DESIGN MODELS FOR RECTANGULAR MICROSTRIP RESONATOR ANTENNAS

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Introduction.

Up to now two methods for the analysis of microstrip resonator antennas were available.

The first one, based on the theory of Wolff and Knoppik (x), allowed us to calculate the resonant frequency of any arbitrary mode. Here, the assumption was made that the open structure, composed of the two conducting surfaces, separated by a layer of dielectric material (fig.1) could be reduced to a structure (with greater physical dimensions : dotted in fig.1), closed by ideal magnetic and electric walls. In this way, just the reactive part of the radiation impedances (equivalent with the edge capacitances of the structure) was taken into account.

The second, made by Munson (xx), gives us a practical method for calculating resonant frequency and impedance, only in the fundamental mode of the resonator (1,0 mode). In this model, the resonator was approximated by a transmission line, with the impedance of a microstripline with the same width (W), loaded at both sides with the radiation impedance of a limited part (of length W_y) of an infinitely supposed slot (fig.2 a,b).

In this paper we compare both models from a design point of view and present a new combined model to obtain more accurate designs.

Design model based on the theory of Wolff and Knoppik.

Considering the edge capacities of the resonator to be equivalent with a corresponding increase of the physical dimensions of the resonator (a "short" open transmission line), the resonant frequencies can be found with the simple closed resonator formula:

$$f_{N_x, N_y} = \frac{c}{2\sqrt{\epsilon_{dyn}}} \cdot \sqrt{\left(\frac{N_x}{W_{x_{eff}}}\right)^2 + \left(\frac{N_y}{W_{y_{eff}}}\right)^2}$$

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where the dynamic permittivity ϵ_{dyn} is defined so that the inhomogeneity of the medium ($\epsilon_r = 1$ above and ϵ_r under the upper surface) is taken into account.

Three different design problems occurring frequently are:

- 1) If one dimension, together with the mode and the frequency of one resonance is known, find the second dimension.
- 2) Find the physical dimensions of a resonator when two frequencies of two different modes are given.
- 3) Two relations between 4 frequencies, corresponding with known modes are given :

$$a.(f_1 - f_c) = b.(f_c - f_2)$$

$$c.(f_3 - f_m) = d.(f_m - f_4)$$

In the first two cases a simple substitution formula can be used in an iteration process ; the last one can be solved with a Newton-Raphson formula. In that case no easy criterion for verifying the possibility of the design exists, except the divergence of the process. However, when used for spacing evenly two modes around a third one ($f_c = f_m$, $d=0$, $a=b=c=1$), starting with values found by the Munson Model a rapid convergence is achieved.

Design model by Munson.

Here the width will be found from the equation:
 $2.\text{Re}(Y_{\text{rad}}(f, W)) = \text{Re}(Y_{in})$. The length will be found from the Smith Chart : $L = (1 - \frac{\text{Im}(Y_{in})}{\pi}) \cdot \lambda_{eff} / 2$ (fig.2c), where λ_{eff} is found with the formulas of Schneider (&), just as in the Wolff-Kroppik model. However, in practice, it turned out that, for radiation impedances greater than 120 Ω the linearization of the radiation admittance as function of W was no longer allowed, but a better approximation had to be used.

We conclude that this easy and rapid design must be used carefully, and that no information is available about possible appearances of higher order modes in the neighbourhood of the fundamental, disturbing both radiation impedance and antenna patterns.

Combined design model.

Considering the disadvantages of each, we propose a combined design model. Here the substrate (PTFE) is, like in the previous models supposed to be lossless. We suppose the resonator loaded with an impedance $Z(S)$, so that :

$$\vec{E}_t = \vec{H}_t \times \vec{n}_{12} \cdot Z(S) \quad (\text{fig.3}) ;$$

for the simplicity $Z(S)$ has been assumed constant over each of the 4 bounds of the rectangular resonator, and zero on the conducting surfaces.

After solving Maxwell's equations with the specified boundary conditions, we obtain the following equations, assuming the resonator symmetrically loaded:

$$\begin{aligned}\tan(k_x W_x + (N_x - \delta_x) \cdot \pi/2) &= -k_x Z_x / (j\omega\mu) \\ \tan(k_y W_y + (N_y - \delta_y) \cdot \pi/2) &= -k_y Z_y / (j\omega\mu) \\ k_x^2 + k_y^2 &= \omega^2 \cdot \mu\epsilon\end{aligned}$$

where Z has to be chosen equal to the radiation impedance of a rectangular slot adjacent to the upper surface of the resonator, because the field produced by those antennas can be explained by the radiation of a narrow slot surrounding the upper surface of the structure (&&).

If we simplify those equations, neglecting the real part of the radiation impedance, it gives already sufficient agreement with experiments (table 1). As we can see from a graphic solution of a very simple case (fig.4), the higher order modes do approach closer to each other, and are, as measured, in general lower than those found by Wolff and Knoppik. Compared with that model, here only effective mode numbers, corresponding with elongations of the structure, are defined (fig.5).

Conclusion.

A more accurate but very simple and easy to use model for the determination of the resonant frequencies has been derived, what will allow us to make more accurate designs, and minimize the influence of the equivalent network inaccuracy, when used in arrays.

References.

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Antenna	Mode	Wolff-Kn.	Comb.	Measured
mm	-	GHz	GHz	GHz
$W_x \times W_y$	$N_x N_y$	f	f	f
27,5 x 102,6	10	3,507	3,509	3,548
"	12	4,126	3,954	3,967
10,4 x 39,7	10	9,214	9,000	9,048
"	12	10,492	9,999	10,253

Table 1 : Resonance frequencies

Fig. 1: Geometry of the resonator antenna

Fig. 3: The loaded resonator

Fig. 2 c: Reflection coefficient

Fig. 5: Equivalent cavity elongations

Fig. 2 a: Fields in the (1,0) mode
b: Equivalent model

Fig. 4: Determination of effective mode numbers (capacitive load, one dimension)