Investigation of Natural Transmission Zeros of Printed Combline Filters Using Electromagnetic Simulators

Jelena Dinkić, Miodrag Tasić, and Antonije Đorđević

***Abstract*—In this paper we exploit a model of combline filters with simple uncoupled transmission lines to investigate the impact of adjacent and nonadjacent couplings to stopband transmission zeros. Furthermore, we present a way to control the slope of the passband skirts by adding appropriate nonadjacent couplings. These effects are also observed in 2D and 3D simulation models and verified experimentally on a combline filter with 5 resonators.**

***Index Terms*—Combline filters, full-wave electromagnetic simulation, multiconductor transmission lines, resonators.**

I. Introduction

Combline structures have been extensively used as bandpass microwave filters owing to their compact size [1] and the ability to keep the parasitic passbands far above the basic passband. A combline filter consists of a set of resonators. In the classical form, each resonator consists of a short-circuited transmission line section, shorter than quarter-wavelength in the passband, and a lumped capacitor. Adjacent resonators are mutually inductively coupled.

Such a structure is usually synthesized following the standard procedures for polynomial filters (all-pole filters), the Chebyshev design being the most common in practice.

However, a combline filter (fabricated in a homogeneous medium) has a natural transmission zero when the length of the transmission-line sections becomes quarter wavelength. Also, by introducing additional couplings among nonadjacent resonators, it is possible to produce other transmission zeros that can be under or above the passband [2, 3, 4].

The goal of this paper is to investigate transmission zeros that naturally occur for a combline structure (other than the quarter-wavelength case), which can be used to improve the filter performance.

II. Equivalent Schemes for Combline Filters

Fig. 1 shows the scheme of one of the basic combline structures, which consists of n coupled transmission lines and $n+2$ capacitors [1]. All transmission lines have a common return conductor (ground). The transmission lines can be printed as microstrips or striplines, or they can be machined in the form of rectangular or circular bars. (These transmission lines can collectively be regarded as a multiconductor transmission line, with n signal conductors and one ground.) Each transmission line is short-circuited at the right end, and terminated in a capacitor at the left end. Ports P1 and P2 are coupled by capacitors to the first and the last line, respectively.

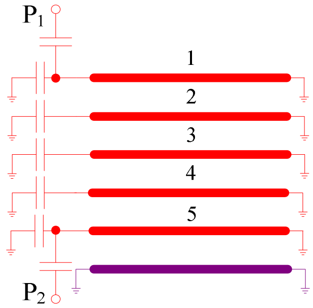

Fig. 1. A combline structure with 5 resonators represented as a multiconductor transmission line.

In the passband, the electrical length of the transmission lines is less than quarter-wavelength, so that the short-circuited transmission lines behave like a set of coupled inductors (coils). Hence, each transmission line and its terminating capacitor constitute one resonator.

In the passband, the major contribution to the losses in the transmission lines usually comes from conductor losses. The dielectric losses are less pronounced. Note that many microwave-circuit simulators improperly take the dielectric losses into account [5], wrongly yielding larger losses than encountered in practice.

Assuming that the dielectric of the multiconductor transmission line is homogeneous, the structure shown in Fig. 1 can be represented by the equivalent scheme shown in Fig. 2a. The multiconductor transmission line is replaced by a set of simple (uncoupled) transmission lines.

If $[Y_c]$ is the characteristic admittance matrix of the multiconductor line shown in Fig. 1, then the characteristic admittances of the simple transmission lines in Fig. 2a are

$$Y_{mm} = c\sum_{n=1}^{N} b'_{mn} = \sum_{n=1}^{N} Y_{cmn}\,,\quad m = 1,\ldots,N\,, \tag{1}$$

$$Y_{mn} = -cb'_{mn} = -Y_{cmn}\,,\quad m,n = 1,\ldots,N,\ m \neq n\,, \tag{2}$$

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where b'_{mn} are the elements of the matrix of the electrostatic-induction coefficients $[B']$, Y_{cmn} are the elements of the matrix $[Y_c]$, and c is the velocity of wave propagation in the dielectric. This equivalent scheme allows us to monitor the influence of individual couplings simply by including or excluding the corresponding transmission lines. For example, Fig. 2b shows the equivalent scheme of the structure shown in Fig. 1 when only adjacent couplings are assumed to exist.

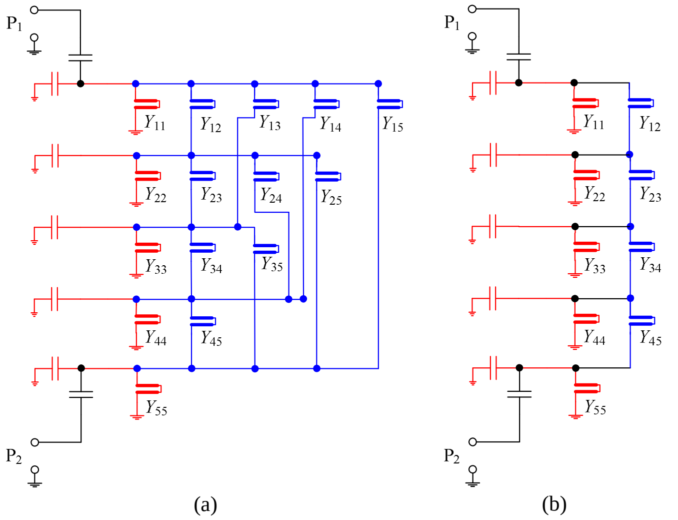


Fig. 2. Equivalent scheme with simple transmission lines: (a) when all couplings are assumed, (b) when only adjacent couplings are assumed.

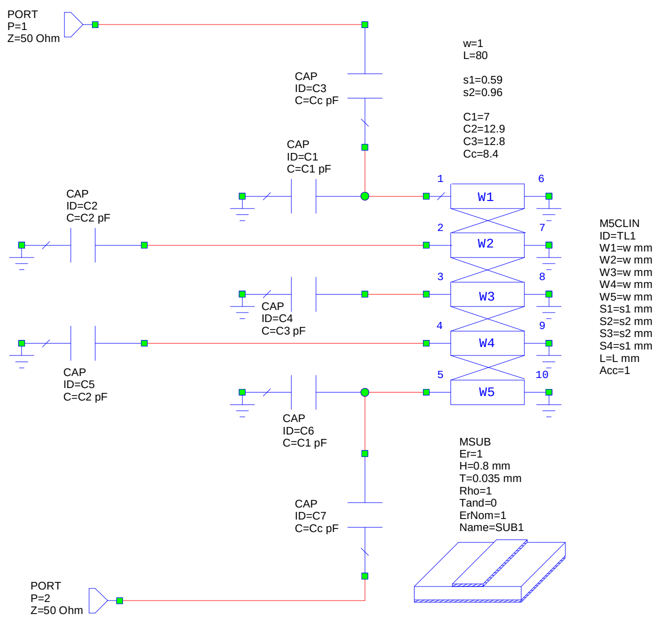


Fig. 3. The 2D model of a microstrip combline filter.

As a concrete example, we consider the microstrip combline filter shown in Fig. 3. For 2D simulations we use Microwave Office [6] (whose kernel for electromagnetic analysis of multiconductor transmission lines is the same as in LINPAR [7]). In order to provide a homogeneous medium, we define the substrate relative permittivity to be 1. The length of the microstrip lines is $L = 80$ mm and the width of all strips is $w = 1$ mm. The spacings between the lines are $s_1 = 0.59$ mm and $s_2 = 0.96$ mm. The capacitances are $C_1 = 7$ pF, $C_2 = 12.9$ pF, $C_3 = 12.8$ pF, and $C_C = 8.4$ pF. The characteristic admittances of the simple transmission lines in Fig. 2a and Fig. 2b are calculated using program LINPAR [7] along with (1) and (2).

In Fig. 4 the parameter s_{21} of the model shown in Fig. 3 is compared with the parameters s_{21} of the equivalent models shown in Fig. 2a and Fig. 2b. The results for the model shown in Fig. 3 and for the model shown in Fig. 2a are in a good agreement for all frequencies. However, they agree with the results for the model shown in Fig. 2b only in the passband. There are significant discrepancies outside the passband, which are due to neglecting the nonadjacent couplings in the model shown in Fig. 2b. This example demonstrates the significance of the nonadjacent couplings.

The parameter s_{21} has a natural transmission zero when the length of the transmission lines is quarter-wavelength (at 937 MHz) because in this case there is no coupling among the resonators. This zero makes s_{21} decay steeper above the passband then below the passband (Fig. 4).

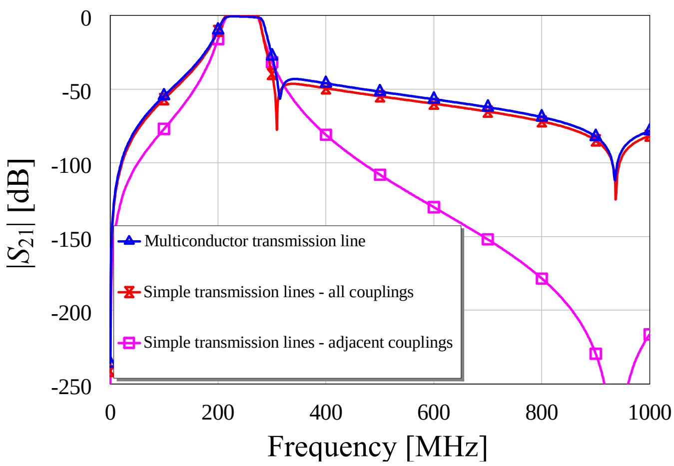


Fig. 4. Simulation results of the models shown in Fig. 2 and Fig. 3.

Fig. 4 demonstrates that the couplings among nonadjacent resonators, which always exist in practice, skew the graph of the transfer function slightly clockwise, making the lower stopband shallower. The upper stopband is deeper (i.e., the filter selectivity is increased), having a transmission zero very close to the passband (at 314 MHz). Note that this transmission zero does not correspond to the frequency when the length of the transmission lines is quarter-wavelength, but it occurs at a lower frequency (three times lower for the considered example). The existence of this phenomenon was suspected in [8], but it was not elaborated or verified.

In the next section we also demonstrate that by adding an appropriate nonadjacent coupling, the leading skirt of the passband can become steeper, whereas the trailing skirt remains only slightly affected.

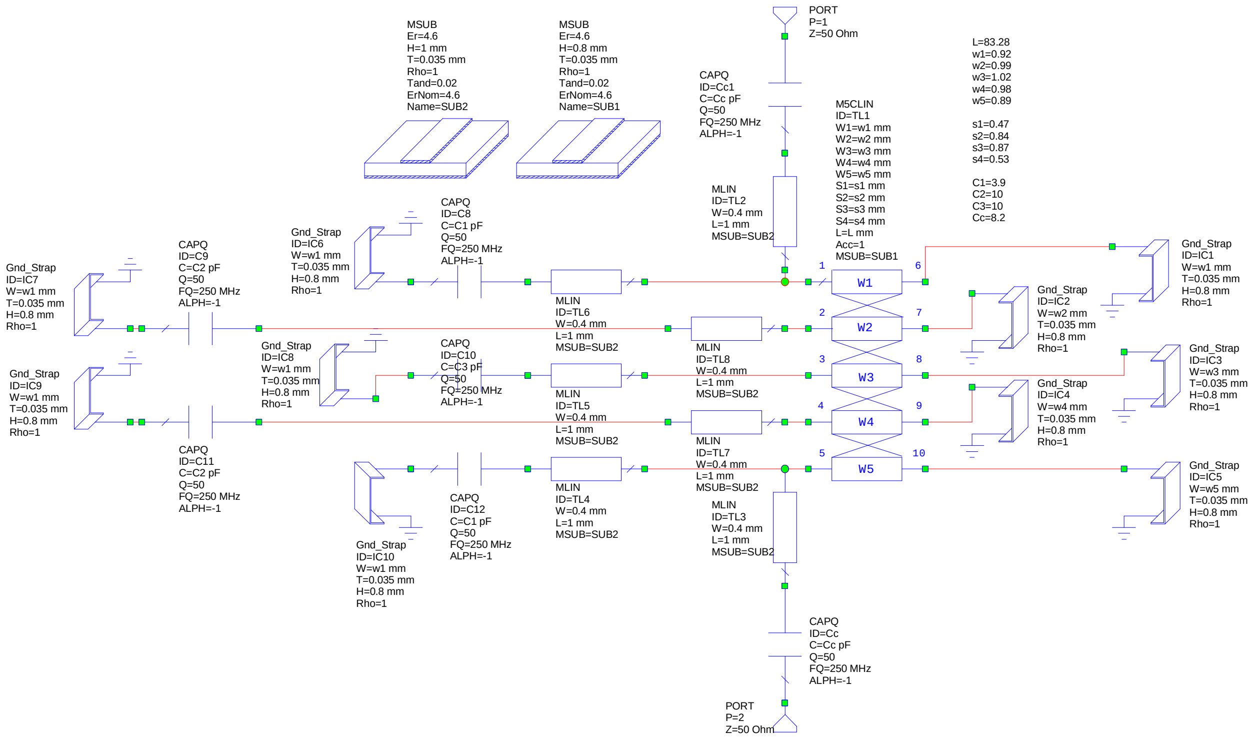


Fig. 5. The 2D model used for optimization in Microwave Office.

III. PRINTED COMBLINE STRUCTURE

In order to further confirm the influence of the additional couplings among the nonadjacent resonators, we have fabricated a combline filter with 5 resonators (as the structure shown in Fig. 1). The substrate is FR-4 now. The filter is optimized for the passband centered at 250 MHz. For the optimization we use the Microwave Office 2D model (Fig. 5). Fig. 6 shows the printed combline filter. The substrate thickness is 0.8 mm, whereas the metallization thickness is 35 µm. The length of the microstrip lines is $L = 83.28\ \mathrm{mm}$, and the widths of the microstrip lines are $w_1 = 0.92\ \mathrm{mm}$, $w_2 = 0.99\ \mathrm{mm}$, $w_3 = 1.02\ \mathrm{mm}$, $w_4 = 0.98\ \mathrm{mm}$, and $w_5 = 0.89\ \mathrm{mm}$. The spacings between the lines are $s_1 = 0.47\ \mathrm{mm}$, $s_2 = 0.84\ \mathrm{mm}$, $s_3 = 0.87\ \mathrm{mm}$, and $s_4 = 0.53\ \mathrm{mm}$, whereas the capacitances are $C_1 = 3.9\ \mathrm{pF}$, $C_2 = 10\ \mathrm{pF}$, $C_3 = 10\ \mathrm{pF}$, and $C_C = 8.2\ \mathrm{pF}$. The schematic shown in Fig. 5 includes models of short circuits (straps) and parasitic effects of SMD components (short transmission lines).

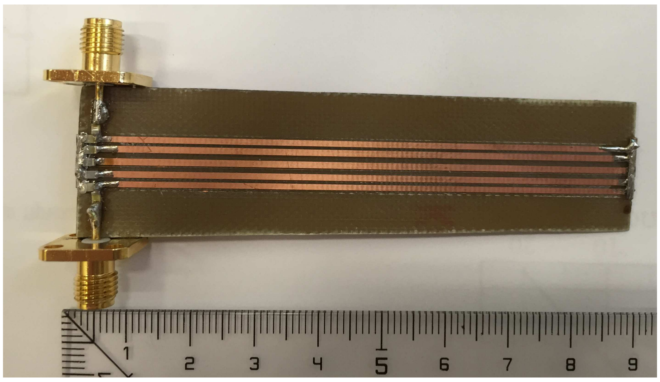


Fig. 6. Printed combline filter.

The parameter s_{21} of the optimized 2D model and the measured results for the fabricated combline structure are shown in Fig. 7.

To demonstrate the influence of the nonadjacent couplings, we introduce an additional coupling. We mount a lumped capacitor $C_{\mathrm{coupling}} = 0.3\ \mathrm{pF}$ between the ports. We made this capacitor from two wires coated with plastic. Fig. 8 shows the combline filter with the additional capacitive coupling. Fig. 9 shows the measured results and the results of the 2D simulation of the combline filter with this additional coupling.

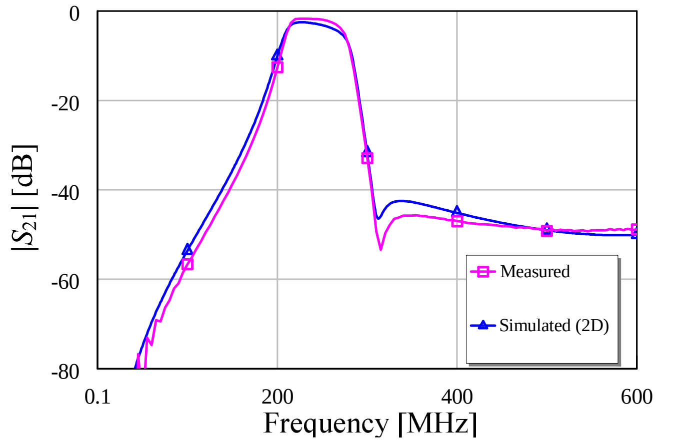


Fig. 7. The parameter s_{21} of the combline filter, measured and simulated (2D) results.

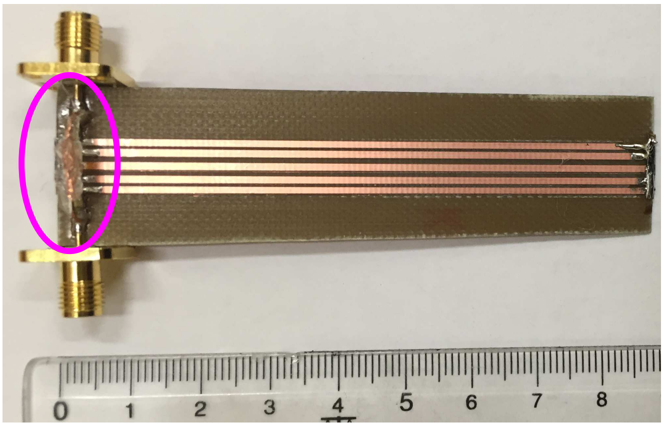


Fig. 8. Printed combline filter with additional coupling between ports (circled in pink).

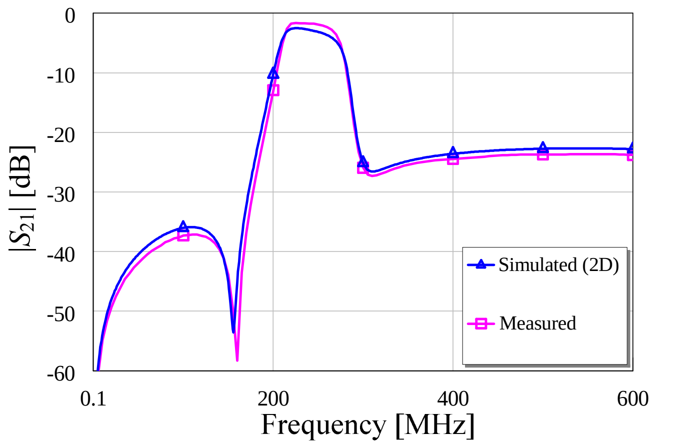


Fig. 9. The parameter s_{21} of the combline filter with additional capacitive coupling between ports, simulated (2D) and measured results.

Fig. 10 compares the measured results before and after inserting the additional coupling. In the latter case, the leading skirt of the passband becomes steeper and the graph of the transfer function is slightly skewed counterclockwise.

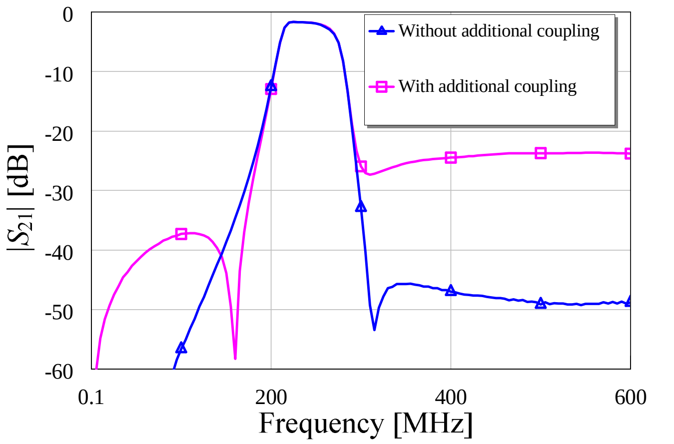


Fig. 10. The parameter s_{21} of the fabricated combline structure without and with additional capacitive coupling between ports.

IV. Full-wave electromagnetic simulation

In order to further confirm the results presented in the previous sections, we perform a full-wave (3D) electromagnetic simulation using software WIPL-D Pro [9]. The model of the combline structure without the additional coupling between two ports, used for this simulation, is shown in Fig. 11. The geometrical modeling is performed using bilinear quadrilateral patches. The surface current is approximated by higher-order polynomials. Losses in the dielectric are modeled assuming a loss tangent of 0.02, whereas the metal conductivity is 19 MS/m. All capacitors are modeled as lumped loadings. Each port is modeled as shown in the inset in Fig. 11, in order to mimic an SMA connector.

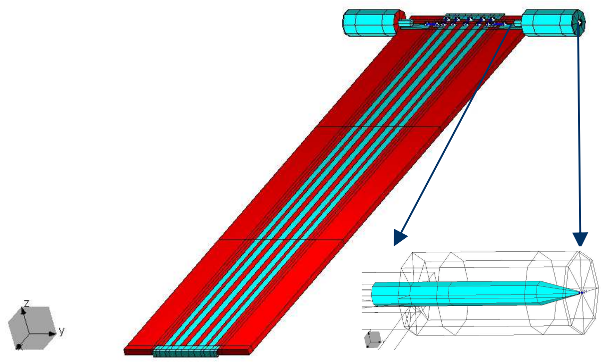
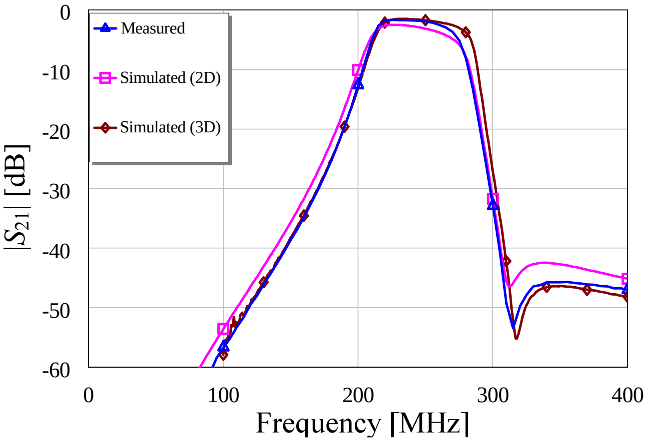
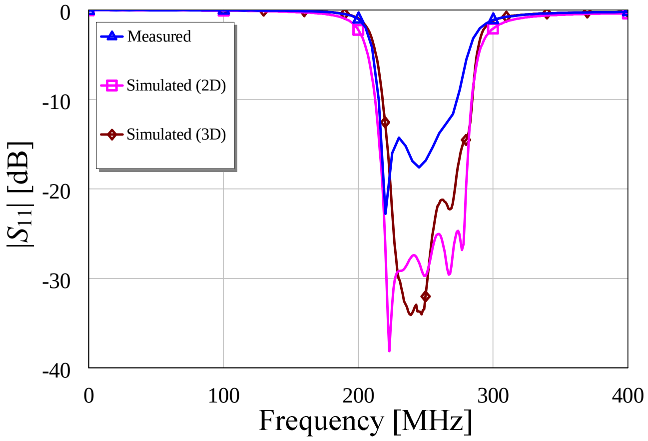


Fig. 11. 3D electromagnetic model of the combline filter.

The measured and simulated results (2D and 3D) are compared in Fig. 12. It can be observed that the 3D simulation provides a better agreement along the leading skirt of s_{21}, whereas along the trailing skirt, close to the passband, the 2D simulation shows better matching (Fig. 12a). Both the 2D and 3D simulations provide a sufficiently good estimation of the s_{11} passband (Fig. 12b). Overall, the WIPL-D Pro 3D simulation is more accurate because it models the discontinuities more precisely than the Microwave Office 2D simulation.



(a)



(b)

Fig. 12. The scattering parameters of the combline filter (without additional capacitive coupling between ports), measured and simulated (2D and 3D): (a) s_{21} and (b) s_{11}.

V. Conclusion

In this paper we investigated the influence of adjacent and nonadjacent couplings in combline structures. We demonstrated that both couplings participate in shaping the transfer function of the filter. The influence of the nonadjacent couplings is particularly pronounced in the stopbands. Naturally occurring parasitic nonadjacent couplings create a transmission zero just above the passband and make the trailing skirt of the passband steeper. Similarly, an intentional insertion of an additional nonadjacent capacitive coupling between the filter ports can make the leading skirt of the passband steeper. These results are confirmed by simulations using 2D and 3D models and by measurements of a fabricated combline filter. In order to clearly prove the concepts, the example shown in the paper was designed for lower frequencies. At higher frequencies, various parasitic effects strongly influence the filter performance and they are hard to be separated. However, the influence of the nonadjacent couplings is also very important for the performance of combline filters designed for higher frequencies. Nonadjacent couplings can be optimized to shape the transfer functions of these filters as well.

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References

[1] V. V. Petrović, D. V. Tošić, A. R. Đorđević, *Microwave Passive Circuits*, Belgrade, Serbia, School of Electrical Engineering, 2010.

[2] J.-S. Zhan, J.-L. Wang, “A simple four-order cross-coupled filter with three transmission zeros,” *Prog. Electromagn. Res. C.*, vol. *8*, pp. *57-68*, 2009.

[3] M. Snchez-Renedo, J. I. Alonso, “Tunable planar combline filter with multiple source/load coupling,” *IEEE MTT-S Int. Dig.*, pp. *771-774*, June 2005.

[4] C.-W. Tang, Y.-C. Lin, C.-Y. Chang, “Realization of transmission zeros in combline filters using an auxiliary inductively coupled ground plane,” *IEEE Trans. Microw. Theory Techn*., vol. *51*, no. *10*, pp. *2112-2118*, Oct. 2003.

[5] A. R. Đorđević , A. G. Zajić , D. V. Tošić, “A note on the modeling of transmission-line losses,” *IEEE Trans. Microw. Theory Techn.*, vol. *51*, no. *2*, pp. *483-486*, Feb. 2003.

[6] www.awrcorp.com

[7] A. R. Djordjević, M. B. Baždar, R. F. Harrington, T. K. Sarkar, *LINPAR for Windows: Matrix Parameters for Multiconductor Transmission Lines*, Boston, MA, Artech House, 1999.

[8] J. G. Hong, M. J. Lancaster, *Microstrip Filters for RF/Microwave Applications*, John Wiley & Sons, 2004.

[9] www.wipl-d.com