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Study of Integrated Antennas-Filters on Dielectric Resonators

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New structures of dielectric resonators, simultaneously performing the functions of antennas and input filters of transceiver-transmitting devices of communication in the microwave wavelength range, are proposed. The new structures have the form of coupled systems of dielectric resonators, some of which are located in the segment of the evanescent waveguide and perform the functions of a band-pass filter, and the other part of the resonators are located in open space and perform the functions of an antenna system. A part of the resonators of the antenna subsystem has the form of dielectric semi-cylinders glued onto a metallic conductive screen and plays the role of directors forming the radiation characteristic. Two cases of the same and different in shape and material dielectric resonators are considered. For the case of dielectric resonators of different types, using the perturbation theory, a system of equations was obtained for the coefficients of the expansion of the scattering field from the known coupled oscillations of the system. Using the developed theory, the frequency dependence of the reflection coefficient of the proposed antenna-filters was calculated. It is shown that the proposed structures are characterized by much more rectangular amplitude-frequency characteristics, which allows to effectively solve the problems of electromagnetic compatibility of several communication systems using adjacent bandwidths. An additional advantage of the considered antenna filters are smaller dissipative losses, which is important for reducing bit errors. Calculated radiation patterns of antennas in the far zone. It is also shown that the radiation characteristics of the antennas do not change, or change only slightly compared with antennas made only with the help of similar structures of dielectric resonators. The obtained simulation results can significantly reduce computation time and optimize complex communication systems that simultaneously perform channel separation and radiation functions.

Key words: different dielectric resonator; scattering; antenna; bandpass filter

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Introduction

When designing antennas, several practical problems must be solved simultaneously: the necessary radiation and loss characteristics are implemented; stability of parameters in a given frequency band; the level of cross-polarization is minimized; spurious bandwidths are eliminated; electromagnetic compatibility with other systems are established [1–10]. Often, these tasks cannot be solved in a single device; in this case, you have to use several consecutively connected elements, each of which performs specified functions, for example, perform frequency filtering, in the input channel. The coupling of several different elements of the input paths leads to an increase in losses, which is an undesirable phenomenon, therefore attempts are being made to combine the functions of the antenna and the input filter in one device. In reality, this goal was not fully achieved in practice.

The purpose of this work is to describe and study a new type of microwave devices built on dielectric resonators (DRs), combining the functions of antennas and filters. Simultaneous utilization of several devices on the DRs in such antenna filters (AF) allows not only to reduce losses, dimensions and cost, but also eliminates the matching problems and generally simplifies tuning.

In this article, we explore the scattering characteristics of microwave input devices, based on DRs, performing functions of the antennas and input filters.

1 Statement of the problem

The purpose of this article is to study the scattering characteristics of a new type of microwave devices – combining the properties of bandpass filters and antennas made on dielectric resonators.

2 Antennas-filters calculation

From the experience of previous studies [11], the structures shown in Fig. 1, have the best efficiency that meet the requirements of the task.

Let consider the general solution of the scattering problem on the DRs structure shown in Fig. 1.

Suppose we have a complex system of $N+M$ DRs, part of which is located in the segment of evanescent metal waveguide (M), and a part of it (N) is located in open space (Fig. 1). The DRs $1,2,...,M$ plays the role of a bandpass filter, and the resonators $1,2,...,N$ plays the role of the antenna system. M-th DR is common to both subsystems. Let a wave $\left(\vec{E}^{+},\vec{H}^{+}\right)$ falls on 1 DR of the system $1,2,...,M$ via regular waveguide.

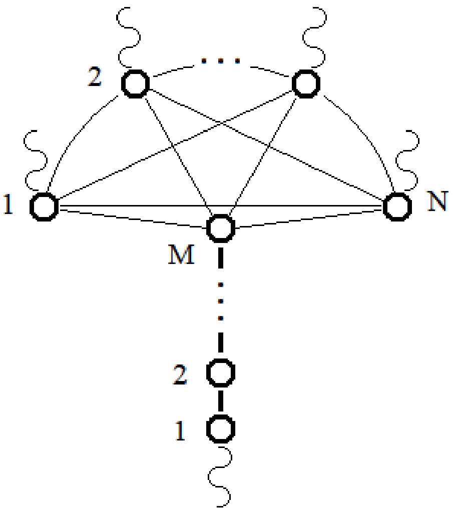

Fig. 1. The structure of coupled DRs of the antenna-filter. The partial DRs are denoted by circles; the coupling of the DRs with transmission line and open space on propagating waves are marked with wavy lines; the coupling between DRs on damped and propagation waves are marked with continuous lines.

We are looking for the solution of the scattering problem on the coupling DR structure:

$$\vec{E}\approx\vec{E}^{+}+\sum_{s=1}^{N+M}a^{s}\vec{e}^{s};\qquad \vec{H}\approx\vec{H}^{+}+\sum_{s=1}^{N+M}a^{s}\vec{h}^{s}, \tag{1}$$

where a^s are the unknown amplitudes ($s=1,2,...,N+M$) and $(\vec{e}^{s},\vec{h}^{s})$ are the s-th eigenoscillation field of the coupled DR system [11]. The eigenoscillation field of such system we represented as a superposition of fields of the isolated resonators $(\vec{e}_u,\vec{h}_u)$:

$$\vec{e}^{s}=\sum_{n=1}^{N+M}b_n^s\vec{e}_n;\qquad \vec{h}^{s}=\sum_{n=1}^{N+M}b_n^s\vec{h}_n.$$

$$\vec{E}\approx\vec{E}^{+}+\sum_{s=1}^{N+M}a^{s}\vec{e}^{s};\qquad \vec{H}\approx\vec{H}^{+}+\sum_{s=1}^{N+M}a^{s}\vec{h}^{s}. \tag{2}$$

Here $\vec{e}_n$ — is the electric field and $\vec{h}_n$ — is the magnetic field of the n-th isolated DR. The matrix $B=\|b_n^s\|$ of the amplitudes of coupled DR oscillations should satisfy the equation system [11] and must be an eigenvector of coupling operator K.

For the AF shown in Fig. 2, 3:

$$K=\left\|0,5i(\tilde{k}_1\delta_{s1}+ +\tilde{k}_{OS}\sum_{t=M}^{M+N}\delta_{st})\delta_{sn}+\kappa_{sn}(1-\delta_{sn})\right\|, \tag{3}$$

where $\tilde{k}_1$ — is the coupling coefficient of the 1th DR with the regular waveguide (Fig. 2, 3, a); $\tilde{k}_{OS}$ is the coupling coefficient of the $M,....(M+N)$ DR with open space; $\kappa_{sn}=k_{sn}$ — is the mutual coupling coefficient of the sth and tth resonators in the evanescent waveguide if $s<M$ and $n<M$ and κ_{sn} is the mutual coupling coefficient of split-cylindrical DRs in the open space if $s\geq M$ and $n>M$ [12]. We assumed also that the antenna array semi-resonators (Fig. 2, 3, a) are coupled only between themselves and the resonator M. Mth DR coupled with the evanescent waveguide through the aperture in the metal screen.

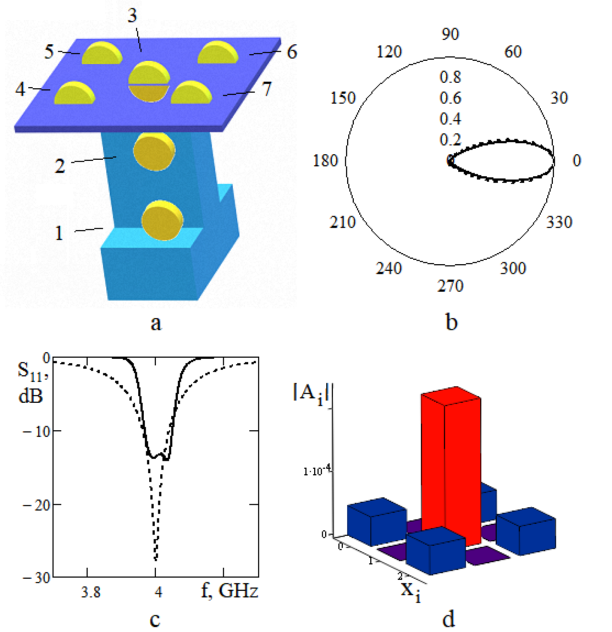

Fig. 2. The design of the AF on 7 cylindrical DRs ($M=3$; $N=4$) with azimuthally symmetric magnetic modes H_{101}^{+} (a) ($\varepsilon_{1r}=36$; $Q^D=2\cdot10^3$; $\Delta=L/2r_0=0,25$; L — is the height, r_0 — is the radius of the cylindrical DR). Far-field radiation pattern in the E-plane - continuous curve; in the H-plane – dotted curve (b). The reflection coefficient of the antenna without a filter (dashed curve) with a filter (solid curve) (c). Module amplitude distribution of the 3-7 DR field (d).

The found eigenvalue λ and eigenvectors of the matrix K

$$B = \begin{bmatrix} b_1^1 & b_1^2 & \dots & b_1^{M+N} \\ b_2^1 & b_2^2 & \dots & b_2^{M+N} \\ . & . & . & . \\ b_{M+N}^1 & b_{M+N}^2 & \dots & b_{M+N}^{M+N} \end{bmatrix}, \qquad (4)$$

were used to solve the scattering problem on a system of coupled DRs.

The reflection coefficient were determined from the ratios [11]:

$$R(\omega) = R_0 - \frac{Q^D}{\det B} \cdot \sum_{s=1}^{M+N} \frac{\det B_1^s}{Q_s(\omega)}, \qquad (5)$$

where R_0 is the reflection coefficient of the structure without DRs;

$$B_1^s = \begin{bmatrix} b_1^1 & \dots & b_1^s \tilde{k}_{11}^{-+} & \dots & b_1^{M+N} \\ b_2^1 & \dots & 0 & \dots & b_2^{M+N} \\ . & \dots & . & \dots & . \\ b_{M+N}^1 & \dots & 0 & \dots & b_{M+N}^{M+N} \end{bmatrix}, \qquad (6)$$

$\tilde{k}_{11}^{-+} = (c_1^- c_1^{+*})/(\omega_0 w_1) = \tilde{k}_1$; $Q_s(\omega) = \omega/\omega_0 + 2iQ^D(\omega/\omega_0 - 1 - \lambda_s/2)$; Q^D — is the dielectric Q-factor of the resonators; $\omega_0 = 2\pi f_0$; f_0 — is the oscillation frequency of isolated microresonators.

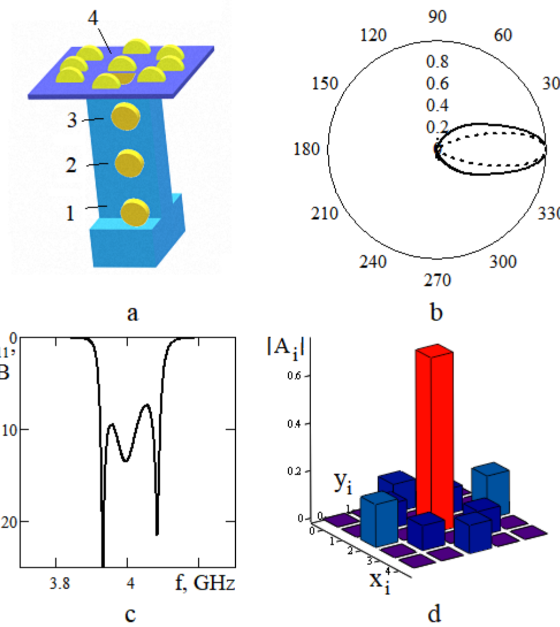

Fig. 3. AF on 12 cylindrical DRs ($M = 4$; $N = 8$) with azimuthally symmetric magnetic modes H_{101}^{+} (a). Far-field radiation pattern in the E-plane - continuous curve; in the H-plane - dotted curve (b). AF reflection coefficient as a function of the frequency (c). Module amplitude distribution of the 4-12 DR field (d).

The field of the antenna in the wave zone was represented in the form:

$$\begin{bmatrix} \vec{E} \\ \vec{H} \end{bmatrix} = \sum_{t=1}^{N} A_t \begin{bmatrix} \vec{e}_t^{\,\infty} \\ \vec{h}_t^{\,\infty} \end{bmatrix}. \qquad (7)$$

Here $A_t = \sum_{s=1}^{N} a^s b_t^s$ — is the amplitude of the t–th DR at scattering; $\begin{bmatrix} \vec{e}_t^{\,\infty} & \vec{h}_t^{\,\infty} \end{bmatrix}$ — is the field of t–th isolated DR in the wave zone.

3 Calculation of the antennas-filters based on different DR

The use of different forms of DR in one device often improves its scattering characteristics. In Fig. 4, 5 shows the design of AF based on cylindrical DR, made of different dielectrics.

We assumed that the fields of the DRs satisfy approximately the orthogonality conditions:

$$\begin{bmatrix} \vec{E} \\ \vec{H} \end{bmatrix} = \sum_{t=1}^{N} A_t \begin{bmatrix} \vec{e}_t^{\,\infty} \\ \vec{h}_t^{\,\infty} \end{bmatrix}. \qquad (8)$$

Using Maxwell's equations for the eigenoscillations of isolated and coupled DRs and the scattering field, as well as the orthogonality conditions (8), after integrating over the volume of each DR, we find a system of the equations for the amplitudes a^s:

$$\sum_{s=1}^{M+N} a^s b_t^s Q_{st}(\omega) = -(c_t^+)^* / P_t^D \qquad (9)$$
$$s, t = 1, 2, ..., M+N,$$

where b_t^s is the same amplitude of t-th resonator of s-th coupled mode of the system on the frequency $\tilde{\omega}^s$;

$$c_t^{\pm} = -\frac{1}{2} \oint_{s_t} \{[\vec{e}_t, \vec{n}]\vec{H}^{\pm}(\vec{r})^* + [\vec{n}, \vec{h}_t]\vec{E}^{\pm}(\vec{r})^*\} ds;$$

is the integral over the surface s_t of t-th DR; $\vec{n}$ — is the normal to the surface of t-th DR; $P_t^D = \omega_0 \frac{\varepsilon''_t}{2} \int_{v_t} |\vec{e}_t|^2 dv$ — is the loss power in the dielectric; $\tilde{\varepsilon}_t = \varepsilon_t - i\varepsilon''_t$ — is the complex dielectric permittivity of the t-th DR. For different DRs

$$Q_{st}(\omega) = 2iQ_t^D(\omega - \tilde{\omega}^s)/\omega_0 + \omega/\omega_0; \qquad (10)$$

$Q_t^D = \omega_0 w_t / P_t^D$ — is the Q-factor of loss in dielectric of t-th DR; $w_t = 1/4 \int_{v_t} (\varepsilon_t |\vec{e}_t|^2 + \mu_0 \left|\vec{h}_t\right|^2) dv$ — is the energy, stored in the dielectric of the t-th DR.

Formally, the solution of the system of equations (9) has the form:

$$a^s(\omega) = -\det B_s(\omega) / \det B(\omega), \qquad (11)$$

where for structure showed on Fig. 4, 5:

$$B_s(\omega)=\left[\begin{matrix} b_1^1 Q_{11}(\omega) & \dots & \left(c_1^+\right)^*/P_1^D \\ b_2^1 Q_{12}(\omega) & \dots & 0 \\ . & \dots & . \\ b_{M+N}^1 Q_{1M+N}(\omega) & \dots & 0 \end{matrix}\right. \left.\begin{matrix} \dots & b_1^{M+N} Q_{M+N1}(\omega) \\ \dots & b_2^{M+N} Q_{M+N2}(\omega) \\ \dots & . \\ \dots & b_{M+N}^{M+N} Q_{M+NM+N}(\omega) \end{matrix}\right];$$

$$B(\omega)= \\ =\begin{bmatrix} b_1^1 Q_{11}(\omega) & \dots & b_1^{M+N} Q_{M+N1}(\omega) \\ b_2^1 Q_{12}(\omega) & \dots & b_2^{M+N} Q_{M+N2}(\omega) \\ . & . & . \\ b_{M+N}^1 Q_{1M+N}(\omega) & \dots & b_{M+N}^{M+N} Q_{M+NM+N}(\omega) \end{bmatrix}.$$

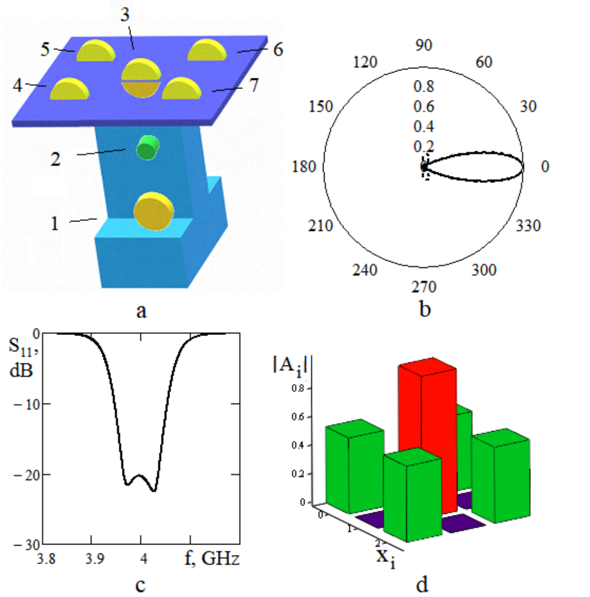

Fig. 4. AF on different cylindrical DRs ($\varepsilon_{1r}=36$; $\Delta_1=0,2$; $Q_1^D=3\cdot10^3$; $\varepsilon_{2r}=82$; $\Delta_2=0,5$; $Q_2^D=1\cdot10^3$). Far-field radiation pattern (b). Reflection coefficient as a functions of the frequency (c). Amplitude distribution of the 3-7 DR field (d).

The reflection coefficient R of different DR system can be obtained from (4), (1):

$$R(\omega)=R_0-c_1^-\sum_{s=1}^{N} a^s(\omega)b_1^s, \tag{12}$$

where R_0 is also the reflection coefficient of the transmission line without DRs.

The field of the antenna can be represented in the form (7), where $A_t(\omega)=\sum\limits_{s=1}^{N} a^s(\omega)b_t^s$.

4 Calculation and study of the AF parameters

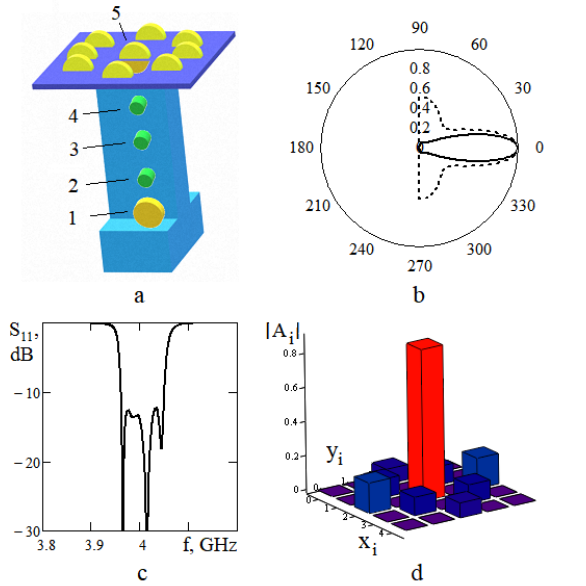

Fig. 5. AF on 13 different cylindrical DRs ($\varepsilon_{1r}=36$; $\Delta_1=0,44$; $Q_1^D=3\cdot10^3$; $\varepsilon_{2r}=82$; $\Delta_2=0,2$). Far-field radiation pattern (b). Reflection coefficient as a functions of the frequency (c). Amplitude distribution of the 3-7 DR field (d).

The developed theory was used to model and analyze the characteristics of proposed AF. The radiation characteristics of the AF, calculated on basis of model (1)-(11), are shown in Fig. 2-5, b. Comparison with the results obtained earlier [11, 12] showed that the radiation characteristics of the AF do not change, or change only slightly compared with antennas made only with the help of similar structures of dielectric resonators. The dependences of the reflection coefficient at the input of devices are in Fig. 2-5, c show a increased squareness for the expansion of the frequency band. Here $S_{11}(\omega)=20\lg|R(\omega)|$; $R(\omega)$ — is the reflection coefficient (5), (12). The distribution of the amplitude modules of the resonators in open space is shown in Fig. 2-5, d. It can be seen that the “active resonator” of the lattice has the maximum amplitude; the amplitude of the resonators playing the role of directors is noticeably smaller. The more efficiently the array is used, the greater the relative amplitudes of the passive resonators (Fig. 2-5, b, d).

It can be seen that the presented calculations predict a noticeable improvement in the frequency characteristics of the AF, which will significantly improve the electromagnetic compatibility of devices when used in telecommunication wireless access systems. The frequency reflection coefficient becomes more rectangular (Fig. 2, 4). The working frequency band is increasing.

5 Discussion and Conclusion

The proposed devices can significantly improve the frequency characteristics of the antennas performed on the DRs.

A scattering theory of the electromagnetic waves on new elements of receiving and transmitting communication systems has been developed. As our studies have shown, the addition of DRs, playing the role of filters, slightly affects the characteristics of the antenna, however, the antenna resonators have a noticeable effect on the frequency scattering characteristics of the reflection coefficient. Setting up such AFs will require additional research on asymmetrical load filters.

The proposed theory can be used to model antennas with account for more subtle diffraction effects. In particular, our calculations did not take into account the influence of the edges of the reflecting metal plane. This problem can be solved by introducing additional diffraction terms into the decomposition (1).

A developed analytical model can be used to directly optimize telecommunication and other devices that are built on different types of dielectric resonators. Obtained results can significantly reduce the computation time for optimizing complex communication systems, containing a large number of DRs.

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Вивчення інтегрованих антен-фільтрів на діелектричних резонаторах

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Запропоновано нові структури діелектричних резонаторів, які одночасно виконують функції антен та вхідних фільтрів приймально-передавальних пристроїв зв'язку сантиметрового діапазону довжин хвиль. Нові структури мають вигляд зв'язаних між собою систем діелектричних резонаторів, частина яких розташована у відрізку позамежного хвилеводу і виконує функції смугового фільтру, а друга частина резонаторів розташована у відкритому просторі і виконує функції антенної системи. Частина резонаторів антенної підсистеми має форму діелектричних полуциліндрів, наклеєних на металевий провідний екран, і грає роль директорів, які формують характеристику випромінювання. Розглянуто два випадки однакових та різних за формою та матеріалом діелектричних резонаторів. Для випадку діелектричних резонаторів різних видів, за допомогою теорії збурень, одержана система рівнянь для коефіцієнтів розкладу амплітуд поля розсіювання за відомими зв'язаними коливаннями системи. За допомогою розвиненої теорії розраховані частотні залежності коефіцієнтів відбиття запропонованих антен-фільтрів. Показано, що запропоновані структури характеризуються значно більш прямокутними амплітудно-частотними характеристиками, що дозволяє ефективно вирішувати завдання електромагнітної сумісності декількох систем зв'язку, які використовують суміжні частотні смуги. Додатковою перевагою розглянутих антен-фільтрів є менші дисипативні втрати вхідних пристроїв, що є важливим для зменшення бітових помилок. Розраховані діаграми спрямованості антен в дальній зоні. Показано, що характеристики випромінювання антен не змінюються, або змінюються не значним образом у порівнянні з характеристиками антен, виконаних тільки за допомогою подібних структур діелектричних резонаторів. Отримані результати дозволяють значно скоротити час розрахунків та оптимізувати складні системи зв'язку, які одночасно використовують частотне розділення каналів та виконують функції випромінювачів.

Ключові слова: різні діелектричні резонатори; розсіювання; антени; смугові фільтри

Изучение интегрированных антенн-фильтров на диэлектрических резонаторах

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Предложены новые структуры диэлектрических резонаторов, одновременно выполняющие функции антенн и входных фильтров приемо-передающих устройств связи сантиметрового диапазона длин волн. Новые структуры имеют вид связанных между собой систем диэлектрических резонаторов, часть которых расположена в отрезке запредельного волновода и выполняет функции полосового фильтра, а другая часть резонаторов расположена в открытом пространстве и выполняет функции антенной системы. Часть резонаторов антенной подсистемы имеет форму диэлектрических полуцилиндров, наклеенных на металлический проводящий экран, и играет роль директоров, формирующих характеристику излучения. Рассмотрено два случая одинаковых и разных по форме и материалу диэлектрических резонаторов. Для случая диэлектрических резонаторов разных видов, с помощью теории возмущений, получена система уравнений для коэффициентов разложения поля рассеяния по известным связанным колебаниям системы. С помощью развитой теории рассчитана частотная зависимость коэффициента отражения предложенных антенн-фильтров. Показано, что предлагаемые структуры характеризуются значительно более прямоугольными амплитудно-частотными характеристиками, что позволяет эффективно решать задачи электромагнитной совместимости нескольких систем связи, использующих смежные полосы пропускания. Дополнительным преимуществом рассмотренных антенн-фильтров являются меньшие диссипативные потери, что является важным для уменьшения битовых ошибок. Рассчитаны диаграммы направленности антенн в дальней зоне. Отмечено также, что характеристики излучения антенн не меняются, или меняются незначительно по сравнению с антеннами, выполненными только с помощью подобных структур диэлектрических резонаторов. Полученные результаты моделирования позволяют значительно сократить время вычислений и оптимизировать сложные системы связи, одновременно выполняющие функции разделения каналов и функции излучателей.

Ключевые слова: различные диэлектрические резонаторы; рассеяние; антенна; полосовой фильтр