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

B. K. SARKAR

***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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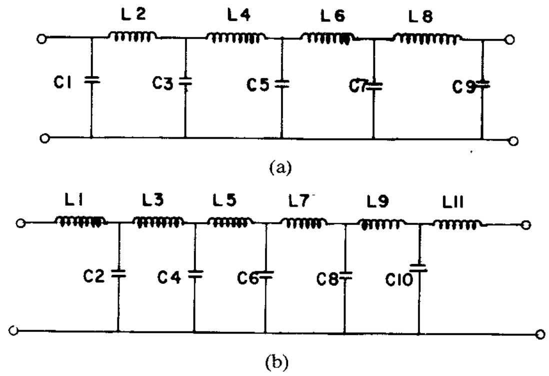

Fig. 1. (a). Low-pass filter with capacitance as the first element. (b) Low-pass filter with inductance as the first element.

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

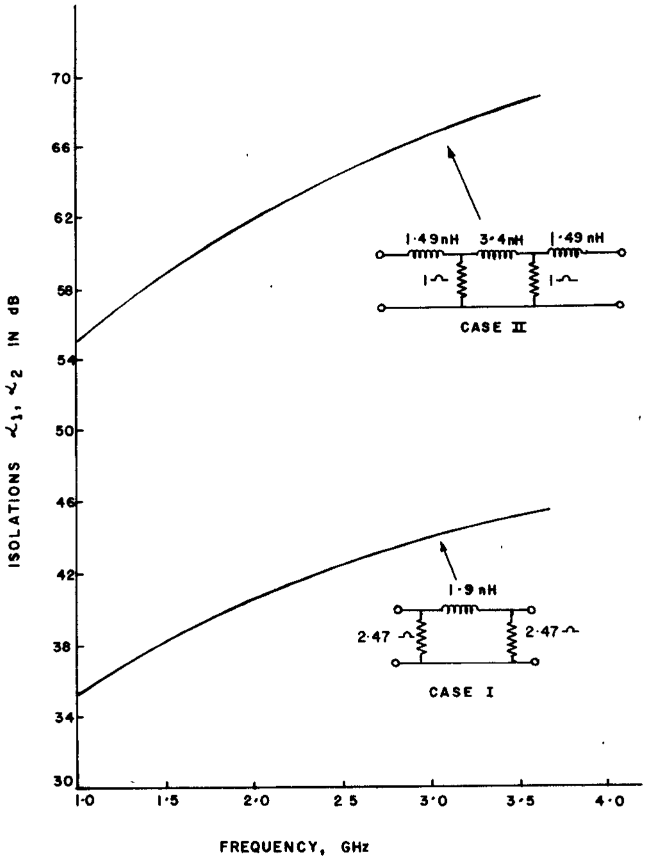

Fig. 2. Isolation versus frequency for passband VSWR $= 1.05$, $f_{co} = 3.3$ GHz, the resistances of the diodes when forward biased (or at high power) for case I and case II, are 2.47 and 1.0 Ω, respectively.

each and one series inductance of 1.9 nH, realized by a short coaxial high-impedance line length.

At forward bias or at high power level, the capacitances C_1 and C_3 will be replaced by the resistances of the diodes. Let us assume that the resistance of each diode is 2.47 Ω. The isolation of the switch/limiter is calculated using $\boldsymbol{ABCD}$ matrices [4] and is given by

$$\alpha_1 = 10 \log \left[451.26 + 2902 f^2 \right] \text{dB} \tag{1}$$

where f is the frequency in GHz.

α_1 is plotted against frequency as shown in Fig. 2

Case II: Inductance as the First Element (Fig. 2)

For the same specifications as in case I, the low-pass filter elements values are $g_1 = 0.6181$, $g_2 = 1.2026$, $g_3 = 1.4130$, $g_4 = 1.2026$, and $g_5 = 0.6181$. Therefore,

$$L_1 = L_5 = \frac{g_1 Z_0}{2\pi f_{co}} = 1.49 \text{ nH}$$

$$C_2 = C_4 = \frac{g_2}{2\pi f_{co} Z_0} = 1.16 \text{ pF}$$

$$L_3 = \frac{g_3 Z_0}{2\pi f_{co}} = 3.4 \text{ nH}.$$

Therefore, the required switch/limiter with passband VSWR of 1.05 will consist of two diodes with capacitances of 1.16 pF each and three series inductances (1.49 nH at the input, 3.4 nH in between the diodes, and 1.49 nH at the output) realized by short coaxial high-impedance line lengths.

At forward bias or high power level, the capacitances C_2 and C_4 are replaced by diode resistances. In order to compare the merits of the circuits in cases I and II, we have to take p-i-n diodes of the same cutoff frequency f_c. Assuming the same value for reverse- and forward-biased series resistances, we have $f_c = 1/2\pi C_1 R_1 = 1/2\pi C_2 R_2$. Therefore, $R_2 = (C_1/C_2) R_1 \cong 1.0\ \Omega$. Where R_1 ($= 2.47\ \Omega$) and R_2 are the forward-biased series resistances of each diode in case I and case II, respectively. The isolation α_2 is calculated using $\boldsymbol{ABCD}$ matrices and is given by

$$\alpha_2 = 10 \log \left[(51 - 205.75 f^2)^2 + 39.5 f^2 (91.44 - 2.98 f^2)^2 \right] \text{dB} \tag{2}$$

where f is the frequency in GHz.

α_2 is plotted against frequency as shown in Fig. 2.

III. Discussions and Conclusion

It is seen from Fig. 2 that the isolation is always greater (20 dB, approximately) in case II compared to case I, though the passband characteristic (with zero- or reverse-biased diode in the case of the switch and at low power level in the case of the limiter) is the same for both the cases. Therefore, a switch/limiter design based on the low-pass filter structure with series inductance as the first element will always offer higher isolation at the same time with low insertion loss.

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On the Possible Use of Microwave-Active Imaging for Remote Thermal Sensing

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Abstract —**Recent results have demonstrated the feasibility of quasi-real-time, active as well as harmless microwave imaging for biomedical purposes. Such a process allows tomographic reconstructions based on differences in the complex permittivity of tissues, the temperature dependence of which can be used for remote thermal sensing. A basic experiment conducted in water at 3-GHz yielded information on spatial resolution and temperature sensitivity. Discussion is devoted to potential capabilities and limitations of this remote-sensing approach in more complicated situations.**

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