MODELISATION OF MICROSTRIP-MICROSLOT APPLICATOR BY EXTENSION OF TRANSMISSION LINE MODEL

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ABSTRACT

In this paper we present a model which allows to characterize a microstrip-slot radiator used in biomedical applications. The approach used is based on the transmission line model of radiating antennas. This usual model is improved by using S.D.A. in order to calculate the microwave parameters of the microstrip line with tuning septums which constitutes the transmission line. The validity of our approach is confirmed by comparison with experimental results.

I - INTRODUCTION

Different types of microstrip radiators or microstrip-slot radiators to induce local hyperthermia operating in the frequency range 0.1 GHz to 4 GHz have been investigated by several authors [1], [2], [3], who have shown their advantages compared with conventional aperture radiators like waveguide. From studies achieved on classical microstrip and microslot structures, we have designed in 1983 a new type of applicator : it is a microstrip-microslot antenna which consists of a gradual transition from a classical microstrip line to a microstrip line with tuning septums ; the applicator is formed by the circular ground plane aperture in contact with the tissue or other lossy medium (figure 1) ; a metallic cylinder put on the microstrip line in the aperture offers the possibility of central cooling by water flowing in this cylinder and then avoid superficial burning of the tissues when this probe applicator is used for microwave hyperthermia on patients [4]. Due to the gradual transition at constant impedance, from the classical microstrip line to the microstrip line with its ground plane opened, the microwave performances of this applicator are very interesting in a large frequency range and so this device can be used both in hyperthermia and radiometry [5].

II - MODELISATION

1) Formulation

Because of the various discontinuities on the microstrip lines and the coupling of the microslot with mediums of complex permittivity, the direct modelisation of this device is very complicated. So we have considered our applicator as a succession of microstrip lines with tuning septums of different geometry (figure 2). So in a first time we have studied the elementary structure shown on the figure 3 :

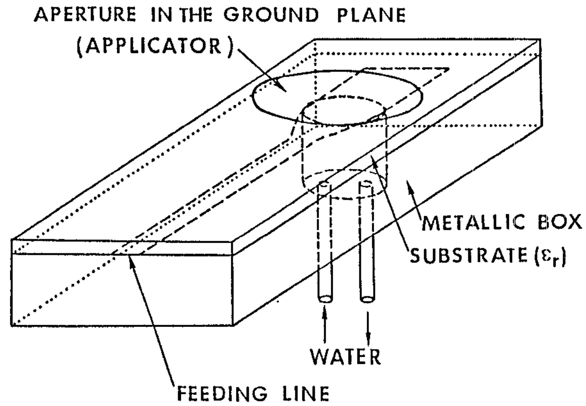

Fig. 1 : Scheme of the microstrip-slot applicator with central cooling system

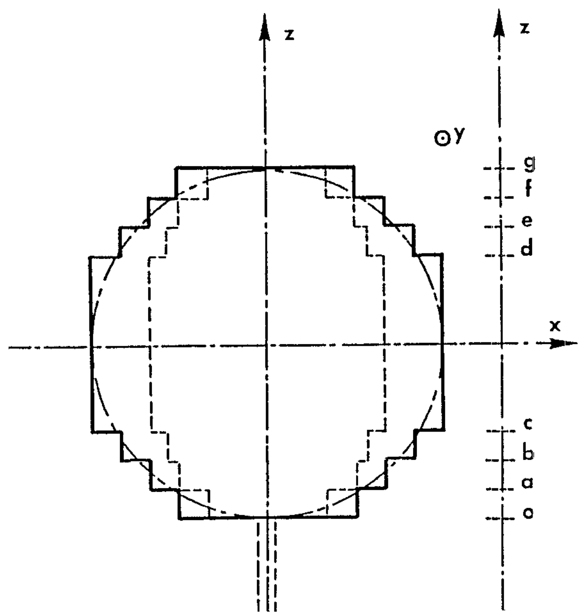

Fig. 2 : Decomposition of our applicator in a succession of microstrip lines with tuning septums of different geometry.

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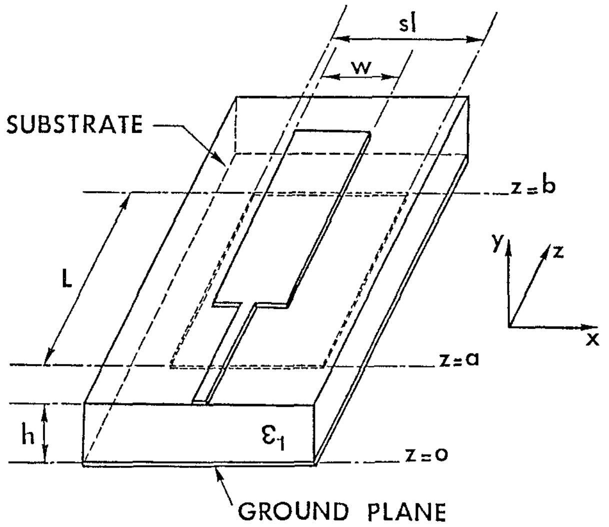

Fig. 3 : Scheme of the studied structure

This slot radiator may also be considered as two discontinuities (in Z = a and in Z = b) connected by a length "L" of microstrip line with tuning septums. As this applicator is put on muscle, it is necessary to determine all the characteristics of the structure shown in figure 4 where human tissues are simulated by a lossy material (ε^*_2).

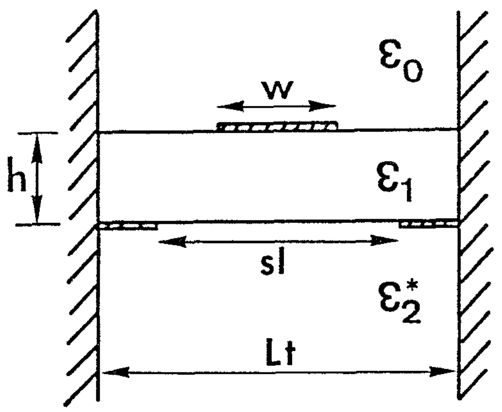

Fig. 4 : Cross section of the structure with a lossy material under the slot ($\varepsilon^*_2 = \varepsilon'_2 - j\,\varepsilon''_2$)

Effective relative permittivity and attenuation of such a transmission line are determined with and without muscle under the slot. The fundamental modes are considered : the even one is called the "strip mode" and the odd one is called the "slot mode". To do so, we use the Spectral Domain Approach.

In the classical S.D.A., field components of hybrid guided waves are written in terms of $\tilde{E}_z(\alpha, y)$ and $\tilde{H}_z(\alpha, y)$ which are the Fourier transforms with respect to x of the axial field components $E_z(x,y)$ and $H_z(x,y)$. Tranverse electromagnetic fields are obtained by conventional formulation. After some mathematical manipulations, the matching conditions at each interface can be written in a hybrid matrix notation as for the lossless case :

$$\begin{vmatrix} E_x \text{ (strip)} \\ E_z \text{ (strip)} \\ J_x \text{ (slot)} \\ J_z \text{ (slot)} \end{vmatrix} = \begin{vmatrix} H \end{vmatrix} \cdot \begin{vmatrix} J_x \text{ (strip)} \\ J_z \text{ (strip)} \\ E_x \text{ (slot)} \\ E_z \text{ (slot)} \end{vmatrix}$$

The solution of this set of equations is then achieved by means of the Galerkin Method. The electric field components in the slot and the current densities on the strip are expanded in terms of suitable series of basis functions. We use Chebyschev's polynomials together with trigonometric functions. These expansions take into account most of the physical aspects like the edge effects on the strip and in the slot. To reduce computer time in searching for the β^* phase constant solution in the complex plane, we use the ZEPLS procedure set up by LAMPARIELLO and SORRENTINO based on the residues theory. Therefore, as we excited the microstrip-microslot radiator by a microstrip line, we must consider only the strip mode.

2) Numerical and experimental results

The variation of ε_{eff} as a function of w the strip width, when the slot width sl is constant, exhibits a maximum which depends on the ratio w/sl for a given substrate (ε_r, h). In this case, the evolution-law w = f (sl) is linear.

The figure 5 represents the evolution of ε_{eff} as a function of w and f (frequency), for a given width sl and a defined substrate (ε_r, h). From these curves, we deduce the network which gives ε_{eff} maximum as a function of frequency with (sl, w) as parameters (figure 6).

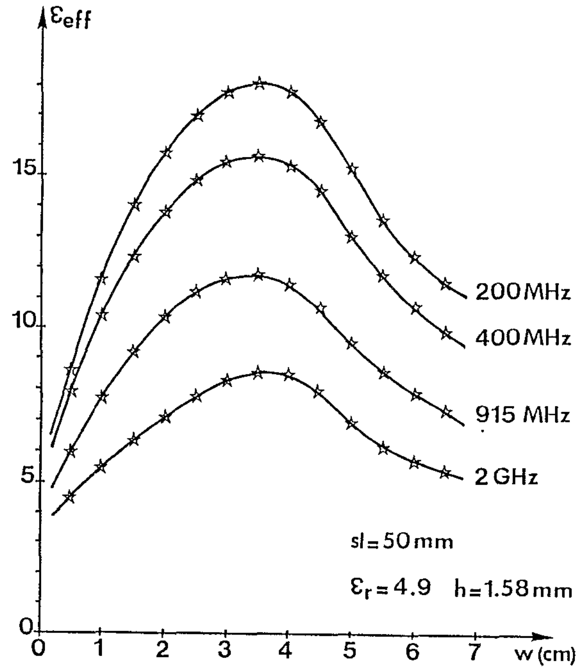

Fig. 5 : Effective relative permittivity frequency behaviours versus size w ; with sl = 50 mm, ε_r=49 and h = 1.58 mm.

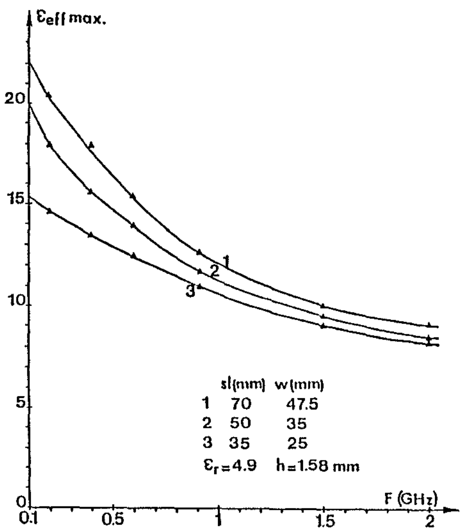

Fig. 6 : Maximum effective relative permittivity frequency behaviours versus couples of parameters (sl, w) ; with $\varepsilon_r = 4.9$ and h = 1.58 mm

The theoretical resonant length of the structure L_{the} corresponds to the first half guided wavelength resonance of the microstrip line with tuning septums : this assumption is based on the hypothesis that the terminal impedances in the planes z = a and z = b present the same reflection coefficient value very close to one.

It is calculated using the relation :

$$L_{the} = \frac{C}{2 \cdot f \cdot \sqrt{\varepsilon_{eff}}} \quad \text{with } C = 3 \times 10^8 \text{ m/s}$$

f is a choosen frequency

ε_{ff} is the effective relative permittivity calculated by S.D.A. for the choosen frequency.

In order to verify theoretical analysis we have made several experimental determinations for various dimensions of the structure of the figure 3.

In this aim, we have measured the resonant frequency of these structures in contact with muscle. We obtain (note that we consider the first resonant frequency in $\lambda g/2$) :

S_L (mm)	35	50	70
W (mm)	21	30	42
L (mm)	70	100	140
f_{exp} (GHz)	0,62	0,4	0,26
ε'_2	49,7	51,8	54,6
σ_2 (℧/m)	0,925	0,85	0,8
εeff theoretical	12,352	15,448	19,164
L_{the}	68,83	95,41	131,78
$\frac{L_{exp} - L_{the}}{L_{exp}}$	1,67%	4,59%	5,87%

with εr = 4,9 , h = 1,58 mm

We can note a good agreement between experimental and theoretical values. This comparison shows the interest of our S.D.A. modelisation to determine the characteristics of a microstrip line with tuning septums in contact with the tissues [6], [7].

III. EXTENSION OF TRANSMISSION LINE MODEL

In order to determine the reflection coefficient parameter of the elementary structure in the plane z = 0 (fig.3), we have extended the "transmission line model" used for the rectangular patch antenna, to the microstrip-slot radiator in contact with the lossy medium(ε^*_2).

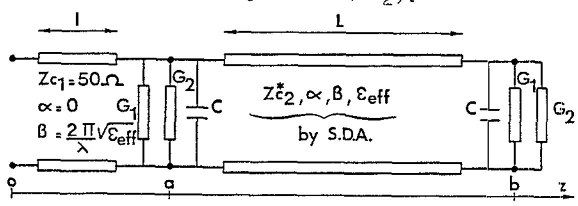

Fig. 7 : Scheme of the transmission line model extended to the microstrip-slot applicator in contact with the lossy medium.

So, this structure can be simulated by the proposed scheme in figure 7 where :

* C is the end line capacitance

* G_1 is the radiation conductance in the free space due to the strip

* G_2 is the conductance due to the field penetration in the space under the slot

* Zc_1 is the characteristic impedance of the quasi TEM mode of the excitation line deduced from the SCHNEIDER and HAMMERSTAD formulations.

* Zc^*_2 is the characteristic impedance of the quasi TEM mode of the microstrip line with tuning septums laid on a lossy material calculated by the S.D.A. modelisation

* l is the length of the excitation line

* L is the length of the microstrip line with tuning septums.

For the determination of the end capacitance, as in the transmission line model, we can say in a similary way

$C = 0{,}01668 \cdot \Delta l/h \cdot w/\lambda g \cdot \varepsilon_{eff}$

with Δl is the extension of the line given by HAMMERSTAD.

h is the height of the substrate

w is the width of the strip

λg the guided wave length = $\lambda o/\sqrt{\varepsilon_{eff}}$

where ε_{eff} is the relative effective permittivity of the microstrip line with tuning septums put on the lossy medium determined by S.D.A.

The two conductances G_1, G_2 simulate the radiation phenomena. G_1 concerns more precisely the radiating in the upper space, this radiation conductance is determined in the same assumptions and conditions as for the patch antenna.

The second conductance G_2 is relative to the penetration of the fields in the lossy medium. In fact, we can consider that the penetration of fields is included in the rigourous analysis of propagation conditions of the microstrip line with tuning septums laid on muscle, so we have neglected this conductance G_2 in a first step. We will confirm the validity of this assumption by comparison with experimentations.

IV. NUMERICAL AND EXPERIMENTAL RESULTS

We present in figure 8 and 9 two cases in which we can compare the experimental results and the theoretical one for the reflexion coefficient of the considered structures (figure 3) in the plane z=0. We can note a good agreement between the experimental and numerical curves.

Furthermore, we have observed that in a numerical point of view, the effective value of the global conductance is not very critical for the determination of the resonant frequency which depends of the capacitance value. This confirms our hypothesis that the value of conductance G_2 can be neglected and the end line reflexion coefficient, essentially due to the value capacitance, is in that case very close to unity.

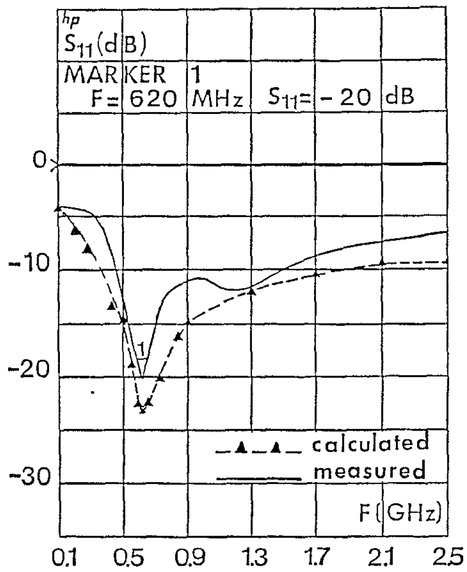

Fig. 8 : Numerical and experimental reflection coefficient behaviour for sl = 35 mm, w = 21 mm, L = 70 mm, ε_r = 4.9 and h = 1.58 mm.

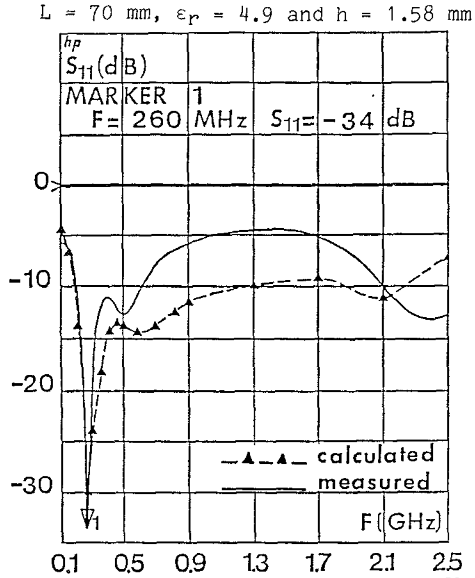

Fig. 9 :Numerical and experimental reflection coefficient behaviour for sl = 70 mm, w = 42 mm, L = 140 mm, ε_r = 4.9 and h = 1.58 mm.

V. CONCLUSION

The microstrip-microslot applicator with a circular aperture in the ground plane we have developped for hyperthermia treatment of tumours and radiometry, presents interesting microwave performances due to the gradual transition between the feeding microstrip line and microstrip line in the aperture.

Empirical considerations allowed us to develop efficient applicators of this type. But in order to optimise our structure, we have developed a new approach to study the frequency evolution for the propagation characteristics of the microstrip line with tuning septums laid on muscle, using the spectral domain approach analysis. An extension of the transmission line model is proposed to characterise the microstrip microslot applicator. Our transmission line model describing the frequency behaviour is validated by theory-experiment comparisons.

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