Mutual Coupling between the Resonators in the Planar Microstrip Antennas
Part I. Principles and Basic Microstrip Resonators

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Abstract – **In recent years there has been considerable interest in microstrip antenna arrays. One of the basic disadvantage of this type of antennas is the presence of surface waves. Because of this reason in this paper are shown the basic principles for excitation and propagation of the surface waves. There is made a comparison between basic constructions of microstrip resonators. The numerical data of reflection and transmission coefficients are evaluated by the method of moments. The results can be used in design of microstrip antenna arrays.**

I. INTRODUCTION

The dielectric coated ground plane of a printed antenna supports a finite number of surface wave modes, which propagate in a direction parallel to the air-dielectric interface. The properties of surface waves in absorbing layers are not well known. Only recently, they have been studied in details by few authors [1], [2].

In particular, it was found that in absorbing dielectric layers, TM surface wave cannot propagate if the frequency exceeds the so-called upper cutoff frequency. At this upper cutoff frequency, a TM-surface wave becomes a regular plane wave incident on an absorbing layer at the Brewster angle [3].

The power launched into the surface waves is power, which will eventually be lost. Hence, the excitation of surface waves lowers the overall radiation efficiency of the antenna. For finite-size substrates, the surface wave power will diffract from the edges of the substrate resulting in a disturbance of the radiation pattern [4]. Furthermore, the excitation of surface waves also results in increased mutual coupling between distant antenna elements, since the surface wave fields decay more slowly with radial distance than the space-wave fields [5]. Because of these reasons, the excitation of surface waves is generally on desirable. Since the dominant TM_0 surface wave of a grounded dielectric layer has a zero cutoff frequency, a microstrip antenna will in generally always excite some surface-wave power.

In the paper, basic types of microstrip antennas resonators are presented which are with reduced mutual coupling. A comparative evaluation in respect of excitation of surface waves, frequency band, and technology is made.

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The surface wave excitation is avoided by specific design of the radiating element and the one of the dielectric substrate geometry.

II. Surface Waves

In this paragraph a short expression of the surface wave propagated in a grounded dielectric slab with perfect conductivity plane is given. The dispersion equation and Pointing's equations are given. Let the dielectric slab is placed in plain (x,z) and y is normal to (x,z). The dielectric slab has a height h orientated along y and it has a permittivity ε and permeability μ. The space over the slab is air with $\varepsilon=\mu=1$.

The both TM and TE surface waves are considered. The TM waves are described by equations

$$H_x = e^{jk_1(y-h)} e^{j\beta_z}, \qquad E_y = -\frac{\beta}{k_0} Z_0 H_x,$$

in the free-space (y>h) and by equations

$$H_x = \frac{\cos(k_2 y)}{\cos(k_2 h)} e^{j\beta_z}, E_y = -\frac{\beta}{k_0} Z_0 H_x,$$

$$E_z = j\frac{k_2}{k_0\varepsilon} Z_0 H_x \tan(k_2 y)$$

inside the layer $0 \le y \le h$. In these equations, $Z_0 = \sqrt{\mu_0/\varepsilon_0}$ is the impedance of the free-space. The wave number in free-space is denoted as $k_0 = \omega\sqrt{\varepsilon_0\mu_0} = 2\pi/\lambda_0$ where ω is the angular frequency of harmonic oscillations and λ_0 is the free-space wavelength.

The wave numbers by the relations are connected

$$\kappa_1^2 + \beta^2 = \kappa_0^2 \text{ and } \kappa_2^2 + \beta^2 = \kappa_0^2 \varepsilon\mu$$

Trough the Helmholtz wave equations. It follows that

$$\kappa_2^2 - \kappa_1^2 = \kappa_0^2(\varepsilon\mu - 1), \ \kappa_1\varepsilon = jk_2\tan(k_2 h).$$

This transforms into the dispersion equations by TE and TM waves are respectively

$$D^{TM}(k_0,\beta) \equiv \sqrt{k_0\varepsilon\mu - \beta^2}\tan\left(h\sqrt{k_0^2\varepsilon\mu - \beta^2}\right) + j\varepsilon\sqrt{k_0^2 - \beta^2} = 0$$

$$D^{TE}(k_0,\beta) \equiv \sqrt{k_0\varepsilon\mu-\beta^2}\, ctg\left(h\sqrt{k_0^2\varepsilon\mu-\beta^2}\right)$$
$$+\, j\mu\sqrt{k_0^2-\beta^2} = 0$$

Above the layer, the time-averaged Poynting vector of TE waves contains the components

$$P_y = \frac{k_1^{'}}{2\omega\varepsilon_0} e^{-2\left[k_1^{''}(y-h)+\beta^{''}z\right]},$$

$$P_z = \frac{\beta^{'}}{2\omega\varepsilon_0} e^{-2\left[k_1^{''}(y-h)+\beta^{''}z\right]}$$

for TM waves and

$$P_y = \frac{k_1^{'}}{2\omega\mu_0} e^{-2\left[k_1^{''}(y-h)+\beta^{''}z\right]},$$

$$P_z = \frac{k_1^{'}}{2\omega\mu_0} e^{-2\left[k_1^{''}(y-h)+\beta^{''}z\right]},$$

$$k_1 = k_1^{'} + jk_1^{''} \quad \text{and} \quad k_2 = k_2^{'} + jk_2^{''}$$

for TE waves.

Thus, in the case of TE waves, the Poynting vector is also perpendicular to the phase front.

III. ANALYSIS METHOD

In this paper, some modifications of radiating elements in an antenna arrays are presented. The goal is reducing power propagated as surface wave. In most configurations, the main task is to be repressing TM_0 surface waves. The propagation of this wave can be reducing by change of the parameters at environments between the resonators.

The electromagnetic simulations are made by the method of moments. Dividing the models in groups is for making the analysis easier. This division will help for the better study of the single characteristics, configurations, and general conclusion between the groups.

The material which is used for dielectric substrate in this analysis is *Rogers RT/Duroid 6010* with ε_r=10.2, $tg\delta$=0.0023 and thickness 1.5mm. The feed of the microstrip resonators is with probe feed. This feeding method is chosen because of its minimal influence toward radiation of elements and easy manufacturing.

1. Basic model

This is the classical model (fig. 1) used for constructing microstrip antenna arrays. It is a base for comparison with another type models.

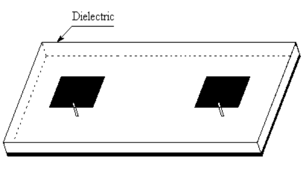

Fig.1. Basic model

In short, the main characteristics of the method with relation to mutual coupling are:
1. the increasing of the permittivity leads to reduce sizes of the resonators, but enhances the mutual coupling and stronger excitation of surface wave;
2. the increasing of the height of the dielectric substrate leads to expand of the frequencies band, but there is stronger surface wave again;
3. the increasing of the distance between the resonators reduces the mutual coupling, but after an optimal value the gain begins to decrease and the diagram pattern starts to disturbance (the power of side lobe is increase).

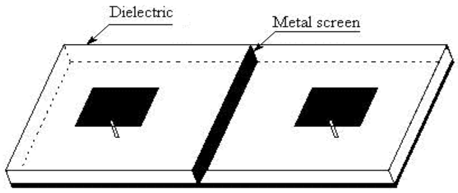

Fig. 2 Reducing mutual coupling between two resonators wit metal wall

The pointed disadvantages limit the area of their applications in the different antenna configurations. By this reason, there was made a simulative investigation on other configurations in the paper. These configurations can be used in different antenna arrays.

2. *Reducing of mutual coupling by screening of resonators each other*

The first shown method of the approach is trough screening of resonators each other with metal wall or metal via-hole. The first way guarantee a better result, but is more difficult for manufacturing. In the second way, the requirement is to choose optimal distance between the via-holes. In the current paper it is accepted for such kind of optimum to be used a distance between the via-holes 2-2,5mm [2]. From a simply technological point of view it is also defined a minimal diameter of the via-holes 6mm.

3. Cell model

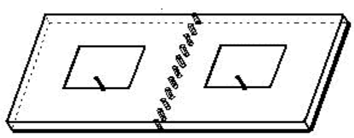

Fig. 3 Reducing mutual coupling between two resonators wit via hole

The other method of decreasing of the mutual influence, considered in the paper, is consisted of separating each resonator from the others by placing in an own cell. In this case, the optimizing task can become very complicated. Because of the thickness of the dielectric substrate the relative dielectric permittivity, and the distance between the resonators is also necessary to be defined, the sizes of the cavity backed, its shape and dielectric permittivity of the material in and out of it.

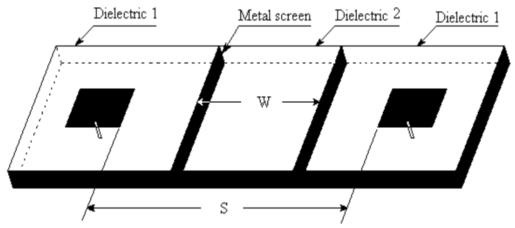

Fig. 4 Reducing mutual coupling between two resonators by dielectric barrier with different permittivity or metal

In fig. 4, it is shown that the screening can be done in immediate proximity to the resonator or in distance, which is defined from the exact requirements. In other words, it is necessary to be optimized the desired effect (wider frequency band, bigger reinforcement, less mutual influence, etc.).

The reducing of mutual coupling between resonators with separating each cell is shown. The variants, are computed, as it is shown below:

1. Dielectric barrier between cell one and cell two is put. In this case the high of dielectric substrate is *h=const*, permittivity ε_r changes from ε_r = 1…20, width *W* changes from W=0…S. The cells are not screening;
2. The single cells are not screening, but dielectric barrier between them is covered with metal. The parameters ε_r and W are the same like these in case one;
3. The cells are screening by metal walls. The parameters ε_r and W are the same like in case one;
4. The area between cells is fully metal. The width of the metal is changing from W=0…S.

In this modification, the distance between two adjacent resonators remains unaltered – S=0,5λ_0. This value of S is preferred because it is used frequently in antenna applications and in that way a maximum gain reaches with it.

IV. Result of the Calculations

Electromagnetic simulations are made with the help of high-frequency electromagnetic simulator working by the moment method. All of the researches are for the same resonance frequency (10,9GHz). With the goal of decreasing the error in the analysis of the results in all of the researches, it is observed the rule – the coefficient of the mutual coupling S12 to be accounted at S11=const. Because of this reason, there is made preliminary

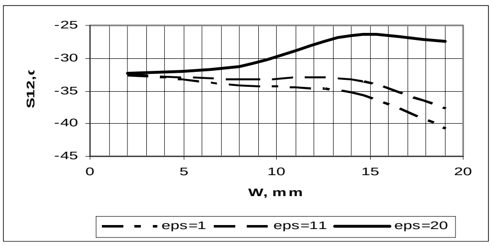

Fig. 6 Mutual coupling without metal strips

impedance matching of all researched structures and the values of S11 are equal or close to the shown in fig.5

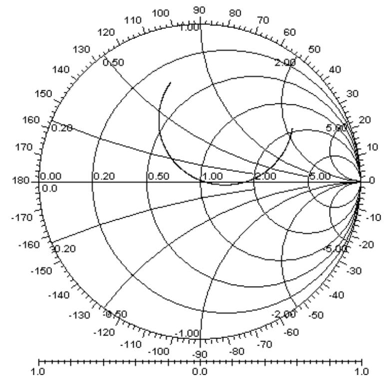

Fig. 5 S11 in Smith chart

It is shown in fig. 6 that when ε_r =1 and ε_r= 11 then W=13-14 S12 (dB) changes negligible but after this value there is an exponential decrease. It can be seen that when ε_r =20 there is a noticeable mutual coupling. S12 has a maximum when W=15 mm.

It can be seen that when each particular cell is put in a private metal screen the parameters S12 do not change essentially. Not only the character of change stays the same but also the values of S12 are reserved.

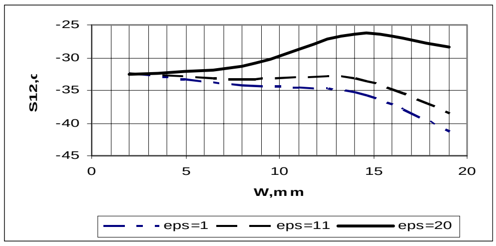

Fig. 7 Mutual coupling with metal strips between dielectric one and dielectric two

Using a metal strip (fig. 7) over the dielectric barrier the graphic has a different form. When W=8 there can be seen a resonance. At this value of W there is a lowest mutual coupling between the resonators. Reducing the mutual coupling between resonators is better than each other cases and it is over 10 dB. These characteristics of the metal strip can be used in designing of microstrip antenna arrays with a low mutual coupling.

Over a defined value of the width of W many parasite

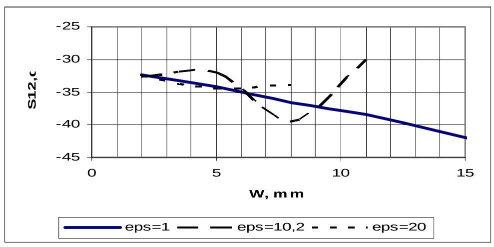

Fig. 8 Using metal strip over dielectric between the cells

resonances can be seen which is typical for this structure. This is the reason for the graphics, when ε_r=10,2 и ε_r =20, not to be finished.

When a metal is put instead of dielectric (fig. 9) it can be seen that the mutual coupling between the resonators reduces with a linear law. When this metal is put, the mutual coupling reduces faster with the increasing of the width of W, as a comparison with the described models. If fig. 8 and fig.9 are compared it can be seen that this type of screening does not have a resonance character and when W=8mm the value of S12 is nearly the same as in the previous case (horizontal metal strip).

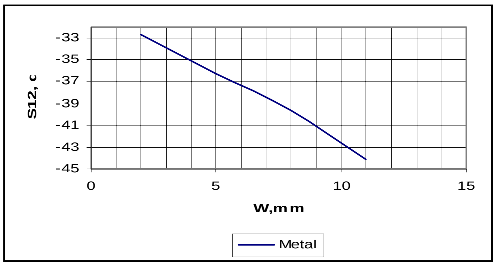

Fig.9. Mutual coupling (S12) between the patches with metal barrier between the cells

On fig. 10 is shown the variation of frequency bandwidth like a function of change in the area W with and without vertical metal strip. When the width W= 7mm all graphics

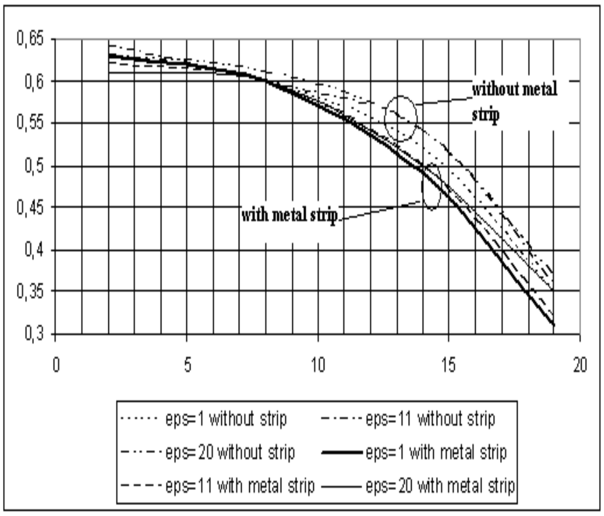

Fig.10 Bandwidth with and without metal strips between dielectric one and dielectric two

are crossed. Over this value, the structures, which has not metal strip, are more bandwidth.

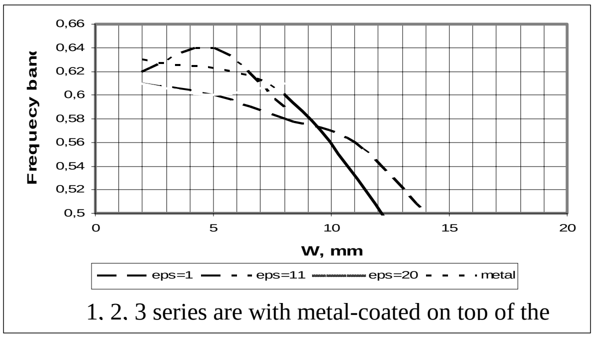

Fig. 11 Bandwidth with metal strip over the dielectric barrier

On fig. 11 are depicted dependencies of bandwidth with metal over the dielectric, which is between two cells and the case when the cells are divided fully with metal. When metal strip is used over the dielectric with width W>7mm there are undesirable parasitical resonances but they are not shown. It is obvious that when the metal strip is used with width W=3-6mm and ε_r =11 the largest bandwidth appears.

V. Conclusion

In the paper is made simulative investigation of mutual coupling between patches with constant distance between them. The changes of the characteristics in the area between resonators exert influence over the mutual coupling. These specific changes can be used of reduce the mutual coupling, surface wave or to increase the bandwidth in determinate area.

In the paper are shown some advantages and disadvantages of the described structures. The resonators can be used in design of antenna arrays with reducing mutual coupling.

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