Computer-Simulated Design of an Active Microwave All-Pass Network

S. E. SUSSMAN-FORT, MEMBER, IEEE

Abstract—A new microwave active all-pass network is proposed. In a computer simulation, it is shown that the network, even with markedly nonideal transistors, can provide a true all-pass response over 8–12 GHz.

I. INTRODUCTION

MICROWAVE filters are most commonly implemented as passive networks of waveguide and transmission-line elements. Active network components have been rarely used in microwave filters, although active filters have enjoyed great popularity in systems at frequencies up to a few megahertz. One reason for this situation is that low-frequency active filters use inexpensive operational amplifiers or transistors whose imperfections can be assumed, for most purposes, to be negligible. At microwave frequencies, however, only bipolar and field-effect transistors are available as gain elements, and these devices can hardly be assumed to be ideal above 1 GHz. Nevertheless, it is clear that microwave active filters would be most desirable as system components for the following important reason.

Microwave active filters would most likely be implemented as networks of high-performance GaAs FET's and lumped and/or distributed circuit elements. In the form of microwave integrated circuits, such filters would be much smaller and lighter than their conventional passive counterparts. Size is obviously an important consideration in the design of modern aircraft and satellite systems, for example.

Over the past 25 years, a wealth of knowledge has become available about low-frequency active filters. It would be most useful if this information could be adapted for use at microwave frequencies. This paper is concerned with a special case of adapting a low-frequency active-filter design to the microwave region. Specifically, we consider an FET filter whose response with ideal elements is that of a second-degree all-pass network with a center frequency of 10 GHz. In a computer simulation, we let the FET's have scattering parameters equal to those specified by the manufacturer for an NEC 388 GaAs FET. This causes the network response to substantially deviate from that of an ideal all pass. We then show that it is possible to optimally adjust the other circuit elements so that the network provides the desired all-pass response from 8–12 GHz. The components in the optimized network are of values which are readily realizable with current MIC technology. (See [1] and [2] for examples of typical, realizable MIC component values.)

Manuscript received May 18, 1979; revised September 11, 1979.
The author is with the Electrical and Systems Engineering Department, Rensselaer Polytechnic Institute, Troy, NY 12181.

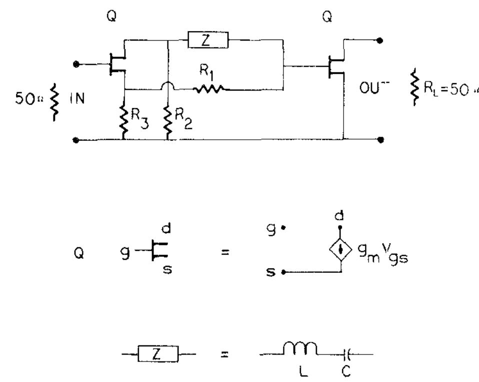

Fig. 1. Active all-pass network (bias circuitry omitted).

II. THE IDEAL CIRCUIT

The active all-pass network that we propose is shown in Fig. 1. It is basically an FET version of a network using bipolar transistors described by Orchard [3] and Calfee [4]. For our network, S_{21} is given by

$$S_{21} = -\frac{2g_m^2 R_3 R_L}{1+g_m R_3} \cdot \frac{Z - R_1R_2/R_3}{Z + R_1 + (R_1+R_2)/(1+g_m R_3)}. \quad (1)$$

With Z as a series connection of an inductor L and a capacitor C, (1) becomes

$$S_{21} = -\frac{2g_m^2 R_3 R_L}{1+g_m R_3} \cdot \frac{s^2 - [R_1R_2/(LR_3)]s + 1/LC}{s^2 + \frac{1}{L}[R_1 + (R_2+R_3)/(1+g_m R_3)]s + 1/LC}. \quad (2)$$

Hence the condition for an all-pass response is

$$R_1R_2/R_3 = R_1 + (R_2+R_3)/(1+g_mR_3).$$

Comparing (2) to the following general form of an all-pass transfer function

$$T(s) = Ke(-s)/e(s), \qquad e(s) = s^2 + \frac{\omega_0}{q}s + \omega_0^2$$

we obtain

$$K = -2g_m^2 R_3 R_4/(1+g_m R_3)$$

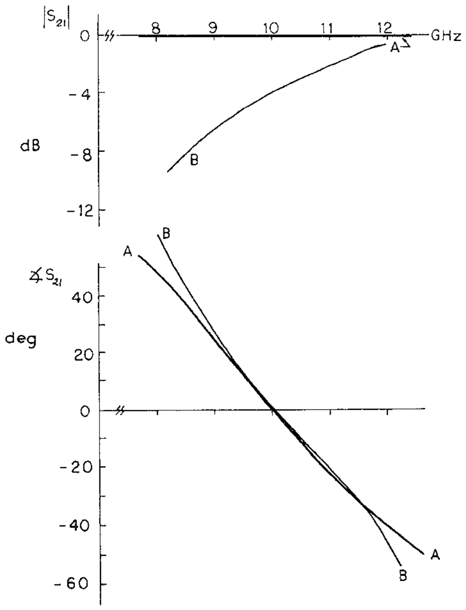

Fig. 2. Comparison of all-pass response with A ideal transistors and B NEC 388's, Case 1. (Phase response B is normalized to provide zero phase shift at 10 GHz.)

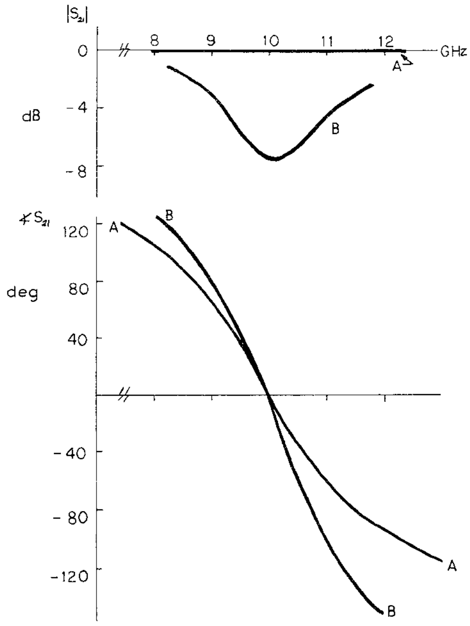

Fig. 3. Comparison of all-pass response with A ideal transistors and B NEC 388's, Case 2. (Magnitude response A has been scaled to provide 0-dB gain; phase response B is normalized to provide zero phase shift at 10 GHz.)

$$\omega_0 = 1/\sqrt{LC}$$

$$q = (R_3/R_1R_2)\sqrt{L/C}\ .$$

A higher degree all-pass network could be obtained, of course, by using a higher degree lossless impedance for Z in Fig. 1. However, it is better practice to realize an active all-pass network as a tandem connection of second-degree, ideally noninteracting sections, as such an arrangement usually reduces the sensitivity of the network to component tolerances.

In the next section, we shall consider two special sets of element values for the ideal, second-degree all-pass network of Fig. 1. These values are as follows.

Case 1

$$K = -1, \quad f_0 = 10\text{ GHz}, \quad q = 1$$
$$R_1 = 50, \quad R_2 = 150, \quad R_3 = 50\ \Omega$$
$$C = 0.106\text{ pF}, \quad L = 2.387\text{ nH}$$
$$g_m = 20\text{ mmho.}$$

Case 2

$$K = -3, \quad f_0 = 10\text{ GHz}, \quad q = 3.$$
$$R_1 = 16.67, \quad R_2 = 50, \quad R_3 = 16.67\ \Omega$$
$$C = 0.106\text{ pF}, \quad L = 2.387\text{ nH}$$
$$g_m = 60\text{ mmho.}$$

III. The Optimization

Using the computer program COMPACT [5], we first determined the effect of substituting NEC 388 GaAs FET's for the ideal transistors in the network of Fig. 1. (COMPACT has a library of tabulated, manufacturer-specified, S parameters for microwave transistors.) In

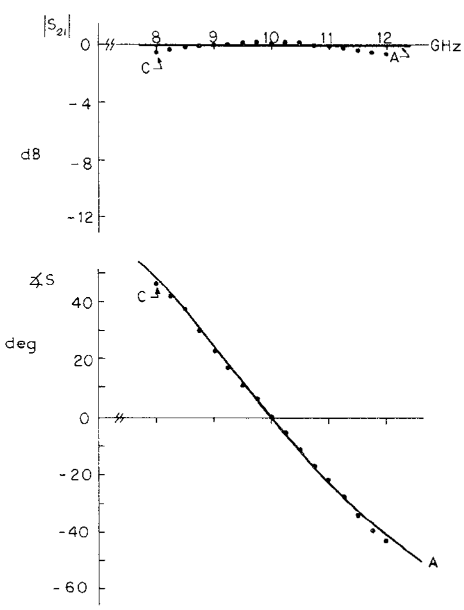

Fig. 4. Comparison of A ideal and C optimized all-pass response, Case 1. (Phase response C is normalized to provide zero phase shift at 10 GHz.)

both Cases 1 and 2, this substitution caused the network response to deviate greatly from that of an ideal all-pass. These results are shown in Figs. 2–3. Using COMPACT, we then attempted to vary the passive elements of our network to achieve a flat magnitude response and the required all-pass phase response. In Case 1, we were able to restore the network to the desired all-pass response over 8–12 GHz. In this frequency range, the magnitude and phase deviation from an ideal all-pass response was less than 0.6 dB and 1.5°, respectively, as illustrated in Fig. 4. The resulting optimized element values were found to be as follows.

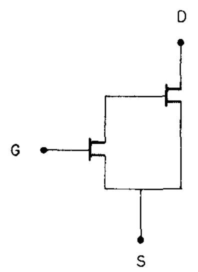

Fig. 5. Composite transistor.

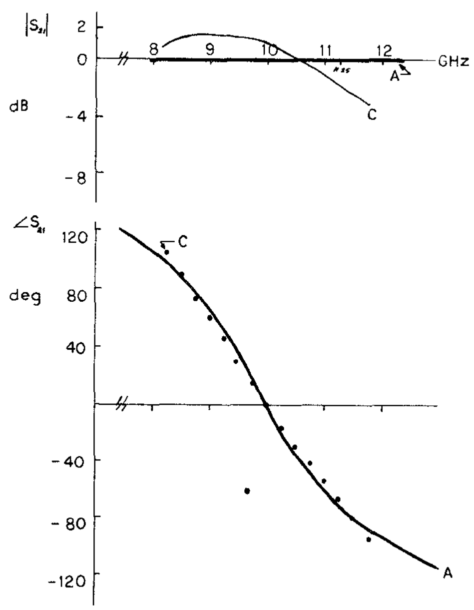

Fig. 6. Comparison of *A* ideal and *C* optimized all-pass response, Case 2. (Magnitude response *A* has been scaled to provide 0-dB gain; phase response *C* is normalized to provide zero phase shift at 10 GHz.)

Case 1

$$R_1 = 64.06, \quad R_2 = 170.88, \quad R_3 = 38.69\ \Omega$$
$$C = 0.169\ \text{pF}, \quad L = 3.433\ \text{nH}.$$

For Case 2, in addition to optimizing the passive element values, it was necessary to replace each FET in our network with the composite transistor of Fig. 5 in order to restore the network to an all-pass response from 8.25 to 11.25 GHz. Over this frequency range, the deviation from all-pass was then less than 1.5 dB and 5.2°, respectively, as illustrated in Fig. 6. The corresponding optimized element values were found to be as follows.

Case 2

$$R_1 = 30.89, \quad R_2 = 49.52, \quad R_3 = 11.05\ \Omega$$
$$C = 0.106\ \text{pF}, \quad L = 2.58\ \text{nH}.$$

IV. Conclusions

It has been shown that a low-frequency active-filter design can be adapted for use in the microwave region. Over a finite frequency range, it has been demonstrated that the passive component values of an active all-pass network can be optimally adjusted to compensate for the distortion in the frequency response introduced by the use of nonideal microwave transistors. The success of this procedure raises significant hope that more of the well-established techniques of low-frequency active-filter design can be effectively utilized at microwave frequencies.

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