Transmission-Line Filters With Capacitively Loaded Coupled Lines

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***Abstract*—Coupled lines with loads at one end, which can create tunable transmission zeros, are studied in this paper. The equation for the transmission zeros is derived from the analysis of even- and odd-mode excitations. Based on this equation, coupled lines with different loads are analyzed and the rules for controlling the transmission-zero frequency are given. The structures are used in the designs of several second-order filters and they are experimentally verified. The use of a skew-symmetric (0°) feed structure in these filters is also discussed and an example is given.**

***Index Terms*—Coupled transmission lines, distributed parameter filters, microwave circuits, transmission-line resonators.**

I. Introduction

THE design procedures of coupled-resonator filters are classic. First, resonators at the design frequency are selected. The couplings between resonators are then designed analytically or manually adjusted to yield the right coupling coefficients. Finally, the input and output tapping points are selected for proper loaded Q. These procedures are well known and can easily be found in the literature [1]–[3]. One way of realization the couplings between resonators are by using coupled lines [4]–[6]. Usually, the lengths of coupled-line sections are a quarter-wavelength long. On the other hand, loaded coupled lines, which are physically less than a quarter-wavelength long, are used in the design of combline filters [7]. The coupled lines in these filters are parts of the resonators and they are also used for establishing the needed couplings. Filter design using coupled lines have been thoroughly researched in these two aspects. However, the transmission zeros created by these coupled-line sections are not yet fully studied. By properly design, these zeros are useful in the rejection of strong interference in the stopband.

In this paper, coupled lines with loads at one end, which can create one or more transmission zeros, are studied. The even- and odd-mode excitations of this symmetric two-port network are analyzed and the equation for transmission zeros is derived. Different loading conditions at one end of coupled lines, including open circuited, short circuited, capacitively loaded, and inductively loaded are studied. The condition and tunability of the created transmission zero for each type of loads are discussed. To demonstrate the applications, several second-order filters using coupled lines with capacitive loads at one end are designed. Finally, in order to show the improvement in the shape factor with a skew-symmetric feed structure, the feed topology of one of designed filters is modified and re-fabricated. The analytic, simulated, and experimental results of all designed filters are then compared.

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II. Transmission-Zero Conditions for Coupled Lines With Different Loads

In a filter design, it will be very helpful if extra transmission zeros can be created without sacrificing the passband response. For example, the extra transmission zeros can be tuned to reject the possible interferences and to improve the stopband rejection of a filter. Furthermore, a low-order filter, which has a smaller circuit size and a lower insertion loss, may meet the stopband rejection specifications with the help of extra transmission zeros. Therefore, it is necessary to study the relations between the extra transmission-zero position and other circuit parameters.

The circuit of coupled lines with loads Z_L at one end is shown in Fig. 1(a). This kind of structure is widely used to provide the necessary coupling between resonators of transmission-line filters. However, its transmission-zero condition is not yet fully studied, especially for inhomogeneous coupled lines. The circuit is a symmetric two-port network and can be analyzed by the classical method of even- and odd-mode excitations, as shown in Fig. 1(b). The even- and odd-mode input impedances are given as

$$Z_{ine} = Z_{0e} \cdot \frac{Z_L + jZ_{0e}\tan\theta_e}{Z_{0e} + jZ_L\tan\theta_e} \tag{1}$$

and

$$Z_{ino} = Z_{0o} \cdot \frac{Z_L + jZ_{0o}\tan\theta_o}{Z_{0o} + jZ_L\tan\theta_o} \tag{2}$$

where Z_{0e} and Z_{0e} are the characteristic impedances of the coupled lines, θ_e and θ_o are the electrical lengths for even- and odd-mode excitations, respectively. S_{21} could be found by the superposition of the even and odd modes in Fig. 1(b) and is given by

$$\begin{aligned} S_{21} &= \frac{\Gamma_e - \Gamma_o}{2} \\ &= \frac{Z_o(Z_{ine} - Z_{ino})}{(Z_{ine} + Z_o)(Z_{ino} + Z_o)} \\ &= \frac{1}{(Z_{ine}/Z_o + 1)} - \frac{1}{(Z_{ino}/Z_o + 1)} \end{aligned} \tag{3}$$

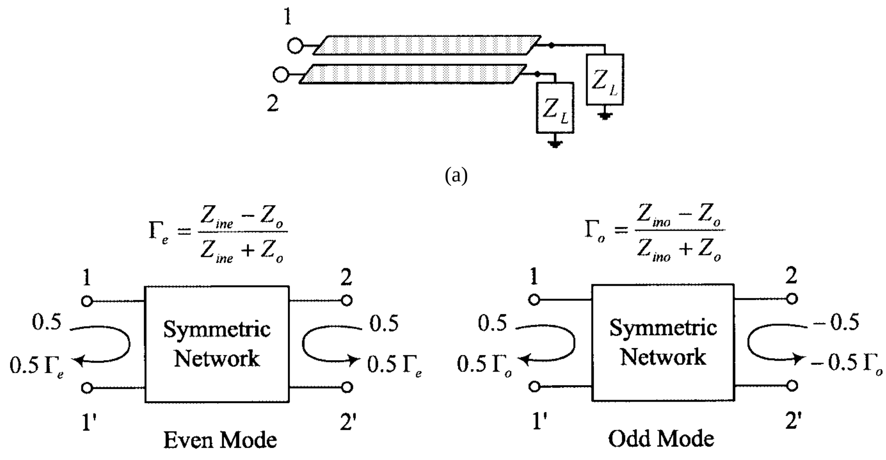


Fig. 1. (a) Coupled lines with loads at one end. (b) Even- and odd-mode analysis of a symmetric two-port network.

where Γ_e and Γ_o are the reflection coefficients for even- and odd-mode excitations and Z_0 is the system reference impedance. Therefore, the necessary and sufficient conditions for a transmission zero ($S_{21} = 0$) is $Z_{ine} = Z_{ino}$. This means

$$Z_{0e} \cdot \frac{Z_L + jZ_{0e} \tan\theta_e}{Z_{0e} + jZ_L \tan\theta_e} = Z_{0o} \cdot \frac{Z_L + jZ_{0o} \tan\theta_o}{Z_{0o} + jZ_L \tan\theta_o}. \tag{4}$$

The circuit with different loads Z_L will be considered and analyzed hereinafter to characterize the created transmission zeros. Although there could be many solutions to this equation, only the lowest nonzero one in frequencies will be studied in this research. The other solutions of the equation could certainly be solved numerically for studying the higher order zeros, but they will not be discussed in this paper.

A. Open-Circuit Termination

To begin with, coupled lines with open circuits at one end are studied. By applying the condition $Z_L = \infty$ to (4), the equation for the transmission zero is reduced to

$$Z_{0o} \cot\theta_o = Z_{0e} \cot\theta_e. \tag{5}$$

For coupled striplines, $\theta_o = \theta_e$, a transmission zero occurs when $\theta_o = \theta_e = \pi/2$. This result concurs with the statement in [8]. For coupled microstrip lines used in filter design, however, the created transmission zero would be at a frequency when $(\theta_o + \theta_e)/2 < \pi/2$ because Z_{0e} and θ_e are larger than Z_{0o} and θ_o, respectively. The transmission zero occurs no longer at the frequency when the average electrical length is equal to $\pi/2$. Fig. 2 is drawn based on solving (5) numerically and shows the relations between the impedance ratio Z_{0e}/Z_{0o}, the odd- and even-mode velocity ratio ν_o/ν_e, and the average electrical length $\theta \equiv (\theta_e + \theta_o)/2$ of coupled microstrip lines when a transmission zero is created. It is clear that, for a given impedance ratio, the velocity ratio must be less than a certain value if a transmission zero is desired. Moreover, this maximum value of the velocity ratio will be reduced if the impedance ratio is decreased. In the circumstance that the impedance ratio is fixed, the transmission-zero position can be tuned to be at a lower frequency by increasing the velocity ratio. Furthermore, when the velocity ratio is fixed, the lower the impedance ratio, the shorter the average electrical length for a transmission zero. In other words, the created transmission zero will move to a lower frequency.

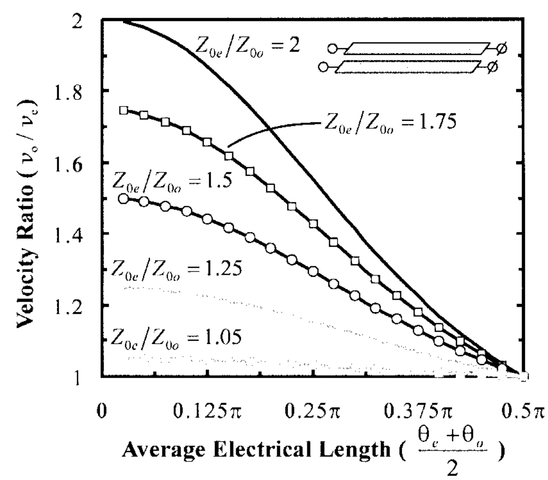


Fig. 2. Transmission-zero conditions for coupled lines with open circuits at one end.

B. Short-Circuit Termination

When one end of coupled lines is shorted to ground, the transmission-zero condition can be found by applying $Z_L = 0$ to (4) and the result is

$$Z_{0e} \cot\theta_o = Z_{0o} \cot\theta_e. \tag{6}$$

From this equation, Fig. 3 is drawn to illustrate the relations between the average electrical length, velocity ratio, and impedance ratio. It is found that if transmission zeros are needed, the velocity ratio should be smaller than a certain value when the impedance ratio is fixed. Besides, two transmission zeros, instead of one, can be created, except the circumstance that the velocity ratio is equal to the upper limit (under this condition, only one transmission zero is found). Moreover, it is clear that the larger the impedance ratio, the higher of the upper limit of the velocity ratio. This figure also shows that when the velocity ratio is fixed and the impedance ratio is reduced, the two created transmission zeros will move away from each

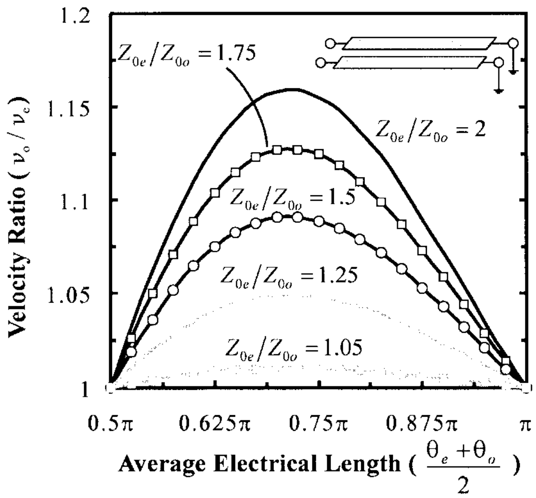


Fig. 3. Transmission-zero conditions for coupled lines with one end shorted to ground.

other and approach to the frequencies that the average electrical length is equal to $\pi/2$ and π. Obviously, this is not the only way to move the transmission zeros. When the impedance ratio is given, similar effect can be found by lowering the velocity ratio. In the case of $\nu_o/\nu_e = 1$ (coupled striplines), Fig. 3 shows that the transmission zeros are at the frequencies when $\theta_e = \theta_o = \pi/2$ and $\theta_e = \theta_o = \pi$. This also agrees with the discussion in [8]. However, for coupled microstrip lines $(\nu_o/\nu_e \neq 1)$, the transmission zeros are no longer at these two frequencies.

C. Capacitive Termination

Next, consider capacitive loads ($Z_L = -jX_C$), (4) can be rewritten as

$$\left[Z_{0e}^2 Z_{0o} + X_C^2 Z_{0o}\right] \cot\theta_o - X_C Z_{0o}^2 = \left[Z_{0e} Z_{0o}^2 + X_C^2 Z_{0e}\right] \cot\theta_e - X_C Z_{0e}^2 \quad (7)$$

which is rather complicated. Two figures are plotted according to this equation. In Fig. 4(a), the impedance ratio Z_{0e}/Z_{0o} is set to be 2.0. Transmission-zero conditions due to different even- and odd-mode impedances ($Z_{0e} = 100\ \Omega$, $Z_{0o} = 50\ \Omega$ and $Z_{0e} = 50\ \Omega$, $Z_{0o} = 25\ \Omega$) with two different load situations ($X_C = 150\ \Omega$ and $X_C = 250\ \Omega$) are drawn. It is found that the trends of these curves are similar to those in Fig. 2. In order to have a transmission zero, Fig. 4(a) shows the acceptable velocity ratios for a given even- and odd-mode impedances and capacitive loads. It should be less than an upper limit that is related to the circuit parameters. The smaller the capacitive reactances or the larger the even- and odd-mode impedances, the lower the limits. When the velocity ratio is fixed, the average electrical length will decrease if the even- and odd-mode impedances are increased or the capacitive reactances are reduced. This means that the frequency of the transmission zero becomes lower under these circumstances. Fig. 4(b) is plotted under the condition that the impedance ratio is selected to be 1.5. It is similar to Fig. 4(a), however, as the upper limits of the velocity ratio are decreased. For capacitively loaded coupled striplines ($\nu_o = \nu_e$), Fig. 4 shows that the transmission zero is no longer

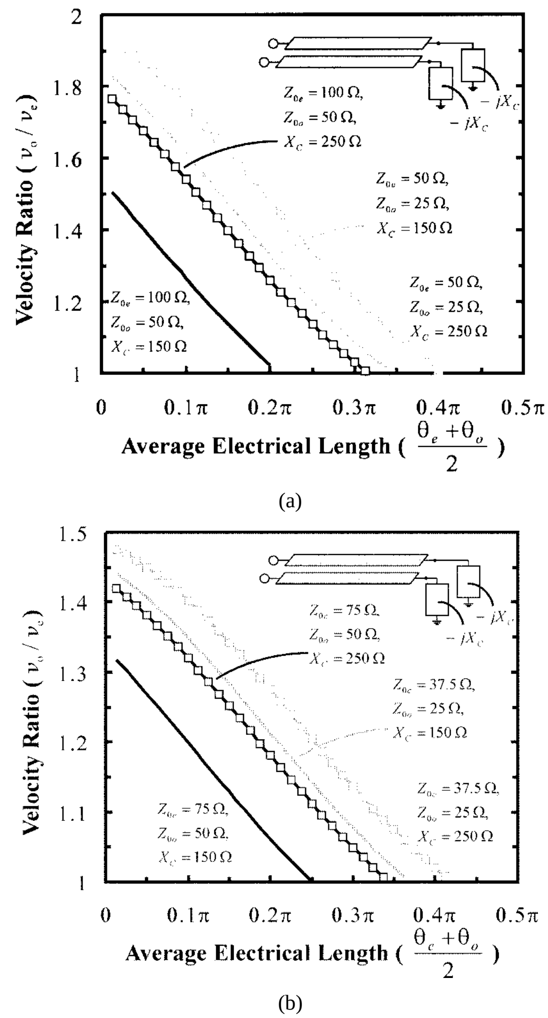


Fig. 4. Transmission-zero conditions for coupled lines with capacitive loads at one end when impedance ratio is: (a) 2 and (b) 1.5.

at the frequency when the average electrical length is $\pi/2$. Its position depends on the even- and odd-mode impedances and the capacitive loads.

D. Inductive Termination

Finally, for the case of inductive loads, the transmission-zero condition can be found by applying $Z_L = jX_L$ into (4) and the result is

$$\left[Z_{0e}^2 Z_{0o} + X_L^2 Z_{0o}\right] \cot\theta_o + X_L Z_{0o}^2 = \left[Z_{0e} Z_{0o}^2 + X_L^2 Z_{0e}\right] \cot\theta_e - X_L Z_{0e}^2. \quad (8)$$

The characteristics of transmission-zero conditions are plotted as Fig. 5(a) and (b), according to different impedance ratios (2 and 1.5, respectively). In Fig. 5(a) and (b), the transmission-zero conditions due to different even- and odd-mode impedances and different inductive loads ($X_L = 10\ \Omega$ and $X_L = 15\ \Omega$) are shown. The trends of the curves in these two figures are similar to those in Fig. 3 (one end of coupled lines is shorted to ground). The transmission zero cannot exist if the velocity ratio is larger than a certain value related to the even- and odd-mode impedances and the inductive loads. It is found that the upper limit will become lower if the impedance ratio is lower, the even- and

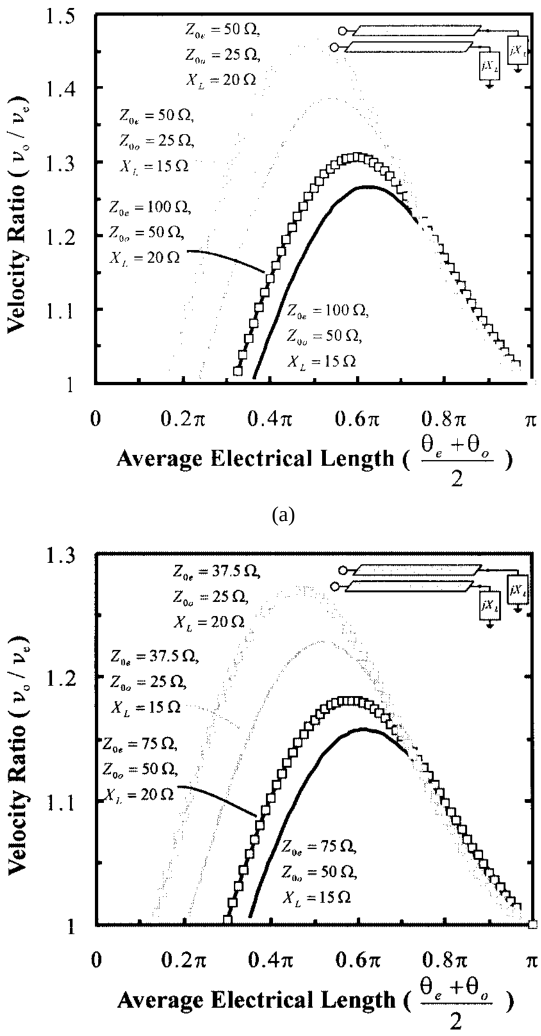


Fig. 5. Transmission-zero conditions for coupled lines with inductive loads at one end when impedance ratio is: (a) 2 and (b) 1.5.

odd-mode impedances are larger, or the inductive reactances are smaller. Two transmission zeros will be found if the velocity ratio is less than the upper limit. These two transmission zeros move away from each other if the impedance ratio is increased, the even- and odd-mode impedances are decreased, or the inductive reactances are increased. For inductively loaded coupled striplines ($\nu_e = \nu_o$), Fig. 5 shows that one of the transmission zeros can be moved to a lower frequency by proper design. However, the other zero is always fixed at the frequency when $\theta_e = \theta_o = \pi$. In fact, this zero always exists and is at this position for the coupled stripline with any uncoupled loads Z_L at one end.

Equations (5)–(8) are transcendental and there are a few parameters involved. Therefore, simple analytic solutions should not be expected. However, this does not mean the equations are only useful for the purpose of analysis instead of design. The designer could always choose to tune only one parameter, while keeping the rest of them fixed for other purposes. For example, the loading capacitance could be used to tune the zeros, while other coupled-line parameters are designed and fixed during the regular filter-design procedures. Although a few iterations

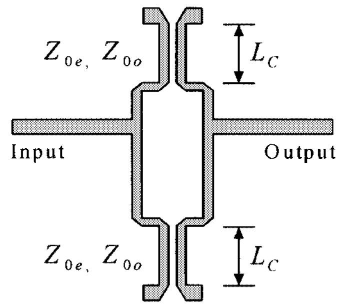


Fig. 6. Circuit configuration of the designed filters.

TABLE I
CIRCUIT PARAMETERS AND TRANSMISSION-ZERO FREQUENCIES OF THE DESIGNED FILTERS

Filter	Circuit Parameters	Zero Frequency		
		Analytic Solution	Simulation Result	Measured Result
A	$W_C = 30$ mils, $Gap = 14$ mils $L_C = 232$ mils, $C = 0.25$ pF	3.25 GHz	3.46 GHz	3.52 GHz
B	$W_C = 30$ mils, $Gap = 10$ mils $L_C = 307$ mils, $C = 0.25$ pF	2.83 GHz	3.02 GHz	3.04 GHz
C	$W_C = 30$ mils, $Gap = 14$ mils $L_C = 157$ mils, $C = 0.25$ pF	3.95 GHz	4.06 GHz	4.14 GHz
D	$W_C = 30$ mils, $Gap = 17$ mils $L_C = 152$ mils, $C = 0.2$ pF	4.42 GHz	4.5 GHz	4.6 GHz

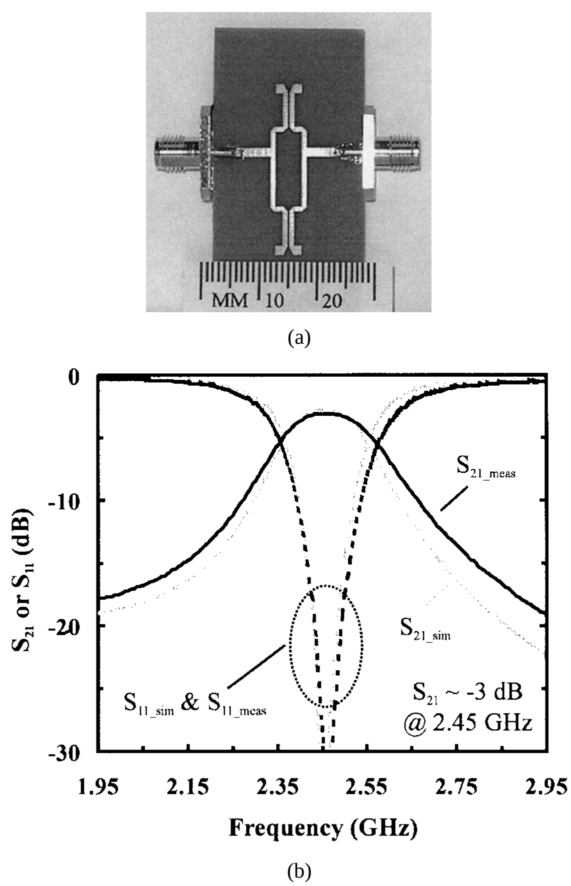


Fig. 7. (a) Photograph and (b) simulated and measured passband responses of Filter A.

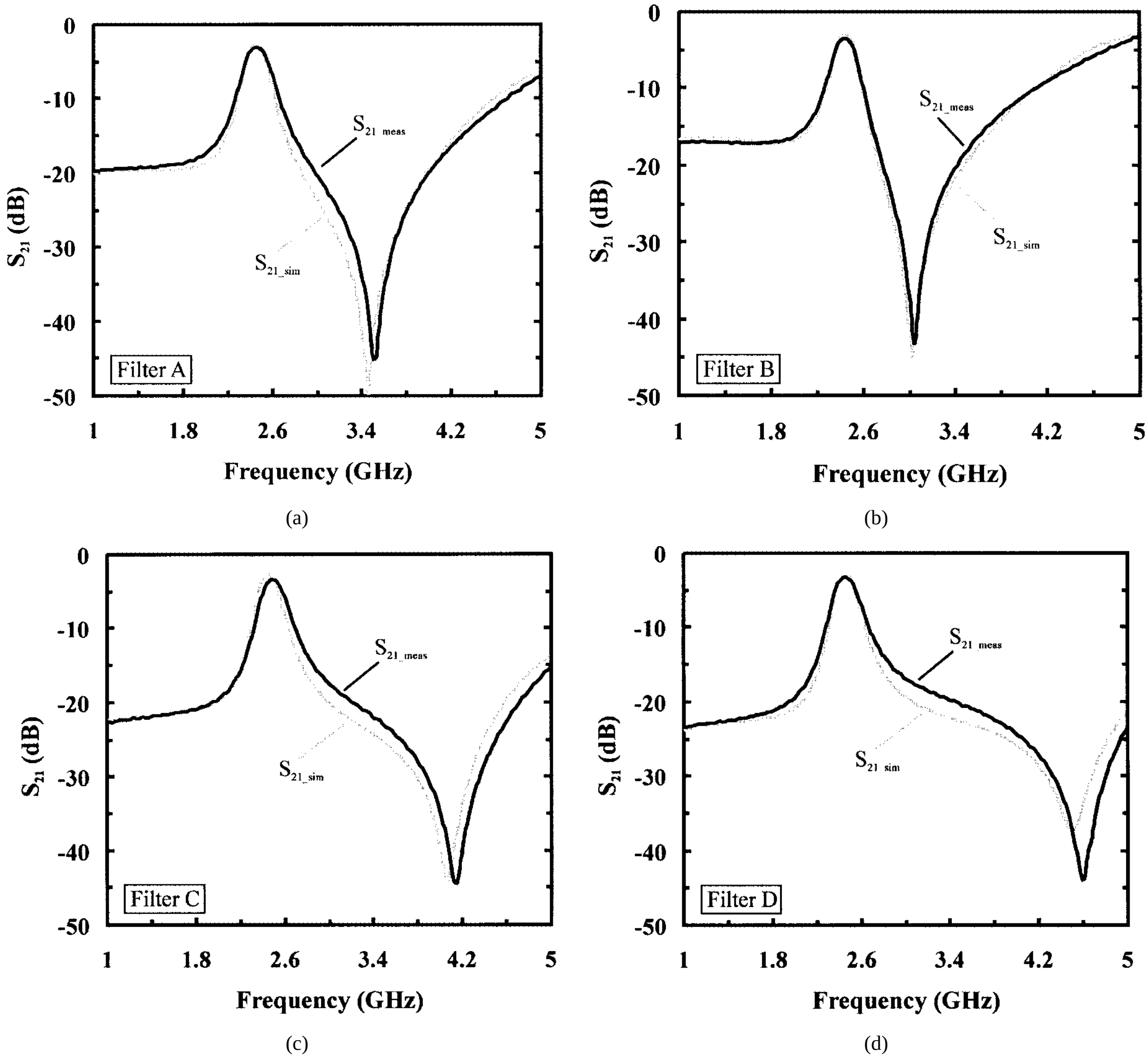


Fig. 8. Simulated and measured out-of-band responses of: (a) Filter A, (b) Filter B, (c) Filter C, and (d) Filter D.

might be needed, the equations and figures still serve as good design guidelines and provide good starting points. Without these, design through trial and error could be time consuming. As discussed in this section, a solution might not even exist at all if it is beyond some limits. A trial-and-error method cannot tell the designer how to avoid this kind of futile search for zeros. On the other hand, the figures cited in this section clearly show the solutions. Some of the applications are shown in Section III.

III. Filter-Design Examples

To demonstrate the applications of the studies in the previous sections, several second-order Butterworth filters using coupled lines with capacitive loads at one end were designed. This is because the tunability of the transmission zero of this type of circuits is better than the one with one end opened or shorted to ground. Moreover, since the capacitive loads can be realized by open stubs, the fabrication of these kinds of circuits is easier than the one with inductive loads. Fig. 6 shows the circuit configuration of these filters. It is well known that a section of transmission line opened at both ends behaves like a resonator when it is approximately a half-wavelength. In Fig. 6, two such kinds of resonators are coupled at their ends. By properly adjusting their coupling and input/output feed locations, this circuit becomes a simple second-order filter. The coupled-line sections control the coupling and, therefore, the bandwidth of the filter. The design procedures can easily be found in the literature [1]–[3].

In Fig. 6, the two coupled-line sections are intentionally folded outward to avoid any coupling between them. This helps to verify that zeros are due to loaded coupled lines, as discussed in the previous section, and not other parasitic couplings. However, in practice, the coupled-line sections can certainly be folded inward to minimize the circuit size. This kind of lower order filters and their variations are suitable for low-loss and compact-size applications because the required circuit elements are few in number. However, there are some problems in such types of filters. For example, the steepness of the skirts near the passband is usually not large enough because of the small number of poles. Besides, the out-of-band rejection also degrades significantly due to the input-to-output directly coupling. Skew symmetric feed structures have been shown to be useful in improving the shape factor of this kind of filters

[9]. An example will be given in Section IV. The transmission zeros created by loaded coupled lines will now be used to improve the out-of-band rejection.

Although the design procedures for these kinds of filters and their variations are well established. The transmission zeros as discussed in the previous section have never been included in the design process. They are treated as uncontrollable by-products. However, they will be considered as important as the other filter parameters in this research. The design procedures begin by selecting a realizable coupled-line section. Open stubs as capacitive loads to the coupled lines are then designed, with the help of Fig. 4 or (7), to yield a zero at the desired frequency. Next, the length of the uncoupled portion of the transmission line in Fig. 6 is determined such that the resonant frequency of the opened transmission line is at the center of the desired passband. Then adjust the coupled-line length and gapwidth to yield the required coupling coefficient for a given filter bandwidth. The above three parameters (i.e., zero position, filter center frequency, and bandwidth) are checked again and retuned if necessary. In a few iterations of the above procedures, a reasonable design could be achieved. Finally, the input/output feed lines are adjusted to complete the filter design.

All filters were designed to have the center frequency at 2.45 GHz, 4% bandwidth, and return loss larger than 15 dB. These filters were fabricated on the FR4 substrates with a relative dielectric constant of 4.3, a loss tangent of 0.015, and a thickness of 31.5 mil. Considering the filter bandwidth and substrate used, high insertion loss is expected. However, this has nothing to do with filter-design techniques and does not hinder the demonstrations of the tunability of the transmission zero. The designed filters are denoted as Filter A–D. The circuit parameters are summarized in Table I. The equivalent values of the capacitive loads are shown instead of the lengths of the open stubs. This is acceptable because a very short open stub could be modeled as a capacitor with very small error at the frequency range of several gigahertz. The analytic solutions of transmission-zero frequencies are calculated by using (7) and are also included in Table I.

Fig. 7 shows the photograph and the simulated and measured passband responses of Filter A. The passband insertion loss of this filter is approximately 3 dB. This is mainly due to the conductor and dielectric losses of the substrate. The passband return loss is larger than 15 dB. The simulated and measured out-of-band responses of Filter A are shown in Fig. 8(a). A transmission zero due to the loaded coupled lines can be clearly observed at 3.52 GHz. The measured data is in good agreement with the simulation. Fig. 8 also shows the responses of the other filters that were designed to have the same passband response as Filter A, but with different transmission-zero frequencies. Since the even- and odd-mode impedances of the coupled-line structures in these filters were intentionally selected to be similar, the difference between the transmission-zero frequencies is dominated by the lengths of the coupled lines and capacitive loads. Among Filters A–C, although the same capacitive loads were used, the transmission-zero frequency of Filter B is lowest because it has the longest coupled lines. The transmission zero of Filter D occurs at the frequency higher than the one of Filter C because it was designed with smaller capacitive loads. Both

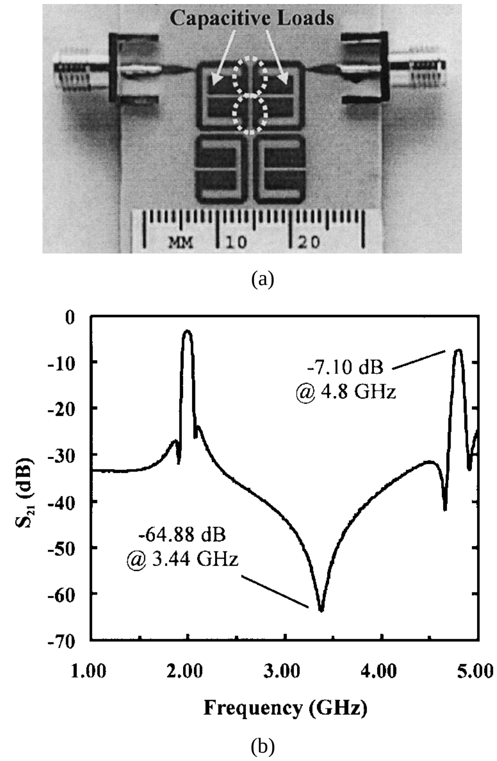


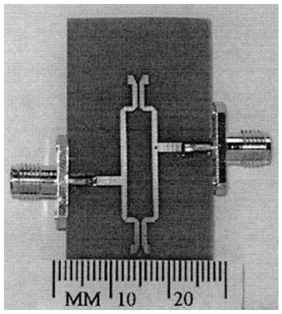
Fig. 9. (a) Cross-coupled miniaturized hairpin filter and (b) its response.

simulated and experimental results of the transmission-zero frequencies of all filters are also summarized in Table I. The difference of the transmission-zero frequency between the analytic and measured results is less than 10%. This is mainly due to the bends in the designed filters and the errors of modeling the open stubs as capacitors.

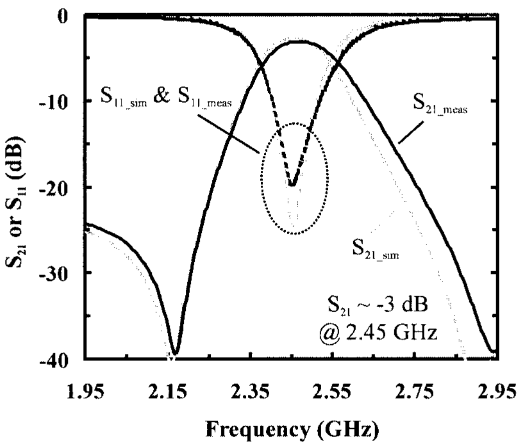
It is also noted that the cross-coupled miniaturized hairpin filter, introduced in [10] and shown in Fig. 9(a), is a good example of this research. The coupling between the input and output resonators was realized by two coupled lines (marked by white dotted circles). The capacitive loads of each coupled-line structure were realized by the other coupled lines with open circuits at one end (indicated by white arrows). Therefore, an extra transmission zero at 3.44 GHz, shown in Fig. 9(b), was found and the stopband rejection was improved significantly. This zero should not be confused with those created by cross-coupling between resonators, which were discussed in many papers and were near and on the opposite sides of the passband around 2 GHz.

IV. Use of a Skew-Symmetric Feed Structure

From [9], it has been known that filters designed with a skew-symmetric (0°) feed structure can have two extra transmission zeros, which are close to and on the opposite sides of the passband. Therefore, the selectivity and stopband rejection of designed filters can be increased significantly. Since a skew-symmetric feed structure is very feasible for the coupling structures, which are in the form of a ring with two splits, it can be applied to the filters designed in the previous section. Fig. 10(a) shows the circuit of Filter E, which has the same circuit parameters of



(a)



(b)

Fig. 10. (a) Photograph and (b) simulated and measured passband responses of Filter E.

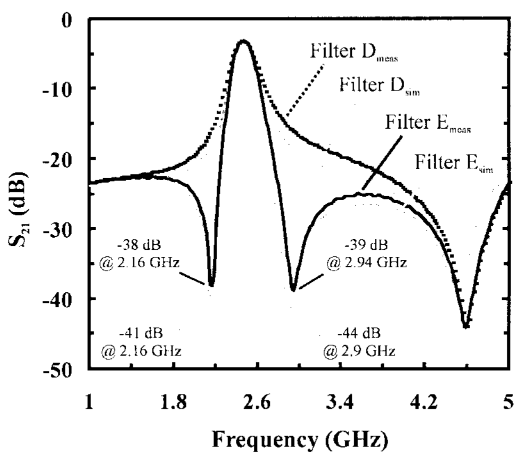


Fig. 11. Simulated and measured out-of-band responses of Filters D and E.

Filter D in Section IV, but with a skew-symmetric feed structure. Fig. 10(b) shows the simulated and measured results of this filter. It is found that the passband insertion loss is approximately 3 dB and the passband return loss is larger than 15 dB in Filter E, which are the same as those of Filter D. The out-of-band responses of Filters D and E are compared and shown in Fig. 11, which clearly shows that the passband response of Filter E has two extra transmission zeros at 2.16 and 2.94 GHz. Combined with the effect of the transmission zero (at 4.6 GHz) created by the capacitively loaded coupled lines, a 25-dB rejection band from 2.82 to 4.94 GHz is achieved. Therefore, both the shape factor and the out-of-band rejection are improved significantly.

V. Conclusions

Coupled lines with loads at one end have been studied in this paper. With the aid of even- and odd-mode analysis, the equation of the transmission-zero condition has been derived. Based on this equation, characteristics of the transmission-zero frequencies of coupled lines with different loads, including open circuit, short circuit, capacitive load, and inductive load have been analyzed. Since analytic solutions are not possible, and several figures have been plotted. They show that transmission zeros depend on the loaded reactance, coupled-line length, even- and odd-mode velocities, and impedances. The tunability of the transmission zero has been discussed in detail. It is also found that transmission zeros do not exist beyond a certain limit. Several filters using coupled lines with capacitive loads at one end have been designed, fabricated, and measured to verify this research and to demonstrate the applications. The transmission zeros have been included in the filter design and are treated as important as other filter parameters. These examples have shown that the zeros are controllable and useful. Finally, a filter with a wide high-rejection stopband has also been designed by using the techniques presented in this paper and a skew-symmetric feed structure.

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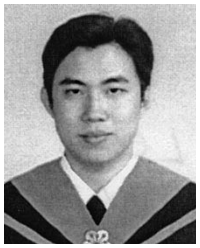

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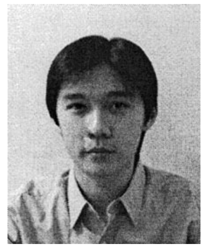

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