ЕЛЕКТРОНІКА, РАДІОТЕХНІКА ТА ТЕЛЕКОМУНІКАЦІЇ

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MICROWAVE FILTERS BASED ON THE STRUCTURES WITH RESONATORS IN PARALLEL CHANNELS AS METAMATERIAL CELLS

Background. Many of the properties of metamaterials are similar to those found in filters with mutually detuned by frequency unrelated resonators. The bridge filters are used as a low-frequency prototypes of such microwave filters. For further development and design of new types of metamaterials it is necessary to establish an analogy between metamaterials and filters with mutually detuned by frequency unrelated resonators.
Objective. Creating a model of metamaterials based on the bandstop microwave filters with mutually detuned resonators and on the low-frequency prototypes.
Methods. Checking the equivalence of metamaterials characteristics and microwave filters with mutually detuned resonators, identifying their inherent laws (oscillation types in parallel channels, location of the attenuation poles above or below the bandwidth), which appear regardless of the types of resonators, studying the possibility of using prototype bridge filters for modeling of metamaterials.
Results. The basic model of an 8-pole network with resonators in parallel channels has been studied in detail. Analytical expressions are obtained for the transmission and reflection coefficients for all inputs of an 8-pole network. The 4-pole networks implemented on the basis of the aforementioned basic model are investigated. Experimental studies are performed that confirm the adequacy of analytical models. An analogy between metamaterials and microwave filters with mutually detuned resonators is established, the possibility of the use as a low-frequency prototype bridge bandpass filters is shown.
Conclusions. Microwave filters with mutually detuned resonators can be used for modeling of metamaterials, and bridge bandpass filters – as low-frequency prototypes, which design techniques are well developed.
Keywords: metamaterials; dielectric resonators; stripline resonators; bandstop microwave filters.

Introduction

More than 25 years ago, the authors theoretically obtained and experimentally confirmed the analytical and circuit models of these devices – resonant directional couplers, bandpass and rejection filters. After a long time, in 2016, an analogy was found between dual-channel notch filters patented by authors and metamaterial cells known as Split Ring Resonators (SRRs).

It turned out that the models proposed and investigated earlier by the authors not only describe the known characteristics of SRR but also allow a number of their properties to be explained, which were considered as “anomalous”. Therefore, the main task is to demonstrate the effectiveness of our approaches in modeling microwave devices based on metamaterial cells.

This paper also considers the possibilities of the models proposed by the authors in one of the most promising areas of research in the microwave range – creation of frequency-selective surfaces with controlled characteristics. A bibliography of authors’ proceedings concerning this subject is given.

Objective

For many years the authors of the submitted work are successfully engaged in the study of microwave devices based on the resonators of different types, which are included in parallel channels and are not connected with each other. The analytical model created for the analysis of such structures proved to be also convenient for analysing the metamaterial cells properties. The purpose of this work is to demonstrate the capabilities of the mentioned analytical model for designing metamaterial cells with improved characteristics.

Split Ring Resonator as the basic metamaterial cell in modern microwave devices

The modern history of metamaterials as artificially created structures with specific properties that have no analogues in nature and consist of identical “cells” began in the late 20^{th} – early 21^{st} centuries [1–10]. Alongside the concept of “metamaterial”, in the scientific and technical literature such terms as “negative dielectric and magnetic permeability”, “superquality” or “extremely high”

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quality factor of SRR (Split Ring Resonator), as well as their polarisability appeared. These terms are present in one form or another in proceedings devoted to various phenomena in metamaterials or devices based on metamaterial cells [11–44]. Without disputing the legality of using new terms and parameters in the aforementioned proceedings, the authors would like to demonstrate the possibility of describing structures similar to SRR based on the so-called bridge prototype filters [45–51]. Below is an overview of our work on this topic over the past 30 years.

The base model of the 8-pole (4-port) network with resonators in parallel channels

The research of bandpass and bandstop filters for different types of resonators was conducted on the basis of generalised 8-pole network. The paper [52] considers the 8-pole network (Fig. 1) formed by two coupled transmission lines (e.g., waveguides) in which an elliptically polarised wave can propagate.

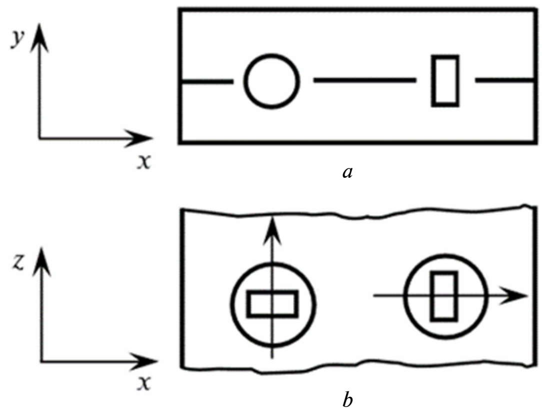


Fig. 1. The 8-pole network formed by two rectangular waveguides coupled by common wide wall. There are dielectric resonators in the holes made in the wall between the waveguides: *a* – end view; *b* – top view

Waveguides are coupled by the holes in common wide wall. In the plane of cross section, that is in the plane perpendicular to the propagation direction of electromagnetic wave, dielectric resonators (DR) are placed, one of which is excited by the transverse magnetic field component h_x and the other – by the longitudinal component h_z. Complex field amplitudes that are reradiated in different directions by resonators in the two waveguides are calculated based on the ratio

$$C_i^{(+/-)} = J\omega m h_i^{(+/-)}/2 \qquad (1)$$

where m is the magnetic moment due to the excitation of the corresponding DR, ω is the circular frequency of the microwave field. In [52] it is shown that

$$C_1^+ = \frac{K_{x1}}{(1+K_{x1}+K_{x2})} + \frac{K_{z1}}{(1+K_{z1}-K_{z2})}; \qquad (2)$$

$$C_1^- = \frac{K_{z1}}{(1+K_{z1}+K_{z2})} - \frac{K_{x1}}{(1+K_{x1}+K_{x2})}; \qquad (3)$$

$$C_2^+ = \frac{\sqrt{[K_{x1}K_{x2}]}}{(1+K_{x1}+K_{x2})} + \frac{\sqrt{[K_{z1}K_{z1}]}}{(1+K_{z1}+K_{z2})}; \qquad (4)$$

$$C_2^- = \frac{\sqrt{[K_{z1}K_{z2}]}}{(1+K_{z1}+K_{z2})} - \frac{\sqrt{[K_{x1}K_{x2}]}}{(1+K_{x1}+K_{x2})}, \qquad (5)$$

where $K_{qi} = \dfrac{j\omega M_q h_{qi}{}^{(+/-)} h_{qi} i^{(-/+)}}{2}$, M_q – parameter that is defined by type of DR oscillation. Knowing the complex amplitudes of resonator radiation fields (2)–(4), we determine the coefficients of transmission and reflection as:

$$T_{12} = 1 + C_1^+;\ T_{13} = C_2^+;\ T_{14} = C_2^-;\ \Gamma_{11} = C_1^-. \qquad (6)$$

Considered 8-pole network is a reciprocal one as the transmission and reflection coefficients are defined by formulae similar to (6) when the generator is connected to the other arms.

Note that 8-pole network design mentioned above is patented by the authors as a directional filter [53].

Bandpass filters. In [52] it is shown that the transmission **T** and reflection **R** coefficients of the waveguide bandpass filter based on the dielectric resonators with different types of oscillations can be easily obtained from the formulae (1)–(6):

$$\mathbf{T} = C_1^+; \qquad (7)$$

$$\mathbf{R} = 1 + C_1^-; \qquad (8)$$

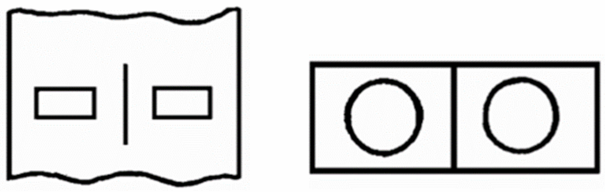
$$\mathbf{T} = \frac{K_2}{(1+K_2)} - \frac{K_1}{(1+K_1)}; \qquad (9)$$

$$\mathbf{R} = 1 - \frac{K_1}{(1+K_1)} - \frac{K_2}{(1+K_2)}, \qquad (10)$$

where K_1, K_2 – coupling coefficients of “magnetic” and “electric” type of oscillations of the transmission lines (in this case – regular waveguide).

Microwave double-resonance filters based on resonators with different types of oscillations have been studied experimentally and theoretically on the basis of formulae (9) and (10) in proceedings [54–58]. The authors have also received copyright certificates for a number of designs of double-resonance bandpass filters based on dielectric and microstrip resonators with different types of oscillations [58–61].

Another way of determining the parameters of double-resonance bandpass filters on mutually detuned resonators (Fig. 2) considered in [62].



a *b*

Fig. 2. Waveguide bandpass filter with dielectric resonators excited by different types of oscillations; *a* – sectional plan view from above; *b* – end view

They are determined on the basis of the scattering matrix parameters:

$$\hat{S} = \begin{bmatrix} 1-\dfrac{\mathrm{K}_m}{1+\mathrm{K}_m}-\dfrac{\mathrm{K}_e}{1+\mathrm{K}_e} & \dfrac{\mathrm{K}_e}{1+\mathrm{K}_e}-\dfrac{\mathrm{K}_m}{1+\mathrm{K}_m} \\ \dfrac{\mathrm{K}_e}{1+\mathrm{K}_e}-\dfrac{\mathrm{K}_m}{1+\mathrm{K}_m} & 1-\dfrac{\mathrm{K}_m}{1+\mathrm{K}_m}-\dfrac{\mathrm{K}_e}{1+\mathrm{K}_e} \end{bmatrix}. \qquad (11)$$

Coupling coefficients from formulae (9), (10) correspond with coupling coefficients of electric K_e and magnetic K_m types of oscillations.

In practice bandpass filters with symmetric centre frequency response are commonly used $(K_1 = = K_2 = \mathrm{K})$. If the resonant frequencies of oscillations do not coincide, the dependence of transmission and reflection coefficients from the frequency can be written as

$$\mathbf{T} = \frac{j2\mathrm{K}\cdot a}{[(1+\mathrm{K}+jx)^2+a^2)]} \qquad (12)$$

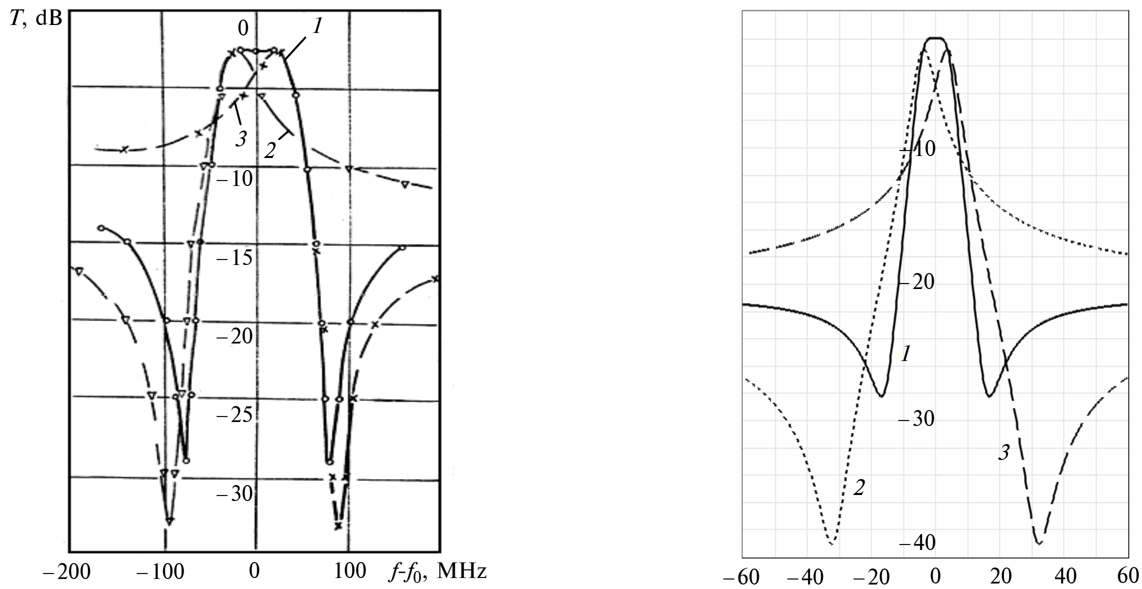
$$\mathbf{R} = \frac{[(1+j\xi)^2-\mathrm{K}^2+a^2)]}{[(1+\mathrm{K}+j\xi)^2+a^2)]} \qquad (13)$$

where ξ – generalised detuning of frequency with respect to f_0 – central frequency of the filter, and a – generalised frequency detuning of "magnetic" f_m and "electric" f_e oscillations with respect to f_0

$$f_0 = \frac{(f_m+f_e)}{2}, \qquad (14)$$

$$a = \frac{2(f_e-f_m)\times Q_0}{(f_e+f_m)} \qquad (15)$$

where Q_0 – own quality factor of resonators (suppose that both resonators have the same quality factor on different types of oscillations).



a *b*

Fig. 3. Experimental (*a*) and calculated (*b*) characteristics of double-resonance bandpass filters with mutually detuned oscillations in the parallel channels: *1* – combined characteristics of the filter with two resonators, *2* – "counter-phase" or "magnetic" oscillations, *3* – "in-phase" or "electric" oscillations

In [54] the influence of out-of-band decoupling between inputs (which corresponds to the transmission coefficient of the filter without resonators) and its influence on the formation of the poles of attenuation are examined in detail. Fig. 3 shows experimental [54] (experiment was conducted 30 years ago) and calculated characteristics of double-resonance bandpass filters with mutually detuned oscillations in the parallel channels. A good coincidence of form of experimental and calculated results show the adequacy of the chosen analytical model of bandpass filter.

It should be noted that in terms of coefficients obtained due to generalised detuning, formulae (12) and (13) are valid not only for dielectric resonators (and various types of oscillations excited in them). They are also suitable for the analysis of various types of resonators (and their oscillations) – microstrip, waveguide-slot, etc.

Analysis of the formulae for coefficients of transmission and reflection of the double-resonance filter resonators with the mutually detuned resonators (oscillations) showed that they are completely similar to formulae for traditional double-resonance filters with cascade-connected coupled resonators. The role of the generalised coupling coefficient of resonators (used when analysing the cascade connection of the coupled resonators) is played by a generalised detuning "*a*" (formula (15) between their resonant frequencies. Hence one of the most important differences and advantages of the filters with mutually detuned resonators (oscillations) is the ability to manage the bandwidth of a filter, changing not the connection between the resonators, but their resonant frequencies. Fig. 4 represents microstrip filter [59] in which the resonant frequency of one of the resonators changes with the change of the control voltage applied to the varicaps connected to it.

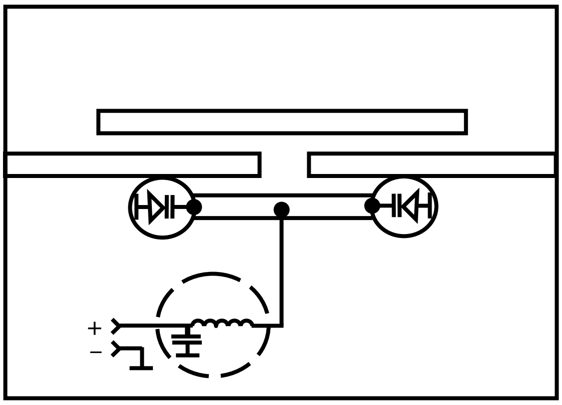
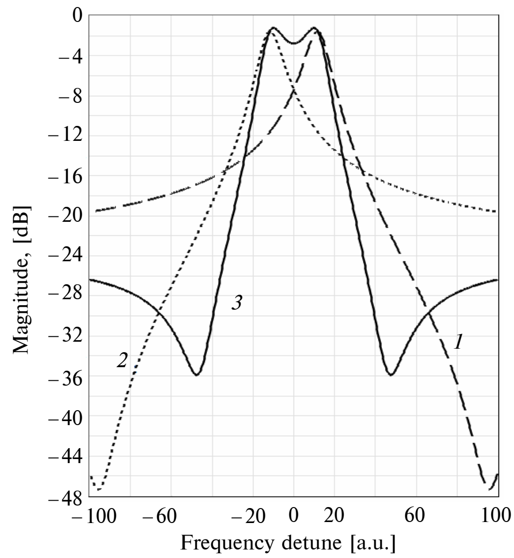
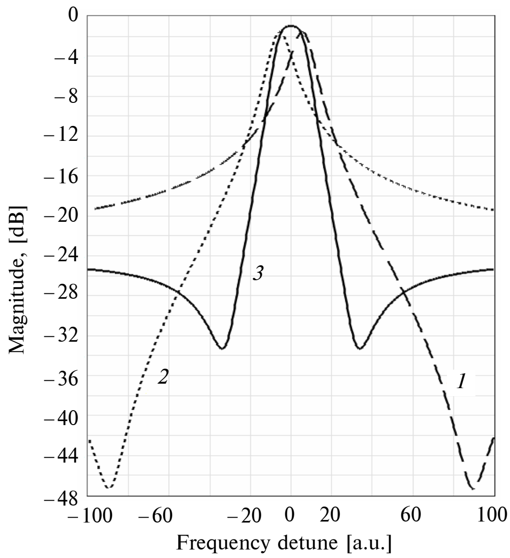


Fig. 4. Double-resonance microstrip bandpass filter with controllable bandwidth

Fig. 5 shows characteristics of double-resonance bandpass filters with different types of oscillations in the parallel channels for the same coupling coefficients of resonators K with the input-output filter (K = 5) and different mutual detuning "*a*" between their resonant frequencies. When "*a*" changes from 12 (Fig. 5, *a*)) to 6 (Fig. 5, *a*)), the bandwidth of the filter at the –3 dB level decreases almost twice. In both cases the level of out-of-band decoupling is –25 dB.



a



b

Fig. 5. Transmission coefficients of double-resonance bandpass filters with different types of oscillations in parallel channels: *1* – characteristics of filter transmission coefficient only with excitation of "electric" type oscillations, *2* – characteristics of filter transmission coefficient only with excitation of "magnetic" type oscillations, *3* – characteristics of filter transmission coefficient with excitation of both types of oscillations simultaneously

The coordinates of the poles of attenuation ζ_p – Fig. 5, *a* and *b* – correspond to calculated ones using the formula (16) from [54] (taking into account the correspondance of detuning "*a*" to the coupling coefficient K_c):

$$\zeta_p = \sqrt{\frac{2K_c \times a}{D}} \qquad (16)$$

where D – level of out-of-band decoupling. For values K, *a* and *D* used in the calculation of characteristics shown in Fig. 5, *a* and *b*, ζ_p is ~ 44 and 32 respectively.

Note another interesting feature of bandpass filters with resonators of different types in parallel channels (Fig. 6). Formula (13) shows that bandpass filter can be perfectly matched (**R** = 0) at the central frequency performing the following ratio:

$$a = \pm\sqrt{(K^2 - 1)}\,. \qquad (17)$$

An important parameter of double-resonance filters with coupled resonators is so-called "critical" value of coupling coefficient in which the characteristics of the transmission coefficient is becoming maximally flat (when connection is bigger than critical in bandpass of the filter, failure appears at the central frequency). From (12) it follows that for double-resonance filters with mutually detuned resonators "critical" value of detuning "a" corresponds to similar coupling coefficient and equals

$$a_{cr} = K + 1\,. \qquad (18)$$

As mentioned detunings "*a*" determined in accordance with formulas (17) and (18) do not match, double-resonance filter with maximally flat characteristics (butterworth type) is not perfectly coincide in central frequency, however with K values, much bigger than 1, a good alignment in the bandpass is achieved as shown in Fig. 7.

In [60] the bandpass filter with waveguide-slotted resonators, the design of filters is shown on Fig. 8.

The filter contains a rectangular waveguide segment along the longitudinal axis of which the metal plate is placed in the E-plane. The plate has resonant slits of linear (longitudinal) and U-shaped form. U-shaped slit is oriented with its ends toward the longitudinal resonant slit. Due to the placement, choice of length and configuration of U-shaped slit, a decrease in the longitudinal length of the filter is provided. The presence of spurious bandpass, due to the influence of lower type oscillations of U-shaped slit, can be compensated by selecting cutoff frequency of rectangular waveguide which is higher than the resonant frequency mentioned type of unwanted oscillations.

Bandstop filters. Proceedings [52–57] considered bandpass filters and focused mainly on their

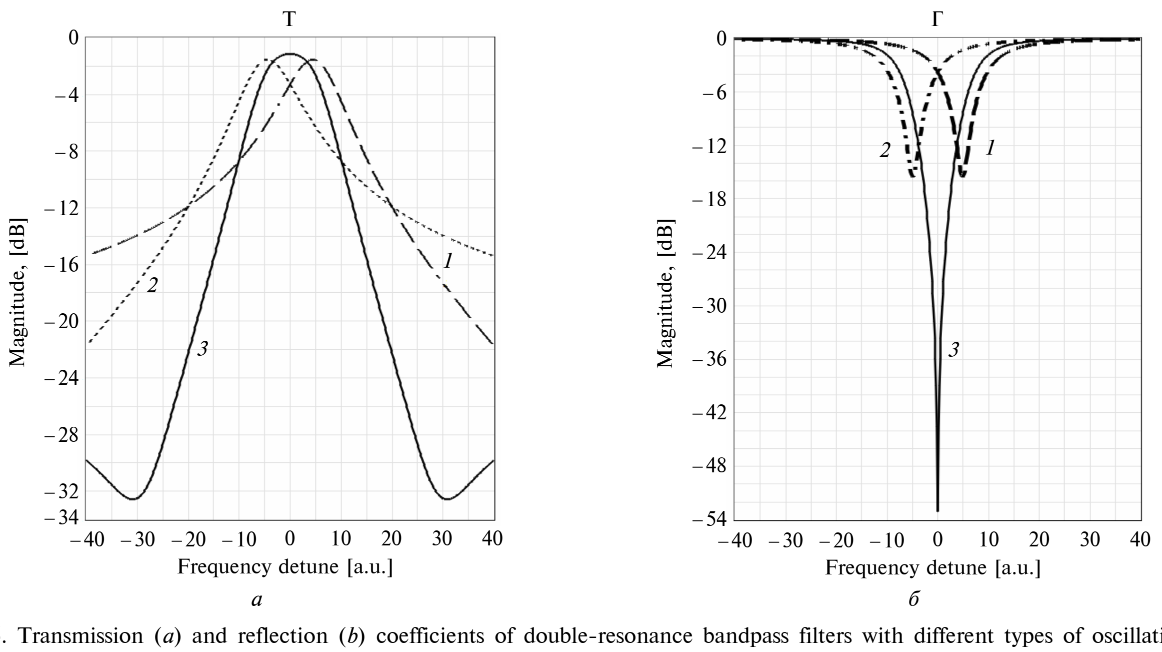
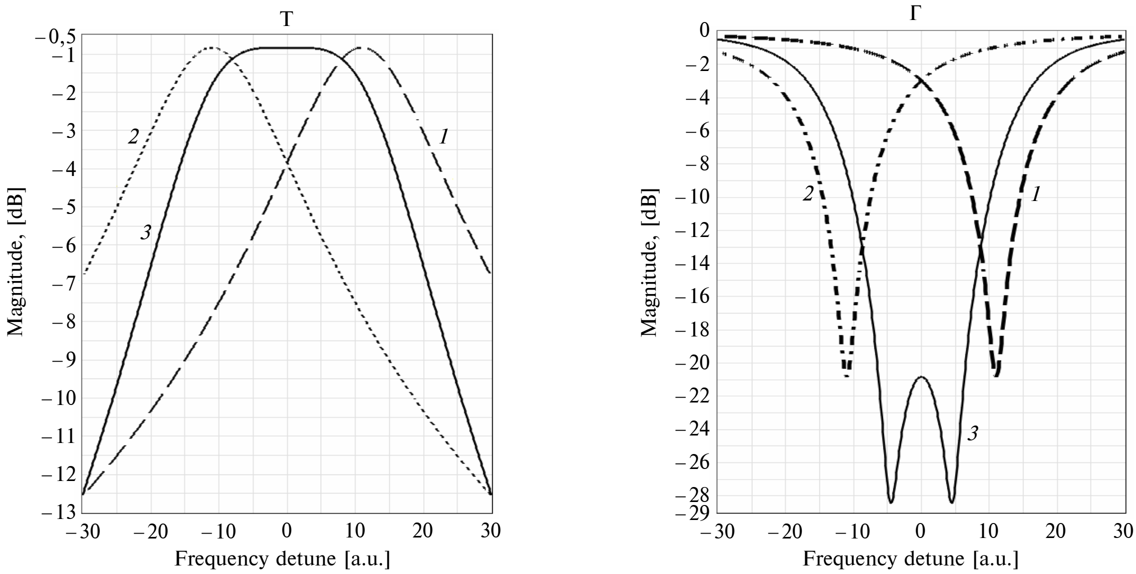
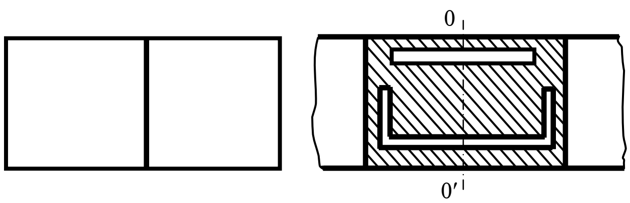


Fig. 6. Transmission (*a*) and reflection (*b*) coefficients of double-resonance bandpass filters with different types of oscillations in parallel channels: *1* – characteristics of filter transmission coefficient only with excitation of "electric" type oscillations, *2* – characteristics of filter transmission coefficient only with excitation of "magnetic" type oscillations, *3* – characteristics of filter transmission coefficient with excitation of both types of oscillations simultaneously; K = 5, *a* = 4.9



a *b*

Fig. 7. Transmission (a) and reflection (b) coefficients of double-resonance bandpass filters with different types of oscillations in parallel channels: *1* – characteristics of filter transmission coefficient only with excitation of "electric" type oscillations, *2* – characteristics of filter transmission coefficient only with excitation of "magnetic" type oscillations, *3* – characteristics of filter transmission coefficient with excitation of both types of oscillations simultaneously; K = 10, *a* = 11 (critical connection). Frequency on Fig. *a* and *b* – in relative units



a *b*

Fig. 8. Waveguide-slotted filter with parallel connected resonators: *a* – end view; *b* – side view

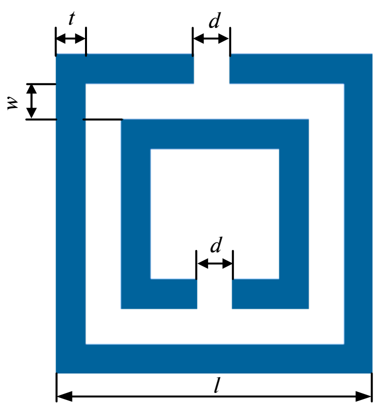


Fig. 9. l = 3 mm, $d = t = w$ = 0.3 mm

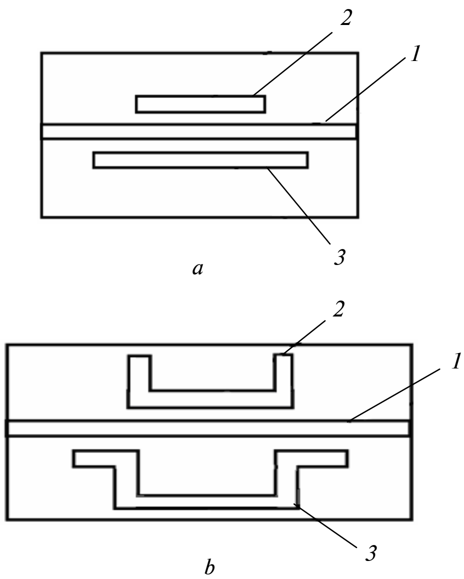


Fig. 10. Microwave bandstop filters *a* – [64], *b* – [65]; *1* – microstrip transmission line, *2* – "half-wave" resonator, *3* – "wave" resonator

filtering properties. Almost after 30 years in [63] it was shown that the bandstop filters with mutually detuned resonators in parallel channels are nothing but metamaterial cells. Therefore, the increased interest in this kind of bandstop filters is justified.

The structure of an elementary metamaterial cell with the dimensions of the external and internal resonators and the gap between them is presented in [64] (Fig. 9).

The filter on Fig. 10, *a* is formed by resonators, length of which multiple $n\lambda/2$ and $(n+1)\lambda/2$ (λ – the wavelength at the resonant frequency of the filter), with n = 1 – "half-wave" and "wave" resonators correspondingly, and resonators of the filter on Fig. 10, *b* – "half-wave" and "wave" (it is clear that they can also have multiplicity n and $(n+1)$).

In [52] it is shown that the resonant frequencies of these resonators may be different and close to the center frequency of the filter (depending on the resonator coupling coefficients of the transmission line). If we compare the length of resonators Fig. 8 (for accuracy – on the lines passing in the middle of the resonators, we will get the following result: the length of smaller ("internal") resonator is 5.7 mm, bigger one ("external") is 11.1 mm. It turns out that the two resonators (including all mentioned in [52] and the error introduced by the capacitive coupling between them) can also be considered as "half-wave" and "wave", and so they can use the same models as filters [64, 14].

It can be shown (see [52], formula (6), section 16.4) that the formulae for the transmission and reflection coefficients of the bandpass and bandstop filters with mutually detuned resonators are "dual", that is, having the formulas (7) and (8) for bandstop filter, we can obtain the following:

$$\mathbf{R} = \frac{K_2}{(1+K_2)} - \frac{K_1}{(1+K_1)}; \tag{19}$$

$$\mathbf{T} = 1 - \frac{K_1}{(1+K_1)} - \frac{K_2}{(1+K_2)}, \tag{20}$$

Introducing K_1 and K_2 as

$$K_1 = \frac{\mathrm{K}_1}{[1+\mathrm{K}_1+j(\xi+\mathbf{a})]} \tag{21}$$

$$K_2 = \frac{\mathrm{K}_2}{[1+\mathrm{K}_2+j(\xi-\mathbf{a})]} \tag{22}$$

it can be shown that in the case of mutual detuning of "**a**" between the resonant frequencies of the used oscillations of resonators, which is determined by the formula

$$\mathbf{a} = \pm\sqrt{(\mathrm{K}_1 \times \mathrm{K}_2 - 1)} \tag{23}$$

transmission coefficient of bandstop filter **T** (formula (20) equals zero, that means a complete, 100-percent rejection of signal is achieved. Note that such effect is not possible to achieve in filters based on the cascade-connected or interconnected resonators.

For equal coupling coefficients $\mathrm{K}_1 = \mathrm{K}_2$ formula (23) was obtained earlier in [52]. Fig. 11, *a* shows calculated (solid curves) and experimental (keys on the curves) characteristics of the waveguide bandstop filter with dielectric resonators whose design is presented in [66] (Fig. 12). Curves *1* and *2* – characteristics of individual resonators, curve *3* – combined characteristic of double-resonance bandstop filter. Fig. 11, *b* shows calculated on the basis of formula (20) transfer rates similar filter coupling coefficients K_1 and K_2 and mutual detuning "**a**" chosen specially to have similar characteristics to those shown in Fig. 9, *a*. Apparently, the good agreement is obtained with $\mathrm{K}_1 = \mathrm{K}_2 = 2.1$, $\mathbf{a} = 1.81$.

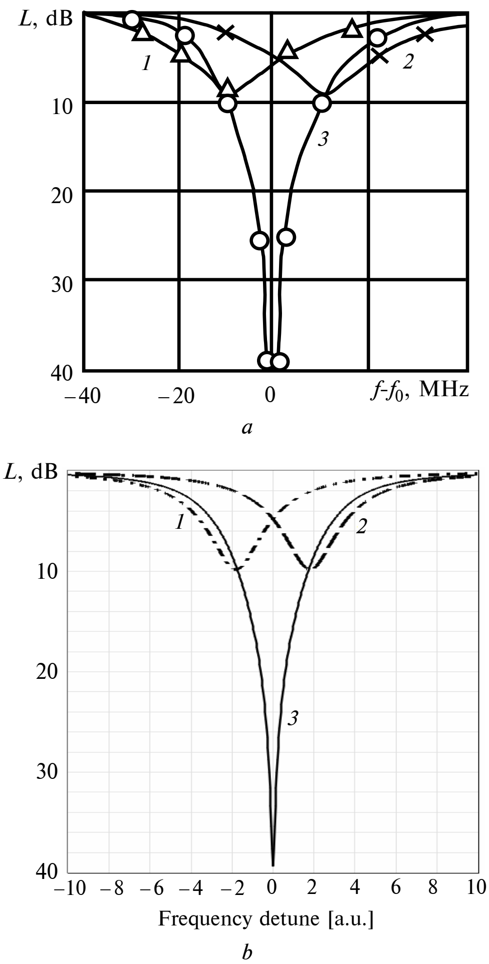


Fig. 11. Experimental (*a*) and calculated (*b*) characteristics of waveguide bandstop filter with dielectric resonators: *1* – characteristics of filter transmission coefficient only with excitation of "electric" type oscillations, *2* – characteristics of filter transmission coefficient only with excitation of "magnetic" type oscillations, *3* – characteristics of filter transmission coefficient with excitation of both types of oscillations simultaneously. Frequency in Fig. *b* – in relative units

Filter resonators in Fig. 10 are excited by different types of oscillations – "electric" and "magnetic”, their resonant frequencies are closely spaced, but not identical, coupling coefficients of resonators with waveguide can be adjusted by changing the depth of their “immersion” into a regular waveguide from the segments of below-cutoff waveguides where they are placed.

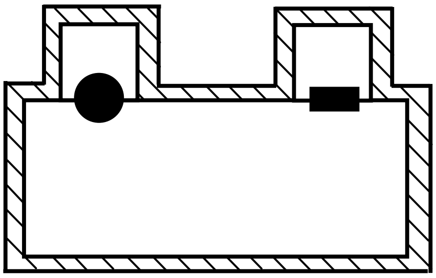


Fig. 12. Waveguide bandstop filter with dielectric resonators

Fig. 13, *a* shows experimental characteristics of waveguide bandstop filter with dielectric resonators with different types of oscillations, resonant frequencies of which are almost the same (filter design is similar to that shown in Fig. 12). Curves *1* and *2* – frequency response of each resonator separately, curve *3* – FR of transmission coefficient of the filter with two resonators. Fig. 13, *b* shows transmission coefficients of the similar filter calculated on the basis of formula (20), which coupling coefficients K_1 i K_2 were selected specifically to have similar characteristics to those illustrated in Fig. 13, *a*. Apparently, a good match is

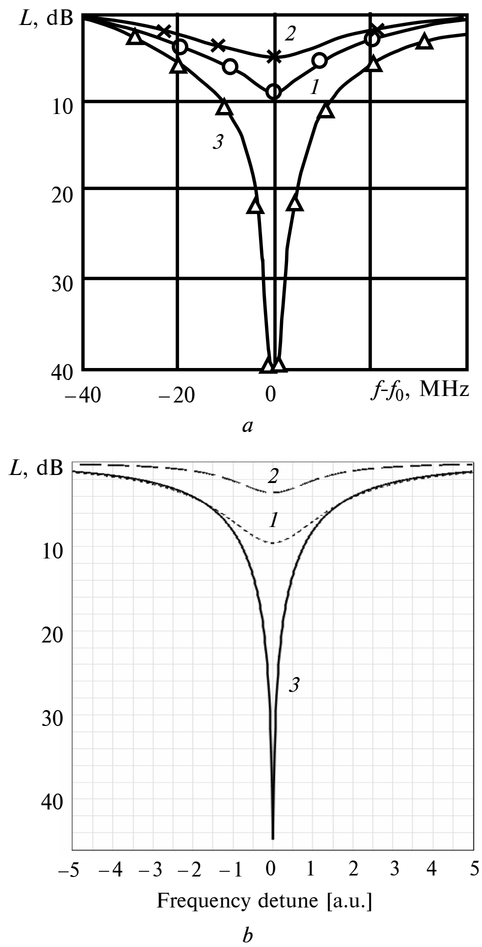


Fig. 13. Experimental (*a*) and calculated (*b*) characteristics of waveguide bandstop filter with dielectric resonators. Frequency in Fig. *b* – in relative units

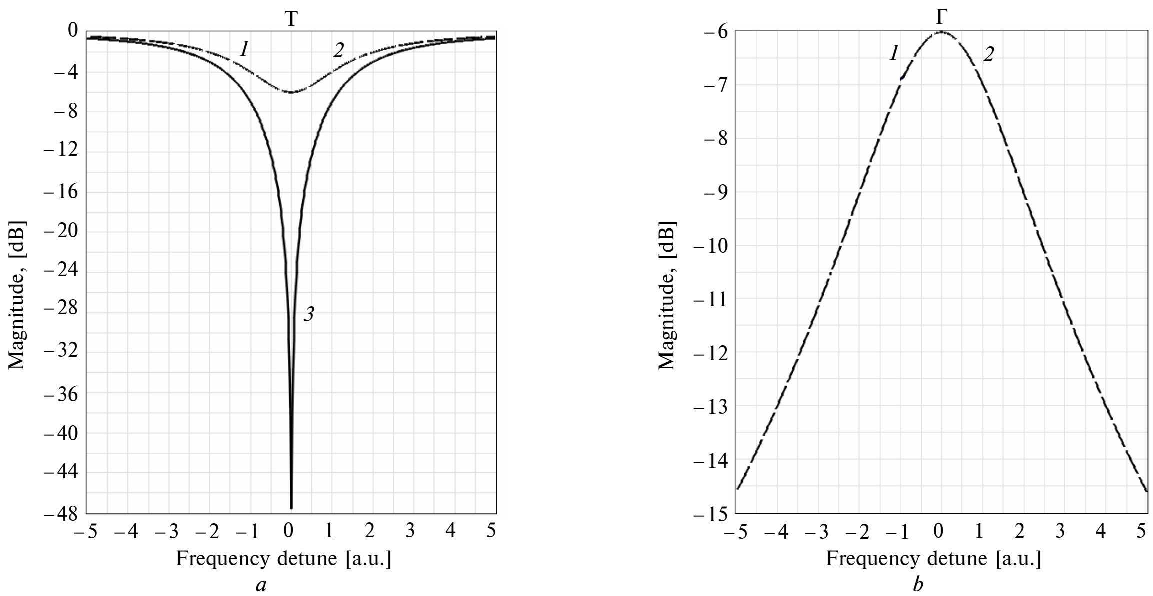


Fig. 14. Transmission (a) and reflection (b) coefficients of band-stop filter with resonators of different types: *1* and *2* – characteristics of individual resonators (match), *3* in the Figure *a* – characteristics of double-resonance filter. Frequencies in Fig. *a* and *b* – in relative units

obtained with $K_1 = 2$, $K_2 = 0.51$, $\mathbf{a} = 0$ (resonant frequencies of resonators, as mentioned above, coincide). Note that with the selected coupling coefficients the ratio (23) is executed as expected.

In [52] it is noted that the narrowest frequency response of transmission coefficient of the bandstop filter will be achieved with the coupling coefficients $K_1 = K_2 = 1$ ($\mathbf{a} = 0$, resonators are excited at the same frequency). At the same time, the loaded quality factor for resonators with each type of oscillations is equal to twice the inherent quality factor. Rejection band at the –3 dB level is just twice wider than the bandwidth of unload dielectric resonator. In addition, this filter is perfectly matched ($\mathbf{T} = 0$), that is, the rejection of the signal is carried out by full absorption of its strength at the resonant frequency. Considered filter can be made based on a construction with spherical dielectric resonator placed at the intersection of regular and below-cutoff waveguide in the area of circular polarization of electromagnetic field. Fig. 14 presents characteristics of the "ideal" bandstop filter ($K_1 = K_2 = 1$, $\mathbf{a} = 0$) calculated in accordance with formulas (19), (20). The absence of the curve *3* in Fig. 14, *b*, corresponding to the reflection coefficient of a filter, is justified by $\mathbf{R} = 0$.

Superlens. In numerous papers on metamaterials, dielectric and magnetic permeability of the medium are regarded as some integral characteristics of dielectric and magnetic permeability ε and μ. At the same time, in [68] appear parameters K_{qi}, which are interpreted as coupling coefficients of a certain type of oscillations (electric or magnetic) with the lines. We can write them down as coupling coefficients of electric and magnetic types of oscillations as K_e and K_m correspondingly and compare with integral characteristics of metamaterials ε and μ. Then in [68] the infinite attenuation in the structure will be achieved with the mutual detuning of frequencies of the orthogonal oscillations "**a**"

$$\mathbf{a} = \pm\sqrt{(K_e \times K_m - 1)}\,, \qquad (24)$$

which fully meets (23), considering that coupling coefficients K_1 and K_2 from formulae (9), (10) correspond coupling coefficient K_e types of electric and K_m magnetic oscillations.

It is possible to assume with a high degree of probability that the ratio $K_e = K_m = 1$, which is inherent to bandstop filter with utmost narrow rejection band (see Fig. 12), corresponds to the condition of "superlens properties" of metamaterials:

$$\varepsilon = \mu = -1 \qquad (25)$$

which in its turn leads to the assumption that the metamaterial unit cell for "superlens" is a "perfect" double-resonance bandstop filter with different types of oscillations, which coincide in frequency.

Unfortunately, this hypothesis has not yet been confirmed experimentally by authors.

Circuit models of filters – lattice (bridge) equivalent circuits

In [63] it is shown for the first time that the metamaterial unit cell is double-resonance system, while the oscillations of individual resonators are not connected. Degeneration of oscillations in the metamaterial cell, properties of individual oscillations and related all sorts of effects are investigated by authors in detail in [68–71]. It is also shown that the widespread in literature opinion of "superquality" (Q – quality) of the metamaterial cells, including so-called Split Ring Resonators (SRR), is not quite correct, since Q is a parameter that characterises the individual oscillations instead of 4-pole network resonant characteristics. It is shown that extremely high Q of metamaterial cells is evident in the area of two different mutually detuned oscillations calculated in accordance with (23), and metamaterial properties – only in the area in which the frequency of cophased oscillations is higher than the frequency of antiphase ones [68]. The figures in [69–71] are experimental results demonstrating the presence of the degenerated oscillations, the possibility of removing the degeneration during oscillation disorder and achieving the extreme values of rejection at two different frequencies.

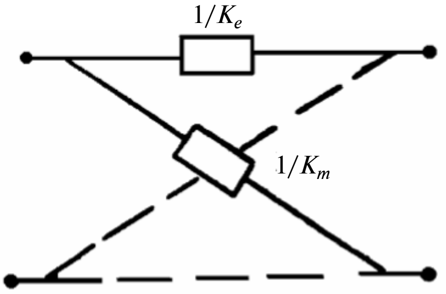


Fig. 15. Lattice (bridge) equivalent circuit of the bandpass filter with mutually detuned electric and magnetic types of oscillations

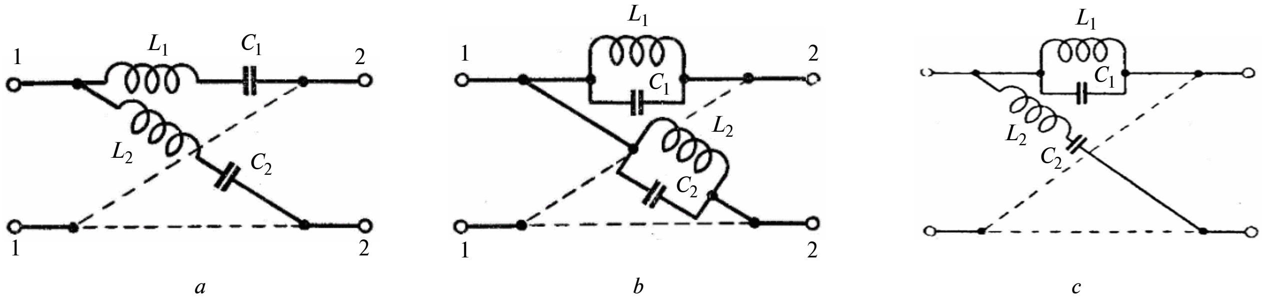


Fig. 16. Bridge equivalent circuits (*a*)–(*b*) of the bandpass filters, (*c*) – of the bandstop filter

Research of cells of metamaterials in terms of oscillations and coupling coefficients is also important in microwave engineering because most of the known proceedings use not "circuit" but rather purely physical terminology including negative dielectric and magnetic permeability. In addition, such approach allows determining the most correct equivalent circuit of the metamaterial unit cell.

In [62] the lattice (bridge) equivalent circuits (see Fig. 12) for the band microwave filters based on mutually detuned resonators in parallel channels (Fig. 15) was used for the first time.

Accordingly, the scattering matrix of such filter is described by formulae in (11). Bridge replacement circuit of the bandpass and bandstop filters are presented in Fig. 16.

If we take into account that when switching from an equivalent circuit of a bandpass filter in Fig. 16, *a* to an equivalent circuit of bandstop filter in Fig. 16, *c* a serial resonant oscillating circuit is "replaced" with parallel circuit and the coupling coefficient K_e will correspond to the coupling coefficient $1/K_e$, then by making a substitution in (9) we get the parameters of scattering matrix of the bandstop filter as

$$S_{11} = \frac{K_e}{(1+K_e)} - \frac{K_m}{(1+K_m)} \tag{26}$$

$$S_{21} = 1 - \frac{K_e}{(1+K_e)} - \frac{K_m}{(1+K_m)} \tag{27}$$

that is, the transmission coefficients of bandpass and bandstop filters with mutually detuned resonators in parallel channels (excluding out-of-band decoupling in bandpass filters) are dual. The transmission coefficient of bandpass filter corresponds to the reflection coefficient of bandstop filter and vice versa.

Note that the obtained values of the elements of the scattering matrix of the 4-pole network (transmission and reflection coefficients) are similar to the values of transmission and reflection coefficients of 8-pole network obtained in [52].

In [63] the main reasons for choosing bridge circuits as prototypes for the filters with mutually detuned resonators are observed:

- both in bridge prototype filters and in filters with mutually detuned resonators energy transmission is carried out by two independent channels;
- resonators in bridge prototype filters (for bandpass filters) are also deranged by frequency, otherwise the signals in the load are "subtracted" (i.e. equal in amplitude and opposite in phase) cancel each other out, that is there will be no signal transmission from input to output;
- at a certain ratio of reactivity of the filters' arms (as with proper selection of the order of relative position of "magnetic" and "electric" oscillations in filters with mutually detuned resonators) "poles of attenuation" [52] can be realised, otherwise there will be no "poles of attenuation";
- and finally, the most important – bridge filters belong to the so-called nonminimum-phase.

Bridge circuits allow you to receive much more wide bandwidth and synthesize the frequency response of the filter regardless of that of the phase-frequency.

Modeling the behavior filters as cells of structures of frequency-selective surfaces

Frequency-selective surfaces (FSS) contain microwave resonators of different types as a main element. Currently, metamaterial unit cells are used as such resonators. Therefore, the research methods used studying bandpass and bandstop filters were also applied to them. Let's consider only one of the important tasks of control of the FSS parameters – "turning on" and "turning off" their selective properties. Fig. 17 *a*) and *b*) shows the characteristics of bandstop filters, which consist of two mutually detuned resonators with different

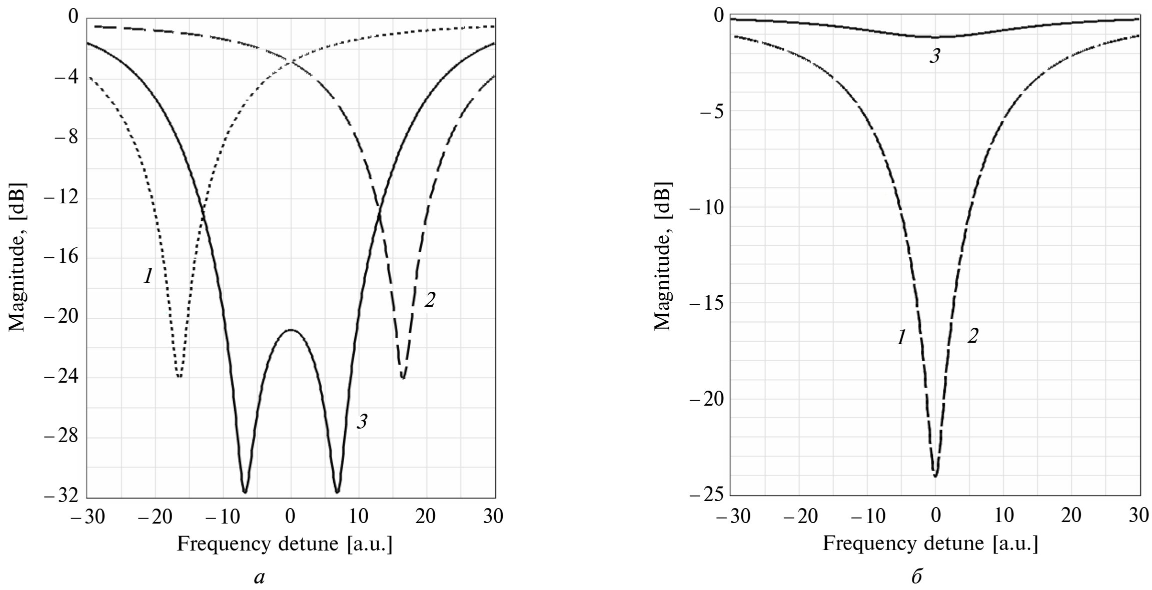


Fig. 17. Characteristics of bandstop filters with coupling coefficients $K_1 = K_2 = 15$, generalised detuning in Fig. *a* — a = 16.5, *b* — a = 0

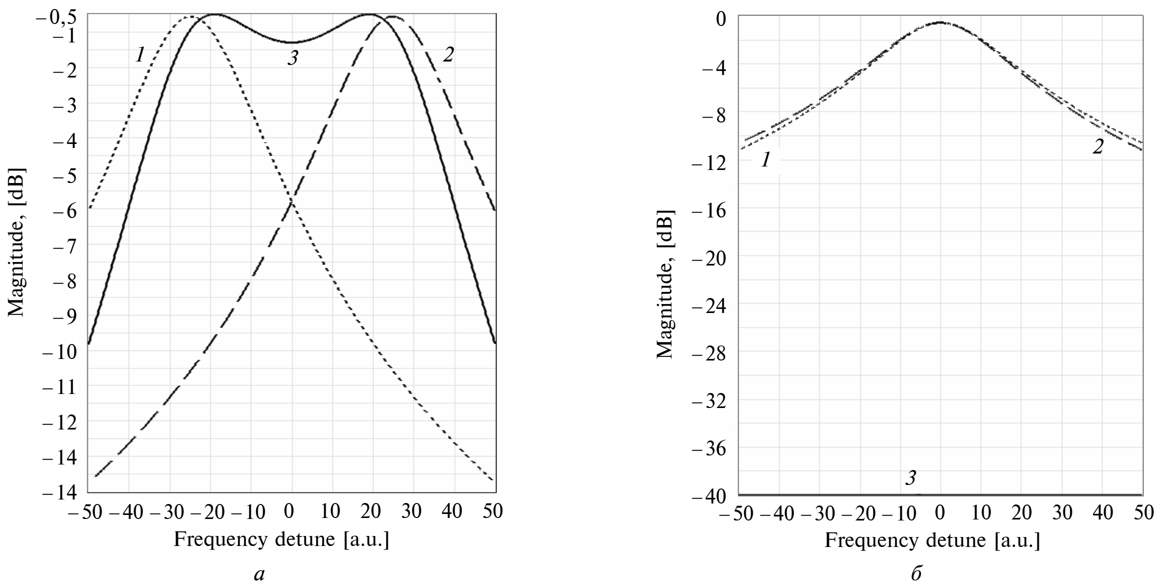


Fig. 18. Characteristics of bandpass filters with coupling coefficients $K_1 = K_2 = 15$, generalised detuning in Fig. *a* — **a** = 25, *b* — **a** = 0

types of oscillations. Curves *1* and *2* — transmission coefficients of individual resonators, curve *3* — combined characteristic of the double-resonance filter. Fig. 17, *a* (generalised detuning **a** = 16.5) stopband at the –20 dB level is eight times wider than the stopband of single-resonance filter at the same –20 dB level. In Fig. 17, *b* (oscillations coincide in frequency, generalised detuning = 0) double-resonance bandstop filter has an attenuation at the –1 dB level, i.e. almost ceases to be a bandstop.

Fig. 18, *a* and *b* show the characteristics of bandpass filters consisting of two mutually detuned resonators with different types of oscillations. Curves *1* and *2* — transmission coefficients of individual resonators, curve 3 — combined characteristic of the double-resonance filter. In Fig. 18, *a* (generalised detuning **a** = 25) bandwidth at the

–3 dB level is at least twice wider than bandwidths of single-resonance filters at the same level. In Fig. 18, *b* (oscillations coincide in frequency, generalised detuning **a** = 0) double-resonance bandpass filter has an attenuation at the –40 dB level (the given level of the out-of-band decoupling), i.e. almost ceases to be a bandpass.

Both in bandpass and bandstop filters the management of mutual detuning of the resonators can be done electronically, for example using varicaps as it is done in the filter in Fig. 4.

Conclusions

Microwave devices based on metamaterials cells are traditionally modeled by 4-pole networks with cascade-connected elements and resonators, the concepts of negative dielectric and magnetic permeability are widely used. The use of lattice (bridge) equivalent circuits as prototypes can not only describe the unique properties of metamaterials but also to optimise their characteristics on the basis of the obtained analytical models. At present, the authors study the properties of frequency selective surfaces and sensors based on metamaterials, the applicability of these approaches to acoustic metamaterials is studied. The scope of our research interests also includes physical processes characterised by the terms "slow light" and "electromagnetically induced transparency".

It should be noted that the considered 8-pole network is "basic" not only for bandpass and bandstop filters but for a number of elementary resonators. The waveguide resonators with dielectric resonators with different modes of oscillations can serve as an example [72–73].

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НАДВИСОКОЧАСТОТНІ ФІЛЬТРИ НА БАЗІ СТРУКТУР ІЗ РЕЗОНАТОРАМИ В ПАРАЛЕЛЬНИХ КАНАЛАХ ЯК КОМІРКИ МЕТАМАТЕРІАЛІВ

Проблематика. Багато властивостей метаматеріалів подібні до тих, які мають фільтри на взаємно розстроєних за частотою і не пов'язаних між собою резонаторах. Як низькочастотні прототипи таких мікрохвильових фільтрів використовуються мостові фільтри. Для подальшої розробки і проектування нових типів метаматеріалів необхідно встановити аналогію між метаматеріалами і фільтрами на взаємно розстроєних за частотою і не пов'язаних між собою резонаторах.

Мета досліджень. Створення моделей метаматеріалів на основі режекторних надвисокочастоних (НВЧ) фільтрів на взаємно розстроєних резонаторах, а також на базі низькочастотних прототипів.

Методика реалізації. Перевірка еквівалентності характеристик метаматеріалів фільтрам СВЧ на взаємно розстроєних резонаторах, виявлення притаманних їм закономірностей (типи коливань у паралельних каналах, розміщення полюсів загасання вище або нижче смуги пропускання), що проявляються незалежно від типів резонаторів, які використовуються, вивчення можливості використання мостових фільтрів-прототипів для моделювання метаматеріалів.

Результати досліджень. Детально досліджено базову модель 8-полюсника з резонаторами в паралельних каналах. Отримано аналітичні вирази для коефіцієнтів передачі та відбиття по всіх входах 8-полюсника. Досліджено 4-полюсники, реалізовані на основі згаданої вище базової моделі. Проведено експериментальні дослідження, які підтвердили адекватність аналітичних моделей. Встановлено аналогію між метаматеріалами і фільтрами СВЧ на взаємно розстроєних резонаторах, показано можливість використання як низькочастотних прототипів мостових смугових фільтрів.

Висновки. Для моделювання метаматеріалів можуть бути використані фільтри НВЧ на взаємно розстроєних резонаторах, а як низькочастотні прототипи – мостові смугові фільтри, методики проектування яких добре опрацьовані.

Ключові слова: метаматеріали; діелектричні резонатори; смужкові резонатори; режекторні фільтри НВЧ.

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СВЕРХВЫСОКОЧАСТОТНЫЕ ФИЛЬТРЫ НА БАЗЕ СТРУКТУР С РЕЗОНАТОРАМИ В ПАРАЛЛЕЛЬНИХ КАНАЛАХ КАК ЯЧЕЙКИ МЕТАМАТЕРИАЛОВ

Проблематика. Многие свойства метаматериалов похожи на те, что имеются у фильтров на взаимно расстроенных по частоте и не связанных между собой резонаторах. В качестве низкочастотных прототипов подобных микроволновых фильтров используются мостовые фильтры. Для дальнейшей разработки и проектирования новых типов метаматериалов необходимо установить аналогию между метаматериалами и фильтрами на взаимно расстроенных по частоте и не связанных между собой резонаторах.

Цель исследований. Создание моделей метаматериалов на основе режекторных сверхвысокочастотных (СВЧ) фильтров на взаимно расстроенных резонаторах, а также на базе низкочастотных прототипов.

Методика реализации. Проверка эквивалентности характеристик метаматериалов фильтрам СВЧ на взаимно расстроенных резонаторах, выявление присущих им закономерностей (типы колебаний в параллельных каналах, расположение полюсов затухания выше или ниже полосы пропускания), проявляющихся независимо от используемых типов резонаторов, изучение возможности использования мостовых фильтров-прототипов для моделирования метаматериалов.

Результаты исследований. Детально исследована базовая модель 8-полюсника с резонаторами в параллельных каналах. Получены аналитические выражения для коэффициентов передачи и отражения по всем входам 8-полюсника. Исследованы 4-полюсники, реализуемые на основе вышеупомянутой базовой модели. Проведены экспериментальные исследования, подтвердившие адекватность аналитических моделей. Установлена аналогия между метаматериалами и фильтрами СВЧ на взаимно расстроенных резонаторах, показана возможность использования в качестве низкочастотных прототипов мостовых полосовых фильтров.

Выводы. Для моделирования метаматериалов могут быть использованы фильтры СВЧ на взаимно расстроенных резонаторах, а в качестве низкочастотных прототипов – мостовые полосовые фильтры, методики проектирования которых хорошо проработаны.

Ключевые слова: метаматериалы; диэлектрические резонаторы; полосковые резонаторы; режекторные фильтры СВЧ.

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