and is suitable for realizing a diplexer. A diplexer can be realized simply by using a microstrip T-junction with two of the proposed single- or dual-mode filters. Here, two dual-mode two-pole filters are selected to realize the proposed diplexer. The central frequencies of the two passbands are 5.25 and 5.57 GHz, the fractional bandwidth of each passband is 0.03, and the passband ripple is 0.01 dB for each filter.

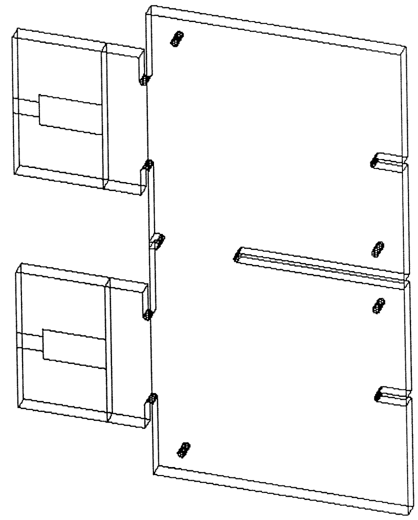
Fig. 10 shows the physical layout and a photograph of the diplexer. Fig. 11 gives the simulated and measured results.

E. Discussion

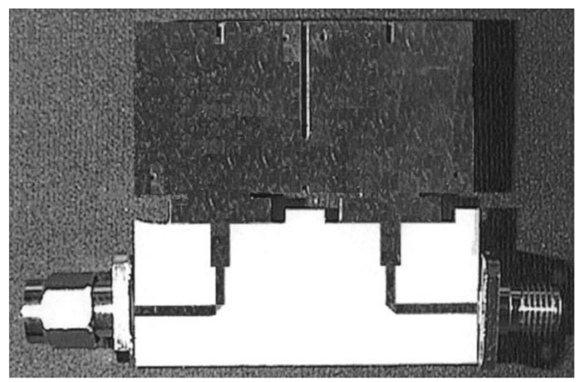
The measured unloaded Q value of the proposed square-shaped dielectric-waveguide resonator is 270 for single-mode resonance and 360 for dual-mode resonance. This value is rather high for a planar circuit cavity and, thus, the filter shows very low passband insertion loss.

The measured central frequencies of all of the filters and the diplexer show drifts to lower frequencies. The primary reason for this frequency drift is the error in the substrate parameters. Consider the Chebyshev filter described in the first paragraph of this section as an example. According to the vendor's data sheet, the dielectric constant of the substrate is 10.2. However, no frequency drift is possible if the dielectric constant of the substrate is set to 10.4 in the simulation. Fig. 12 gives the simulated and measured results for the four-pole single-mode Chebyshev filter.

The passband return loss exhibits a little degradation in some filters because of the fabrication tolerance of the PCB process. For example, the diameter of the drilled hole is not as expected.



(a)



(b)

Fig. 8. Circuit layout and photograph of the four-pole dual-mode filter. (a) Circuit layout. (b) Photograph.

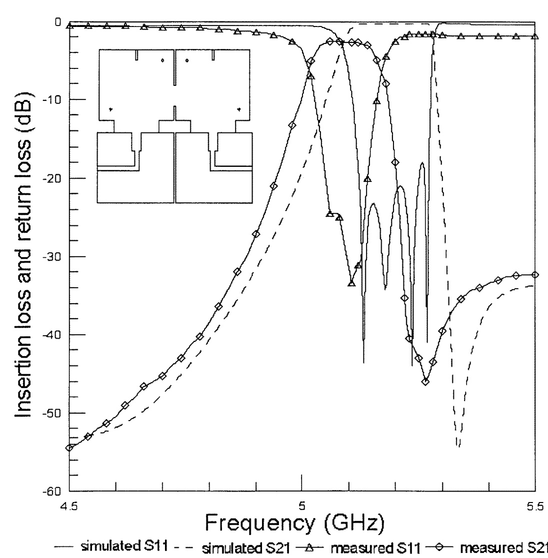
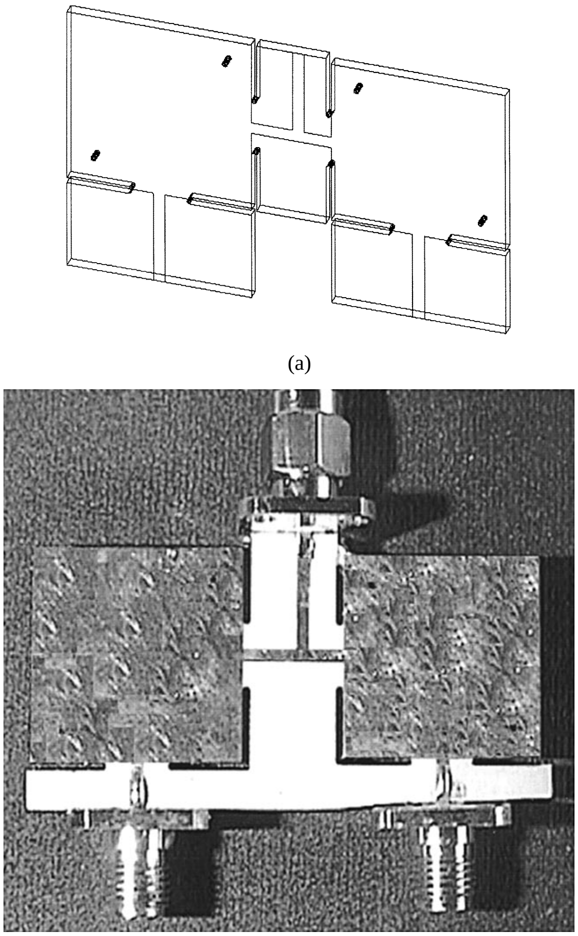


Fig. 9. Simulation and measurement performances of the filter in Fig. 8.



(a)

(b)

Fig. 10. Circuit layout and photograph of the diplexer. (a) Circuit layout. (b) Photograph.

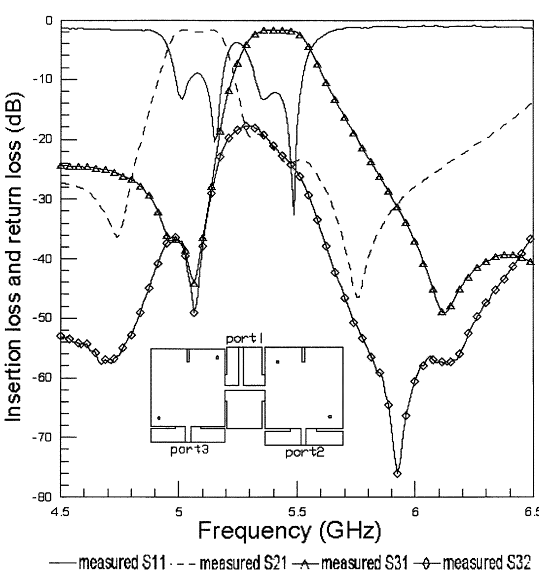


Fig. 11. Simulation and measurement performances of the diplexer in Fig. 10.

The designed diameter is 0.6 mm, whereas the measured diameter is 0.8 mm. The process, used to realize all of the filters discussed in this study, is the most popular process by which PCB

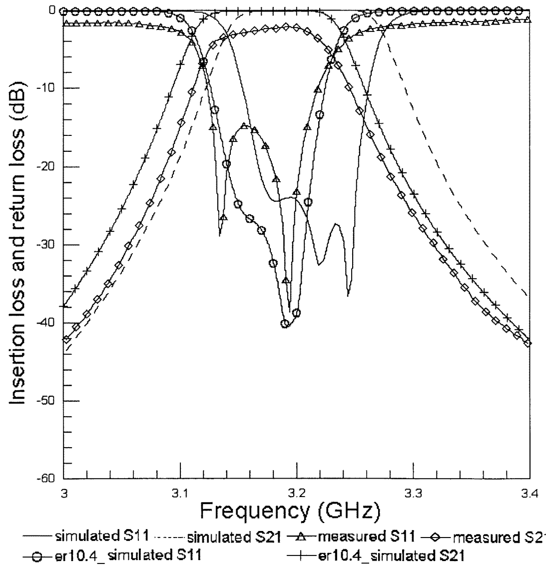


Fig. 12. Simulation results with a different substrate dielectric constant. Comparing them to the measured response of the four-pole single-mode Chebyshev filter.

circuits are mass produced. The PCB process may be specifically tuned for our filters at a higher cost to yield better performance.

V. CONCLUSIONS

This paper has presented a square-shaped dielectric-waveguide resonator, which can realize dual-mode and cross-coupled filters. The measured passband insertion loss was low because the unloaded resonator Q values were quite high. The design procedures have been detailed. Some example filters have been realized with performances correlated well with simulation results.

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