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Abstract available at: http://ieeexplore.ieee.org/xpl/freeabs_all.jsp?arnumber=1420743

This paper appears in: Microwave Theory and Techniques, IEEE Transactions on
Issue Date: April 2005
Volume: 53 Issue:4
On page(s): 1164 - 1167
ISSN: 0018-9480
References Cited: 5
Cited by : 1
INSPEC Accession Number: 8375533
Digital Object Identifier: 10.1109/TMTT.2005.845725
Date of Current Version: 18 April 2005
Sponsored by: IEEE Microwave Theory and Techniques Society

Low-Reflection Bandpass Filters with a Flat Group Delay

Antonije R. Djordjević and Alenka G. Zajić

Abstract—**Recently, low-reflection lowpass RLC filters with a linear phase characteristic have been reported. The present paper is an extension of this work to bandpass filters, including their realizations with lumped-element and distributed-parameter networks. A topology particularly suitable for microstrip techniques is proposed. It yields a compact design that enables miniaturization and is easy for manufacturing and tuning.**

Index Terms—**Bandpass filters, impedance matching, linear phase filters.**

I. Introduction

RECENTLY, a class of lowpass filters has been proposed that has a quasi-Gaussian amplitude characteristic and a linear phase (i.e., a flat group delay), as well as a low reflection both in the passband and stopband [1]. The filters were derived from a hypothetical distributed-parameter network. A brief summary of these lowpass filters is given in Section II and the work from [1] is extended to include numerical data for a practical design.

Similar bandpass filters have certain applications [2]. Hence, in this paper, we investigate low-reflection bandpass filters with a flat group delay. Section III presents a design procedure for bandpass lumped-element and distributed-parameter networks that have a linear phase, along with a low reflection at all frequencies. Section IV presents experimental results for a microstrip implementation and compares them with simulation data.

II. Lowpass Prototypes

Fig. 1 shows a topology of the basic network for lowpass prototypes. The network in Fig. 1 consists of identical cascaded cells, where only the first and the last series branch are different from other series branches. Resistors can exist in all branches (complete filter) or only in series branches (incomplete filter).[1] We assume that the nominal impedances of both ports are Z_c. The incomplete filter has fewer resistors than the complete filter, a higher attenuation at very high frequencies in the stopband, and a worse reflection coefficient.

For both the complete and incomplete filters, $L_1 = L_2/2$, while L_2 and C satisfy the relation

$$\sqrt{L_2/C} = Z_c . \tag{1}$$

The number of cells, N ($N = 3$ in Fig. 1), is selected to control the phase characteristic up to a required frequency limit (f_u) or, equivalently, up to a required insertion loss level (a_u).

We introduce the standard normalization of filter elements [3, p.139]. A normalized inductance is given by $L_n = L\omega_{3dB}/Z_c$ (where $\omega_{3dB} = 2\pi f_{3dB}$ is the 3 dB angular frequency), a normalized capacitance is given by $C_n = Z_c C\omega_{3dB}$, and a normalized resistance is $R_n = R/Z_c$.

Filter optimization can be performed with respect to various

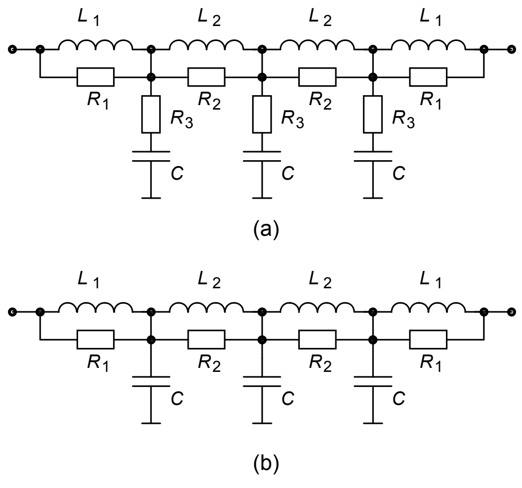

Fig. 1. Lowpass prototype of RLC filters: (a) complete filter and (b) incomplete filter ($N = 3$).

Manuscript received February 26, 2004, revised May 1, 2004. This work was supported in part by the Serbian Ministry of Science, Technology, and Development.

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[1] This topology is dual to the one presented in [1]. The topology presented in [1] starts and ends with shunt branches, and the incomplete filter has resistors only in shunt branches. The topology shown in Fig. 1 is chosen because it results in a compact printed-circuit design of bandpass filters.

TABLE I
OPTIMAL LOWPASS FILTERS

N	C_n	q_c	q_i	f_u / f_{3dB}	a_u [dB]	$\tau_d f_{3dB}$	$\lvert\Delta\tau_d f_{3dB}\rvert$	a_r [dB]
2	1.00	0.082	0.190	1.8	12	0.32	0.07	15/13
3	0.84	0.105	0.230	2.15	17	0.40	0.05	19/14
4	0.72	0.127	0.255	2.4	20	0.46	0.05	20/14
5	0.65	0.128	0.270	2.55	23	0.52	0.05	21/15

criteria. One can control the linearity of the phase characteristic, the group delay, the matching, etc. For example, for the complete filter, the normalized resistances of the optimal cell that has a maximally flat group delay are $R_{2n} = 1/q_c$ and $R_{3n} = q_c$, where $q_c \approx 1/5$ in an infinite array of cells. For the incomplete filter, the optimal cell has $R_{2n} = 1/q_i$, where $q_i \approx 1/3$ in an infinite array of cells. For a finite number of cells (N), the optimal parameters q_c and q_i are somewhat smaller.

A set of practical values, obtained after a combined optimization of the phase characteristic and the group delay, is given in Table I, for various N.[2] The second column shows the normalized capacitance, C_n. Due to (1), $L_{2n} = C_n$, whereas $L_{1n} = C_n / 2$. The normalized resistances in the first and last branches are $R_{1n} = 1$. All these data are identical for the complete and incomplete filters. The optimal parameters that define the normalized resistances R_{2n} and R_{3n} for the complete and the incomplete filters are given in the third and fourth column of the table, respectively.

The next column lists the normalized frequency up to which the phase characteristic is linear within an arbitrarily chosen error of $\pm 3°$. The following column shows the insertion loss at this frequency. The next column gives the normalized group delay at low frequencies and it is followed by the normalized tolerance of the group delay up to f_u. Finally, the last column shows the worst-case return loss (a_r) for the complete/incomplete filters.

As an example, the phase characteristic of an incomplete filter is practically linear ($\pm 3°$ maximal deviation) up to an insertion loss of 17 dB for N=3 (i.e., up to $f_u = 2.15 f_{3dB}$). The return loss at any frequency is better than 14 dB.

The filter characteristics can be further optimized by letting all element values vary independently. The tuning can increase the bandwidth in which the group delay remains flat or it can improve the matching. The spread of element values remains very low, even for high filter orders.

III. BANDPASS FILTER REALIZATIONS

We want to synthesize bandpass filters that have a flat group delay up to a prescribed attenuation level and within a specified frequency band. We start from lowpass prototypes described in Section II and implement the standard lowpass-to-bandpass filter transforms [3, p.157]. Each coil of the lowpass filter, of normalized inductance L_n, is replaced by a series LC resonant circuit with $L_{sn} = L_n / B_n$ and $C_{sn} = B_n / L_n$, where $B_n = \Delta\omega_{3dB} / \omega_0$ is the normalized 3 dB bandwidth, $\Delta\omega_{3dB}$ is the 3 dB bandwidth, and ω_0 is the central angular frequency of the bandpass filter. (The normalization for the bandpass filter is similar as for the lowpass filter, except that ω_0 is used instead of ω_{3dB}.) Each capacitor of the lowpass filter, of normalized capacitance C_n, is replaced by a parallel resonant circuit with $C_{pn} = C_n / B_n$ and $L_{pn} = B_n / C_n$. Denormalized values of the elements of the bandpass filter are finally: $L_s = L_n Z_c / \Delta\omega_{3dB}$, $C_s = \Delta\omega_{3dB} / (L_n Z_c \omega_0^2)$, $C_p = C_n / (\Delta\omega_{3dB} Z_c)$, and $L_p = \Delta\omega_{3dB} Z_c / (C_n \omega_0^2)$.

In this section, we follow an example of an incomplete filter with $N = 3$ and $B_n = 0.1$. By transforming the lowpass filter shown in Fig. 1(b), we obtain the bandpass filter shown in Fig. 2. Normalized elements of this filter are given in Table II.

We also develop a transmission-line realization of the incomplete filter. We replace each resonant circuit by a quarter-wavelength transmission line. A series circuit (L_s, C_s) is replaced by an open-circuited line whose characteristic impedance is $Z_{cs} = \frac{4}{\pi}\sqrt{\frac{L_s}{C_s}} = \frac{4}{\pi}\omega_0 L_s = \frac{4}{\pi}\frac{L_n}{B_n} Z_c$ (i.e., the normalized characteristic impedance is $Z_{csn} = \frac{4}{\pi}\frac{L_n}{B_n}$). A parallel resonant circuit (L_p, C_p) is replaced by a short-circuited line whose characteristic impedance is $Z_{cp} = \frac{\pi}{4}\sqrt{\frac{L_p}{C_p}} = \frac{\pi}{4}\frac{1}{\omega_0 C_p} = \frac{\pi}{4}\frac{B_n Z_c}{C_n}$ (i.e., .the normalized characteristic impedance is $Z_{cpn} = \frac{\pi}{4}\frac{B_n}{C_n}$). An incomplete

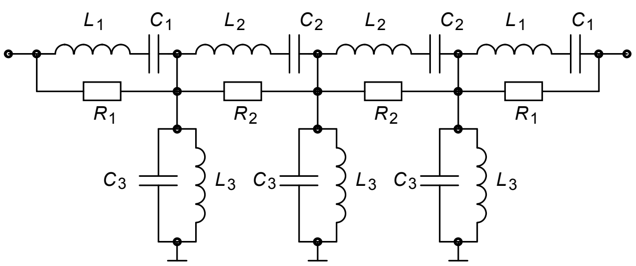

Fig. 2. Bandpass lumped-element incomplete filter ($N = 3$).

[2] In the optimization, all cells have identical element values.

TABLE II
NORMALIZED ELEMENTS OF THE INCOMPLETE BANDPASS FILTER

N	L_{1n}	C_{1n}	L_{2n}	C_{2n}	L_{3n}	C_{3n}	R_{1n}	R_{2n}
3	4.2	0.238	8.4	0.119	0.119	8.4	1	4.35

bandpass filter of this kind is shown in Fig. 3 (for $N = 3$). For the same bandwidth as for the lumped-element filter shown in Fig. 2, the normalized characteristic impedances in Fig. 3 are $Z_{c1n} = 10.7$ and $Z_{c2n} = 0.093$. (Note that $Z_{c1n} Z_{c2n} = 1$.) The normalized resistances are the same as in Table II.

Figure 4 shows the corresponding scattering parameters and group delay for the lumped-element and transmission-line realization of the incomplete bandpass filter, as computed using *Microwave Office*[3]. All capacitors, inductors, and transmission lines are assumed lossless. The characteristics of the lumped-element filter are asymmetric on the linear frequency scale. The characteristics of the transmission-line filter are symmetric, but at the expense of having parasitic passbands because the scattering parameters are periodic functions of frequency.

IV. MICROSTRIP IMPLEMENTATION AND EXPERIMENTAL RESULTS

Our next goal is to design a compact microstrip bandpass filter and exploit advantages of printed-circuit techniques. One problem with the structure shown in Figure 3 is that the transmission lines in series branches have a very high characteristic impedance (Z_{c1}). Hence, we replace each line by two open-circuited lines (of two times lower characteristic impedance, $Z_{c1}/2$) connected in series, as shown in Fig. 5. This results in wider printed traces for these lines. The short-circuited transmission lines have a low characteristic impedance (Z_{c2}), i.e., their conductors are very wide.

The lower conductor of each transmission line whose characteristic impedance is Z_{c2} can be grounded at all points. Such lines are directly produced as wide microstrip lines. We note that the three transmission-line conductors joined by the dashed lines in Fig. 5 are at the same potential at the junction point. These three conductors can be merged. Hence, we can produce each transmission line of characteristic impedance $Z_{c1}/2$ as a narrow microstrip line whose ground conductor coincides with the top ("hot") conductor of the short-circuited transmission line. This results in a multilayer microstrip structure that is shown in Fig. 6. The bottom metallization layer is ground. The low-impedance microstrip lines are printed on a thin substrate (middle metallization layer). The high-impedance microstrip lines are printed on a thick substrate (upper metallization layer). The top view of the structure is shown in Fig. 7. Vias are used to bring up connections for SMD resistors, which are soldered on the top of the printed structure. The wide traces in the middle layer are connected to the ground by another set of vias.

To properly take into account couplings among resonators, the filter was analyzed as a multiconductor transmission line using programs from [4] and [5]. These couplings reduce the bandwidth of the filter. This is advantageous as the span of

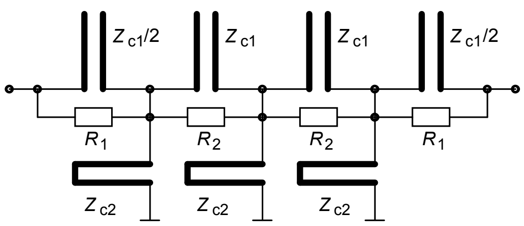

Fig. 3. Bandpass transmission-line incomplete filter ($N = 3$).

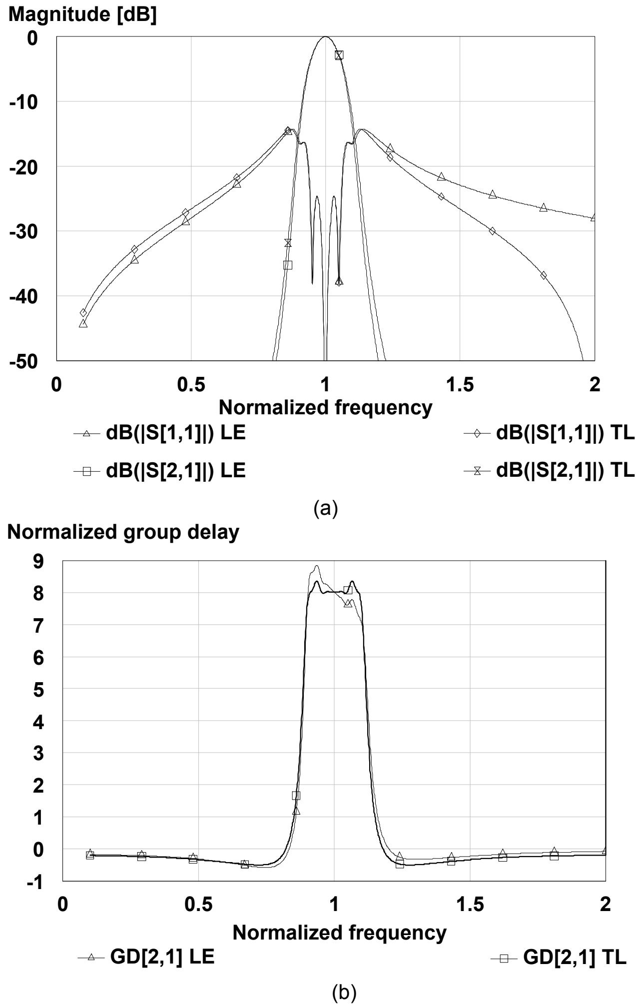

Fig. 4. Bandpass filter characteristics: (a) scattering parameters, and (b) group delay. LE refers to the lumped-element filter, and TL refers to the transmission-line filter.

[3] Microwave Office 2001, Applied Wave Research, Inc.

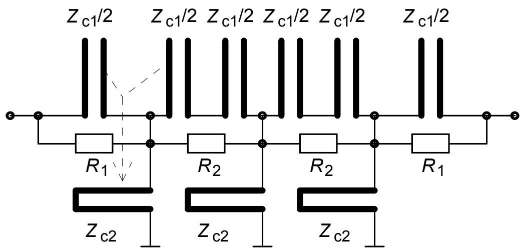

Fig. 5. Modification of the structure shown in Fig. 4.

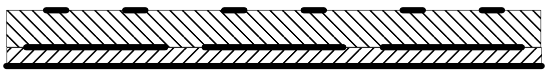

Fig. 6. Cross section of multilayer structure.

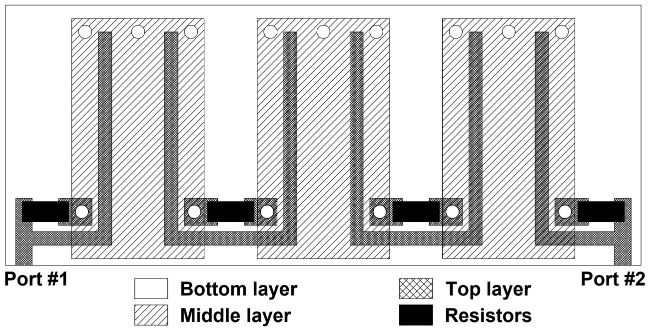

Fig. 7. Simplified layout of microstrip implementation (not drawn to scale).

characteristic impedances is lowered for a given bandwidth.

As an example, we present results for a filter whose central frequency is $f_0 = 900\ \mathrm{MHz}$, the relative bandwidth is $B_\mathrm{n} = 0.24$, and the nominal impedance is $Z_\mathrm{c} = 50\ \Omega$. The thickness of the lower substrate is 0.254 mm. The thickness of the upper substrate is 0.508 mm. The relative permittivity of both substrates is 2.33. The width of each high-impedance trace is 0.43 mm and the spacing between each pair of adjacent traces is 0.8 mm. The width of each low-impedance trace is 3.0 mm and the spacing is 2.0 mm. Standard metal-film SMD resistors (1206 size) are used, with $R_1 = 51\ \Omega$ and $R_2 = 180\ \Omega$.

The low-impedance and high-impedance lines have slightly different effective permittivities. To compensate for the different electrical lengths, the high-impedance lines should be terminated at their open ends by small capacitances (0.4 pF). In addition, to equalize velocities of modal propagation, we place a compensating mutual capacitance (0.04 pF) between each pair of adjacent high-impedance lines, at their open ends. Hence, each pair of adjacent high-impedance lines is terminated in a π-network of capacitors. These capacitors can be technically created in a variety of ways. We use the standard technique of increasing the width of the traces and reducing the spacing between them near the open ends of high-impedance lines.

Fig. 8 shows the simulated and measured response of the filter. Very good agreement has been achieved between the

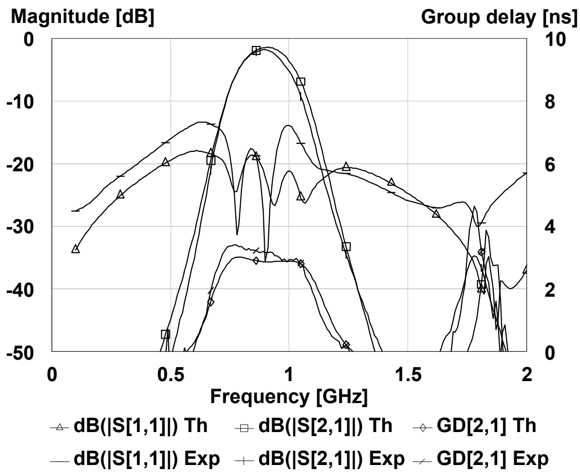

Fig. 8. Computed (Th) and measured (Exp) response of a microstrip filter.

two sets of results. The filter has an insertion loss of about 1.8 dB at the central frequency due to the conductor and dielectric losses. (The conductor losses dominate.) The first parasitic passband of the filter is at about twice the central frequency (1.8 GHz). This passband can be further suppressed by more carefully equalizing velocities of the guided modes. The next parasitic passband is at three times the central frequency (2.7 GHz). (This passband can be attrtibuted to the periodicity of the scattering parameters of the filter shown in Figure 5.) The filter is fairly insensitive to manufacturing tolerances and easy for tuning.

Acknowledgment

Authors thank Dr. Dejan Tošić for his useful suggestions during the preparation of the manuscript.

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