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ANTENNA ON MICROSTRIP LINE WITH ORTHOGONALLY PLACED DIELECTRIC RESONATOR

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Single-element antenna that uses microstrip line as a feeder and cylindrical dielectric resonator orthogonally oriented relative to the line as a radiating element has been investigated. The complete mathematical model of proposed design consisting of a number of analytical expressions for main antenna characteristics is obtained and verified. The analytical relation for antenna return loss versus stub length and coupling coefficients of dielectric resonator with the feeding line and open space is derived. The assessment of the potential possibility and conditions of perfect antenna matching with feeding line is carried out. The influence of main parameters of a dielectric resonator antenna on its characteristics is examined. The numerical analysis of different antenna parameters proving the obtained analytical expressions is performed. The results of theoretical analysis are in good agreement with experimental data.

Introduction

A dielectric resonator (DR) as a radiating element was first investigated in [1]. Currently, dielectric resonator antennas (DRA) are presented by a large number of designs and are widely used as small-sized antennas in microwave devices. The DRAs that use microstrip line as a feeder and DRs as radiators have wide applications. Two basic excitation methods have been proposed for such antennas: using slot, etched in ground plane of microstrip line [2], and placing DR beside a strip on the surface of dielectric substrate of the line [3]. In both cases, DR is placed so that its axis of symmetry is orthogonal to ground plane of microstrip line. The excitation method different from both ones stated above was proposed in [4]. It implies that the axis of symmetry of cylindrical DR lies in a ground plane of microstrip line. Such an arrangement of a DR relative to microstrip line was called as orthogonal.

In the case of orthogonal arrangement of DR in microstrip line, the influence of construction metal surfaces on characteristics of entire radiating system including the Q-factor and operating frequency is noticeably less relative to slot excitation. The advantage of DRAs using orthogonal orientation is construction simplicity: it is not necessary to adjust the parameters of coupling element (slot) and coordinates of the placement point of the DR.

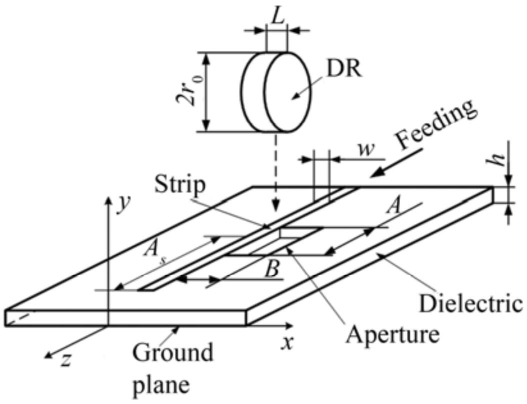
Regardless the excitation method used in such antennas, dielectric resonator is placed in a segment of feeding line open-ended on one side. The last one forms a stub, which length has impact on antenna parameters and characteristics and can be used to control them. The existing approach to analyzing dielectric resonator antennas consists in considering DR as lumped obstacle, which impedance is a complex value in certain section of the line. The impedance is calculated numerically, and the stub length is chosen so that its reactance could compensate the reactance of the DR on resonant frequency [5]. Such an approach does not allow obtaining general dependences characterizing the influence of main elements of a device on its parameters and characteristics. At the same time, such general expressions would permit to analyze mutual influence of each pair of parameters on one another. Additionally, the replacement of the numerical analysis by analytical expressions would shorten the duration of calculation.

Formulation of the problem

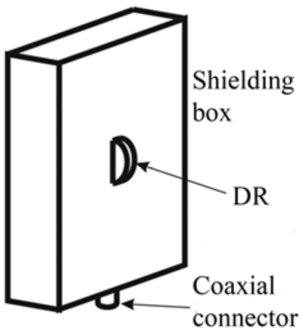
The single-element DRA with orthogonally oriented resonator is shown in Fig. 1. The half of DR's volume is situated inside the shielding box and coupled with feeding line, and the rest is situated outside the construction and provides radiation into surrounding space (Fig. 1*b*). In this work, the impact of basic parameters of construction (parameters of DR and microstrip line, location point of DR, stub length) on parameters and characteristics of the device (operating frequency, bandwidth, reflection coefficient, coefficient of efficiency) is studied.

The purpose of the present work is to obtain and verify a number of analytical expressions for main antenna characteristics, which, being put together, present the full mathematical model for DRA of proposed design. They are as follows: coupling coefficient of cylindrical DR with microstrip line in the case of orthogonal orientation of a DR relative to microstrip line; return loss of the antenna as a function of stub length and coupling coefficients; antenna coefficient of efficiency as a function of the same parameters.

In addition, the assessment of the potential possibility and conditions of perfect matching with feeding line in the proposed design will be considered.



(a)



(b)

Fig. 1. Design (*a*) and appearance (*b*) of DRA using orthogonal orientation of DR relative to microstrip line.

Coupling coefficient of a cylindrical DR with an edge microstrip line

To locate DR in microstrip line, it is required to remove parts of both dielectric substrate and ground plate, which are adjacent directly to the strip (Fig. 1*a*). To simplify the problem, let us consider the interaction of a cylindrical DR with infinite microstrip line, part of which is completely removed on one side of the strip (Fig. 2). The prerequisites for such a simplification are, firstly, low disturbance of natural oscillations of the DR by the removed part of the line, and, secondly, the fact of concentration of the major part of traveling wave energy in a small volume around the strip. Let the microstrip line with removed part, as described above, (Fig. 2) be called as edge microstrip line (EML) in the text below. Therefore, to provide the analysis of interaction of a cylindrical orthogonally oriented DR with microstrip line, the system comprising of DR in EML is enough to be studied.

To provide the calculation of parameters of the system, it is sufficient to compute coupling coefficients of the cylindrical DR with the line and with open space (k_w, k_{os} respectively) [6, 7], which can be obtained from general power expressions:

$$\frac{1}{Q_{ext}} = \frac{P_\Sigma}{P} = \frac{P_w}{P} + \frac{P_{os}}{P} = k_w + k_{os};$$
$$P_\Sigma = P_w + P_{os};\ P = 2\pi f \cdot w(f);$$
$$k_w = \frac{P_w}{P};\ k_{os} = \frac{P_{os}}{P},$$

where Q_{ext} denotes the external Q-factor of the dielectric resonator; P is the power stored in the resonator volume; P_Σ represents the total power radiated by the dielectric resonator; P_w defines the power radiated by the dielectric resonator into the transmission line; P_{os} is the power radiated by the dielectric resonator into open space; k_w, k_{os} are the DR coupling coefficients with the transmission line and open space respectively; f is the oscillation frequency of the dielectric resonator; $w(f)$ denotes the frequency dependent energy stored in the volume of the DR over the period of oscillations.

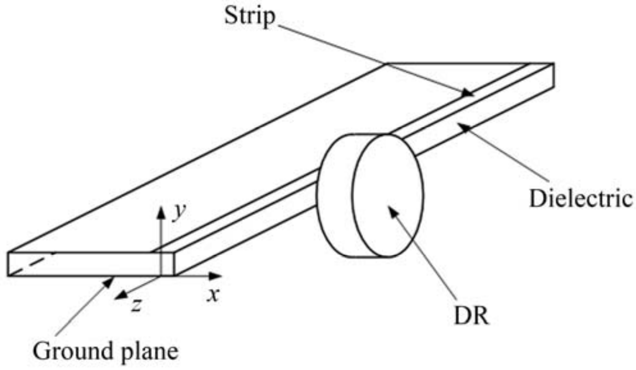


Fig. 2. Orthogonal orientation of the DR in edge microstrip line.

When calculating the parameters of the system, at the first stage the parameters of DR should be computed [7], at the second one the values of mentioned above coupling coefficients are defined, at the third one the characteristics of the whole system are determined. Therefore, the problem of analysis of the DR coupled with microstrip line is reduced to a study of these coefficients as functions of x coordinate in the case of the orthogonal orientation of the DR relative to the line [4]. The expression for k_{os} was obtained in [7] and is considered to be known. For the coupling coefficient of the DR with EML the following expression was obtained:

$$k_w = \frac{w_0^2 r_0^2 L q_\perp^2 p_\perp^2 p_z^2 (k_1^2 - k_0^2)^2 f^2(0)}{\pi^2 f_0^3 \varepsilon_0 \varepsilon_{1r} \mu_0^2 (2p_z - \sin 2p_z)} \times \\ \times \frac{1}{J_1^2(p_\perp) - J_0(p_\perp) J_2(p_\perp)} H_x^2(x_0, y_0, z_0); \qquad (1)$$

$$f(\zeta)=[p_z \sin p_z \cos(\zeta q_z)-\zeta q_z \sin(\zeta q_z)\cos p_z]\times$$
$$\times\frac{J_0(p_\perp)J_2(\sqrt{1-\zeta^2}q_\perp)-J_0(\sqrt{1-\zeta^2}q_\perp)J_2(p_\perp)}{(p_z^2-\zeta^2 q_z^2)[p_\perp^2-q_\perp^2(1-\zeta^2)]}$$

where $H_x(x_0, y_0, z_0)$ is the x component of the magnetic field strength of the line in the point of geometrical centre of the resonator; other parameters are defined by geometrical sizes of the DR and its dielectric constant as described in [7].

To verify (1), the obtained theoretical values of coupling coefficient k_w were compared with the experimental data as shown in Fig. 3. The magnetic field strength of the transmission line $H_x(x_0, y_0, z_0)$ was computed using three numerical techniques [4]: finite-difference method, finite element method and method of conformal mapping.

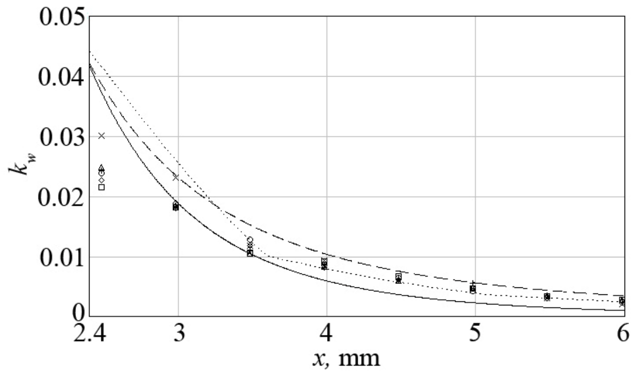


Fig. 3. The coupling coefficient of the dielectric resonator with edge microstrip line as a function of x coordinate: (solid line) finite-difference method; (dotted line) finite element method; (dashed line) method of conformal mapping; markers represent the measured results.

When turning from the electrodynamic system shown in Fig. 2 to antenna (Fig. 1*b*), a segment of microstrip line with DR placed beside it is complemented with metal box shielding parasitic radiation of microstrip line and DR. The half of DR's volume is situated inside the shielding box and interacts with transmission line, while another part of the DR is outside and produces radiation into open space.

After the shielding box has been installed, the depth of resonance rises, which is caused by less level of outside radiation of the DR. It was established that coupling coefficients of the DR with transmission line and with open space in cases of absence and presence of metal box obey the following relations:

$$k_w = k_w^0;\ k_{os} = \frac{1}{2}k_{os}^0,$$

where $k_w{}^0$, $k_{os}{}^0$ are coupling coefficients without the shielding box in the structure depicted in Fig. 2; k_w, k_{os} are coupling coefficients in the case if metal shielding box is present as shown in Fig. 1*b*.

The impact of the stub length on the characteristics of the DRA

The scattering matrix of a cylindrical dielectric resonator orthogonally oriented relative to infinite matched with a source microstrip line can be expressed as follows [8]:

$$S=\frac{1}{1+K_w+K_{os}+j\xi}\begin{bmatrix} K_w & 1+K_{os}+j\xi \\ 1+K_{os}+j\xi & K_w \end{bmatrix}, \quad (2)$$

where $K_w = P_w / P_t$, $K_{os} = P_{os} / P_t$ are coupling coefficients of the dielectric resonator with microstrip line and open space respectively; $\xi = Q_d(f/f_0 - f_0/f) \approx \approx 2Q_d(f/f_0-1)$ is relative frequency detuning; P_w, P_{os} are powers of waves, radiated by the dielectric resonator into microstrip line and open space respectively; P_t denotes the power of heat loss in the volume of the dielectric resonator; Q_d represents the Q-factor of the dielectric resonator material; f denotes the operating frequency; f_0 is the resonant frequency of the dielectric resonator on the TE_{101} mode.

If microstrip line of finite length is loaded on any complex impedance, the reflection coefficient of such structure (Fig. 1*a*) can be determined by the following expression [9]:

$$r = S_{11}+S_{12}S_{21}\frac{R\cdot e^{-j2\Theta}}{1-S_{22}R\cdot e^{-j2\Theta}}, \quad (3)$$

where R defines the reflection coefficient of complex impedance load; $\Theta = 2\pi l/\lambda$; $l = A_s$ is stub length; λ denotes the wavelength in the transmission line.

Substituting S_{ij} from (2) into (3), the expression for reflection coefficient of the antenna as a function of relative detuning and stub length can be obtained:

$$r(\xi,\Theta)=e^{-j2\Theta}\times$$
$$\times\frac{R-K_wR+\alpha+K_{os}R+j(\xi R+\beta)}{1+K_w-R\alpha+K_{os}+j(\xi+R\beta)}, \quad (4)$$

where $\alpha = K_w \cos 2\Theta$; $\beta = K_w \sin 2\Theta$. In (4), the value $R=1$ corresponds to open-ended line (open stub) and $R=-1$ to short circuited line.

To determine the frequency (or detuning ξ), at which reflection coefficient is minimal, the equation $\partial/\partial\xi\,|r(\xi,\Theta)|=0$ should be solved. There are short-circuited and open-circuited stubs cases when $R=1$ and $R=-1$ respectively, that are of interest. For these cases, the point of function $|r(\xi,\Theta)|$ minimum under condition of fixed Θ defines the optimal value of relative frequency detuning:

$$\xi_{opt} = -RK_w \sin 2\Theta \qquad (5)$$

Using expression (5), the relative shift of resonant frequency as a function of the stub length can be determined as:

$$\frac{f_r - f_0}{f_0} = \frac{\xi_{opt}}{2Q_d} = -\frac{K_w R}{2Q_d} \sin 4\pi \frac{l}{\lambda}. \qquad (6)$$

Substituting (5) into (4), it is possible to obtain the least value of the reflection coefficient of short-circuited or open-circuited structure as a function of stub length. In the case of microstrip line, it is more suitable to use the open-circuited stub (R=1). Then, taking into account the relation $\Theta = 2\pi l / \lambda$, we can obtain the required expression for reflection coefficient as a function of stub length l in the following form:

$$r(l/\lambda) = r[\xi_{opt}(2\pi l/\lambda), 2\pi l/\lambda] = \\ = \frac{R - K_w R + K_w \cos(4\pi l/\lambda) + K_{os} R}{1 + K_w - RK_w \cos(4\pi l/\lambda) + K_{os}} e^{-j4\pi l/\lambda}. \qquad (7)$$

A study of the antenna shown in Fig. 1*a* with resonant frequency $f_0 = 2.6$ GHz and coupling coefficients $K_w = 4.5$, $K_{os} = 5.33$ has been performed. The expressions for computing these coefficients are given respectively in [4] and [7]. The value of Q-factor of the dielectric resonator material was chosen as $Q_d = 1300$. The antenna parameters and its characteristics in the case of $\lambda/2$ short-circuited stub which is equivalent to $\lambda/4$ open-circuited one are given in [10].

To verify the expression (7), the numerical experiment was made. For each value of stub length, the studied system was modeled using finite element method with following determination of the reflection coefficient. Results of the numerical antenna modeling are presented in Fig. 4. The choice of this numerical technique was conditioned by its simplicity comparing with full-size experiment and absence of errors while measuring the stub length. As far as the results of modeling the studied DRA coincide with the ones obtained from the experiment carried out for a few certain values of stub length, the numerical experiment (i.e. computer modeling) can be considered as valid approach to the analytical expressions verification.

As follows from relations (5)—(7), the frequency of minimum return loss, as well as the width of the resonant curve $|S_{11}|$ (antenna bandwidth) depends on stub length l. These conclusions are confirmed by Fig. 4, where the following values are plotted for each quantity of the stub length. Results of numerical simulation of the reflection coefficient (7) are plotted by solid curve. Measured values of reflection coefficient $|S_{11}|$ are presented by circles. The relative shift of antenna operating frequency from resonance $(f_r - f_0)/f_0 \cdot 100\%$ calculated according to (6) is shown by dashed line. The dotted line illustrates the relative width of resonant curve $\Delta f / f_0 \cdot 100\%$ at level of -3 dB, where Δf is determined by solving the equation $20\log|r(f)| = -3$ for each value of stub length.

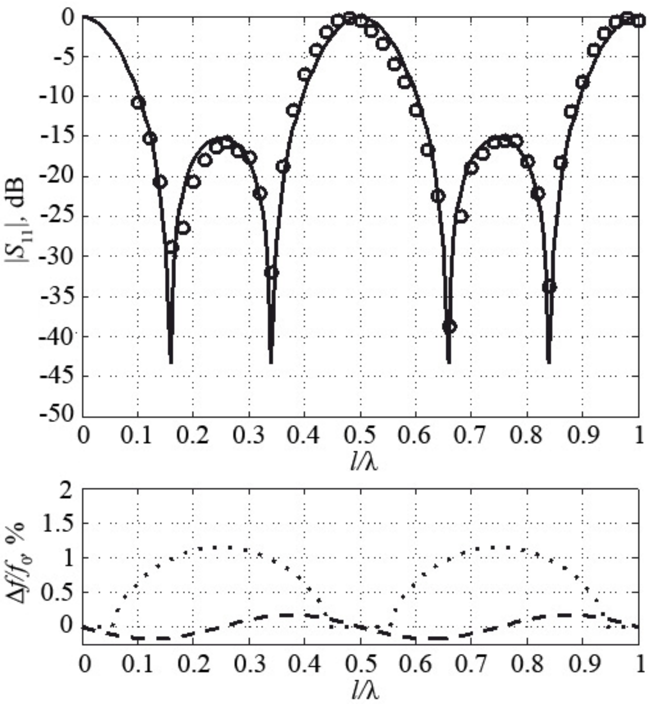


Fig. 4. Characteristics of microstrip resonator antenna as functions of the relative stub length l/λ: (solid line) reflection coefficient; (dotted line) relative bandwidth $\Delta f / f_0$; (dashed line) shift of resonant frequency $(f_r - f_0)/f_0$.

It can be seen from Fig. 4 that maximal value of bandwidth is observed at $l = (2n-1)\lambda/4$, where $n = 1, 2, \ldots$. However, changing the stub length relative to basic value $\lambda/4$, it is possible to decrease the return loss considerably and thereby enhance antenna matching with slight reduction of bandwidth.

Equating (7) to zero, it is possible to find such values of stub length, where perfect matching of the device is observed:

$$l/\lambda = \arccos[(K_w - K_{os} - 1)R / K_w]/(4\pi).$$

For $R = 1$ and stated above values of coupling coefficients for the studied antenna $(l/\lambda)_{opt} = 0.151$. Then, the minimal values of $|S_{11}|$ occur at $l/\lambda = (l/\lambda)_{opt}$; $l/\lambda = 1/2 - (l/\lambda)_{opt}$; $l/\lambda = 1/2 + (l/\lambda)_{opt}$; $l/\lambda = 1 - (l/\lambda)_{opt}$. At these points the return loss in dB converge to $-\infty$. The graphs of $|S_{11}|$ for two values of stub length $l/\lambda = 0.25$ and $l/\lambda = 1/2 - (l/\lambda)_{opt} \approx \approx 0.34$ are depicted in Fig. 5.

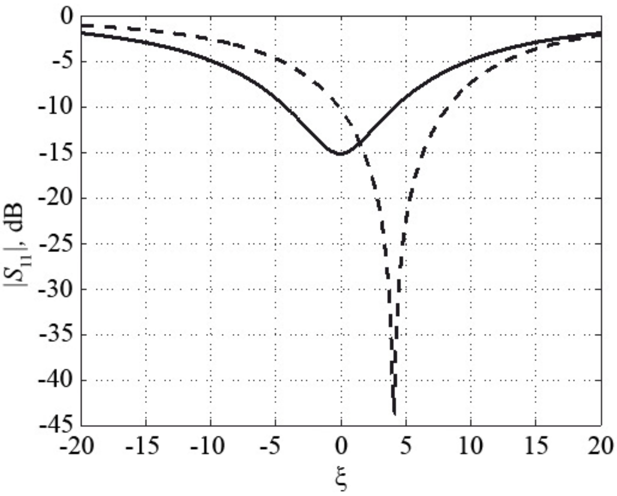


Fig. 5. Modulus of reflection coefficient for different stub lengths: (solid line) the basic value $l = \lambda/4$; (dashed line) the value corresponding to perfect matching $l = 0.34\lambda$.

Let us analyze the expression (7) for different values of coupling coefficients. Equating $|r(\xi,\Theta)|$ to zero, we can obtain the expression describing the relation between K_w and K_{os}, which causes zero return loss (in natural units):

$$K_w = \frac{R\left(1+K_{os}\right)}{R-\cos 2\Theta}.$$

In the case when $R=1$ and $l=\lambda/4$, the last expression gives the following value:

$$K_w^{opt} = \frac{1+K_{os}}{2}.$$

The calculated return loss of the standing wave antenna as a function of stub length for constant K_{os} and different values of K_w is presented in Fig. 6. As can be seen, there are no points of perfect matching if $K_w < K_w^{opt}$. Value $K_w = 0$ corresponds to zero coupling when dielectric resonator can be considered as removal from the transmission line. Since in this case $|S_{11}| = 0$ dB, the complete reflection of the incident wave is observed. For $K_w = K_w^{opt}$, there are two points of perfect matching disposed at $l = \lambda/4$ and $l = 3\lambda/4$. If $K_w > K_w^{opt}$, splitting the curve around each of these points occurs, so that there are four points of perfect matching.

For the antenna under study $K_w > K_w^{opt}$. To reduce K_w to the value K_w^{opt} and thereby provide perfect matching, the DR should be moved away from the strip along axis x as shown in Fig. 7*a*.

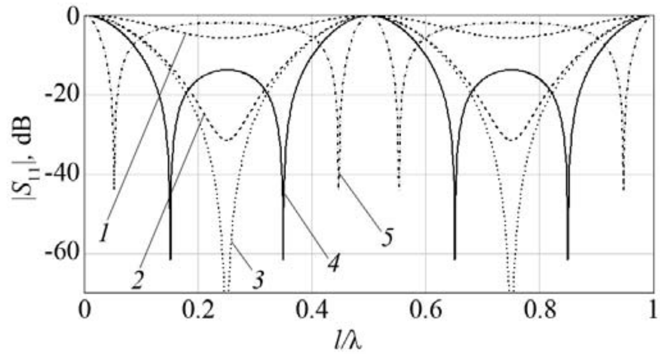
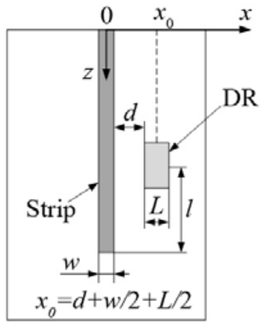
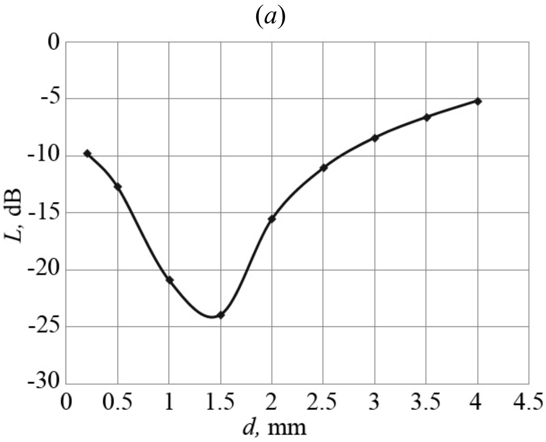


Fig. 6. Return loss of the standing wave antenna as a function of stub length for $K_{os} = 5.33$ and different values K_w: (*1*) $K_w = 1$; (*2*) $K_w = 3$; (*3*) $K_w = (1+K_{os})/2 = 3.165$; (*4*) $K_w = 4.8$; (*5*) $K_w = 30$.



(*a*)



(*b*)

Fig. 7. Matching the antenna: (*a*) general appearance of the scheme; (*b*) return loss L as a function of distance d between dielectric resonator and the strip.

Return loss values at operating frequency calculated by finite element method are plotted in Fig. 7*b*. It can be seen that optimal distance $d \approx 1.5$ mm.

The expression for coefficient of efficiency of DRA as a function of coupling coefficients and stub length was also obtained:

$$\eta = \frac{2K_wK_{os}}{(1+K_w+K_{os})^2(1+K_w+K_{os}-K_wR\cos 2\Theta)^2}\times$$
$$\times\{[(1-R\cos 2\Theta)(1+K_w+K_{os})+K_w\sin^2 2\Theta]^2 +$$

$$+(K_w \cos 2\Theta + R + RK_{os})^2 \sin^2 2\Theta] .$$

The coefficient of efficiency as a function of coefficient K_w for different values of K_{os} is shown in Fig. 8.

Analyzing the results presented in Fig. 8, one can see that with an increase of K_w relative to K_w^{opt} the coefficient of efficiency slowly decreases. As K_w decreases relative to K_w^{opt}, coefficient of efficiency falls much faster, reaching zero in $K_w = 0$. Therefore, the range $K_w < K_w^{opt}$ is characterized by low efficiency and inability of perfect matching. Thus, this range is unacceptable for operation. Coefficient of efficiency can reach the values 80—90% and rises while the coupling coefficient of DR with open space grows.

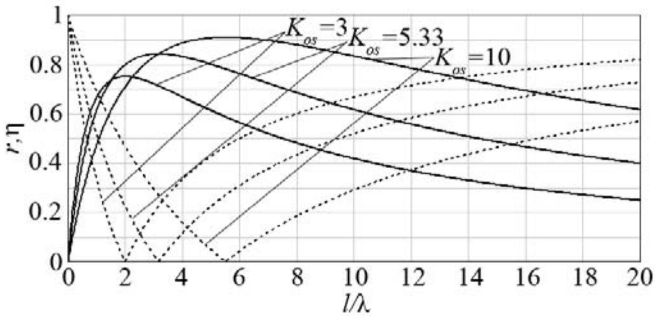


Fig. 8. Reflection coefficient r (dashed line) and coefficient of efficiency η (solid line) as functions of coupling coefficient of dielectric resonator with transmission line K_w for different values of coupling coefficient of DR with open space K_{os} .

Conclusion

Single-element antenna that use microstrip line as a feeder and cylindrical dielectric resonator, orthogonally oriented relative to the line, as a radiating element have been proposed and comprehensively investigated. The analytical expressions for the coupling coefficient of a cylindrical DR with edge microstrip line are obtained. If the DR is placed in immediate vicinity of the strip, the coupling coefficient can slightly differ from calculated value and operating frequency sustains a slight rise. It is mainly caused by disturbance of the field of natural oscillations of the DR by the strip.

When the feeding structure is placed into metal enclosure (shielding box), in order to calculate frequency characteristics of the device, it is enough to leave coupling coefficient of the dielectric resonator with microstrip line without change and to halve coupling coefficient of DR with open space.

There is complex interconnection between stub length and coupling coefficients of the dielectric resonator with transmission line and open space. As the stub length rises or falls, operating frequency of the DRA rises or falls respectively.

The frequency shift is in linear dependence on coupling coefficient K_w. Moreover, bandwidth decreases if stub length deviates from basic value $\lambda / 4$. There is optimal value of coupling coefficient of a dielectric resonator with the line, which linearly depends on coupling coefficient of a dielectric resonator with open space. Owing to this fact, it is possible to decrease return loss and thereby enhance matching by moving the DR away from the strip (if $K_w > K_w^{opt}$).

The main benefits of the antenna considered in the present work are small size, high efficiency in millimeter wave band and higher power capability comparing with DRA using slot excitation. The main application of such antennas is mobile and stationary terminals for wireless telecommunication systems operating in high frequency bands.

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