Because a simple, closed-form expression for C'_{fo} as a function of gap length g is not available, the value of g which satisfies (2) has to be found iteratively as follows. For a particular value of g, the odd-mode fringing capacitances C'_{fo}, which occur in (10) or (11), are found from Getsinger's Fig. 4, and the corresponding gap capacitance C_g, calculated. Next, the frequency f_0, which satisfies (2), is found graphically or numerically. The process is repeated to produce a graph of resonant frequency f_0 as a function of gap length g. The value of g which corresponds to the desired resonant frequency can then be found from the graph.

III. Experimental Results

The validity of the assumptions and approximations in the procedure have been tested for a resonator located between two adjacent resonators, in a 2-percent bandwidth interdigital linear-phase filter of degree 6. The measured resonant frequency was compared to the theoretical frequency, computed using (11), for two gap lengths and Table I shows that good agreement was obtained between theory and experiment.

IV. Conclusion

A simple procedure has been presented for computing the gap width of the loosely coupled rectangular cross-sectional interdigital resonator to obtain a specified resonant frequency. The method uses a simple formula for the gap capacitance of the resonator, based on Getsinger's odd-mode fringing capacitance data [1].

Experimental results suggest that the accuracy of the method is of the order 1 percent. From Table I it is observed that the theoretical resonant frequency is higher than the measured frequency, which indicates that the theoretical gap capacitance C_g is less than the actual capacitance. This could imply that the corner capacitances are not negligible, as assumed in approximation (8).

The technique is expected to give best results for structures that use loosely coupled bars, such as narrow-band filters. Tightly coupled bars would have different even- and odd-mode fringing capacitances, thus invalidating approximation (9). However, the procedure should be adequate for most practical filter designs, since tightly coupled bars are usually associated with wide-band filters which are relatively insensitive to deviations from the ideal circuit element values. A practical resonator design approach is to design the gap width so that the theoretical resonant frequency is 1–2 percent higher than the desired frequency, and to provide a capacitive tuning screw at the gap. The gap width should be slightly larger than required when the tuning screw is flush with the cavity wall. Therefore, the resonant frequency can be reduced to the desired value by introducing the tuning screw into the gap, thus reducing the effective gap width and increasing the gap capacitance.

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A Design Method for Noncommensurate Broad-Band Matching Networks

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***Abstract*—A simple design method for broad-band ladder network impedance transformers having noncommensurate section lengths and predictable passband response is presented. Limitations on section impedances imposed by constructional constraints are more easily met than with commensurate networks, and harmonically related passbands are largely avoided. An example is presented.**

I. Introduction

The short-step impedance transformer [1] and other broad-band transmission-line networks based on rational insertion loss functions employ commensurate (equal) element length sections. As a consequence of this constraint, the frequency response is highly periodic. Moreover, having specified the desired frequency response, overall network length and circuit topology (i.e., number of series/shunt elements etc.), the designer has little flexibility with the range of element characteristic impedances required.

For example, in a particular application a designer might consider the use of a series-cascaded commensurate network because of its relative design and constructional simplicity. Having specified the passband response and overall network length, the minimum number of sections needed and thereafter the section impedances can be determined. Because of the short element lengths, a wide range of section impedances usually results. The designer must then determine whether such a design can be realized in practice, in the transmission-line type desired. (A typical range feasible in microstrip, for example, is 20–110 Ω, while in slotline 55–300 Ω, or coax 10–100 Ω.) If the design is not practical, then using commensurate elements, only a network of greater complexity and/or length will provide the desired passband performance.

On the other hand, noncommensurate networks have the advantage of allowing greater design flexibility because the constraint that all elements have the same electrical length is removed. A further advantage of such designs is that the periodic recurrence of higher order passbands is largely avoided. However, the circuit transfer function can no longer be described in terms of rational functions. To date, no general theory has been presented for the synthesis of noncommensurate circuits with prescribed gain functions.

In this paper, a design procedure is outlined enabling a noncommensurate network to be derived from a commensurate prototype. The procedure involves the use of a transformation which keeps the fundamental passband frequency response of the derived noncommensurate circuit almost identical to that of the commensurate prototype. This is achieved while providing the designer with greater flexibility in the choice of element impedance levels.

II. Design Procedure

Having considered the conventional commensurate design and found it to be unsuitable, a noncommensurate network may be derived by considering pairs of sections of the commensurate

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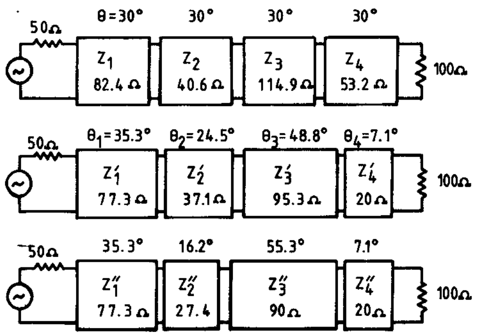


Fig. 1. Designs for 50–100-Ω impedance transformers having a VSWR of 1.36 over the frequency range 250–750 MHz. (a) Commensurate. (b) and (c) Noncommensurate.

TABLE I
TRANSMISSION MATRIX ELEMENTS FOR TWO-SERIES TRANSMISSION-LINE SECTIONS

	Commensurate	Non-Commensurate
A	$\cos^2\theta - \frac{Z_1}{Z_2}\sin^2\theta$	$\cos\theta_1 \cos\theta_2 - \frac{Z_1'}{Z_2'}\sin\theta_1 \sin\theta_2$
-jB	$(Z_1 + Z_2)\sin\theta\cos\theta$	$Z_1' \sin\theta_1 \cos\theta_2 + Z_2' \sin\theta_2 \cos\theta_1$
-jC	$\left(\frac{1}{Z_1} + \frac{1}{Z_2}\right)\sin\theta\cos\theta$	$\frac{1}{Z_1'} \cdot \sin\theta_1 \cos\theta_2 + \frac{1}{Z_2'}\sin\theta_2 \cos\theta_1$
D	$\cos^2\theta - \frac{Z_2}{Z_1}\sin^2\theta$	$\cos\theta_1 \cos\theta_2 - \frac{Z_2'}{Z_1'}\sin\theta_1 \sin\theta_2$

design. These pairs are modified by the designer to meet realizability constraints, and in such a way that the transmission matrix of the pair remains unchanged at the passband center frequency. Pairs of sections are operated on because the problem of equivalencing the commensurate and noncommensurate circuits for this case is relatively simple. It has been shown [3] that this method results in almost the same broad-band power transfer ratio for the derived circuit.

The advantage of this method is, therefore, that the same circuit topology and approximately the same passband frequency response is maintained, but the circuit may be easier to realize.

The A, B, C, D parameters of the transmission matrix for a pair of commensurate sections (Fig. 1(a)) and a pair of noncommensurate sections (Fig. 1(b)) are given in Table I, Z and θ being element characteristic impedances and electrical lengths, respectively.

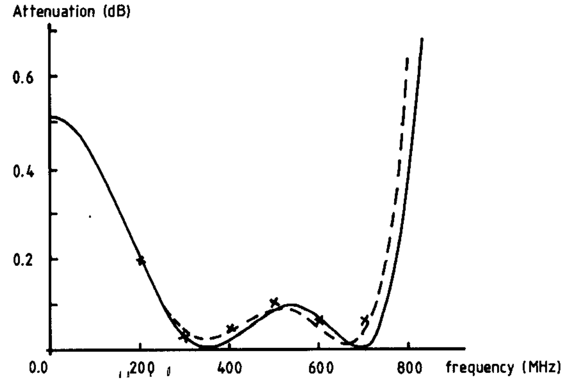
From the commensurate network parameters (Z_1, Z_2, θ), four new parameters $(Z_1', Z_2', \theta_1, \theta_2)$ can be found for the noncommensurate case which, at one frequency, result in the same transmission matrix. Since the network is reciprocal, there are three independent simultaneous equations and, therefore, one degree of freedom. By the appropriate choice of θ_1 or θ_2, the impedances Z_1' and Z_2' can be made easier to realize than Z_1 and Z_2. The solution of the equations of Table I is straightforward. From the solution it can be shown [3] that the new lengths θ_1 and θ_2 are such that

$$\theta_1 + \theta_2 \approx 2\theta \tag{1}$$

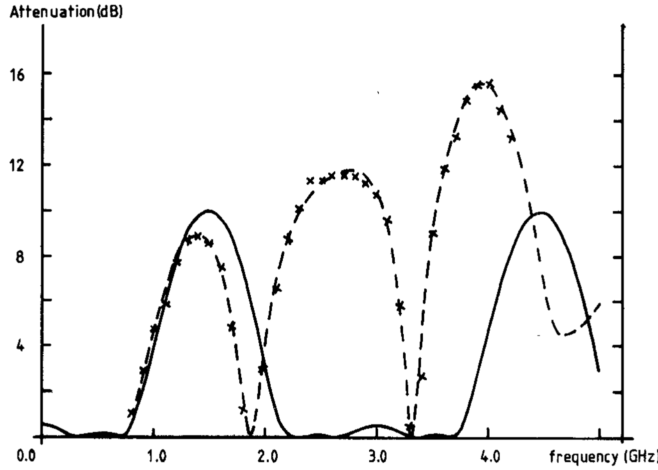
and furthermore

$$\theta_1 + \theta_2 \leqslant 2\theta. \tag{2}$$

From the Taylor series expansion of the A, B, C, D parameters, with respect to frequency, it can also be shown [3] that because of



(a)



(b)

Fig. 2. Comparison of frequency responses. ——theoretical response, commensurate design (Fig. 1(a)). ---theoretical response, noncommensurate design (Fig. 1(c)). xxx measured response, practical realization of noncommensurate design (Fig. 1(c)). (a) Passband. (b) dc to 5 GHz.

(1), the difference in the first-order terms of the expansion for the commensurate and noncommensurate cases is small. This accounts for the wide-band behavior of the equivalence.

III. PRACTICAL EXAMPLE

Consider the design of a 50–100-Ω impedance transformer for operation over the frequency range 250–750 MHz with a VSWR of less than 1.36 (insertion loss less than 0.1 dB). Suppose that the transformer is to be realized in coaxial form using an outer conductor diameter of 14.3 mm. Furthermore, the total network length should be less than 200 mm.

Following conventional techniques [1], [2], a commensurate network having Chebyshev response may be derived (Fig. 1(a)), which meets the frequency response and length specifications. For reasons of mechanical rigidity of the inner conductor, and machining considerations, a minimum inner conductor diameter of 3 mm is desirable. This puts an upper limit of 90 Ω on element characteristic impedances. With this limitation, the network of Fig. 1(a) is not realizable. This could be overcome by increasing the number of elements and/or network length. Because the network length is constrained in this case, a design using shunt or series stubs would need to be used.

An alternative which does not require an increase in circuit complexity is to derive a noncommensurate circuit from Fig. 1(a) by the method outlined in the previous section.

The pair of sections Z_1, Z_2 are first modified and then Z_3, Z_4 are modified to give the circuit of Fig. 1(b). The pair of sections Z_2, Z_3 are then further modified to give the final circuit in Fig. 1(c). (These circuits were obtained assuming an arbitrary minimum section impedance of 20 Ω). In terms of the physical realizability constraints, this circuit is now realizable in the desired form.

Fig. 2 illustrates the theoretical frequency responses of the original commensurate network (Fig. 1(a)), the derived noncommensurate circuit (Fig. 1(c)), together with the measured response of an experimental circuit designed to realize the noncommensurate circuit. For clarity, the passband response is illustrated on an expanded scale in Fig. 2(a), while the response from dc to 5 GHz is shown in Fig. 2(b).

IV. Conclusion

A simple technique has been presented for aiding in the design of realizable compact broad-band circuits where conventional cascaded commensurate designs may be impractical or difficult to implement. The technique ensures that the derived noncommensurate network has a fundamental passband frequency response almost identical to that of the prototype. Furthermore, the noncommensurate circuit does not exhibit wide, harmonically related passbands. The design method has been experimentally tested and verified.

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Design Consideration for High-Isolation Coaxial Broad-Band p-i-n Diode Switches and Limiters

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***Abstract*—Broad-band coaxial p-i-n diode switches and limiters are realized using a low-pass filter structure in which shunt capacitances are realized by the capacitances of reverse- or zero-biased p-i-n diodes. Usually, the design considerations are given only for insertion-loss state. No design guideline exists in the literature to optimize isolation for these type of switches and limiters. This paper shows that using low-pass filter structure with series inductance as the first element, higher isolation without increasing insertion loss can be achieved.**

I. Introduction

Broad-band microwave switches and limiters are realized in a coaxial low-pass filter structure which employs p-i-n/limiter diodes to secure the shunt capacitances [1] as shown in Fig. 1(a). The inductances of the filter are realized by short sections of high-impedance coaxial line lengths. The design procedures are well documented in the literature [2]. But all the literature deals only with design considerations to obtain low insertion loss. Nowhere in the literature is found the design consideration for these type of switches and limiters for high isolation, which at the same time have low insertion loss. This paper deals with the design consideration for high isolation for coaxial broad-band switches and limiters.

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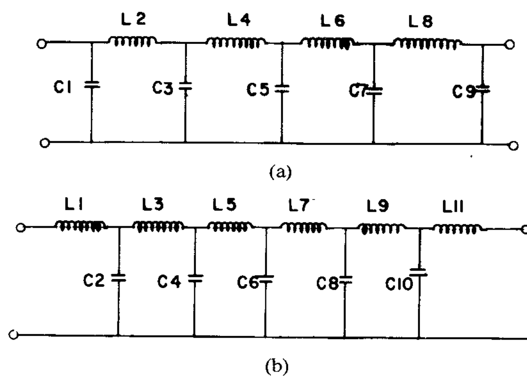

Fig. 1. (a). Low-pass filter with capacitance as the first element. (b) Low-pass filter with inductance as the first element.

II. Design Consideration for High Isolation

To make the switches and the limiters broad band, the filter structure can be chosen with shunt capacitance as the first element as in Fig. 1(a) (which is usually used [1]) or series inductance as the first element as in Fig. 1(b). The number of diodes to be used in the filter is the same as the number of shunt capacitances. Therefore, there will be two additional inductance elements in the filter structure with inductance as first element compared to the filter structure with capacitance as the first element [Fig. 1(a) and (b)]. Though the number of elements is not the same, both the filters can be designed with the same passband insertion loss. At forward bias (in case of switches) or at high power level (in case of limiters), the diodes behave as a low-resistance elements which make the filter structure no longer a low-pass filter and the resulting network offers isolation. The structure of Fig. 1(b) gives more isolation due to two additional inductances compared to structure of Fig. 1(a). This will be clear from the example given below. For the sake of simplicity, the cases of two diodes switches/limiters are taken.

Case 1: Capacitance as the First Element (Fig. 2)

Let us assume that

$$\text{Passband VSWR} = 1.05$$
$$\text{Cutoff frequency } f_{co} = 3.3 \text{ GHz}$$
$$\text{Characteristic impedance } Z_0 = 50\ \Omega.$$

Therefore, the filter elements values are [2], [3]

$$g_1 = 0.4861, \quad g_2 = 0.8259, \quad \text{and} \quad g_3 = 0.4861.$$

Therefore, $C_1 = C_3 = g_1/\pi 2 f_{co} Z_0 = 0.469$ pF

$$L_2 = g_2^{Z_0}/2\pi f_{co} = 1.9 \text{ nH}.$$

Therefore, the required switch/limiter with passband VSWR of 1.05 will consist of two diodes having capacitances of 0.469 pF

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