УДК 621.396, 621.3.049

Synthesis of Extremely Wide Stopband E-plane Bandpass Filters

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The article presents the results of the development of a novel procedure for the direct synthesis of bandpass E-plane evanescent mode filters. Unlike previously developed synthesis procedures the proposed new and quick synthesis technique makes it possible to obtain the dimensions of all elements of the filter topology, which either do not require further time-consuming optimization, or optimization is reduced to re-application of the proposed technique with specially changed performance requirements. The developed technique was adequate in the development of the proposed E-plane filters, built on segments of the antipodal finline with the significant overlap of its ridges in the evanescent mode rectangular waveguide. Under this conditions the reduction of resonators relative to half the wavelength reaches 80 %. Proposed E-plane implementation of the evanescent mode rectangular waveguide filter allows significantly expanding the bandwidth, increase the attenuation introduced in it and at the same time ensure the repeatability of the characteristics of the filters without any action to adjust them. The effectiveness of the proposed approach to the implementation of the filter and its calculation was demonstrated in the development and experimental study of a 21 GHz filter, which in terms of parameters (loss of about 1 dB, stopband up to the fourth harmonic of the central frequency of passband) meets high requirements for electrical characteristics and cost. It is shown that the developed method of synthesis of such filters remains relevant in the synthesis of filters with a relative bandpass width of up to 40 %.

Key words: bandpass filter; millimeter wave generators; RF hybrid integrated circuit; stopband; evanescent mode waveguide

DOI: 10.20535/RADAP.2021.84.22-29

Introduction

Stopband width is one of the most important parameter of microwave bandpass filters (BPF). In typical BPFs on half-wave resonators with central frequency f_0, the second passband is located at frequencies that do not exceed $(1.3 \div 1.4)f_0$, which is also accompanied by a small attenuation in the stopband. Such characteristics are often unacceptable for a number of today's applications: increasing the nominal values of the first intermediate frequencies in receivers to several gigahertz requires expanding the stopband of the image channel filters, insufficient protection of broadband circuit analyzers from local oscillator harmonics generates parasitic responses. The need for broadband filtering arises in the development of modern passive radar systems, the development of compact satellite terminals that operate on a single antenna, and so on. The need for filters with ultra-wide stopbands has led to a significant amount of papers concerned to this problem [1–4]. Among these filters, evanescent mode waveguide filters play a special role, as their stopband can reach several octaves, and the high quality of resonators allows to successfully use these filters at frequencies of centimeter and millimeter wavelengths [5,6]. The waveguide-planar implementation of such filters (E-plane filters), proposed in [7, 8] and further developed in a number of works [9, 10], can significantly increase the flexibility of structures and reduce their cost. As the E-plane implementation of evanescent mode waveguide filters excludes any procedures regarding their tuning the role of calculation accuracy increases. The final dimensions of these filters elements can be determined only in the procedure of their optimization, considering the diffraction on the entire structure. This is internally connected to the principles of construction of such filters. Therefore, the cost of developing filters largely depends on the initial values of the topology elements sizes, which can be determined within the appropriate synthesis procedure. The aim of this work was to develop a method of evanescent mode waveguide filters synthesis, which would require minimal effort in further optimization of the found dimensions.

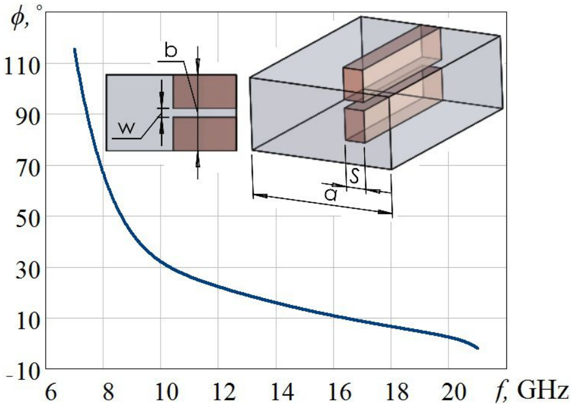
1 Synthesis of evanescent mode waveguide filters

Problems with the synthesis of evanescent mode waveguide filters are related to the length of their resonators l. In the approximation of the circuits theory, it can be found from the relation

$$\beta l = \frac{1}{2}\left(\varphi_i + \varphi_{i+1}\right), \qquad (1)$$

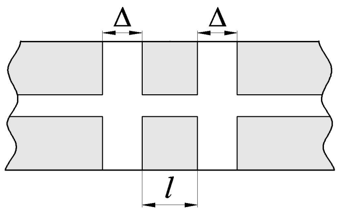
where β is the phase constant of the resonator line, and φ_i, φ_{i+1} are the phases of the reflection coefficients from the elements limiting of the resonator line. The typical frequency dependence of the phase of the reflection coefficient is shown in Fig.1a. Calculations were performed for the resonator line in the form of an double ridged waveguide by mode-matching technique. The dimensions are shown in the figure; the evanescent mode section is considered semi-infinite. It is seen that in the frequency range, where the rectangular waveguide is evanescent, i.e. where the resonator should be formed on the segment of double ridged waveguide, the phase is positive and close to zero, which causes a small length of the resonator. Thus, the resonator (Fig.1b) at a frequency of f_0=12.53 GHz (the dimensions of the components are the same as shown in Fig.1a), with the evanescent waveguide sections $\Delta = 10$ mm has a length l equal to 0.932 mm, which is $0.031\lambda_g$. This is what creates significant problems in the development of the synthesis methods. Classical methods for the synthesis of microwave BPFs involve the use of half-wave and quarter-wave resonators. For them, the parameters of the slope of the reactance x_i were found, which are included in the calculation formulas of the synthesis procedures [11]. Therefore, their direct use in the calculation of evanescent mode waveguide filters with significantly reduced resonators lengths is impossible. Secondly, there is a question concerned the possibility of applying relation (1) to calculate the lengths of resonators, because the values of φ_i and φ_{i+1} belong to two separate filter elements, and the extremely small distance between them violates their isolation. The latter may call into question the very possibility of developing an adequate method of synthesis of these filters, as this procedure involves the direct finding of each of its individual element.

These difficulties are reflected in different approaches to the procedure of synthesis of the considered filters. In the first works, the connections between the resonators were calculated from an equivalent circuit built for the first evanescent mode of waveguide by circuit theory, and the capacitances of the resonators were found in the quasi-static approximation. As the result were samples of filters that required refinement, which was performed experimentally. Subsequently, efforts were focused on the development of detailed electrodynamic models of the filter, and the sizes were determined in the procedures of their optimization in order to achieve the required characteristics. This also applies to the few works [12] in which the estimation of the initial dimensions was carried out within the developed synthesis procedures. The first work in which an acceptable accuracy of the dimensions of the elements of a BPF on the segments of the double ridged waveguide in an evanescent mode rectangular waveguide without the application of the optimization procedure was achieved was work [14]. In it, the general approach developed in [13], taking into account the frequency dependence of the inverter parameters, was used to calculate the considered filters. Note that the quality of the proposed synthesis procedure is indicated in the title of the work: "Dimensional synthesis. . . ", which shows that the proposed method of synthesis produces the final sizes, rather than the original data for further refinement.



The phase of reflection coefficient looking from ridged guide into evanescent mode waveguide

(a)



The phase of reflection coefficient looking from ridged guide into evanescent mode waveguide
(b)

Fig. 1. Ridged guide based evanescent mode resonator and its electrical characteristics

The paper assumes that the dimensions of the resonators can be calculated by relation (1), and the parameters of K-inverters are modified by including parts of resonator lines in the inverters. Since the modified values of the inverter parameters require some initial values, the synthesis procedure includes an iterative process in which the lengths of the resonators and the inverter parameters are determined at each step. In general, this method of synthesis allows you to find all the dimensions of the filter elements accurately enough, as confirmed by the simulation of the filter frequency response in the software package of electrodynamic analysis and experimentally, but the procedure is quite complex and requires a significant number of calculations. In addition, there are no estimates of the passband width and resonator sizes at which the technique is operational.

The basis of the proposed method of synthesis is the calculation of the parameters of the resonators reactive conductivity slope, which should take into account the fact that part of the resonator consists of limiting inhomogeneities. The necessary calculations can be performed based on the equivalent circuit of the resonator shown in Fig.2.

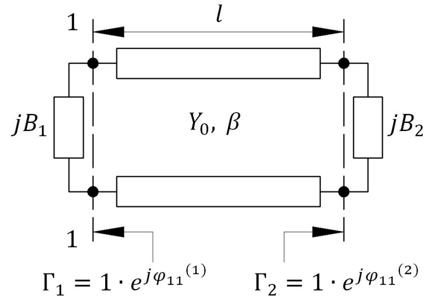


Fig. 2. Equivalent circuit model of resonator

According to Fig.2, the resonator consists of a line segment of length l bounded by reactive inhomogeneities jB_1, jB_2, which create modulus-unit reflection coefficients whose phases are equal to $\varphi_{11}^{(1)}$, $\varphi_{11}^{(2)}$, respectively. For the depicted model

$$\overleftarrow{B}_1 = -\tan\frac{\varphi_{11}^{(1)}}{2}\ \overrightarrow{B}_2 = -\tan\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right), \qquad (2)$$

where $\overleftarrow{B}$ and $\overrightarrow{B}_2$ denote the immittances on the left and right of the plane $1-1$. The total conductibility in this model is equal to

$$B = \overleftarrow{B} + \overrightarrow{B}_2 = -Y_0\left[\tan\frac{\varphi_{11}^{(1)}}{2} + \tan\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)\right] = = Y_0\frac{\sin\Theta}{\cos\dfrac{\varphi_{11}^{(1)}}{2}\cos\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)}, \qquad (3)$$

where $\Theta = \beta l - \frac{1}{2}\left(\varphi_{11}^{(1)} + \varphi_{11}^{(2)}\right)$.

The resonance condition $B = 0$ gives a resonant frequency that satisfies the condition $\Theta = 0$. To calculate the parameter of the resonators reactive conductivity slope we note that in order to minimize the effect of dispersion in the resonator line, it is advisable instead of the original definition of this parameter given in [11], use a modified value:

$$b = \frac{1}{2\lambda_{g0}} \cdot \left.\frac{\partial B}{\partial\left(\dfrac{1}{\lambda_{g0}}\right)}\right|_{\lambda_g = \lambda_{g0}}, \qquad (4)$$

where $\lambda_g 0$ – is the wavelength at the resonant frequency. Converting (4) is easy to obtain

$$b = \left.\frac{\beta_0}{2}\frac{\partial B}{\partial\omega} \cdot V_g\right|_{\omega=\omega_0}, \qquad (5)$$

where β_0, V_g – phase constant and group velocity in the line at the resonant frequency. Differentiating (3), we find

$$\frac{\partial B}{\partial\omega} = Y_0\left.\frac{\cos\dfrac{\partial\Theta}{\partial\omega}\left[\cos\dfrac{\varphi_{11}^{(1)}}{2} \cdot \cos\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)\right] - \sin\dfrac{\partial\Theta}{\partial\omega}\left[\cos\dfrac{\varphi_{11}^{(1)}}{2} \cdot \cos\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)\right]}{\left[\cos\dfrac{\varphi_{11}^{(1)}}{2} \cdot \cos\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)\right]^2}\right|_{\omega=\omega_0} \qquad (6)$$

At the resonant frequency $\sin\Theta|_{\omega=\omega_0} = 0$, $\cos\Theta|_{\omega=\omega_0} = 1$, $\left.\cos\left(\frac{\varphi_{11}^{(2)} - 2\beta l}{2}\right)\right|_{\omega=\omega_0} = \left.\cos\frac{\varphi_{11}^{(1)}}{2}\right|_{\omega=\omega_0}$,

$$\left.\frac{\partial\Theta}{\partial\omega}\right|_{\omega=\omega_0} = \left.\frac{\partial\beta}{\partial\omega}l - \frac{1}{2}\frac{\partial}{\partial\omega}\left(\varphi_{11}^{(1)} + \varphi_{11}^{(2)}\right)\right|_{\omega=\omega_0} = \left.\frac{l}{V_g} - \frac{1}{4\pi}\frac{\partial}{\partial f}\left(\varphi_{11}^{(1)} + \varphi_{11}^{(2)}\right)\right|_{\omega=\omega_0}.$$

Substituting these values in (5) and in (6), we find

$$b = Y_0\frac{1}{\cos\frac{\varphi_{11}^{(1)}}{2}}\left[\frac{\pi}{\lambda_{g0}}l - \left.\frac{V_g}{4\lambda_{g0}}\frac{\partial}{\partial f}\left(\varphi_{11}^{(1)} + \varphi_{11}^{(2)}\right)\right|_{\omega=\omega_0}\right]. \qquad (7)$$

Note that for evanescent mode waveguide filters $\varphi_{11}^{(1)} \approx \varphi_{11}^{(2)} \approx 0$, so for an arbitrary i-th filter resonator we have

$$b \approx Y_0\left[\frac{\pi}{\lambda_{g0}}l - \left.\frac{V_g}{4\lambda_{g0}}\frac{\partial}{\partial f}\left(\varphi_{11}^{(1)} + \varphi_{11}^{(2)}\right)\right|_{\omega=\omega_0}\right]. \qquad (8)$$

The latter relation is working for the calculation of the considered filters, as it allows to find J-parameters and, accordingly, the modules of reflection coefficients of evanescent mode sections from the regular lines on which the resonators are built, and hence their dimensions (lengths). Here are the corresponding phases of the

Tab. 1 The comparison of filters dimensions from work [14] and this work

Filter element	Size from work [14], mm	Size from this work, mm
l_{c1}	4.9702	4.8125
l_{c2}	11.6017	11.2524
l_{c3}	12.9767	12.693
l_{r1}	9.7559	10.7333
l_{r2}	8.7594	9.6407

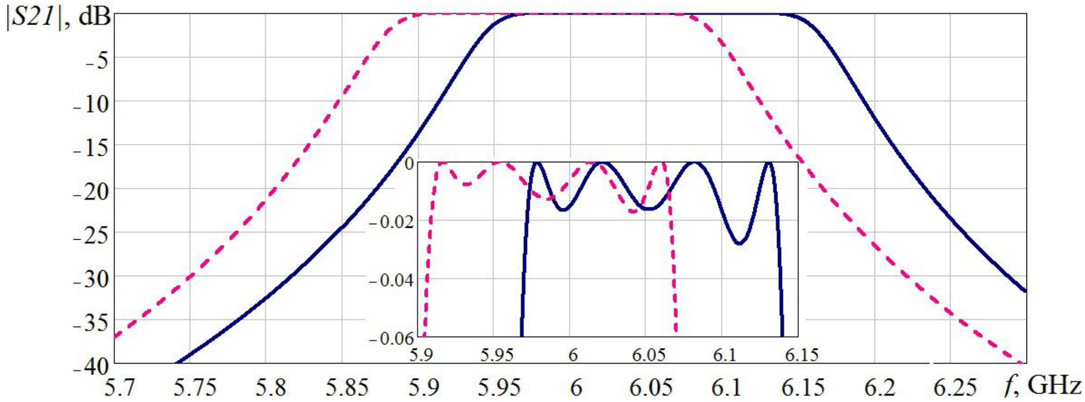


Fig. 3. Simulated characteristics of filters with sizes from [14] (solid curve) and sizes calculated in this work (dashed curve)

reflection coefficients. Once this is done, the resonator lengths can be found from the resonance condition $\Theta = 0$. Note that the initial values of the parameters b_i require some values of length l_i. They are taken equal to half the wavelength in the resonator line. Therefore, as in [14], the completion of the synthesis procedure requires a number of iterations, but the intermediate steps according to (8) are very simple, and the iterative process converges quickly. We also emphasize that, as in [14], it is assumed everywhere that the lengths of the resonators can be calculated by formula (1) (or under the condition $\Theta = 0$), i.e. the synthesis procedure was possible in principle.

For comparison the dimensions of the components of four-resonator evanescent mode waveguide filters with the segments of single ridged waveguide in evanescent mode waveguide, synthesized according to the method [14] and method, proposed in this work, are shown in Tabl. 1. According to the notations accepted in [14], here l_{ci} – lengths of evanescent mode sections, l_{ri} – lengths of resonators. The filters have a center frequency f_0=6 GHz, bandwidth Δf=158 MHz, return losses 25 dB. The filters are made in a chamber with dimensions $a \times b$=12.5×5.625 mm. The width of the ridged waveguide ridge S=6.25 mm, the slot size w=0.86 mm. The simulated frequency characteristics of the filters with the sizes specified in Tabl. 1 are presented in Fig. 3. Here is a detailed comparison of the characteristics in the passband. It is seen that the proposed method of synthesis (dashed line) leads to a slightly better result, although both filters obviously meet the requirements for characteristics with sufficient for practice accuracy.

It should be emphasized, that although the proposed method of synthesis of evanescent mode waveguide filters, as well as the method [14], give results that agree well with the experimental data and look closed, their applicability is limited by condition (1) for determining the resonator lengths. And since the calculation of filters in accordance with these procedures cannot verify the validity of this, strictly speaking, they cannot be considered universal. The relative deviation of resonant frequencies f_0, calculated by formula (1), and f_{0s}, calculated electrodynamically, depending on the relative resonator length $\frac{l}{\lambda_{g0}}$ is shown in the Fig. 4.

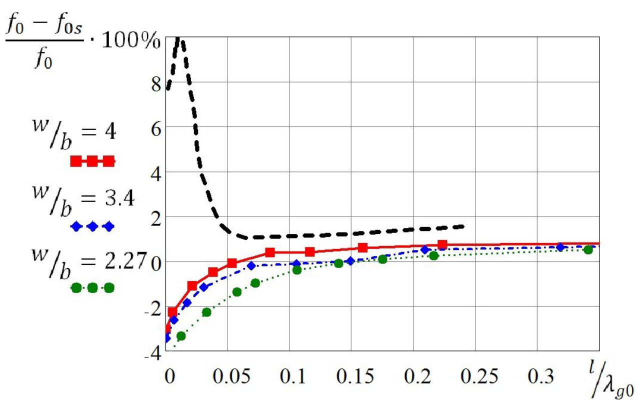
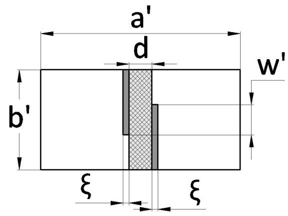


Fig. 4. Deviation of resonant frequencies f_0, calculated from (1) and frequencies f_s, calculated electrodynamically versus normalized resonator lengths

The calculations were performed for three values of the normalized gap width $\frac{w}{b}$ between the ridges of the double ridged waveguide; the resonator was considered weakly coupled. It is clear that even for sufficiently wide slots, the error in the resonant frequency remains small up to the values of $\frac{l}{\lambda_{g0}} \sim 0.05$. This explains the effectiveness of the proposed technique, confirmed by developing of significant number of filters performed by the authors.

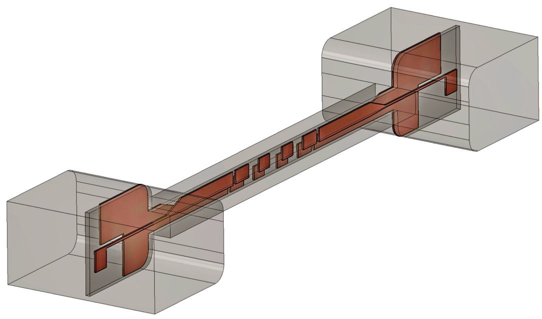
2 Application to the development of E-plane filters with Extremely Wide Stopband

E-plane implementation of evanescent mode waveguide filters was proposed and realized in [7]; electrodynamic analysis of the entire topology of the filter, made on the basis of the finline (FL), was first performed in [8]. Since the attempt to significantly increase the width of the attenuation band leads to reducing the size of the gap between the ridges of the FL, its significant expansion cannot be achieved here, because with small slot the sensitivity to its size is very significant and it is impossible to produce it with the required accuracy. The solution was found in [15] and consisted in using an antipodal finline (AFL) as an electrodynamic basis for resonators, the critical frequency of which can be easily reduced to the required value by overlapping the ridges. Although this feature of AFL in these works was not fully used (BPF was implemented on an empty double ridged waveguide with antipodal placement of bulk ridges), the developed filter had a stopband up to the fifth harmonic of the central frequency of the passband. Moreover, the theoretical part of these works was limited to calculations of the dispersion characteristics of AFL, so the method of filter synthesis used by the authors did not go beyond the calculations, performed in the first works on the development of such filters. The proposed method of synthesis can be successfully applied to the development of E-plane filters on AFL, because due to the small volume, occupied by the field in the space between adjacent ridges, diffraction corrections to size should be negligible. Therefore the length of resonators can be calculated by (1) to their very small size (see the dependence shown in Fig. 4 drawn with dashed line). As an example of application the developed technique was used for calculation of the four-cavity Chebyshev filter with the ripples level 0.2 dB in a passband 19.5–22.5 GHz. The structure uses a dielectric substrate with a thickness of d=127 µm and a dielectric constant ε=2.2 with bilateral metallization. The thickness of the metallization layer ξ=17 µm. The topology of regular AFL is shown in Fig. 5a. Input and output waveguides have standard dimensions $a \times b$=11×5.5 mm; the dimensions of the evanescent waveguide are $a' \times b'$=3×1.5 mm. The ridges of the AFL overlap on a segment whose relative size is $\frac{w'}{b'}$ =0.2. As input and output transformers, specially designed high-quality transitions are used, which consist of the developed longitudinal-probe transitions between a rectangular waveguide and a microstrip line (MSL) and transitions between MSL and AFL (Fig. 5b). Their use avoids the influence of inaccurate setting of the filter structure relative to the boundaries of the evanescent waveguide sections.



The topology of antipodal finline

(a)



The sketch of the designed filter

(b)

Fig. 5. The topology of regular antipodal finline and filter based on it

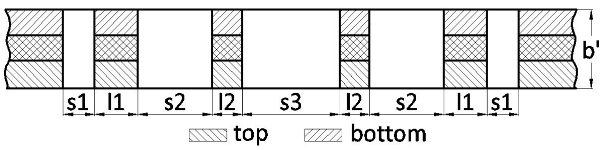


Fig. 6. The planar structure of the designed filter sketch

The topology of filter is depicted in the Fig. 6. Fig. 7 shows the simulated frequency response of synthesized filter (dashed line). It can be seen that the shift of central frequency is 0.37 % and the deviation of passband width is minus 2.6 %. The simulated frequency response of filter with optimized dimensions is shown in this figure by solid line. The comparison of direct synthesis dimensions and optimized dimensions is shown in Table 2. The photograph of the filter is shown in the Fig. 8. To measure the frequency response of the filter in a wide frequency band, the filter structure was connected by transformers designed to operate in the corresponding operating frequency bands of the waveguides of standart cross sections (7.2 × 3.4 mm; 5.2 × 2.6 mm; 3.6 × 1.8 mm). The measurement results are shown in Fig. 10 by points.

Tab. 2 The comparison of direct synthesis dimensions and optimized dimensions

Filter element	S_1	S_2	S_3	l_1	l_2	w'	a'	b'	d
Dimensions by synthesis, mm	0.325	1.23	1.503	1.507	1.184	0.3	3	1.5	0.127
Dimensions by synthesis, mm	0.316	1.211	1.484	1.493	1.167	0.3	3	1.5	0.127

Tab. 3 The relative deviations of the center frequency and passband width from the required values for filters with different relative passband width

$\frac{(\Delta f)_0}{f_0}$	0.05	0.1	0.15	0.2	0.25	0.3	0.35	0.4
$\frac{f_0 - f_{0s}}{f_0} \cdot 100\%$	-2.81	-3.2	-3.08	-3.24	-3.11	-2.37	-2.08	-1.85
$\frac{(\Delta f)_0 - (\Delta f)_s}{(\Delta f)_0} \cdot 100\%$	-3.4	-2.6	-2.2	-0.9	-1.24	-1.73	-2	-2.25

The filter meets the requirements of high attenuation in a wide frequency band (up to the harmonics of the central frequency of the passband).

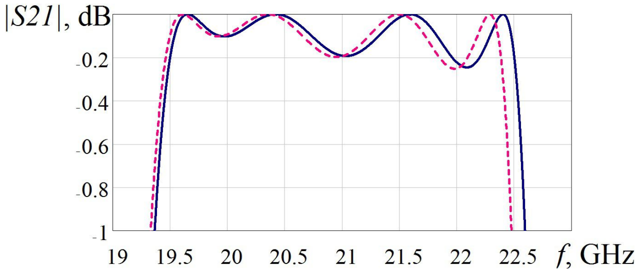


Fig. 7. Simulated characteristics of the designed filter: synthesized filter (dashed curve), filter with optimized dimensions (solid curve)

An important characteristic of the microwave BPF synthesis technique is the restriction on the relative frequency passband width of the synthesized filters. The results of the studies are shown in Table. 3. We have consistently calculated eight 7-cavity filters with relative passband width $\frac{(\Delta f)_0}{f_0}$ =0.05÷0.4 with the same center frequency f_0=10 GHz and frequency response requirements in the passband. The table shows the relative deviations of the center frequency and passband width from the required values. It is seen that the deviations remain acceptable up to the values $\frac{(\Delta f)_0}{f_0}$ =0.4. The simulated frequency response of the filter with the relative passband width $\frac{(\Delta f)_0}{f_0}$ =0.4 (f_{min}=8 GHz, f_{max}=12 GHz) is shown in Fig. 9 by solid curve. Here, the dashed line shows the frequency response of the filter with changed dimensions, which were obtained as a result of second application of the procedure of synthesis of the filter with artificially changed requirements for the characteristic. These requirements were formulated based on the differences between the desired characteristics and the characteristics of the filter with the dimensions obtained after the first application of the synthesis (the so-called false object method).

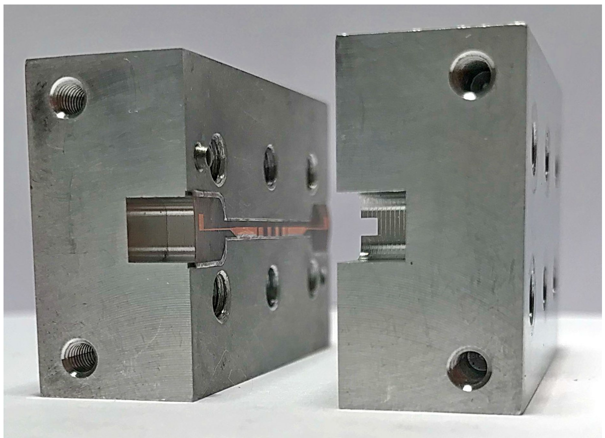


Fig. 8. The photograph of filter

It is seen that the filter calculated in this way fully meets the requirements for it. Thus, it can be concluded that the proposed method of synthesis is an effective for calculating the considered filters with a wide range of passband width requirements. This method avoids time-consuming optimization to find the final dimensions of the elements of the filter topology.

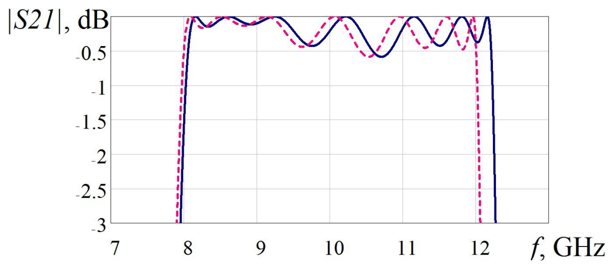


Fig. 9. Simulated frequency responses of 7-cavity filters with 40% passband width. The shown responses are presented for filter with dimensions obtained in proposed synthesis procedure (solid curve) and obtained after repeated synthesis with changed requirements (dashed curve)

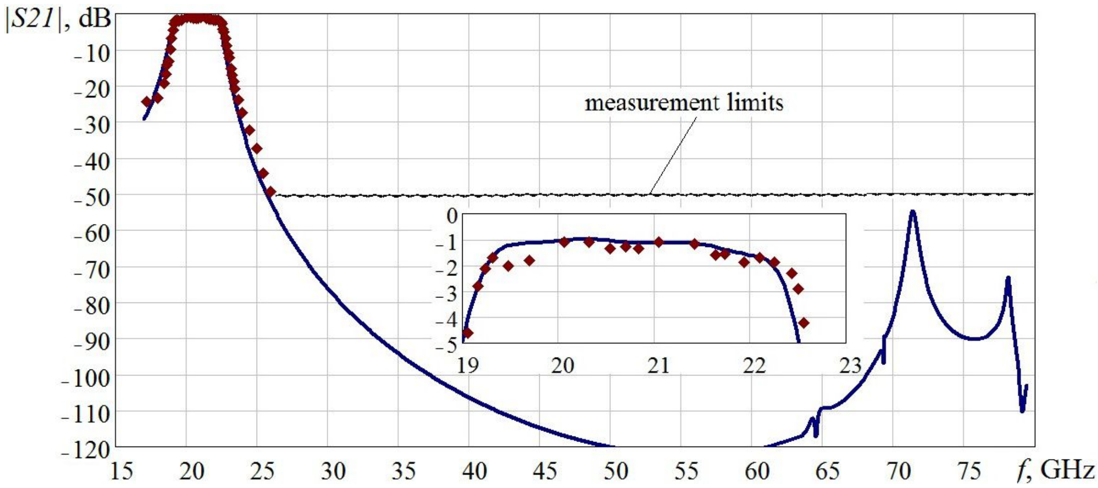


Fig. 10. Measured performance of the designed filter (points). The simulated characteristic of filter (solid curve), shown for comparison

Conclusion

The proposed method of direct synthesis of bandpass filters on the ridge waveguide segments in evanescent mode rectangular waveguide makes it possible to find the sizes of all filter elements with an accuracy that does not require a expensive optimization procedure. The technique can be classified as a modification of the classical Cohn filter synthesis procedure, which consists in calculating such values of conductivity inverters that would satisfy the values of the reactive conductivity slope parameters of resonators with lengths significantly less than half the wavelength. It was shown in this work that the proposed technique provides satisfactory results in the synthesis of filters, the resonator lengths of which are only one fifth of the half wavelength. It has been shown that the proposed technique is effective in the development of bandpass filters of this type with a relative bandwidth of at least 40 %. It is also shown that the proposed synthesis technique is adequate for the synthesis of E-plane filters based on sections of antipodal finline with significant overlapping of the ridges. These high-tech filters are capable for providing stopband widths of up to several octaves in the millimeter wavelength range. This conclusion is confirmed by the results of an experimental study of a calculated and manufactured sample of a filter with a passband in the frequency range of 21 GHz.

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Хвилеводно-планарні смугові фільтри із широкою смугою загородження

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В статті наведені результати розробки процедури прямого синтезу смугопропускаючих фільтрів, побудованих на планарних метало-діелектричних структурах, розташованих в позамежовому прямокутному хвилеводі. Розроблена методика синтезу дозволить отримати розміри всіх елементів топології фільтра, які або не потребують подальшої трудомісткої оптимізації, або ж оптимізація зводиться до повторного застосування запропонованої методики із штучно зміненими вимогами до характеристики, які базуються на даних про відхилення характеристики синтезованого фільтра від вимог до неї. Розроблена методика виявилася адекватною при розробці запропонованих хвилеводно-планарних фільтрів, побудованих на відрізках антиподальної хвилеводно-щілинної лінії у позамежовому прямокутному хвилеводі за умови значного перекриття її гребенів, при якому скорочення резонаторів відносно половини довжини хвилі досягає 80 %. Зазначена хвилеводно-планарна реалізація фільтра на позамежовому прямокутному хвилеводі дозволяє значно розширити смугу частот непропускання, збільшити внесене в ній загасання і одночасно забезпечити повторюваність характеристик фільтрів без будь-яких дій по їх настроюванню. Ефективність запропонованого підходу до реалізації фільтра і його розрахунку продемонстрована на розробці і експериментальному дослідженні фільтра діапазону 21 ГГц, який за сукупністю параметрів (втрати порядку 1 дБ, ширина смуги загородження до четвертої гармоніки центральної частоти смуги пропускання) відповідає високим вимогам щодо електричних характеристик і технологічності. Показано, що розроблена методика синтезу подібних фільтрів залишається актуальною при синтезі фільтрів із відносною шириною смуги частот пропускання до 40 %.

Ключові слова: смугові фільтри; міліметровий діапазон; гібридно-інтегральні схеми НВЧ; смуга загородження; позамежовий хвилевід

Синтез волноводно-планарных фильтров со сверхширокой полосой заграждения

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В статье приведены результаты разработки процедуры прямого синтеза полоснопропускающих фильтров, построенных на планарных металло-диэлектрических структурах, расположенных в запредельном прямоугольном волноводе. Разработанная методика синтеза позволяет получить размеры всех элементов топологии фильтра, которые или не нуждаются в дальнейшей трудоемкой оптимизации, или оптимизация сводится к повторному применению предложенной методики с искусственно измененными требованиями к характеристике, основанными на данных об отклонении характеристики синтезированного фильтра от требуемой. Разработанная методика оказалась адекватной при разработке предложенных волноводно-планарных фильтров, построенных на отрезках антиподальной волноводно-щелевой линии в запредельном прямоугольном волноводе. При значительном перекрытии ее гребней сокращение резонаторов относительно половины длины волны достигает 80 %. Указанная волноводно-планарная реализация фильтра на запредельном прямоугольном волноводе позволяет значительно расширить полосу частот заграждения, увеличить вносимое в ней затухания и одновременно обеспечить повторяемость характеристик фильтров без каких-либо действий по их настройке. Эффективность предложенного подхода к реализации фильтра и его расчету продемонстрирована на примере разработки фильтра диапазона 21 ГГц, который по совокупности параметров (потери порядка 1 дБ, ширина полосы заграждения до четвертой гармоники центральной частоты полосы пропускания) соответствует высоким требованиям к электрическим характеристикам и технологичности. Показано, что разработанная методика синтеза подобных фильтров остается актуальной при синтезе фильтров с относительной шириной полосы частот пропускания до 40 %.

Ключевые слова: полосовые фильтры; миллиметровый диапазон; гибридно-интегральные схемы СВЧ; полоса заграждения; запредельный волновод