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OPTIMAL SYNTHESIS OF STUB MICROWAVE FILTERS

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ABSTRACT

Context. Microwave stub filters are widely used in radio engineering and telecommunication systems, as well as in technical information protection systems due to simplicity of design, possibility of realization in microstrip design and manufacturability in mass production. For synthesis of stub filters nowadays traditional methods based on transformation of low-frequency prototype filters on *LC*-elements into filtering structures on elements with distributed parameters are used. The transformations used are approximate and provide satisfactory results for narrowband stub filters. In this connection there is a necessity in development of direct synthesis methods for stub filters, excluding various approximations and providing obtaining of amplitude-frequency characteristics with optimal shape for any bandwidths.

Objective. The purpose of the study is to develop a method for direct synthesis of stub band-pass filters and low-pass filters with Chebyshev amplitude-frequency response in the passband.

Method. The procedure of direct synthesis includes the formulation of relations for filter functions of plume structures, selection of approximating functions of Chebyshev type for filter functions and formation of a system of nonlinear equations for calculation of parameters of filter elements.

Results. A method for the direct synthesis of stub bandpass and lowpass filters with Chebyshev response is developed.

Conclusions. Scientific novelty of the work consists in the development of a new method of direct synthesis of l stub filters. The method, in contrast to approximate traditional methods of synthesis of microwave filters, is exact, and the obtained solutions of synthesis problems are optimal.

The experiments confirmed the performance of the proposed method and the optimality of the obtained solutions. Prospects for further research suggest adapting the method to the synthesis of filter structures with more complex resonators compared to stubs.

KEYWORDS: synthesis, band-pass filter, low-pass filter, plume, scattering parameters, filter function, approximation.

ABBREVIATIONS

BPF is a band pass filter;
LPF is a low-pass filter;
FF is a filter function;
PB is a passband
SB is a stopband;
AFR is an amplitude-frequency response.

NOMENCLATURE

Θ is an electric length, rad;
ρ_s is a stub impedance, Ω;
ρ_c is a wave impedance of transmission lines, Ω;
ρ_0 is a port impedance, Ω;
R_i is a normalized wave impedance;
S_{ik} is a scattering parameter of the filter structure element;
$\hat{S}_{ik}$ is a filter scattering parameter;
α is a attenuation, dB;
f_0 is a center frequency, GHz;
f_c is a ripple cutoff frequency, GHz;
T is a transmission matrix.

INTRODUCTION

Microwave filters are the most important component of modern microwave systems of information processing, transmission, reception and protection. Among the designs of microwave filters a special place is occupied by stub filter structures, due to the simplicity of their structure and the possibility of its realization in planar microstrip design.

Synthesis of microstrip stub filters is usually carried out by the traditional method using low-frequency *LC*-prototype filters [1–6]. To transform the prototype filter into a filtering structure composed of elements with distributed parameters, approximate procedures are used, which do not provide optimal frequency characteristics of the filters.

The object of study. The object of research are procedures and methods of synthesis of filtering structures on elements with distributed parameters.

The subject of study. The subject of the study is methods of synthesizing stub filters with Chebyshev characteristic.

The purpose of the work. Based on the initial data for the synthesis, including the values of the center frequency f_0, relative bandwidth $2\Delta f/f_0$, ripple level α_c in the passband and attenuation level α_s in the fence band, it is required to determine the values of the wave impedances ρ_i of the filter elements, at which the ideal equal-wave AFR with a given level of ripple is ensured and the technological requirements for the value of the wave impedances are fulfilled.

1 PROBLEM STATEMENT

Realization of the method of direct synthesis of stub filters determines the necessity of development of a number of computational procedures, providing the

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compilation of the transfer function of the filter in a form convenient for synthesis, selection of the approximating function for the amplitude-frequency response and calculation of the parameters of the filter elements on the basis of the condition of physical realizability.

2 REVIEWS OF THE LITERATURE

The traditional method of synthesizing microwave stub filters is based on the transformation of a low-pass prototype filter composed of *LC*-elements into a structure consisting of elements with distributed parameters [1–6]. The prototype filter has a ladder structure, characterized by the cutoff frequency f_c=1 Hz and *g*-parameters representing the values of inductances and capacitances normalized by the resistance of the loads.

The transition from the low-frequency prototype filter to the filter, the scheme and values of *LC*-elements of which provide the cutoff frequency and bandwidth required by the synthesis task, is carried out according to the known formulas of frequency conversion [1–5].

Fulfillment of the requirement for physical realizability of the filter in the form of a planar structure is carried out by modifying the prototype filter circuit by introducing into it inverters that convert branches included in series into branches included in parallel.

To transform the prototype filter circuit into a filtering structure on elements with distributed parameters, the Richards frequency transformation [1–5] is used, according to which the inductance *L* is replaced by short-circuited and capacitance *C* by open-circuited sections of the transmission line on the basis of the relations $j\omega L=j\Omega\mathrm{tg}(\theta)$, $j\omega C=j\Omega\mathrm{tg}(\theta)$, где $\Omega=\mathrm{tg}(\theta)$ a frequency variable introduced by Richards, $\theta=\omega l/v$– electrical length of the transmission line segment, l – section length, ν – wave propagation velocity in the line. In turn, the concentrated *LC*-resonators are replaced by distributed resonators by equating their goodness-of-fit or admeasurement steepness [1–5]. Quarter-wave sections of transmission lines are used as inverters [1–5]. Equivalent transformations in the transmission line of serial branches into parallel ones are realized by means of Kuroda identities [1–6].

It should be noted that the frequency characteristics of resonators on short-circuited and open-circuited transmission line segments approximate with sufficient accuracy the characteristics of concentrated *LC*-resonators only near resonance, while quarter-wave line segments are characterized by perfect inversion only at the center frequency corresponding to the $\theta=\pi/2$. Thus, the conventional synthesis procedure is approximated by matching the performance of the prototype low-pass *LC*-filter and the corresponding stub filter in a narrow frequency band.

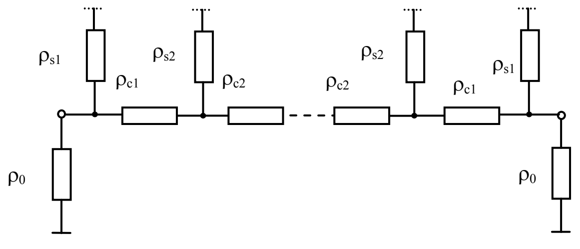
In [7–11], digital filters were proposed to be used as prototype filters for the synthesis of microwave filters. Type of transfer functions of digital filters $K_d(z)$, $z=\exp(j\omega\tau)$, composed by means of the discrete Laplace transform is identical to the representation of transfer functions of circuits composed of transmission line segments of the same length. In [9, 10] the method of calculation of circuit parameters on commensurate segments of the transmission line by the function $K_d(z)$ digital prototype filter [12]. The method includes selecting a suitable circuit configuration for realizing the function $K_d(z)$ and compilation of an autoregressive model for estimating the values of the coefficients of the transfer function of the circuit by the values of the coefficients of the function $K_d(z)$ of the prototype filter. The estimation is performed by the method of least squares. This synthesis method is an approximation. It should also be noted that the circuit configuration chosen for synthesis may not be optimal and may contain redundant elements in its structure.

Thus, there is a need to develop methods for direct synthesis of stub filters.

For stub filter structures with Chebyshev characteristic the method of direct synthesis is proposed in [14, 15]. The method includes the main stages of classical synthesis of electric filters [13]: selection of approximating function for a given filter structure; compilation of Hurwitz polynomial for physically realizable transfer function; determination of type and parameters of filter elements by transfer function. The stage of filter realization in this method is rather complicated and time-consuming. Therefore, there is a need to improve the efficiency of this method by simplifying the computational procedure at the stage of filter realization.

3 MATERIALS AND METHODS

Fig. 1 shows the investigated filter structures composed of combinations of quarter-wave stubs with wave impedance ρ_s and quarter-wave sections of transmission lines with wave impedance ρ_c. The input and output ports of the filters have wave impedance ρ_0. BPFs use short-circuited stubs, LPFs use open-circuited stubs. Filter structures have symmetry.



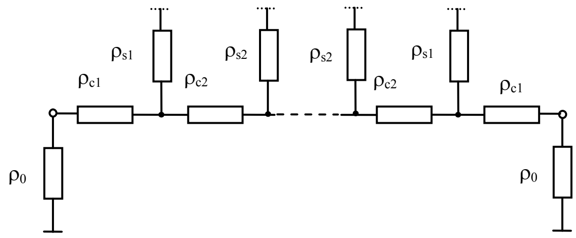


Figure 1 – Stub filter structures

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It is rational to analyze the n-cascade filter structure using the modified transfer matrix (T-matrix) by the formulas [14]:

$$\left.\begin{aligned} &\hat{T}=\begin{bmatrix}\hat{B} & -\hat{A}_{11}\\ \hat{A}_{11} & -\hat{A}\end{bmatrix}=\prod_{i=1}^{n}\begin{bmatrix}B_i & -A_{11_i}\\ A_{11_i} & -A_i\end{bmatrix},\\ &\hat{A}_{21}=\prod_{i=1}^{n}A_{21_i},\end{aligned}\right\} \tag{1}$$

where $\hat{A}=\dfrac{\hat{A}_{11}^2-\hat{A}_{21}^2}{\hat{B}}$, $A_i=\dfrac{A_{11_i}^2-A_{21_i}^2}{B_i}$.

On the basis of (1) the scattering parameters are determined $\hat{S}_{ik}=\hat{A}_{ik}/\hat{B}$ filter and its components $S_{ik_n}=A_{ik_n}/B_n$, i,k=1,2.

Let us represent the transfer matrices of the filter elements as follows:

– for the transmission line segment

$$\left.\begin{aligned} &T_c(\theta)=\cos(\theta)E+j\frac{\sin(\theta)}{2R_c}\begin{bmatrix}R_c^2+1 & -\left(R_c^2-1\right)\\ R_c^2-1 & -\left(R_c^2+1\right)\end{bmatrix},\\ &A_{21_c}=1;\end{aligned}\right\} \tag{2}$$

– for open-circuited stubs

$$\left.\begin{aligned} &T_s(\theta)=\cos(\theta)E+j\frac{\sin(\theta)}{2R_s}\begin{bmatrix}1 & 1\\ -1 & -1\end{bmatrix},\\ &A_{21_s}=\cos(\theta);\end{aligned}\right\} \tag{3}$$

– for short-circuited stubs

$$\left.\begin{aligned} &T_s(\theta)=\sin(\theta)E-j\frac{\cos(\theta)}{2R_s}\begin{bmatrix}1 & 1\\ -1 & -1\end{bmatrix},\\ &A_{21_s}=\sin(\theta).\end{aligned}\right\} \tag{4}$$

Here $j=\sqrt{-1}$ – imaginary unit, θ – electrical length of transmission line segments and stubs, $R=\rho/\rho_0$ – normalized wave impedance of lines, E – unit matrix.

Selective properties of the filter are characterized by its transfer function

$$\left|\hat{S}_{21}(\theta)\right|^2=\frac{1}{1+\left|F(\theta)\right|^2} \tag{5}$$

Here $F(\theta)$ is FF. For symmetric structures it is imaginary and is defined by the relation:

$$F(\theta)=j\hat{A}_{11}(\theta)/\hat{A}_{21}(\theta). \tag{6}$$

Analysis by formulas (1), (2), (4) of the filter structures in Fig. 1, consisting of n_s closed stubs and n_c of transmission lines gives the following relation for BPF:

$$\left.\begin{aligned} &\hat{A}_{11}(\theta)=j\sin(\theta)^{n_s-1}\gamma\sum_{k=0}^{m}a_k\cos(2k\theta),\\ &\hat{A}_{21}(\theta)=\sin(\theta)^{n_s},\end{aligned}\right\} \tag{7}$$

where $m=n_c/2$, $\gamma=\cos(\theta)$ at n_c even; $m=(n_c+1)/2$, $\gamma=1$ at n_c odd.

Accordingly, FF for BPF will take the following form:

$$F(\theta)=\frac{j\gamma\sum_{k=0}^{m}a_k\cos(2k\theta)}{\sin(\theta)}. \tag{8}$$

It should be noted that the function (8) does not depend on the type of circuits of the structures in Fig. 1, but is determined by the number of transmission lines in these structures.

For LPFs whose structures are composed of n_s of open stubs and n_c of transmission lines the result of analysis by (1), (2), (3) is the formula

$$\left.\begin{aligned} &\hat{A}_{11}(\theta)=j\sin(\theta)\sum_{k=0}^{m}a_k\cos(2k\theta),\\ &\hat{A}_{21}(\theta)=\cos(\theta)^{n_s},\end{aligned}\right\} \tag{9}$$

where $m=(n_s+n_c-1)/2$.

According to (9) FF for LPF will take the following form:

$$F(\theta)=\frac{j\sin(\theta)\sum_{k=0}^{m}a_k\cos(\theta)^{2k}}{\cos(\theta)^{n_s}}. \tag{10}$$

One of the main tasks of the synthesis is to compose an approximating function on FF of the investigated structure, providing the given requirements on AFR of the filter. For BPF with FF (8) the approximating function can be a Chebyshev function of the following form

$$T(\theta)=\mathrm{ch}\left[n_s\mathrm{arch}\left(\frac{\mathrm{ctg}(\theta)}{\mathrm{ctg}(\theta_c)}\right)+n_c\,\mathrm{arch}\left(\frac{\cos(\theta)}{\cos(\theta_c)}\right)\right], \tag{11}$$

where θ_c – angle at the filter bandwidth boundary in terms of ripple level.

Let us represent this function as follows:

$$T(\theta)=\frac{A^{(+)}(\theta)+A^{(-)}(\theta)}{2\sin(\theta)}, \tag{12}$$

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where $A^{(\pm)}(\theta)=\left(\frac{S}{C}\cos(\theta)\pm Q(\theta)\right)\left(\frac{\cos(\theta)}{C}\pm Q(\theta)\right)^{n_c}$,

$Q(\theta)=\sqrt{\cos(\theta)^2/C^2-1}$, $S=\sin(\theta_c)$, $C=\cos(\theta_c)$.

Ratio (12) can be written in a form similar to FF (8)

$$T(\theta)=\frac{\beta\sum_{k=0}^{m}\alpha_k\cos(2k\theta)}{\sin(\theta)}, \qquad (13)$$

where $m=n_c/2$, $\beta=\cos(\theta)/C^{n_c+1}$ at n_c even; $m=(n_c+1)/2$, $\beta=0,5/C^{n_c+1}$ at n_c odd.

Table 1 presents a series of relations for the coefficients in (13) for a range of values of n_c.

For LPF with FF (10) the Chebyshev approximating function can be written in the form

$$T(\theta)=\mathrm{ch}\left[n_s\,\mathrm{arch}\left(\frac{\mathrm{tg}(\theta)}{\mathrm{tg}(\theta_c)}\right)+n_c\,\mathrm{arch}\left(\frac{\sin(\theta)}{\sin(\theta_c)}\right)\right]. \qquad (14)$$

Let us represent this function as follows

$$T(\theta)=\frac{A^{(+)}(\theta)+A^{(-)}(\theta)}{2\cos(\theta)^{n_s}}, \qquad (15)$$

where $A^{(\pm)}(\theta)=\left(\frac{C}{S}\sin(\theta)\pm Q(\theta)\right)^{n_s}\left(\frac{\sin(\theta)}{S}\pm Q(\theta)\right)^{n_c}$,

$Q(\theta)=\sqrt{\sin(\theta)^2/S^2-1}$.

Transforming (15), we obtain

$$T(\theta)=\frac{\sin(\theta)\sum_{k=0}^{m}\alpha_k\cos(2k\theta)}{\cos(\theta)^{n_s}}, \qquad (16)$$

where $m=(n_s+n_c-1)/2$.

In Table 2, where $\eta=2S^{n_c+n_s}$, for a range of values n_c, n_s are presented expressions for the coefficients of the rows in (16).

The process of filter synthesis begins with determining the number of elements of the selected circuit on the basis of the following initial data: f_0 – center frequency corresponding to the electric length $\theta_0=\pi/2$; α_c – attenuation in terms of ripple level in the passband, *dB*; f_c – bandwidth limit frequency; α_s – attenuation at frequency f_s in the stopband, dB.

The number of elements is determined by the formulas derived from the approximating functions (13), (16):

– for BPF

$$n_c=\frac{\mathrm{arch}\left(\sqrt{\frac{10^{\frac{\alpha_s}{10}}-1}{10^{\frac{\alpha_c}{10}}-1}}\right)-\mathrm{arch}\left(\frac{\mathrm{ctg}(\theta_s)}{\mathrm{ctg}(\theta_c)}\right)}{\mathrm{arch}\left(\frac{\cos(\theta_s)}{\cos(\theta_c)}\right)}, \qquad (17)$$

Table 1 – Coefficients of the Chebyshev function of the stub BPF

n_c	n_s	α_0	α_2	α_4	α_6
1	2	$-1+S+2S^2$	1+S		
2	1 3	$-1+2S^2+S^3$	1+S		
3	2 4	$S^2+3S^3+2S^4$	$-1+S+5S^2+3s^3$	1+S	
4	3 5	$1-2S^2+2S^3+4S^4+S^5$	$-2+6S^2+4S^3$	1+S	
5	4 6	$S^2+5S^3+7S^4+5S^5+2S^6$	$2S^2+10S^3+13S^4+5S^5$	$-1+S+7S^2+5S^3$	1+S
6	5 7	$1+4S^2+3S^3+S^4+6S^5+6S^6+S^7$	$2-6S^2+6S^3+19S^4+9S^5$	$-2+8S^2+6S^3$	1+S

Table 2 – Coefficients of the Chebyshev function of the stub LPF

n_c	n_s	α_0	α_2	α_4	α_6
1	2	$-(1+C)(1+C-4C^2)/\eta$	$-(1+C)^2/\eta$		
2	1	$(-2+4C^2+2C^3)/\eta$	$-2(1+C)/\eta$		
2	3	(1+C)2(3–3C–8C2+14C3)/2η	(1+C)2 (2–C–4C2–C3)/η	(1+C)3/2η	
3	2	$(1+C)^2(2-4C+C^2+4C^3)/\eta$	$(1+c)^2(3-4C-3C^2)/\eta$	$(1+C)^2/\eta$	
3	4	$(1+C)^3(-7+17C+C^2-39C^3+48C^4)/4\eta$	$-(1+C)^4(11-32C+20C^2+16C^3)/4\eta$	$(1+C)^4(-5+8C+3C^2)/4\eta$	$-(1+C)^4/4\eta$
4	3	$(-2+7C^2-5C^4+13C^5+20C^6+7C^7)/\eta$	$-(1+C)^3(7-18C+6C^2+18C^3+2C^4)/2\eta$	$(1+C)^3(-2+3C+2C^2)/\eta$	$-(1+C)^3/2\eta$

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– for LPF

$$n_c = \frac{\operatorname{arch}\left(\sqrt{\dfrac{10^{\frac{\alpha_s}{10}}-1}{10^{\frac{\alpha_c}{10}}-1}}\right) - \delta \operatorname{arch}\left(\dfrac{\operatorname{tg}(\theta_s)}{\operatorname{tg}(\theta_c)}\right)}{\operatorname{arch}\left(\dfrac{\operatorname{tg}(\theta_s)}{\operatorname{tg}(\theta_c}\right) + \operatorname{arch}\left(\dfrac{\sin(\theta_s)}{\sin(\theta_c}\right)}, \quad (18)$$

where $\theta_c = \pi f_c / 2 f_0$, $\theta_s = \pi f_s / 2 f_0$, $\delta = \pm 1$.

According to the values calculated by (17), (18) n_c and $n_s = n_c + \delta$ the coefficients of polynomials α_k in (13), (16) are calculated.

Wave impedances of the filter circuit elements are determined by solving the system of nonlinear equations

$$a_k(R) - \varepsilon \alpha_k = 0, k = 1, \ldots m, \quad (19)$$

where $a_k(R)$ – coefficients of the polynomials in (8), (10); R – vector formed from the sought wave impedances, $\varepsilon = \sqrt{10^{\alpha_c/10} - 1}$ – ripple value in the filter passband.

The system (19) is solved by Newton's method. At each iteration the coefficients $a_k(R)$ are calculated by decomposition $\hat{A}_{11}(\theta) = T_{21}(\theta)$ from (1) into Fourier series by $\cos(2k\theta)$. To solve the system (19) can also be used methods of solving optimization problems with constraints in case of the need to take into account technological tolerances on the value of line impedances.

4 EXPERIMENTS

For approbation and substantiation of reliability and efficiency of the proposed method, the results of synthesis of broadband BPF and LPF are presented and the obtained results are compared with the data of calculations by methods based on prototype filters.

Stub BPF synthesis is performed for the following data: f_0=2 GHz; α_c=0.1 dB (ε=0.1526); α_s=40 dB; f_s=3,5 GHz and for two variants of calculation: f_c=1.5 GHz at relative bandwidth $2\Delta f/f_0$=0.5 (50%) and f_c=1.0 GHz, at $2\Delta f/f_0$=1.0 (100%).

For these data, the type and number of elements in the circuits are determined by (17): for the circuit in Fig. 1a, four transmission lines are obtained n_c=4 and five stubs' n_s=5; for the scheme in Fig. 1b – n_c=4, n_s=3. For f_c=1,5 GHz for the specified filter structures the coefficients of the rows in (13) took the following values: α=543.100; 764.645; 234.412.

For the circuit in Fig. 1a the required vector of normalized resistances is: $R=R_{s1}$, R_{c1}, R_{s2}, R_{c2}, R_{s3}. From the solution of the system (19) we can obtain a number of normalized values of R, depending on the initial approximation R_0. For example, for the given values of the coefficients of the polynomial, the following is obtained:

– R=0.4889; 1.0160; 0.5417; 1.3839, 0.7353 to an initial approximation R_{0k}=0.5;

– R=0.4745; 1.0847; 0.6855; 1.3575, 0.5407 for R_{0k}=1.0;

– R=0.4471; 1.2613; 1.0229; 1.8692, 0.7695 for R_{0k}=1.5, k=1,...,5.

For the circuit in Fig. 1, b the required resistances are: $R=R_{c1}$, R_{s1}, R_{c2}, R_{s2}. At the initial approximation R_{0k}=0.5 these resistances obtained the following values: R=0.3301; 0.0378; 0.2412; 0.6866.

By the traditional synthesis method using the formulas [1, 4] compiled when converting the LC prototype filter to a stub filter, the calculation for the circuit in Fig. 1a with the n_c=4, n_s=5 gave the following values for the resistances: R=0.5674; 0.7732; 0.2883; 0.7175; 0.2931

Fig. 2 shows the AFR BPF in the form of dependencies $S_{21}(f)_{dB} = -20 lg\left|\hat{S}_{21}(f)\right|$. In this figure and in the following figures, curve 1 corresponds to the calculation by the developed method, curve 2 – to the traditional methods. AFRs in the filter bandwidth are given in the center of the figures.

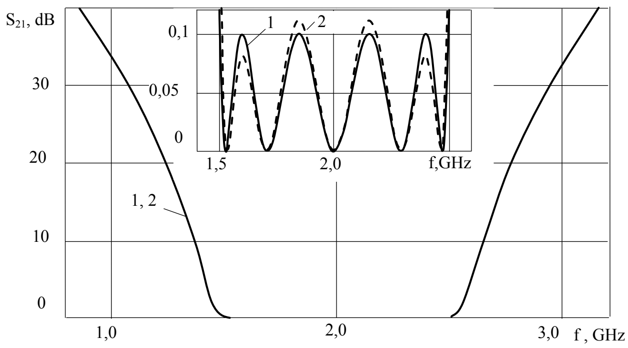

Figure 2 – AFR BPF with relative bandwidths 50%

In Fig. 3 shows the AFRs for fc=1.0 GHz at relative bandwidth $2\Delta f/f0$=1.0. For these data the coefficients of the rows in (13) took the following values: α=10.6568; 13.6568; 9.6568, For these values from the solution (19) at initial approximation R_{0k}=0.5 received:

– R=2.4384; 0.7909; 1.4946; 0.6894; 1.0649 for the scheme in Fig. 1a;

– R=0.5972; 0.6199; 0.3776; 0.500 for the scheme in Fig. 1b.

Calculation by the traditional method according to formulas from [1,4] for the scheme in Fig. 1a gives R=2.2977; 0.7732; 1.1815; 0.7175; 1.2161.

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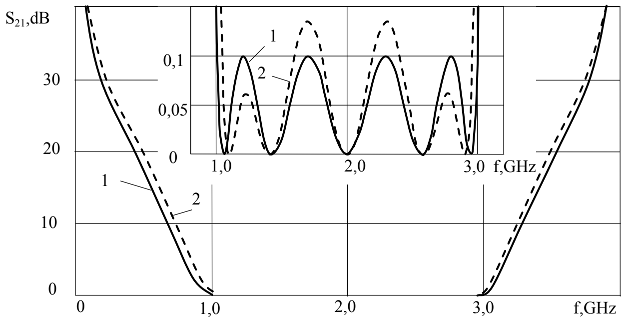

Figure 3 – AFR BPF with relative bandwidths 100%

Fig. 4 shows the results of BPF synthesis with relative bandwidth 2Δ*f*/*f*0=0.4 and attenuation α_c=0.5 dB in the passband. In this figure, curve 2 is plotted based on the transfer function of the prototype digital filter when the stub filter structure formed by of eight stubs and seven transmission lines transmission lines was selected for synthesis [9]. In the simulation, this structure was represented by seven sections. Each section consisted of two stubs connected by a section of transmission line. As a result of the synthesis, the resistances of the elements of the sections, represented as follows (ρ_{s1}, ρ_{s2}, ρ_{c1}), took the following values at port resistance 50 Ω: (27.1; 98.6; 49.5), (98.9; 96.3; 89.5), (47.5; 53.7; 92.0), (56.3; 77.1; 64.8), (75.8; 94.0; 31.7), (85.9; 59.9; 27.3) [9].

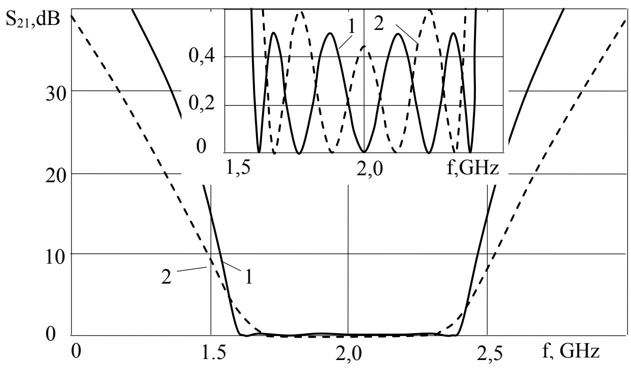

Figure 4 – AFR BPF with relative bandwidths 40%

AFR LPF with relative bandwidth $2\Delta f/f_0$=1.0 c are shown in Fig. 5.

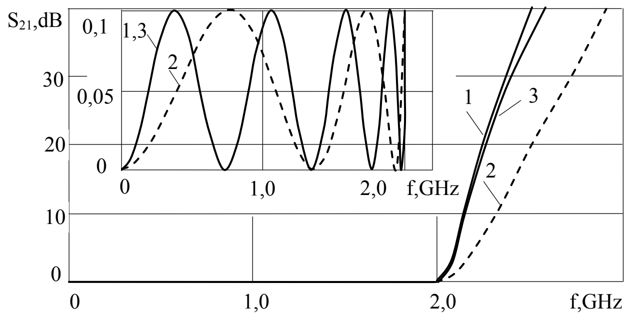

Figure 5 – AFR LPF with relative bandwidths 100%

Curve 1 in Fig. 4 corresponds to the calculation developed by the direct synthesis method for a structure of five stubs and four transmission lines with normalized impedances *R*=0.229; 0.89; 0.334; 1.05; 0.249.

When synthesizing LPF with f_0=2 GHz; α_c=0.1 dB, coefficients of the rows in (16) at $2\Delta f/f_0$=1.0 take the following values:

– α=118.1684; –203.4698; 138.4215 –62.9662; 20.5030. for the scheme in Fig. 1a at n_c=4, n_s=5;

– α=142.9481; –246.6830; 166.6154; –75.8302; 24.0208 for the scheme on 1b at n_c=5, n_s=4.

The corresponding values of resistances obtained from (19):

– *R*=0.8417; 1.8135; 0.4828; 1.9622; 0.4622 for the scheme in Fig. 1a;

– *R*=1.7378; 0.8095; 2.5761; 0.6376; 2.6621 for the scheme on 1b.

By the traditional method using Richards transform, inverters and Kuroda identities it is obtained for scheme 1a: *R*=2.872; 1.5342; 0.6241; 1.837; 0.5063 [1, 6].

Curves 1,3 in the figure are obtained by the developed synthesis method, curve 1 corresponds to the scheme in Fig. 1a, curve 3 corresponds to the scheme in Fig. 1b.

5 RESULTS

As follows from Figs. 2–4, curves 1 obtained by the developed synthesis method have an ideal equal-wave shape and fully satisfy the synthesis assignment.

AFR BPF of schemes in Fig. 1a and 1b, at equal number of connecting transmission lines are identical, their frequency characteristics at n_c=4 are presented in Fig. 2–3 by one curve 1.

The shape of curves 2 in Figs. 2–3, obtained by the conventional synthesis method according to the low-pass *LC*-prototype filter using ideal inverters, does not correspond to the equal-wave characteristic, and the attenuation in the passband exceeds the specified value. The mismatch grows with the increase of the passband, at the same time its narrowing occurs.

AFR BPF, presented in Fig. 4, curve 2, is calculated for the stub filter structure of 15 elements by the transfer function of the prototype digital filter. This characteristic does not fully satisfy the conditions of the synthesis task. These conditions, curve 1, were met when applying the developed synthesis method to the filter structure with a much smaller number of elements, equal to 9 elements.

The AFR LPFs in Fig. 4, calculated for the schemes in Figs. 1a and 1b by the developed method (curves 1 and 3), have an equal-wave shape in the passband, matching the graphical accuracy. However, outside the passband, the scheme in Fig. 1a has a higher steepness of the AFR decay in the fence band.

Curve 2 in Fig. 4, corresponding to the traditional synthesis method using Kuroda identities, also has an equal-wave AFR in the passband, but since only stub resonators were involved in the formation of the AFR, this characteristic has a smaller number of ripples in the passband and a much smaller slope steepness in the obstruction bandwidth.

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6 DISCUSSIONS

The presented calculation results confirm the validity and efficiency of the developed direct method of loop filter synthesis. Unlike approximate traditional synthesis methods, the results of synthesis by the developed method fully correspond to the technical specification, providing a strictly equal-wave frequency response with a given level of ripple in the passband and a given value of attenuation in the barrier band for filters with narrow and wide passbands. It should also be noted that in the developed method all elements of filtering structures are involved in the formation of the AFR of filters, so at the stage of realization of filters their microstrip structures will be more compact in comparison with the structures realized according to the results of traditional synthesis procedures.

From the presented results of synthesis by the developed method, it follows that the BPF based on the schemes in Fig. 1a and 1b have matching AFR at the same number n_c of transmission lines. The advantage of the scheme in Fig. 1b is the smaller number of elements. However, the values of wave impedances of elements in this scheme may not meet the conditions of technological realizability, in particular, for narrowband filters. For microstrip structures at wave impedance of ports ρ_0=50 Ω the permissible values of wave impedances of lines lie in the range of 15–150 Ω. These conditions are largely satisfied by the values of the wave impedances of the elements obtained as a result of synthesis for the circuit in Fig. 1, a.

Frequency response of the LPF based on the schemes in Fig. 1a and 1b with the same number of elements have an equal-wave shape in the passband, which coincides with the graphical accuracy. As for the fence band, the AFR of the scheme in Fig. 1a has a steeper decline. This scheme also corresponds more to the technological conditions on the realized in the process of synthesis value of wave impedances of its elements. Thus, when choosing a scheme for the BPF or LPF, the scheme in Fig. 1a should be preferred.

The proposed method is simple in program implementation, since the basis of the program algorithm is formed by known, well-developed computational procedures from the mathematical support of the Fourier transform and the solution of systems of nonlinear equations [16]. If necessary, in the program algorithm at the stage of parametric optimization, it is easy to add restrictions on the values of wave impedances of the elements of the synthesized filter circuits when composing the target function.

CONCLUSIONS

The actual problem on improvement of design methods of microwave stub filters is solved.

The scientific novelty. Scientific novelty of the obtained results consists in the fact that for the first time a method of direct synthesis is proposed, which allows obtaining optimal amplitude-frequency characteristics of stub filters directly from the filtering function by means of simple computational procedures of decomposition of functions into Fourier series and solving systems of nonlinear equations.

The practical significance. The practical value of the obtained results consists in the fact that the filter structures synthesized by the developed method are optimal both by the number of elements contained in them and by compliance with the conditions of the technical task for synthesis. The program realization of the developed method does not cause difficulties, and its use in computational experiments requires insignificant computational and time resources. Application of the developed method and program in computer-aided design systems will allow to improve the quality of design solutions in the synthesis of stub filters.

Prospects forfurther research. Prospects for further research consist in extending the proposed approach to the solution of problems of synthesizing a wide range of diverse filtering structures composed of commensurate sections of transmission lines.

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ОПТИМАЛЬНИЙ СИНТЕЗ ШЛЕЙФНИХ МІКРОХВИЛЬОВИХ ФІЛЬТРІВ

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АНОТАЦІЯ

Актуальність. Мікрохвильові шлейфні фільтри знаходять широке застосування в радіотехнічних і телекомунікаційних системах, а також у системах технічного захисту інформації завдяки простоті конструкції, можливості реалізації в мікросмужковому виконанні та технологічності при масовому виготовленні. Для синтезу шлейфових фільтрів наразі використовують традиційні методи, засновані на перетворенні низькочастотних фільтрів-прототипів на LC-елементах на фільтрувальні структури на елементах із розподіленими параметрами. Використовувані перетворення є наближеними і забезпечують задовільні результати для вузькосмугових шлейфових фільтрів. У зв’язку з цим виникає необхідність у розробленні для шлейфних фільтрів методів прямого синтезу, що виключає різного роду наближення і забезпечує отримання амплітудно-частотних характеристик з оптимальною формою для будь-яких смуг пропускання.

Мета. Метою дослідження є розробка методу прямого синтезу шлейфних смугово-пропускних фільтрів і фільтрів нижніх частот із чебишевською амплітудно-частотною характеристикою в смузі пропускання.

Метод. Процедура прямого синтезу містить у собі складання співвідношень для функцій фільтрацій шлейфових структур, підбір апроксимувальних функцій чебишевського типу для функцій фільтрацій і формування системи нелінійних рівнянь для обчислення параметрів елементів фільтрів.

Результат. Розроблено метод прямого синтезу шлейфних смугово-пропускних і фільтрів нижніх із чебишевською характеристикою.

Висновки. Наукова новизна роботи полягає в розробленні нового методу прямого синтезу шлейфних фільтрів. Метод, на відміну від наближених традиційних методів синтезу мікрохвильових фільтрів, є точним, а одержувані розв’язки задач синтезу – оптимальними.

Проведені експерименти підтвердили працездатність пропонованого методу й оптимальність одержуваних рішень. Перспективи подальших досліджень передбачають адаптацію методу на синтез фільтрувальних структур зі складнішими порівняно зі шлейфами резонаторами.

КЛЮЧОВІ СЛОВА: синтез, смугово-пропускний фільтр, фільтр нижніх частот, шлейф, параметри розсіювання, функція фільтрації, апроксимація.

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