

TECHNICAL UNIVERSITY OF WROCLAW
INSTITUTE OF TELECOMMUNICATIONS AND ACOUSTICS

Report No. I-28/PRE-21/93

NUMERICAL MODELLING OF TEM CELLS

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Doctoral thesis

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Keywords:

electromagnetic compatibility,
electromagnetic field,
numerical methods,
computer modelling,
TEM cell

WROCLAW 1993

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Notations

a	- half of the cell width,
a_{mn}, b_{mn}, c_{mn}	- auxiliary coefficients,
b	- half of the cell height,
b_1	- distance between the cell floor and the septum,
b_2	- distance between the cell ceiling and the septum,
c	- light velocity,
C	- capacity,
C_0	- cell unit capacity,
$\mathbf{C}$	- TLM connection matrix,
D	- septum displacement along the x axis, with respect to the symmetry axis,
$\bar{E}$	- electric field strength,
E_x, E_y, E_z	- components of the electric field strength,
f	- frequency,
f_{res}	- resonant frequency,
$(f_c)_{mn}$	- cutoff frequency for mn mode,
g	- gap between the septum and the outer wall,
G_0, Y_0	- characteristic admittances of TLM stub lines,
$G(\mathbf{r}, \mathbf{r}')$	- Green function,
$G(x, y, x_0, h)$	- Green function in the solution of Poisson equation,
h	- distance between the septum and the floor of the cell,
h_0	- effective height of an object,
$\bar{H}$	- magnetic field strength,
H_x, H_y, H_z	- components of the magnetic field strength,
i, j, k, m, n	- indexes,
$\mathbf{i}$	- unit vector pointing the direction of propagation,
K	- electric potential of the septum,
k	- propagation constant; number of conductors,
L	- inductance,
L, L_1, L_2, L_k, L_i	- directing lines,
l	- length of a cell,

l_{res}	- resonant length of a cell,
$p(x)$	- approximate charge distribution,
$p(i)$	- pulse response,
$p'(i)$	- recalculated pulse response,
$P_n^i(\mathbf{r})$	- base functions,
$q(\mathbf{r}), q_i(\mathbf{r})$	- charge distribution functions on metal surfaces,
Q_n^i	- coefficients of charge distribution functions,
Q_n^j	- vector of Q_n^j coefficients,
$r, \theta, \theta_1, \theta_2,$	- variables in a polar system of co-ordinates,
R	- number of equations,
$\mathbf{r}$	- leading vector of the observation point,
$\mathbf{r}'$	- leading vector of the source point,
$\mathbf{r}_m$	- vector of the middle of the segment ΔL_m ,
s	- set of points at which $p(x)$ is equal to exact charge distribution; summation index,
S	- TLM scattering matrix; matrix of potential coefficients,
S_{mn}^{ij}	- submatrix of the potential coefficients matrix,
S_{mn}	- element of the potential coefficients matrix,
t	- time,
U_m^i	- vector of exciting coefficients,
v	- phase velocity,
v_n	- pulse velocity in a TLM network,
$V_0, V_1, V_2, V_i, V(\mathbf{r})$	- electric potentials,
V^s	- scattered impulse in a TLM network,
V_k^i	- voltage vector at node i of a TLM network, at a time step k ,
w	- half of the septum width,
x, y, z	- variables in a Cartesian system of co-ordinates,
Z_0	- cell characteristic impedance,
Z_f	- wave impedance,
$\mathbf{1}_n$	- unit vector,
β_n	- propagation constant of a TLM network,

Δl	- unit length in a TLM network,
ΔL_n^i	- segment of the surface L_i ,
Δt	- unit time in a TLM network,
ε_0	- free space dielectric permittivity,
ε_r	- relative dielectric permittivity,
ϕ, ϕ_1, ϕ_2	- electric potentials,
η_0	- free space wave constant,
μ_0	- free space magnetic permeability,
μ_r	- relative magnetic permeability,
ρ	- reflection coefficient,
σ	- conductivity,
Ψ^e	- scalar potential for TM modes,
Ψ^h	- scalar potential for TE modes,
ω	- pulsation.

1. Introduction

In the contemporary world, where the number of electric and electronic devices is growing very rapidly, problem of electromagnetic compatibility is playing more and more important role. All devices, that produce electromagnetic fields may therefore be treated as possible sources of unwanted emission. Even devices that the electromagnetic emission is the purpose of acting, such as radio and TV transmitters or wireless telephone sets can be treated as sources of interferences from the point of view of the user of some other electronic equipment, that is sensitive on EM fields. The problem is well known to everyone who tried to listen to the FM radio while working with IBM PC compatible machine made in Taiwan.

Therefore, there is a need to investigate emission and susceptibility characteristics of all electronic devices, i.e. to check the level of electromagnetic field produced by a device as well as the level of EM interference to which the device operates properly. Because of the great importance of the problem, there are many local and international regulations (PN, CISPR, VDU, MIL), that specify both levels, according to the class of a device. Of course, the most sharp are regulations concerning military and aircraft equipment because of the importance of their infallible work.

There are a few methods of testing the emission and susceptibility levels of electronic devices: free space measurements, measurements using anechoic chambers or so called TEM (Transverse ElectroMagnetic) cells. All standards are applied to free space measurements. These measurements demand, however, rather expensive test stands and are sensitive to outer interferences, because there is no isolation from environment signals influence. Anechoic chambers solve the problem as far as the unwanted signals are concerned, but they do not operate properly at low frequencies (a few tens of megahertz) because of absorbers characteristics. Apart from it they are also extremely expensive and - last but not least - rather large.

Because of the disadvantages of free space and anechoic chambers the best solution seems to be applying TEM cells, which in low and medium frequency range are impossible to replace. They have one disadvantage: levels measured in TEM cells are not simply comparable with results obtained with other methods or even with other types of TEM cells. The extensive research works are carried out in leading scientific centres of the world to enable such a comparison [33,40,61].

The concept of TEM cell was first introduced by M.L. Crawford [6], and therefore the first (and still used) type of the TEM cell is sometimes called the Crawford cell. Next, many different solutions applying Crawford idea were proposed (for example the TEM cell concept developed in Asea Brown-Boveri EMC [19,34,61]).

A TEM cell acts as a wide band transformer of a linear phase and amplitude, transforming field strength into high frequency voltage, or high frequency voltage into field strength. Using TEM cell one can obtain a standard, well-controlled electromagnetic fields for vulnerability tests, and get an information about fields generated by equipment under tests (EUT). The cell is also a shielded room that provides electric insulation of the EUT with respect to the environment. Applying a TEM cell there is no need to use measuring probe antennas, which are usually the sources of many limitations like bandwidth, polarisation, directivity, etc.

The investigations concerning TEM cells are under interest in leading countries in the world (USA, Japan, Great Britain, Switzerland). The investigations on the TEM cells can be divided into following parts:

- TEM cell designing and improving [2,6,7,8,19, 34,41,59,68],
- theoretical analysis of TEM cells and determination of their electrical parameters [5,20,22,25,26,27,30,35,36,37,38,39,54,55,63,64,66,68],
- ways of applying TEM cells in EMC measurements (special attention is put on the comparing results with other techniques) [2,7,9,10,14,40,50,52,62],
- ways of using TEM cells for other purposes (such as antenna calibrations, bioeffects dosimetry studies) [31,32,42,67].

Above parts are of course related to each other, because for example analysis techniques are applied in cell design as well, as in developing measurement procedures, error analysis, etc.

In Poland the only place where such research is carried out is the Institute of Telecommunications and Acoustics (ITA) of the Technical University of Wrocław [16].

A Crawford cell has been constructed under government-sponsored Telecommunication Project concerning establishing a specialised laboratory in the buildings of the Technical University of Wrocław for investigating the electromagnetic interference (EMI) caused by the Information Technology Equipment (ITE) [16,17,18]. The cell has been designed on the bases of special curves published by National Bureau of Standards, USA [7].

Present contribution is the continuation of that work concerning application of TEM cells in electromagnetic compatibility measurements. A special attention is put on theoretical prediction of physical phenomena inside the cell.

The aim of this work is to develop methods and computer algorithms that enable both theoretical analysis of electromagnetic environment inside TEM cells, and determining cell electrical parameters as a function of its type, geometry, and the presence of EUT. The following thesis of the work has been formulated:

It is possible to use computer simulation to investigate electromagnetic field distribution and determine cell parameters both in the case of an empty cell, and in the case of a cell with an EUT inside it.

The author proves the above thesis by developing the set of algorithms and computer programs that enable such an analysis. In Chapter 2 a more detailed description of TEM cells is presented, together with a short review of the methods used till now for TEM cells analysis. In Chapter 3 a model of TEM cell is presented which enables determining the field distribution and cell parameters, under assumption that only TEM mode waves are propagated inside the cell. Both two- and three-dimensional models have been developed. The analysis of EUT influence on cell characteristic impedance is presented. At the end of the chapter an example of applying described techniques to design new GTEM cell is presented. In Chapter 4 the method for further analysis (TLM method) is presented in detail together with a short

comparison to other techniques. Chapter 5 discusses higher order modes in TEM cells, and finally the full analysis of electromagnetic field inside TEM cells (both Crawford and GTEM type) is presented in the Chapter 6. In this chapter electromagnetic fields distributions for different frequencies as well as resonant frequencies of TEM cells are discussed. Chapter 7 is a recapitulation of the contribution. Also the ways of further investigation concerning TEM cells and their applications in EMC measurements are pointed out. Chapter 8 is a list of references. At the end of the work the appendix concerning calculation of surface integrals of the $1/r$ function is placed.

2. Methods of theoretical analysis of TEM cells

2.1. TEM cell parameters

There are a few models of TEM cells that are used in EMC investigations. The most popular is the standard TEM cell (sometimes called the Crawford cell), and a cell with enlarged bandwidth called GTEM (Gigahertz TEM) cell sometimes called also the Brown-Boveri cell.

The Crawford cell is presented in Fig. 2.1. It is basically a piece of uniform, rectangular, 50-ohm transmission line. To enable connections through standard coaxial connectors, the cell has been tapered on both sides. The inner conductor is a flat metal plate attached to external walls by dielectric supports. The access to the cell's interior is possible through a shielded door placed on one side of the walls.

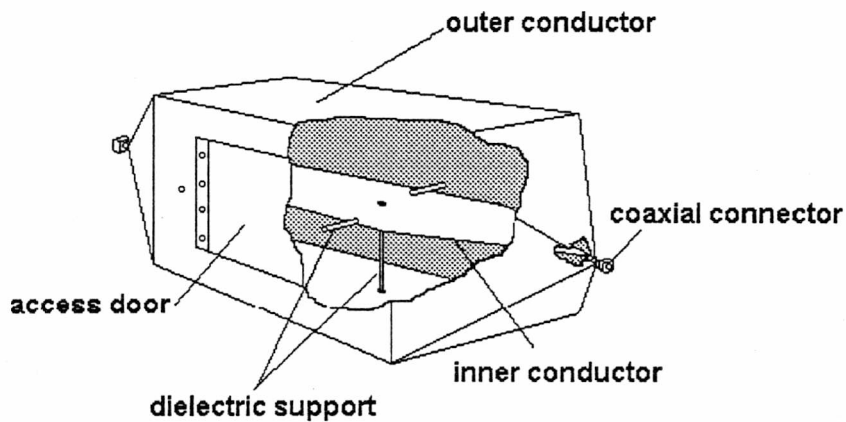


Fig. 2.1. The Crawford cell

In the first Crawford cell constructed in the National Bureau of Standards (NBS), Boulder, Colorado, USA, the inner conductor (septum) has been placed in the middle of the cell height. Such a cell is called a symmetrical TEM cell. Next investigations have proved, that applying asymmetric cells allows to investigate objects of larger size, with only little worse electric parameters of the cell.

The cell is equipped with mains filters to feed the equipment under test (EUT) placed inside it. There is also a baffle that facilitates connection of signal cables. The baffle is screened to eliminate exposure to environmental radiation or energy leakage from the cell interior.

The GTEM cell (Fig. 2.2) is a different solution. In order to eliminate wave resonances, the cell itself is a long tapered section of a transmission line. Instead of normal 50-ohm load at the end of the structure, the distributed load has been applied. For low frequency range it consists of special resistor system (with resultant resistance equal to 50 ohms), for high (radio) frequencies the radio absorbing materials has been used. In order to deal with medium frequencies (just below the range of absorbers acting), the system of resistors is made in a form of resistors matrix, with proper

impedance distribution. Thanks to above solutions the reflections from the end of the cell are minimized and the bandwidth of the cell is enlarged.

Electrically, the TEM cell is a TEM transmission line. A characteristic feature of the TEM wave is that the electric and magnetic components are perpendicular to each other, and to the direction of propagation. The E and H fields are almost homogenous nearly within the entire working space between the inner and outer conductor. They simulate the field distribution in free space. Because the TEM mode has no lower frequency limit, the TEM cell can be used at frequencies as low as required for the case under study. The only limitation is the efficiency of magnetic shielding that depends on material used for building cell walls.

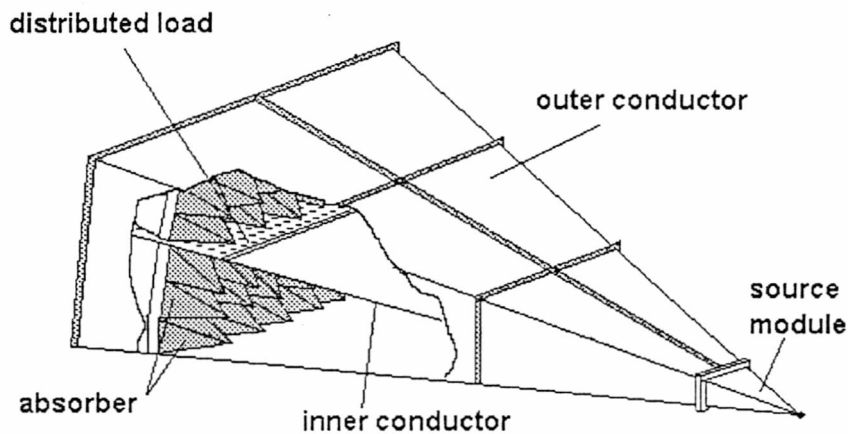


Fig. 2.2. The GTEM cell

The main parameters that characterise TEM cell are:

- characteristic impedance,
- maximum usable frequency,
- the working volume.

As it was said before, usually cell characteristic impedance is assumed to be 50 ohms in order to enable connecting standard test equipment (cables, loads, generators, spectrum analysers, time-domain reflectometers, etc.).

Maximum usable frequency is a frequency at which resonances and higher order modes occur. Of course, the higher maximum frequency is, the better.

The working volume is limited by the electromagnetic field irregularities. Usually it is assumed, that field levels inside the working volume should not differ from point to point more than 1 dB, but other tolerance levels are also in use (for example 3 dB).

In fact, the two latter parameters are dependent: increasing the cell dimensions in order to increase the working volume leads to decreasing the cell frequency bandwidth, and vice versa: increasing the frequency bandwidth can be obtained by decreasing cell size (and of course the working volume). The compromise between the two factors can be achieved in asymmetric cells mentioned above.

Apart from described above, some other parameters are sometimes used to describe TEM cells: the standing wave ratio (SWR) or the insertion loss. Those parameters are especially useful when characterising cells with absorber loading [7].

The electric parameters described above can be treated as supplementary, because they are determined by the electromagnetic field distribution inside the TEM cell. This in turn depends on the frequency used, the cell geometry and the material filling the cell. Also putting the EUT into the cell working volume can significantly change the field distribution. Therefore the electromagnetic field distribution inside the cell can be treated as a basic cell parameter.

Described quantities are usually obtained in the experimental way. The lack of this method is the fact that measurements can be done AFTER the cell is built, while they are needed BEFORE the cell is designed. Therefore theoretical modelling of the TEM cells is an extremely important task (taking into account the fact of great costs of each model of TEM cell). Also the ability to investigate the processes of the waves propagation in the built cell can be very useful during making the measurements, and interpreting obtained results.

2.2. Review of known solutions

Exact analysis of a TEM cell taking into consideration all important factors is extremely difficult and therefore is not available. Existing elaborations propose solutions for some special cases, assuming far going simplifications. Possible simplifications are as follows:

- it is assumed, that only TEM mode is propagated in the cell - under this assumption the problem reduces to the electrostatic one;
- calculations are made for empty cells only, field perturbations caused by the presence of EUT are assumed to be negligible;
- the analysis is reduced to two-dimensional one which is equivalent to evaluating infinite transmission line;
- only simple geometrical configurations are taken into account (for example the symmetric line).

Methods of analysis presented below are set in an order showing successive removing of the simplifications (which is equivalent to making the description closer to reality). The problem geometry is shown in Fig. 2.3.

2.2.1. Symmetric cell

For easy case of symmetric TEM cell analytical solution of the problem of calculating electric field patterns has been given by Tippet and Chang in [57]. Analysis has been done using conformal transformation method. The electromagnetic field pattern has been calculated numerically. Results have been compared with experimental data obtained by Crawford [6]. Full agreement of theoretical and measured results has been achieved. The same method has been applied to calculate capacitance per unit length of cell treated as the infinite transmission line (it is known that capacitance is invariant in relation to the conformal transformation).

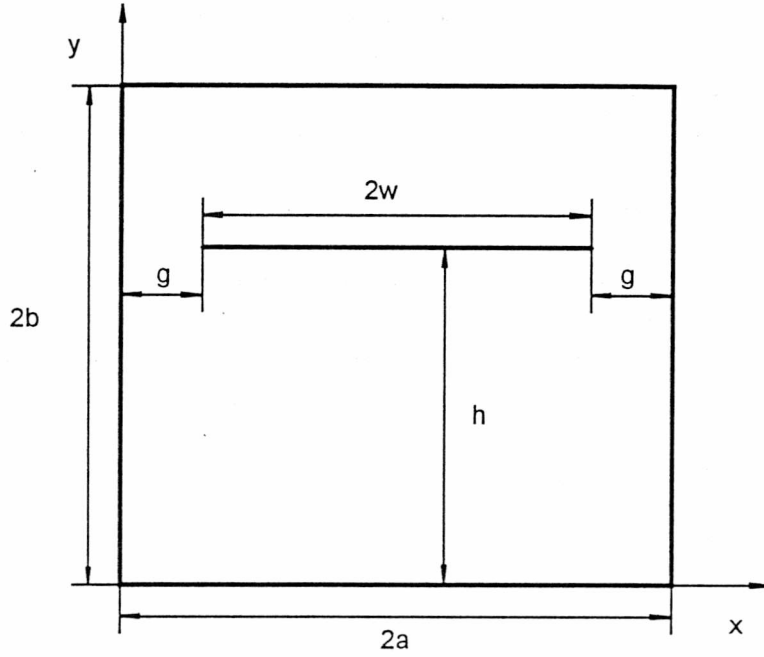


Fig. 2.3. Cross section of a TEM cell

It is impossible to show direct dependence of cell capacitance versus its geometry. Because of that an approximate formula is used:

$$\frac{C_0}{\varepsilon_0} \cong 4 \left[\frac{w}{b} + \frac{2}{\pi} \ln \left(1 + \coth \frac{\pi g}{2b} \right) \right] \quad (2.1)$$

The comparison of values calculated using formula (2.1) and obtained by numerical analysis for various cell geometries shows that it can be used under following conditions:

$$\frac{a}{b} \geq 1, \quad \frac{w}{b} \geq \frac{1}{2} \quad (2.2)$$

On the basis of calculated capacitance, it is possible to calculate characteristic impedance of an equivalent transmission line using the formula:

$$Z_0 = \frac{1}{v C_0} \quad (2.3)$$

where $v = \frac{1}{\sqrt{\mu_0 \epsilon_0}}$ is the phase velocity.

Then:

$$Z_0 \cong \frac{\eta_0}{4 \left[\frac{w}{b} + \frac{2}{\pi} \ln \left(1 + \coth \frac{\pi g}{2b} \right) \right]} \quad (2.4)$$

where $\eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}}$

The Tippet-Chang solution for symmetric cell is used as a reference for more sophisticated methods describing more complicated situations.

2.2.2. Asymmetric cell

Analysis of an asymmetric cell (Fig. 2.4) has been given in [68]. Calculations have been made assuming that only TEM mode exists and using Poisson equation. The analytic formula existing for microwave circuits has been utilised.

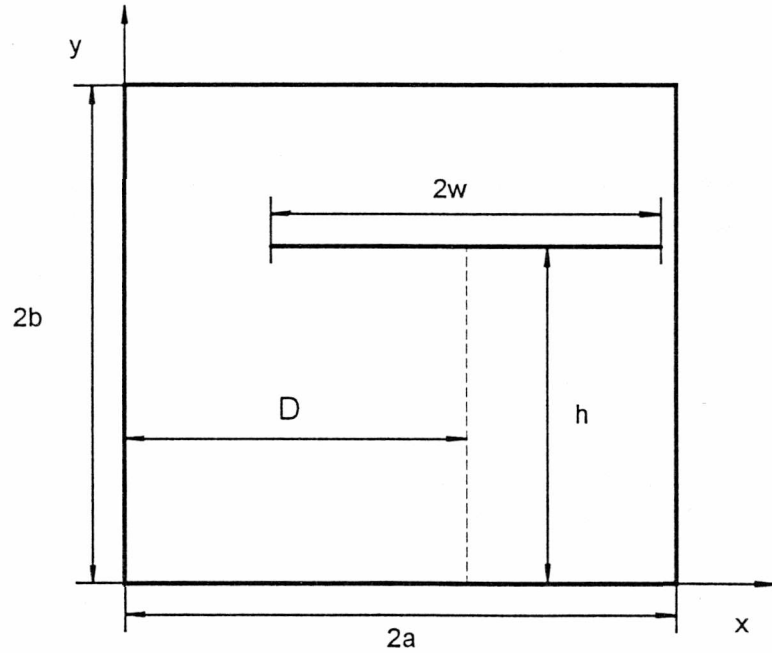


Fig. 2.4. Cross section of an asymmetric cell [68]

Green functions can be written as:

$$G(x, y, x_0, h) = \begin{cases} \sum_{n=1}^{\infty} A_{n1} \operatorname{sh}\left(\frac{n\pi y}{2a}\right) \sin\left(\frac{n\pi x}{2a}\right) & (0 \leq y \leq h) \\ \sum_{n=1}^{\infty} B_{n1} \operatorname{sh}\left(\frac{n\pi}{2a}(2b-y)\right) \sin\left(\frac{n\pi x}{2a}\right) & (h \leq y \leq 2b) \end{cases} \quad (2.5a)$$

where

$$A_{n1} = \frac{\frac{2}{n\pi\epsilon_0} \sin\left(\frac{n\pi x_0}{2a}\right) \operatorname{sh}\left(\frac{n\pi}{2a}(2b-h)\right)}{\operatorname{sh}\left(\frac{n\pi b}{a}\right)} \quad (2.5b)$$

$$B_{n1} = \frac{\frac{2}{n\pi\epsilon_0} \sin\left(\frac{n\pi x_0}{2a}\right) \operatorname{sh}\left(\frac{n\pi}{2a}h\right)}{\operatorname{sh}\left(\frac{n\pi b}{a}\right)} \quad (2.5c)$$

Capacitance per unit length can be calculated using variational formula:

$$C_0 = \frac{\left[\int_{x_0} p(x) dx \right]^2}{\int \int_{x_0 x_0'} p(x) G(x, h, x_0, h) p(x_0) dx_0 dx} \quad (2.6)$$

The charge distribution in the above expression is assumed to be:

$$p(x) = \frac{1}{2w} \left[1 + \frac{1}{(2w)^s} |x - D|^s \right] \quad (D - w \leq x \leq D + w) \quad (2.7)$$

where: D - a displacement of the middle of the inner conductor respecting to the symmetry axis,
 s - a set of points for which $p(x)$ is equal to the exact charge distribution.

Next, the characteristic impedance can be calculated from the expression (2.2). Applying Harrington procedure, the electric potential at any point of the cell section and next the electric field have been calculated. In order to verify obtained results, calculations have been performed for the symmetric line treated as a special case of the asymmetric one. The obtained results have been compared to the exact analysis of Tippet and Chang. A good agreement has been achieved.

2.2.3. Cell with an object (loaded cell)

The simple way of taking into account the presence of the equipment under test (EUT) inside the cell has been proposed by Crawford and Workman in [7]. The authors have noticed that putting EUT into the cell seemingly shortens distance between its plates, which results in proportional increasing of the electromagnetic field strength inside. Thus, new - correct values of the field strength can be calculated using new effective distance $b' = b - h_o$, where h_o is an object effective height. Because h is not equal to the physical height of the object, Crawford and Workman have developed an experimental way of finding h_o , measuring the cell characteristic impedance before and after putting EUT into it. b' can be calculated using transformed equation (2.3). However, one must remember that this equation is true only for symmetric cells. Therefore, it has been proposed that in order to measure new characteristic impedance Z_0' two identical objects should be placed in both parts of the cell. It is also possible to put into one of those parts a special metal box (see Fig. 2.5).

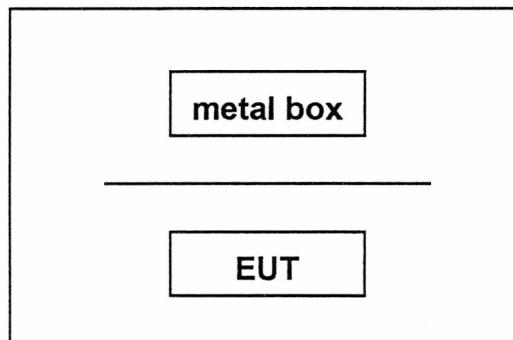


Fig. 2.5. Cross section of the cell with the second EUT modelled by perfectly conducting box

On the bases of the idea shown above Seshardi, Das and Sinha have proposed in [46] the solution taking into account the presence of EUT in only one part of the cell. This is equivalent to solving problem for the asymmetric cell. The EUT presence can be taken into consideration by changing (reducing) of the height of proper part of the cell. It means that the hypothetical cell is analysed with dimensions changed according to physical ones.

As the base of calculations the Laplace equation $\nabla^2 \phi = 0$ has been used. For symmetric line the analytical solution in spherical co-ordinates takes form:

$$\phi = K + \sum_{m=1,3,5\dots} A_m r^{\frac{m}{2}} \cos\left(\frac{m\theta}{2}\right) \quad (2.8)$$

where: K - potential of the inner conductor,
 A_m - unknown coefficients.

Above solution has been generalised for asymmetric line case by dividing cell into two parts, for which solutions are assumed to be:

$$\phi_1 = K + \sum_{m=1,3,5\dots} A_m r^{\frac{m}{2}} \cos\left(\frac{m\theta_1}{2}\right) + \sum_{m=2,4,6\dots} B_m r^{\frac{m}{2}} \sin\left(\frac{m\theta_2}{2}\right) \quad (2.9)$$

$$\phi_2 = K - \sum_{m=1,3,5\dots} A_m r^{\frac{m}{2}} \cos\left(\frac{m\theta_1}{2}\right) + \sum_{m=2,4,6\dots} B_m r^{\frac{m}{2}} \sin\left(\frac{m\theta_2}{2}\right) \quad (2.10)$$

Unknown coefficients A_m, B_m can be found numerically from boundary conditions using the method of successive integration or of least square collocation (both these are based on residue methods).

2.2.4. Asymmetric cell - numerical solution using method of moments

The different idea of calculating Crawford cells parameters has been given by Janiszewski and Karwowski in [25,26]. The authors have decided to use integral equations in the analysis. It allows to avoid methodological difficulties connected with solving differential equations for any surfaces in space.

The usage of the integral model is equivalent to changing problem in a way that instead of solving the Laplace equation for the whole space, the source distribution at

the boundaries is found, and next using easy integral formula the potential at any point of space can be calculated. Additionally, numerical solving of integral equations is much easier than direct solving of the Laplace equation.

The integral equation for the function $q(\mathbf{r})$ describing the charge distribution along a directing line L of the cylindrical surface loaded to potential V_0 takes the form:

$$\int_L q(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' = V_0 \quad \mathbf{r}, \mathbf{r}' \in L \quad (2.11a)$$

where

$$G(\mathbf{r}, \mathbf{r}') = -\frac{1}{2\pi\epsilon_0} \ln|\mathbf{r} - \mathbf{r}'| \quad (2.11b)$$

$\mathbf{r}, \mathbf{r}'$ denote respectively leading vectors of observation and source point.

For the case of two surfaces the electric potential on each of them is the superposition of the potential produced by charges on both surfaces. This statement can be written in the following way:

$$V_1 + V_2 = (V_0)_1 \quad \text{on } L_1 \quad (2.12a)$$

$$V_1 + V_2 = (V_0)_2 \quad \text{on } L_2 \quad (2.12b)$$

where $(V_0)_1$ and $(V_0)_2$ denote assumed values of the potential of both surfaces, while V_1 and V_2 denote partial potentials from charges on respective surfaces.

Combining V_1 and V_2 with respective functions of source distribution gives a set of equations:

$$\begin{aligned} \int_{L_1} q_1(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' + \int_{L_2} q_2(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' &= (V_0)_1, \quad \mathbf{r} \in L_1 \\ \int_{L_1} q_1(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' + \int_{L_2} q_2(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' &= (V_0)_2, \quad \mathbf{r} \in L_2 \end{aligned} \quad (2.13)$$

This system of equations can be algebraized using well-known Galerkin scheme, which leads to obtaining a set of linear equations. Solving this set of equations results in obtaining the charge distribution on the metal surfaces (i.e. the septum and the walls of the TEM cell). When the charge distribution is known, the electric field potential, capacitance per unit length and the characteristic impedance can be easily calculated.

On the bases of above idea, two-dimensional and three-dimensional modelling of loaded cells has been developed in this work (Chapter 3).

2.2.5. Frequency limit of a cell

Ways of finding TEM cell parameters presented in points 2.2.1 - 2.2.4 assume that only TEM mode exists inside the cell. Thus, it is necessary to determine up to what frequency such a model is true.

First - obvious - solution [6] is to apply well known formulas for rectangular waveguide. The formula for calculating cutoff frequency for any mode is:

$$(f_c)_{mn} = \frac{c\sqrt{(b^2m^2 + a^2n^2)}}{2ab} \quad (2.14)$$

for first (TE₁₀) mode we have:

$$(f_c)_{10} = \frac{c}{2a} \quad (2.15)$$

In the work [57] more adequate formula (this time exactly for TEM cell geometry) has been given:

$$(f_c)_{10} = \frac{75}{a} \sqrt{1 + \frac{4ab}{\pi b_1 b_2 \ln\left(\frac{8a}{\pi g}\right)}} \quad [\text{MHz}] \quad (2.16)$$

Then the resonant frequency of the cell can be calculated from:

$$f_{res} = \sqrt{(f_c)_{10}^2 + \left(\frac{c}{2l_{res}}\right)^2} \quad (2.17)$$

where l_{res} denotes the resonant length of the cell

In [63] Wilson and Ma have given approximate expressions for first eight cutoff frequencies of symmetric TEM cells and first three cutoff frequencies of asymmetric TEM cells. Expressions have been obtained assuming small gaps between the walls and the inner conductor of a cell. Unluckily, method applied requires each mode to be investigated separately. This leads to rather large number of formulas and the author has decided not to present them in this work. In [64] Wilson has extended the analysis into higher order mode field distribution.

In [22] the boundary element method (BEM) has been applied to find cutoff frequencies of a TEM cell. As the bases the Helmholtz equation with proper boundary conditions has been taken:

$$\nabla^2 \Psi^{h(e)} + k^2 \Psi^{h(e)} = 0 \quad (2.18)$$

where Ψ^h and Ψ^e - scalar potentials for modes TE and TM (respectively),
 k - the propagation constant

The problem has been reformulated according to BEM technique, and solved numerically. Obtained results have been verified by applying the method to rectangular waveguides (good agreement), but when taking results presented by Hill [20], results for TE_{01} mode in TEM cells are not correct.

Another way of analysis has been proposed in [27], where the Transmission-Line Matrix (TLM) method has been used to calculate TEM cell cutoff frequencies (empty cell were considered).

2.2.6. TEM cell analysis along the direction of propagation

Solutions shortly described in previous points concerne two-dimensional problem: the cell analysis is reduced to analysis of a single cross section perpendicular to the direction of propagation. The analysis of cell parameters along the direction of propagation allows to complete the description with the additional dimension. It is a step into full three-dimensional analysis of TEM cells.

In [66] the influence of measured object on cell parameters has been investigated. The original way of analysis has been applied, concerning cell modelling using a transmission line with varying impedance (Fig. 2.3).

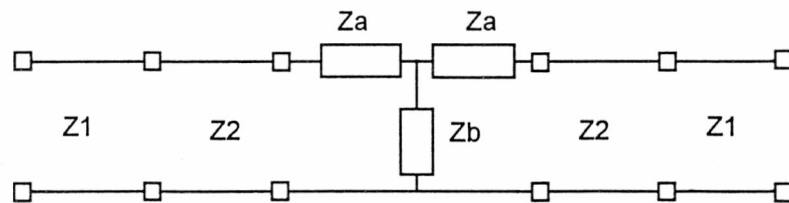


Fig. 2.6. TEM cell modelled by a transmission line

The T-type four terminal network seen in Fig. 2.6 models the investigated object as an ununiformity in the transmission line. The uniform parts of the line are modelling the other parts of the Crawford cell: Z_1 - tapered sections, Z_2 - main section of the cell. In order to obtain values of Z_a and Z_b impedances the theory of small obstacles (dual to the theory of small apertures) has been used. The EUT is described in terms of scattering matrix, which can be next transformed into equivalent impedance matrix.

The analysis has been done for TEM, TE_{10} and TE_{11} modes, but in [66] only results for TEM mode have been presented. The object has been modelled by conductive spheres of various diameters, which have simple analytical description.

Input impedance as a function of frequency has been calculated as well as field distortion due to EUT presence (as a function of EUT position). The results have been verified experimentally and a very good agreement of results has been obtained.

2.2.7. Three-dimensional analysis of a Crawford cell

In [54] and [55] Tanaka, Honma and Kaji have presented three-dimensional analysis of a TEM cell using mixed boundary element method. Earlier, the authors tried to make use of conventional boundary element methods but it caused the necessity of taking into account many special cases and of solving set of equations with a great number of unknowns. Applying mixed BEM has allowed to reduce the number of equations to "only" 632.

The main purpose of the work was to find the characteristic impedance along the TEM cell. In order to check the method, first simple three-dimensional model of the main (rectangular) section of the Crawford cell was analysed. Next the analysis was extended to the tapered sections.

In the analysis only basic (TEM) mode has been considered. Thanks to it the problem can be reduced to solving a three-dimensional Laplace equation. The following boundary conditions for electric potential ϕ have been assumed:

$$\phi = 1.0 \text{ on the inner conductor,}$$

$$\phi = 0.0 \text{ on the outer walls,}$$

$$\frac{\partial \phi}{\partial n} = 0.0 \text{ on the planes of symmetry and end planes of the cell.}$$

As one can see it is equivalent to taking into account only 1/8 part of the cell (which must be a symmetric one).

The obtained results (the characteristic impedance distribution along the cell) have been compared with results of two-dimensional analysis and with results of measurements using time domain reflectometer (TDR) published in [6]. Numerical results differ partially from measured values which has been suggested by the authors to be a result of simplifications used in the model. The main problem pointed out is the

fact that the 50 ohm load has not been applied in the model. Apart from it the physical construction of the cell (dielectric support) could affect the results of measurements.

Apart from the differences it can be stated that the presented method can be a start point to more adequate analysis of Crawford cells. Therefore some ideas of the work have been used in the present contribution (see Chapter 3).

3. TEM mode in TEM cells

First, TEM cells have been modelled assuming, that only TEM mode is propagated inside. In such a way the basic and the most important mode of cell operation has been investigated.

It is extremely important to find out which of cell parameters and characteristics are related to the TEM mode (i.e. the mode to operate with the cell was designed), and which are the result of the presence of higher order modes. Knowing the nature of - for example - resonances inside the cell (which are limiting the useful bandwidth of the cell) it is much easier to eliminate them using one of known methods [8]. Therefore the ability to investigate TEM mode separately is of great importance.

The TEM mode investigation has been divided into two parts. First, a two-dimensional model was applied. Such an attitude is equivalent to treating TEM cell as an infinite TEM line, i.e. eliminating effects of longitudinal resonances, tapered sections and imperfect load influence. Next, a simplified three-dimensional model is presented which is especially useful in determining the equipment under test (EUT) influence on cell characteristic impedance distribution.

3.1. Method of calculations

It is well known, that the TEM mode, i.e. the mode, in which both E and H vectors of the electromagnetic field are perpendicular to the direction of the wave propagation, can be propagated only in structures build of two or more conductors [47]. Assuming, that such a structure is infinite along the direction of the wave propagation one can prove [47,69], that the electric field in the cross section is non-rotational. It means, that in order to calculate the distribution of such a field, a scalar potential can be used, that satisfies the Laplace equation. From this it is obvious, that methods used in electrostatics can be easily applied. Of course, infiniteness of the structure means, that we deal with two-dimensional analysis, in this case treating a TEM cell as an infinite transmission line. Such attitude towards the problem causes the lack of information concerning field distribution along the cell axis. Unfortunately the simple way of three-dimensional analysis does not exist. The following way of treating the problem is popular [54,55]: when we assume, that electric and magnetic field profiles of any surface remain generally unchanged in the direction of propagation, and TEM wave dominates even for small changes in the field distribution, then we can assume, that such electromagnetic field can be approximated by three-dimensional electrostatic field, satisfying three-dimensional Laplace equation. However one must remember that such an attitude means in fact far going simplification, and therefore must be treated with much care.

Methods used to calculate electrostatic fields can be formulated in two ways: the problem can be investigated on the basis of differential equations (differential model) or integral equations (integral model). The differential model concerns solving the Laplace equation for electric potential V in unbounded region, with additional boundary conditions on the metal surfaces. The integral model is the integral Fredholm equation of the first order for the functions describing charge distribution on the appropriate surfaces. In the first case solving the differential model is equivalent to solving the initial problem. In the second (integral model), after solving the integral

equation it is necessary to make some additional calculations in order to find the desired potential. Although calculating the electric field distribution produced by the sources of known distribution causes no methodological difficulties, the necessity of making such calculations may seem to be a disadvantage of the integral model. In fact the situation is quite different. One must take into account, that direct method of solving the problem for surfaces of any shape is still unknown. So, the only way is to apply numerical methods. There are few well known and described numerical methods:

- finite-differences method,
- finite-elements method,
- variational method,
- the method of moments.

Theoretically all mentioned methods can be applied to solve equations with differential, integral, or mixed operators. The efficiency of each of them depends on the type of the problem. Finite-differences and finite-elements methods are usually used to solve the Neuman and Dirichlet (or mixed) problems in the bounded region of the space (two- or three-dimensional). In the case of the TEM cell, especially when investigating the three-dimensional case, problem cannot be treated as an internal one. Also investigating any hole in the metal wall of the cell needs to taking into account unbounded region to find a solution. Both mentioned methods are not the tool for this purpose, because of the necessity of covering the whole region of interest with the dense enough network with finite number of nodes. So in external problems that "infinite region" must be bounded in some way. Applying to open surfaces the direct Neuman condition [54,55] doesn't seem to result from any logical factor.

Applying the variational method to solving differential model (i.e. the Laplace equation) of the problem is possible, when one can give set of functions, on which appropriate function satisfying boundary conditions and warranting appropriate behaviour of the potential in infinity, will be minimalized. For the surface of any shape it is rather hopeless task. One can meet similar difficulties trying to apply the method of moments - this time the problem concerns basis functions.

Difficulties mentioned above do not occur, when the problem is formulated on the basis of integral equations. Applying integral model is equivalent (as it was said before) to reformulate the problem in such a way, that instead of solving directly Laplace equation for the potential in unbounded space, one can find the source distribution on the boundary of this region, and then applying simple formula, calculate potential at any point of the space. So, the base of this formulation of the problem is the dimension reduction. Apart from it, numerical solving of appropriate integral equation is much easier, then solving Laplace equation.

3.2. Two-dimensional analysis

3.2.1. General remarks

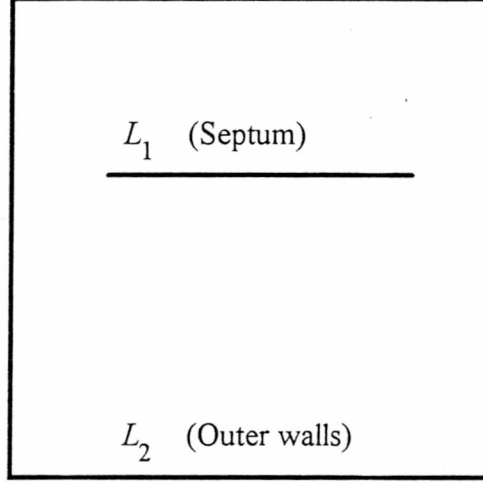


Fig. 3.1. Cross section of a TEM cell

The situation is shown in Fig. 3.1. Let's consider generally a set of k parallel conductors with directing lines $L_1, L_2, \dots, L_k$, loaded to potentials $(V_0)_1, (V_0)_2, \dots, (V_0)_k$. The densities and source distributions on every surface must satisfy boundary conditions, i.e. the potential of the electric field produced by those sources must be $(V_0)_1, (V_0)_2, \dots, (V_0)_k$ on the proper surfaces. Taking $V_1, V_2, \dots, V_k$ as part potentials from charges on respective surfaces one can write:

$$\sum_{i=1}^k V_i = (V_0)_j \quad \text{on } L_j, \quad j = 1, 2, \dots, k \quad (3.1)$$

V_i is connected with proper function of the charge distribution $q_i(\mathbf{r})$ by the integral equation:

$$V_i = \int_L q_i(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') dl' \quad (3.2)$$

where G function is defined by formula:

$$G(\mathbf{r}, \mathbf{r}') = -\frac{1}{2\pi\epsilon_0} \ln|\mathbf{r} - \mathbf{r}'| \quad (3.3)$$

The end of the observation point vector $\mathbf{r}$ is placed on L_j , the end of the source point vector $\mathbf{r}'$ - on L_i . In this way the system of k coupled integral equations including unknown functions $q_1(\mathbf{r})$, $q_2(\mathbf{r})$, ..., $q_k(\mathbf{r})$ is obtained:

$$\begin{aligned} \int_{L_1} q_1(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \int_{L_2} q_2(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \dots + \int_{L_k} q_k(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' &= (V_0)_1, \quad \mathbf{r} \in L_1 \\ \int_{L_1} q_1(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \int_{L_2} q_2(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \dots + \int_{L_k} q_k(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' &= (V_0)_2, \quad \mathbf{r} \in L_2 \\ \dots & \\ \int_{L_1} q_1(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \int_{L_2} q_2(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' + \dots + \int_{L_k} q_k(\mathbf{r}') G(\mathbf{r}, \mathbf{r}') d\mathbf{l}' &= (V_0)_k, \quad \mathbf{r} \in L_k \end{aligned} \quad (3.4)$$

This set of equations can be algebraized taking approximation for charge distribution functions in form of the linear combination:

$$q_i(\mathbf{r}) = \sum_{n=1}^{N_i} Q_n^i P_n^i(\mathbf{r}), \quad i = 1, 2, \dots, k \quad (3.5)$$

of base functions:

$$P_n^i(\mathbf{r}) = \begin{cases} 1, & \mathbf{r} \in \Delta L_n^i \\ 0, & \mathbf{r} \notin \Delta L_n^i \end{cases} \quad (3.6)$$

ΔL_n^i are the fragments of curves L_i (each curve is divided into N_i segments; n denotes the number of the segment). Substituting (3.5) into (3.4), and demanding that the set must be true at N_i points of each curve L_i - in the middle of each ΔL_n^i segment we obtain the system of $R = \sum N_i$ linear equations for unknown Q_n^i coefficients.

The set of equations can be written in a matrix form:

$$\begin{bmatrix} S_{mn}^{11} & S_{mn}^{21} & \dots & S_{mn}^{k1} \\ S_{mn}^{12} & S_{mn}^{22} & \dots & S_{mn}^{k2} \\ \dots & \dots & \dots & \dots \\ S_{mn}^{1k} & S_{mn}^{2k} & \dots & S_{mn}^{kk} \end{bmatrix} \cdot \begin{bmatrix} Q_n^1 \\ Q_n^2 \\ \dots \\ Q_n^k \end{bmatrix} = \begin{bmatrix} U_m^1 \\ U_m^2 \\ \dots \\ U_m^k \end{bmatrix} \quad (3.7)$$

S_{mn}^{ij} ($i, j=1, 2, \dots, k$) is a $(N_i \times N_j)$ - element matrix of potential coefficients, Q_n^j - a N_j -element vector of searched coefficients Q_n^j of the approximation (3.5) of the function $q_j(\mathbf{r})$, and U_m^i is a N -element vector of "exciting coefficients":

$$U_m^i = \begin{bmatrix} (V_0)_i \\ (V_0)_i \\ \dots \\ (V_0)_i \end{bmatrix} \quad (3.8)$$

Matrixes S_{mn}^{ii} describe an influence between segments belonging to the same conductor, S_{mn}^{ij} ($i \neq j$) - between segments of different conductors.

Solving (3.7) is equivalent to finding the approximate charge distribution functions $q_1(\mathbf{r})$, $q_2(\mathbf{r})$, ..., $q_k(\mathbf{r})$. Then the potential at any point of the space can be obtained from the formula:

$$V(\mathbf{r}) = \sum_{i=1}^k \sum_{n=1}^{N_i} Q_n^i \int_{\Delta L_n^i} G(\mathbf{r}, \mathbf{r}') dl' \quad (3.9)$$

Knowing the electric potential distribution one can easily calculate the electric field strength as:

$$\bar{E} = -\text{grad} V \quad (3.10a)$$

Magnetic field $\mathbf{H}$ can be found from the relationship between the electric and magnetic field vectors in TEM mode [47]:

$$\mathbf{i} \times \bar{\mathbf{E}} = \pm Z_f \bar{\mathbf{H}} \quad (3.10b)$$

where $\mathbf{i}$ denotes unit vector pointing the direction of propagation, and Z_f is the wave impedance of the free space.

Elements of the matrix $\mathbf{S}$ are calculated using the formula:

$$S_{mn} = \int_{\Delta L_n^i} G(\mathbf{r}_m, \mathbf{r}') dl' \quad (3.11)$$

where $\mathbf{r}_m$ - a vector of the middle of the segment ΔL_m^i .

Assuming subtle segmentation of the curves, each of ΔL_n^i segments can be replaced with the sector of the straight line. In this case integral (3.11) can be calculated analytically [26]:

$$S_{mn} = \frac{\Delta L_n^i}{2\pi\epsilon_o} \left\{ 1 - \ln\left(\frac{\Delta L_n^i}{2}\right) - \frac{1}{4} a_{mn}^+ \ln(c_{mn}^+) - \frac{1}{4} a_{mn}^- \ln(c_{mn}^-) \right. \\ \left. + b_{mn} \left[\arctg\left(\frac{-a_{mn}^+}{b_{mn}}\right) - \arctg\left(\frac{a_{mn}^-}{b_{mn}}\right) \right] \right\} \quad (3.12)$$

where

$$a_{mn}^{\pm} = 1 \pm \frac{2}{\Delta L_n^i} [\mathbf{l}_n \cdot (\mathbf{r}_m - \mathbf{r}_n)] \quad (3.13a)$$

$$b_{mn} = \frac{2}{\Delta L_n^i} \sqrt{|\mathbf{r}_m - \mathbf{r}_n|^2 - [\mathbf{l}_n \cdot (\mathbf{r}_m - \mathbf{r}_n)]^2} \quad (3.13b)$$

$$c_{mn}^{\pm} = (a_{mn}^{\pm})^2 + b_{mn}^2 \quad (3.13c)$$

For $m=n$:

$$S_{mm} = S_{nn} = \frac{\Delta l_n^i}{2\pi\epsilon_0} \left[1 - \ln \left(\frac{\Delta l_n^i}{2} \right) \right] \quad (3.14)$$

Δl_n^i denotes the length, $\mathbf{1}_n$ - the unit vector, and $\mathbf{r}_n$ points the middle of the section ΔL_n^i .

3.2.2. Results of two-dimensional modelling

The technique described above has been applied to calculate electric parameters of the Crawford cell constructed in the Institute of Telecommunication and Acoustics of the Technical University of Wroclaw, in the EMC Laboratory (under the government project CPBR no. 8.13 KASK). The dimensions (according to the Fig. 3.1) are:

$$a = 0,6 \text{ m}; \quad b = 0,6 \text{ m}; \quad h = 0,8 \text{ m}; \quad w = 0,99 \text{ m}.$$

Those dimensions are the result of using special curves, published by [7] in order to help TEM constructors. The characteristic impedance of the cell with such dimensions calculated with simple approximate formula (2.4) given by Tippet and Chang [58] is equal to 51,9 ohms.

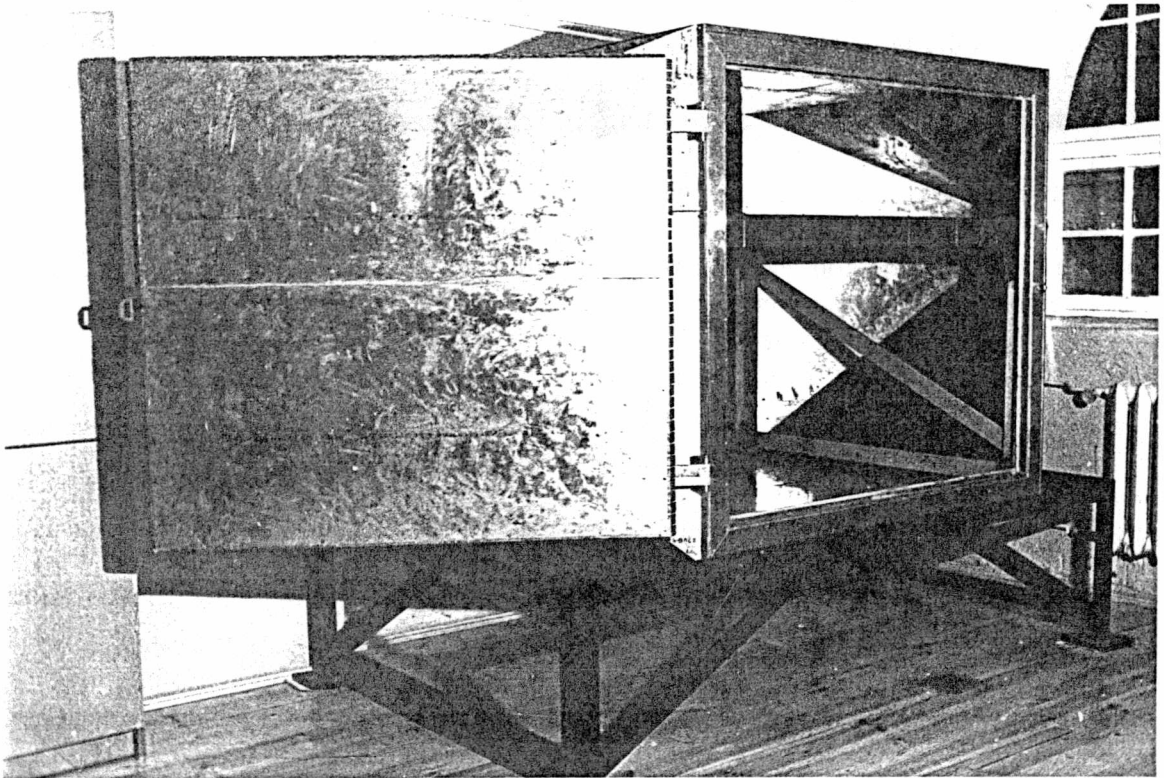


Fig. 3.2. TEM Crawford cell constructed in ITA

The segmentation has been made using the Tchebycheff polynomials technique - it allows to obtain more exact results with less (compared with equal segmentation) number of segments, because in regions, where the field and the charge distribution structures are more complicated, the segmentation is more subtle, and vice versa. The inner conductor potential is assumed to be 1 V , the outer walls are grounded. With this assumption charge on 1 meter of the inner conductor is equal to cell's capacitance per unit length C_0 [25] (the cell is considered as a TEM transmission line). Knowing C_0 it is easy to calculate the characteristic impedance from:

$$Z_0 = \frac{1}{cC_0} \quad (3.15)$$

The characteristic impedance obtained numerically for the ITA cell is $50,643$ ohms, while the purpose of constructors was to obtain exactly 50 ohms. So the difference is about 1% , which is in the range of tolerance.

From above results it is obvious, that numerical modelling gives better results than applying approximate formulas (those formulas were obtained under rather strong assumptions concerning very small gap between the inner conductor and the walls of the cell).

The main purpose of this work is to investigate the electromagnetic field distribution inside a TEM cell in order to determine primarily the field regularity in the working volume. Calculating the strength of the field components at any point of the working volume creates no serious problems, when the charge distributions on the septum (internal electrode) and on the cell walls are known. What rises considerable problems, is the presentation of the results. This problem is related to the vectorial nature of the investigated quantities. In engineering practice there is only one presentation method that rises no objections. It consists in covering the whole region of interest with dense enough network of nodes, and in assigning to each of them two scalar components - in the case of electric field - E_x and E_y values at a given point of the field. This method however results in rather poor visualisation. Visualising the field lines gives the information about the field structure but not about values. Apart from it there are no fast algorithms of obtaining the field lines picture for the field of the unknown structure. Therefore in this work the author decided to use two ways of presentation of the results of calculation. In the case of TEM mode one should keep in mind, that the transverse electric field is a potential field. Of course the presentation of the calculated results for a scalar value, that the electric potential is, is much easier than for a vectorial function. It should be mentioned, that in this case no important information is lost, because from the potential distribution it is easy to conclude on the distribution and strength of the field. Such attitude is applied by many authors [19,25,26]. Additionally the author decided to present the $|E_x|$ and $|E_y|$ distribution on separate pictures. This is caused by the fact, that in the case of investigating higher order modes of the electromagnetic field (Chapters 4, 5) calculated fields are no longer potential ones, and there is a need of comparing the field distribution for TEM and non-TEM ranges of the cell operation. Because the electric field strength is much easier to obtain (formula 3.10a) the author decided to use E_x and E_y values for the field presentation. Of course in all cases it is also possible to obtain corresponding H-field

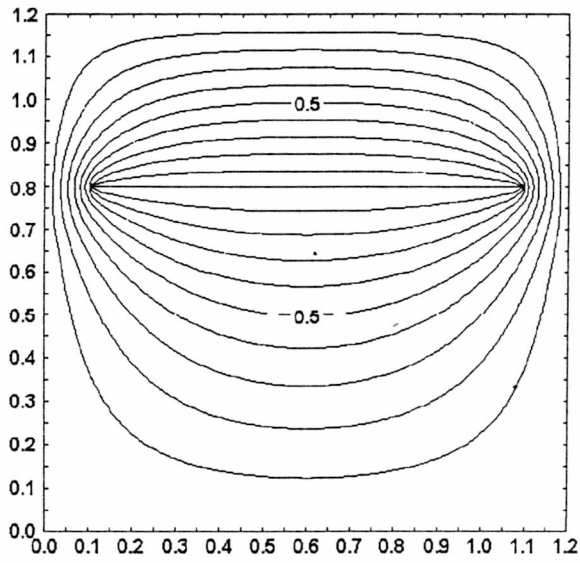


Fig.3.3a. Potential distribution in an empty TEM cell - equipotential lines

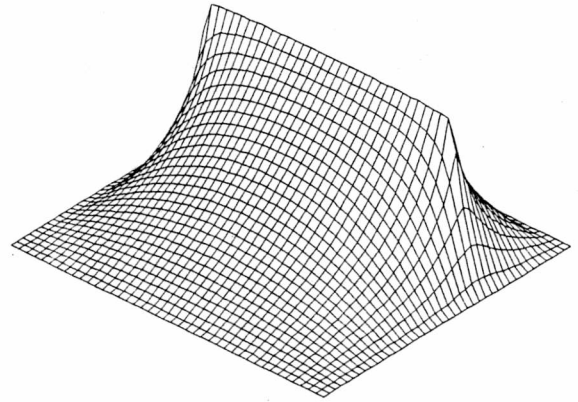


Fig.3.3b. Potential distribution in an empty TEM cell

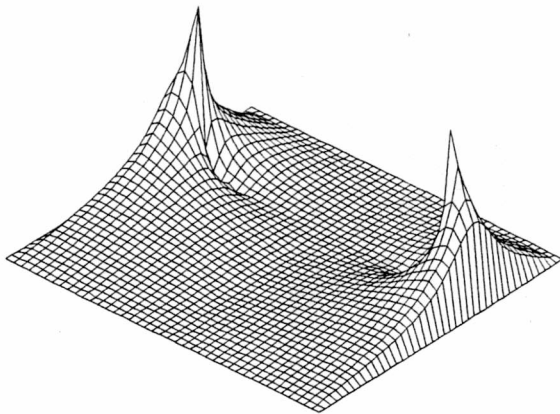


Fig.3.3c. Field distribution in an empty TEM cell - $|E_x|$ component

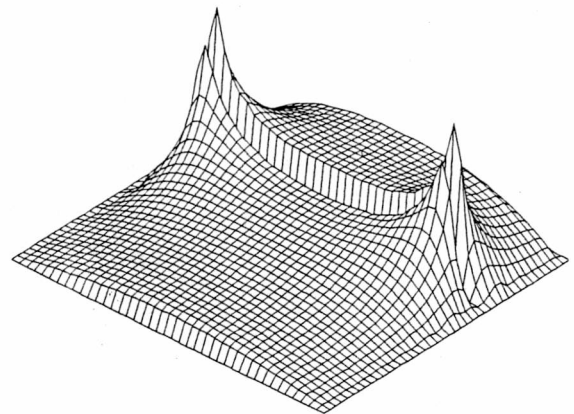


Fig.3.3d. Field distribution in an empty TEM cell - $|E_y|$ component

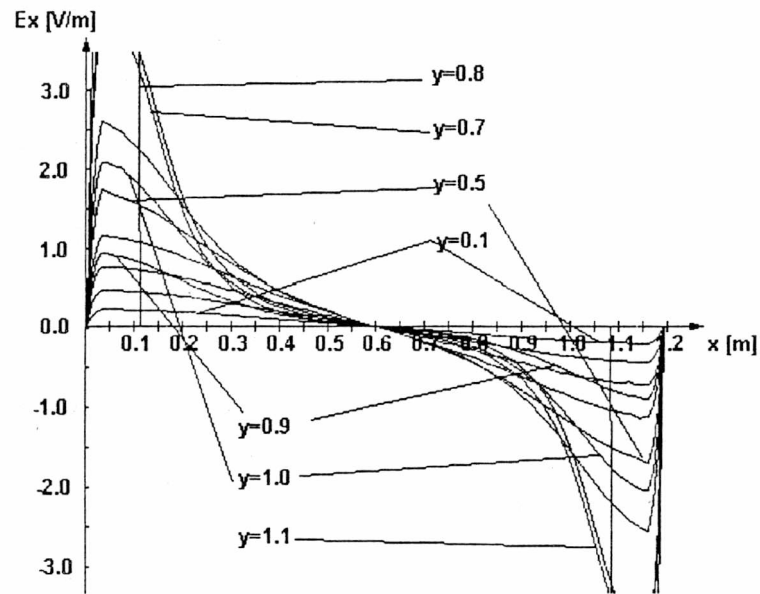


Fig.3.3e. Field distribution in an empty TEM cell - values of the E_x component

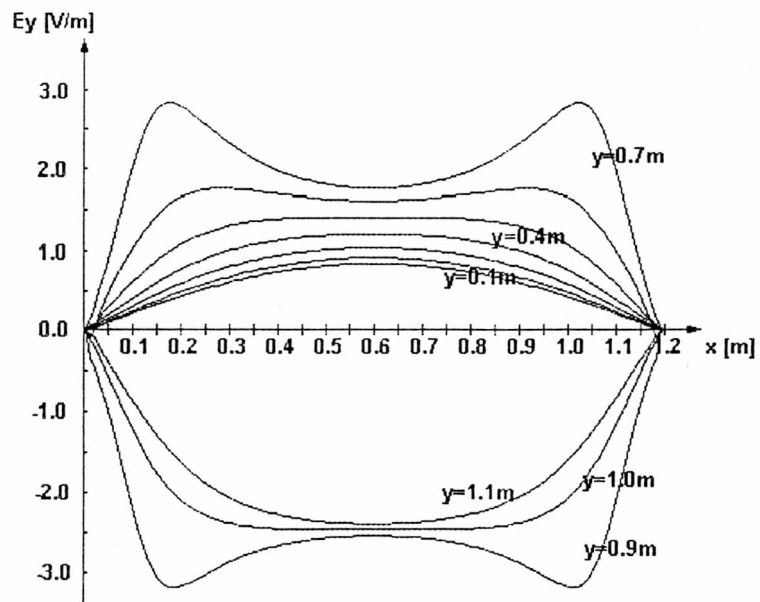


Fig.3.3f. Field distribution in an empty TEM cell - values of the E_y component

quantities (in this case from 3.10b). Apart from it the electric field is commonly used by the other authors [64], which gives the possibility of making some comparisons.

Results of modelling of the empty ITA cell are shown in Fig. 3.3. Figure 3.3a and 3.3b show the electric potential distribution. The reader can compare the results with presented in other works [25,26]. In Figures 3.3c and 3.3d respectively the $|E_x|$ and $|E_y|$ fields distributions are presented. It is worth to notice, that $|E_x|$ vanishes on the walls lying along x axis and $|E_y|$ - along y axis, which points out that boundary conditions are satisfied (tangential electric field component is equal to zero on metal surfaces). As it was expected the field distribution is more uniform in the region above than in the region below the inner conductor. Of course the field strength is also bigger in the region above septum, because of the shorter distance between electrodes. Unfortunately the region above septum is much smaller, which results in much smaller work volume in this part of the cell. The results of analysis very well demonstrate the main problem of TEM cells constructors - necessity to achieve a compromise between the working volume size and the field regularity. Increasing the working volume size by constructing a bigger TEM cell results in decreasing the usable frequency bandwidth. Additionally, for the empty cell the E_x and E_y field distribution is presented in the form of a family of curves visualising the x -axis dependence with a parameter that is the value of y (Fig. 3.3e, 3.3f). It means the field distribution in the horizontal section was shown on different heights of the cell. One can see that E_x values are equal to zero in the middle of the cell. The E_y values in the middle of work volume (i.e. in the middle of the distance between the floor and the septum, $y=0.4m$) are nearly equal to $1.25 V/m$, which is the value obtained by simply dividing the potential difference on both conductors ($1V$) by the distance between conductors ($0.8m$). It means that for the case of empty cell the simple way of estimation of the field magnitude commonly used [7] is valid.

Figures 3.4 to 3.6 show the EUT influence on the field distribution inside the cell. Taking into account presence of EUT results with the serious problem. In the method applied it is necessary to assume the electric potential of the object. In this work two possible solutions are investigated.

First, the EUT potential has been assumed to be $0V$, which is equivalent to grounding the object. It seems to be a good assumption when considering EUT working with the outer power supply (for example personal computer). As the EUT the rectangular cylinder has been taken. Of course modelling objects of any shape is possible with the method. The Fig. 3.4 and 3.5 present results of calculations for this particular case. The EUT is placed respectively in the middle of the lower and upper part of the TEM cell. One can see that the field distribution changes dramatically. There is almost no electric field between the floor (or the ceiling) and the object - it is obvious: both the EUT and the walls are grounded. From the other hand, the field strength between the EUT and the septum strongly increases. It means that EUT is no longer placed in the field predicted for empty cell, and therefore any comparison between susceptibility levels obtained in TEM cells are not directly comparable with those obtained in free space. There are two ways of taking this fact into account: applying special field probes to monitor the field distribution (it should be mentioned that such probes are not commercially available up till now), or predicting field distribution theoretically, to which this work is devoted. Corresponding computer programs for PC are available from the author.

Next, the case of not grounded object has been considered. It corresponds to the situation, when the EUT has its own power supply (for example a notebook PC computer with a battery). The object potential is treated as an additional unknown

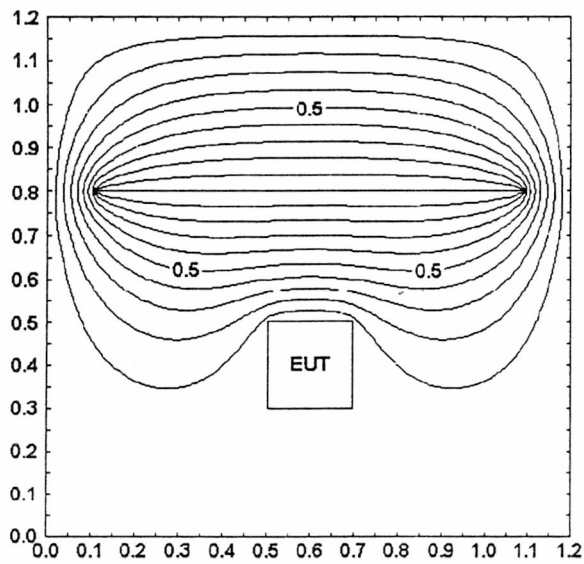


Fig.3.4a. Potential distribution in a loaded TEM cell - equipotential lines

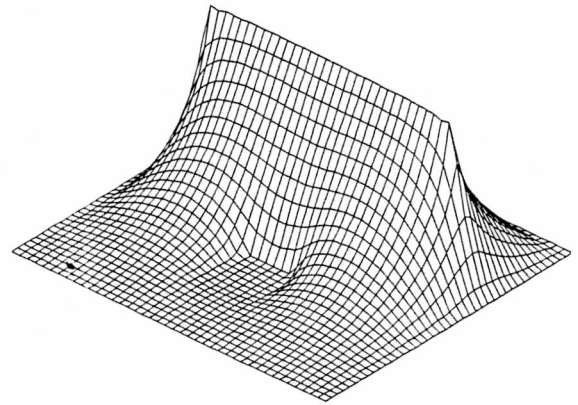


Fig.3.4b. Potential distribution in a loaded TEM cell - grounded EUT in the lower part

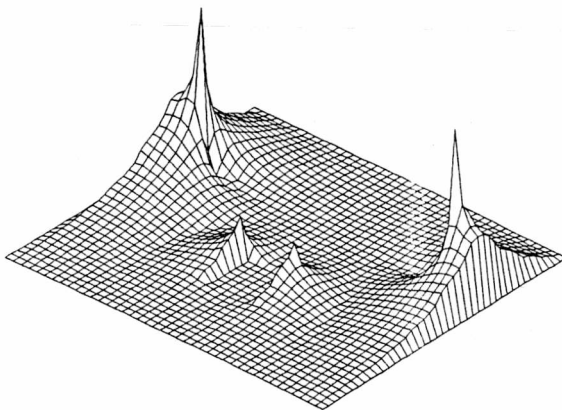


Fig.3.4c. Field distribution in a loaded TEM cell - $|E_x|$ component

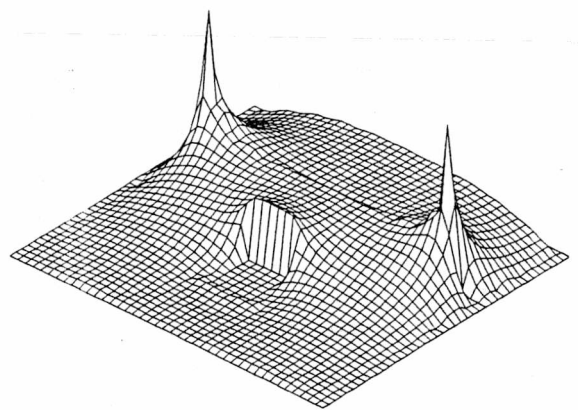


Fig.3.4d. Field distribution in a loaded TEM cell - $|E_y|$ component

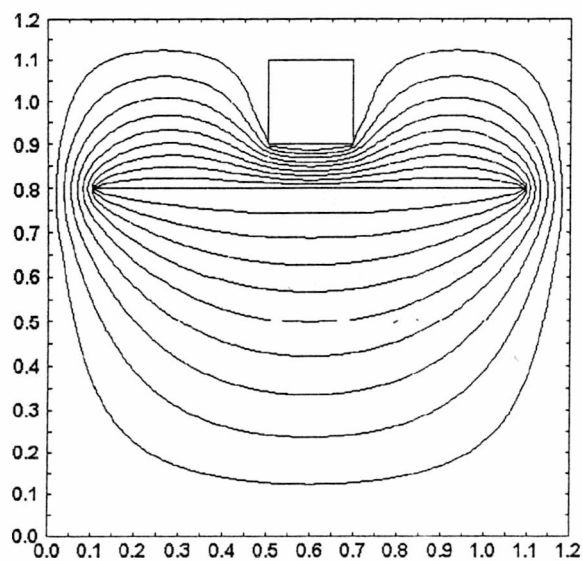


Fig.3.5a. Potential distribution in a loaded TEM cell - equipotential lines

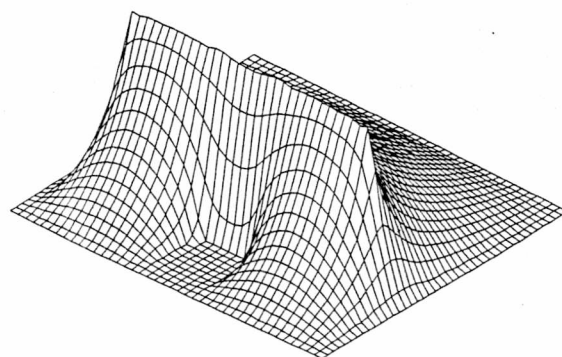


Fig.3.5b. Potential distribution in a loaded TEM cell - grounded EUT in the upper part

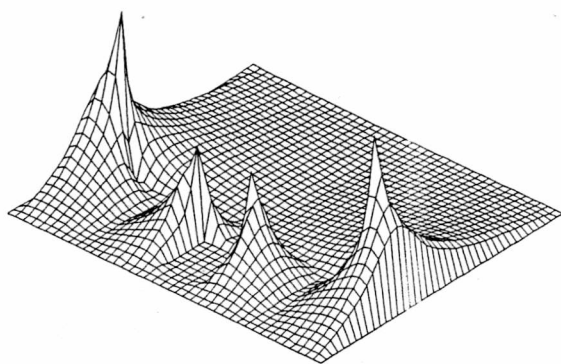


Fig.3.5c. Field distribution in a loaded TEM cell - $|E_x|$ component

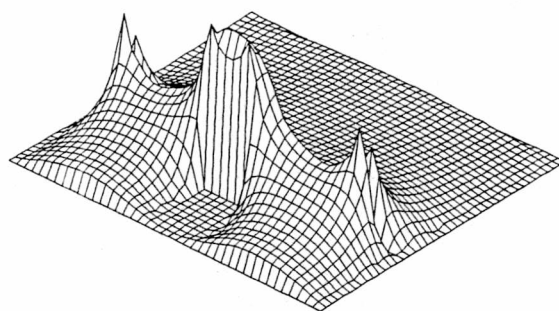


Fig.3.5d. Field distribution in a loaded TEM cell - $|E_y|$ component

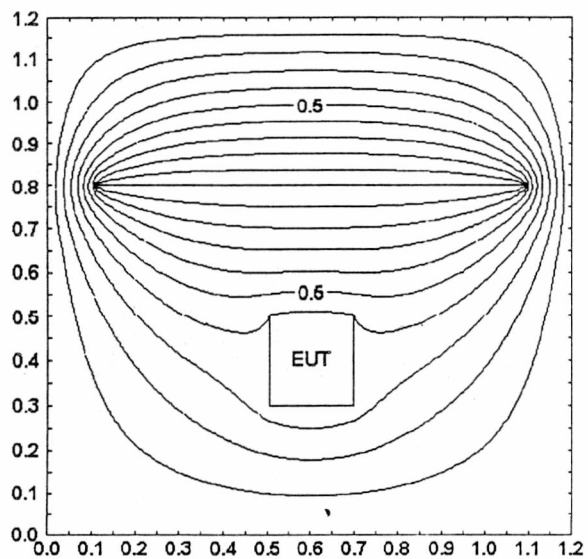


Fig.3.6a. Potential distribution in a loaded TEM cell - equipotential lines

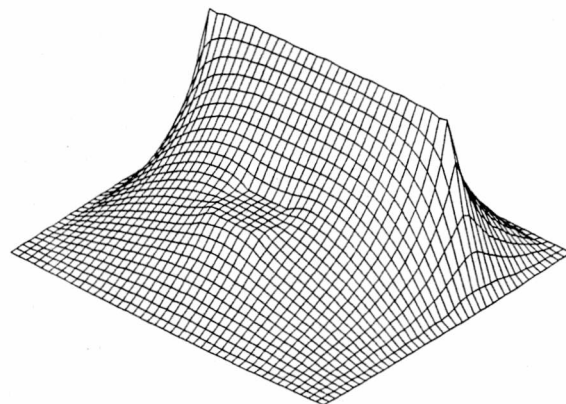


Fig.3.6b. Potential distribution in a loaded TEM cell - EUT potential calculated

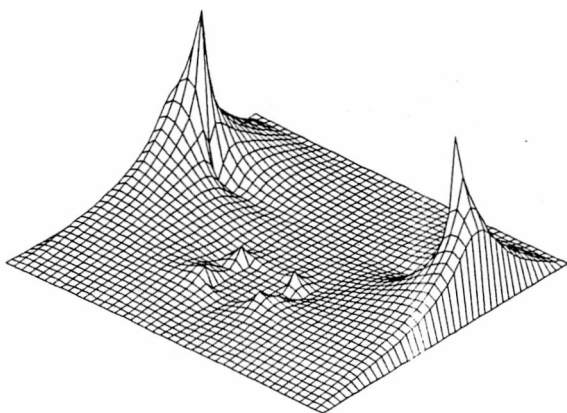


Fig.3.6c. Field distribution in a loaded TEM cell - $|E_x|$ component

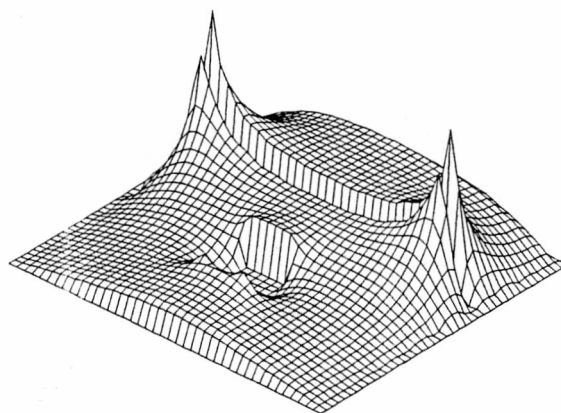


Fig.3.6d. Field distribution in a loaded TEM cell - $|E_y|$ component

quantity in the set of equations (3.7). In order to make the problem solvable the set of equations must be completed with an additional equation. To satisfy this condition the whole charge on the EUT is assumed to be zero, because there is no electrical connection through which the charge could be change. Additional equation takes the form:

$$\sum_{n=1}^{N_3} Q_n^3 \Delta L_n^3 = 0 \quad (3.16)$$

The results of calculation are presented in Fig. 3.6. The calculated object potential is equal to 0.38 V . It is easy to notice that the field distribution is much less destroyed than in the case of grounding the object. The only change is a small increasing of the $|E_y|$ field level (in comparison to empty cell). The conclusion is that as long as the size of EUT is small, and the EUT is not to be grounded, the simple way of recalculating the field strength given by Crawford [7], can be applied. Of course this must be verified experimentally. The verification will be possible when the field probes for measuring fields in the near zone will be available. The investigation on building such probes is in progress in the Radio Department of ITA.

The source of additional difficulties is the change of the cell characteristic impedance due to the presence of the EUT. Putting a metal box into working volume changes via the field coupling TEM cell parameters. The table below lists the characteristic impedance values for the cases investigated before.

Table 3.1. Changes of the characteristic impedance of ta TEM cell due to the EUT presence

EUT	Characteristic impedance [Ω]
No (Empty cell)	50.643
Grounded, lower part of the cell	46.349
Grounded, higher part of the cell	37.798
Not grounded, lower part of the cell	50.106

The characteristic impedance is practically not affected only in the last case (not grounded EUT).

3.3. Three-dimensional analysis

The experience with the Crawford cell leads to a conclusion that two-dimensional modelling is not enough in order to determine cell parameters exactly. The main problem is to obtain reliable value of the characteristic impedance. For the empty cell

two-dimensional model seems to be good enough, but when investigating the EUT presence taking two-dimensional analysis is equivalent to assuming that the EUT is infinitely long along the z axis (the same as electrodes of the cell treated as an infinite TEM line). Apart from it, when using 2D method it is impossible to find out why characteristic impedance of the Crawford cell is frequency dependent, which is known from measurements. Changes of the impedance value in relation to the frequency are related to non-uniformity of the impedance along the cell [7,54]. In order to investigate this problem, and to check usefulness of results from the Table 3.1 the three-dimensional model of the TEM (Crawford) cell must be applied. Such a model was proposed in [54,55]. The authors used the boundary element method (BEM) which resulted in very large set of linear equations to be solved. The method was applied to the symmetrical NBS TEM cell. The symmetry allowed to reduce the dimension of the problem eight times, which resulted in solving 632×632 set of equations. Applying this method to asymmetric cells (four times reduction), with an object inside, would result in dealing with about 1400×1400 elements matrices, which is not possible using PCs, and makes worse the accuracy of the results.

Therefore in this work the method similar to the one used for two-dimensional analysis has been developed. It allows to achieve the same as in [54] level of accuracy using only 348×348 matrix (for empty cell).

For the three-dimensional case in place of curves L_i one must take metal surfaces modelling the electrodes of the cell. The integrals in (3.2), (3.4), (3.9) and (3.11) become now the surface integrals. As the kernel of the equation (3.2) the following function is assumed [47]:

$$G(\mathbf{r}, \mathbf{r}') = \frac{1}{4\pi\epsilon_0} \cdot \frac{1}{|\mathbf{r} - \mathbf{r}'|} \quad (3.17)$$

which is the result of the three-dimensional solving of the Poisson equation.

The functions $q_j(\mathbf{r})$ can now be treated as functions of surface charge density on each electrode. In order to apply numerical method all metal surfaces must be divided into small flat quadrangle elements. Next, the procedure is the same as in chapter 3.2.

The calculation of elements of the matrix $\mathbf{S}$ bases on calculating the function $1/|\mathbf{r}|$ integral along the flat piece of surface. Such a piece can always be decomposed into parts that are rectangles or rectangular triangles, so the problem is to calculate the integral along the rectangle or triangle. There are two methods of dealing with this problem: numerical and analytical. The numerical method concerns on subdividing the element into small sub-elements which are assumed to have a constant value of the $1/|\mathbf{r}|$ function - as this value the value from the middle of the sub-element can be taken. Next the integration is replaced with summation. Such a method is equivalent to the method of rectangles used in numerical integration of Rieman integral. The analytical method concerns on applying standard methods of surface integration of functions of two variables. The procedure is complicated and cannot be presented in the form of simple formulas similar to (3.12 -14). The integration scheme which requires taking into account many special cases is presented in the Appendix A. The integration formulas were obtained for normalised set of coordinates with assumption that the element lies in the plane xy with the middle at the $(0,0,0)$ point (rectangle), or with the right angle in the $(0,0,0)$ point (triangle). So, applying this scheme to calculation of elements S_{mn} requires special transformation of the coordinate system.

Results of analytical and numerical integration have been compared in order to verify analytical formulas, as well as to check the accuracy of calculations using numerical method. Generally the difference between values obtained from both methods are on the level of better than 0.01 per cent while using the number of subdivisions in the numerical method 10×10 . The exceptions are diagonal elements S_{mm} , where differences are even bigger then 20 per cent. It means, that numerical method is not suitable for computing diagonal elements of the matrix. There is no simple way of solving this problem, because even using the number of subdivisions 100×100 for each element does not make the situation much better. The reason is simple: there are residua while calculating S_{mm} elements and the numerical method is not well for dealing with values going to infinity. Because of the reasons mentioned above, during calculations both methods were used simultaneously in order to control each other.

Because of the problem symmetry, the electric charge on corresponding elements of each $1/4$ of the cell is the same. So, the number of unknowns in (3.7) is reduced four times. It should be pointed out that the fact, that unknowns are equal, does not mean that elements of matrix are equal. Elements S_{mn}' of the new, reduced matrix are obtained through summation of corresponding four elements of S matrix.

Another problem is how to solve the linear set of equations (3.7) with rather large number of unknowns (about 300). The special method for large matrices [3] has been used, which is equivalent to Gauss method with partial choice of the main element. Luckily the obtained matrices have good numerical properties, which means there are no significant errors due to the finite precision of computer calculations.

The way of calculation of cell characteristic impedance is the same as in p.3.1, i.e. first capacitance per unit length is obtained from inner conductor charge distribution, and next Z_0 is calculated from (3.15) for each segment of the cell's cross section.

3.3.1. Model of an empty cell

In order to check the method described above, first the calculations for NBS cell [6], with dimensions $a=0.75m$, $b=0.75m$, $w=0.619m$ have been done, which enables comparison of results with those presented in [54,55], and with measured values obtained using time domain reflectometer [7].

The segmentation has been done using Tchebycheff polynomials technique in order to achieve higher accuracy with the less number of elements. The segmentation of $1/4$ of the NBS empty cell is shown in Fig. 3.7 (the picture has been obtained from the program that is used to prepare data for matrix S calculation). As one can see, additional section of the cell simulating the load has been added. It is because the simple truncating of the cell at the end of tapered section causes some errors [55]. In order to better simulate 50 ohm load the following algorithm has been used: for the last cross section of the cell the charge distribution corresponding to 50 ohm characteristic impedance is calculated using two-dimensional method. Next, calculated values are used as additional parameters for the (3.7) equation set, which of course is respectively reduced. In this way the 50 ohm characteristic impedance at the load is forced.

The results of modelling are presented in Fig. 3.8. The picture presents characteristic impedance distribution measured (a), calculated using BEM [54,55] (b), and calculated using present method (c).

One can see, that differences in the shape are of course the biggest at the end of the cell. It is probably caused by the difficulties with the load modelling mentioned above. In [54,55] for the load cross section the Neuman condition has been used.

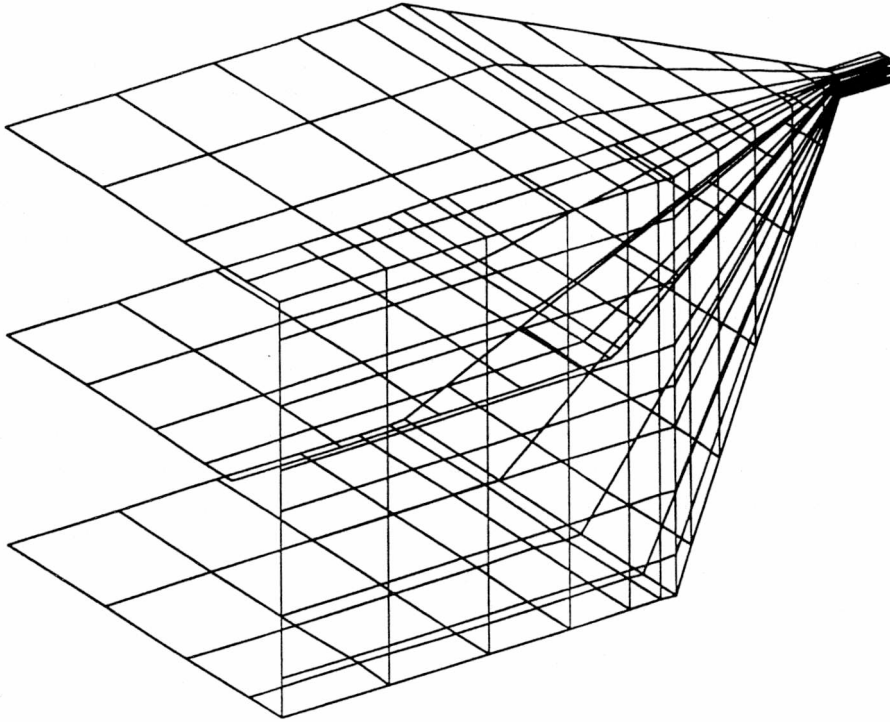


Fig.3.7. Segmentation of the NBS cell (1/4 of the cell volume)

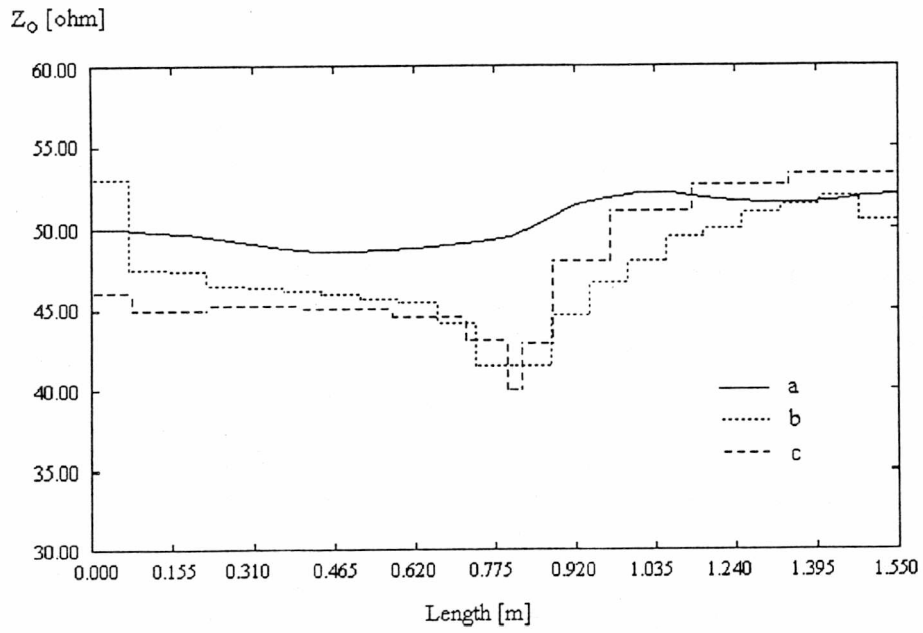


Fig.3.8. Characteristic impedance distribution along the NBS cell
a - measured,
b - boundary element method [55],
c - present method

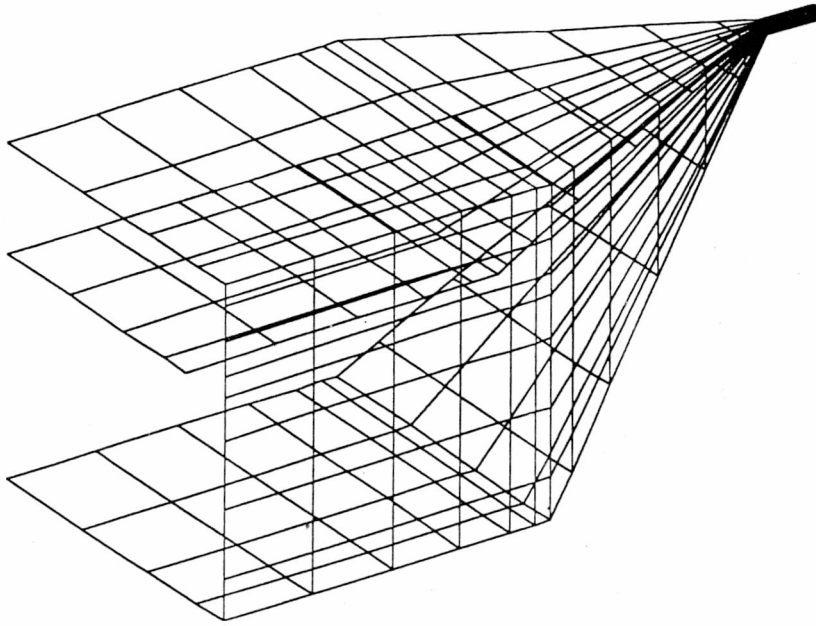


Fig.3.9. Segmentation of the ITA asymmetric cell (1/4 of the volume)

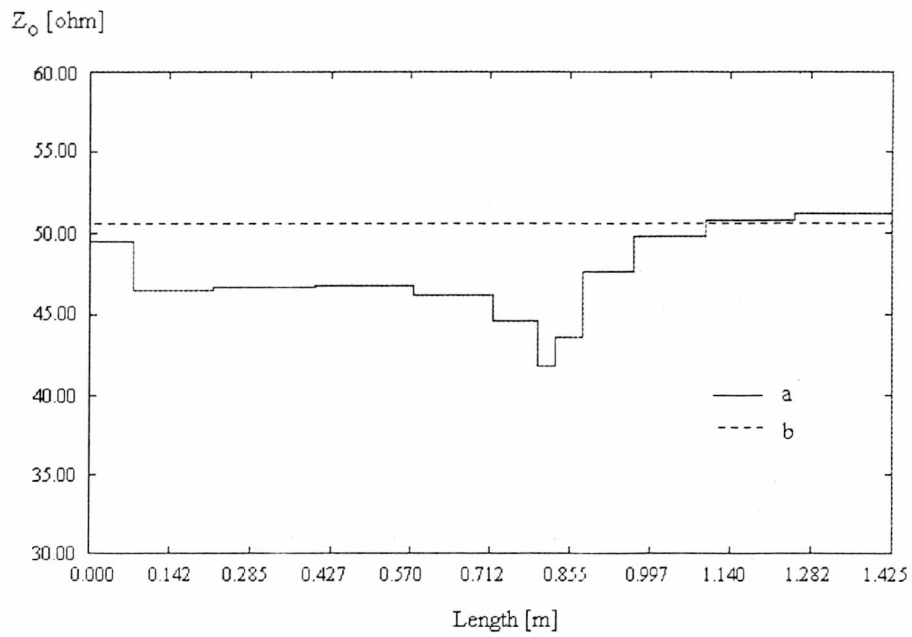


Fig.3.10. Characteristic impedance distribution along the ITA cell (half of the cell)
a - calculated using three-dimensional model,
b - calculated using two-dimensional model

Some difference can be also noticed in the place of transition between the main part of the cell and its tapered section. It is probably because the segmentation used in [55] was not subtle enough in this part of the cell, so the notch was not detected. Similar calculations were done for the ITA cell (see p.3.1). The cell segmentation is shown in Fig. 3.9. The calculated characteristic impedance is presented in Fig. 3.10. A line representing impedance obtained two-dimensional model is also shown (dashed line). One can notice that the values of characteristic impedance near the middle of the cell are almost the same. It means, that for the work volume of the cell two-dimensional model is accurate enough to determine the cross section parameters (potential distribution).

The parameter that is easy to measure is the input impedance of the cell which is function of frequency. For ITA cell input impedance has been measured in the frequency range from 10 kHz to 110 MHz. The same was then calculated treating the cell as the cascade of transmission lines with characteristic impedance calculated for each section, and assuming 50 ohm load at the end of the cell (see Fig. 3.11).

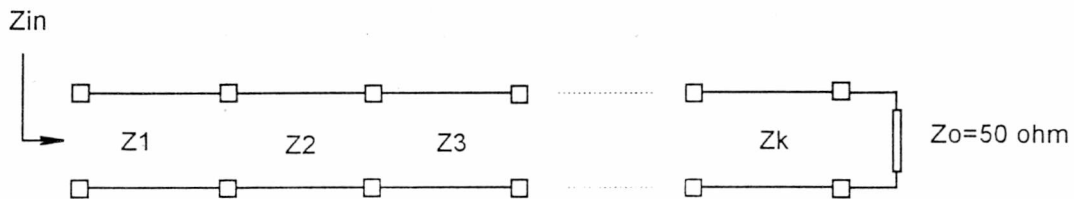


Fig. 3.11. TEM cell as a cascade of transmission lines

In order to calculate input impedance well-known formula [69] for transmission lines was sequentially applied:

$$Z_1 = Z_c \frac{Z_2 + jZ_c \tan \beta l}{Z_c + jZ_2 \tan \beta l} \quad (3.18)$$

where Z_1 - input impedance of the section of transmission line,
 Z_2 - impedance "seen" at the end of the section of transmission line,
 Z_c - characteristic impedance of the section of transmission line,
 l - length of the section of transmission line,

$$\beta = \frac{2\pi}{\lambda} = \frac{2\pi f}{c},$$

c - light velocity in vacuum,

λ - wavelength,

f - frequency.

Results of input impedance calculations are shown in Fig. 3.12. With dotted line the measured values are presented. In this point the agreement of results is not good. It is probably caused by the fact that assumptions about slow changes in potential distribution may not be correct, especially in tapered sections of the TEM cell, which are responsible for resonance effects.

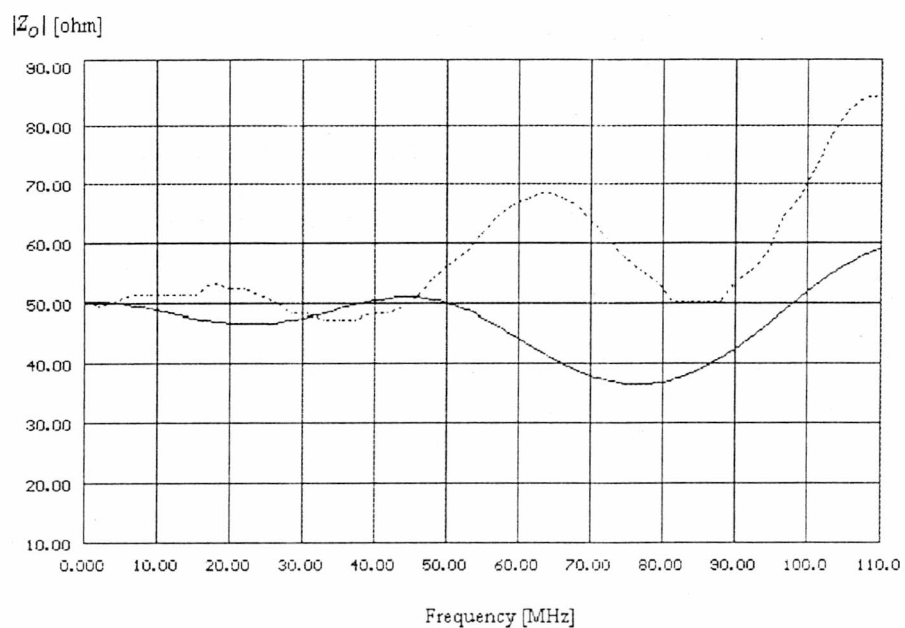


Fig.3.12a. Module of the input impedance of the ITA cell
(dotted line - values measured using TD reflectometer)

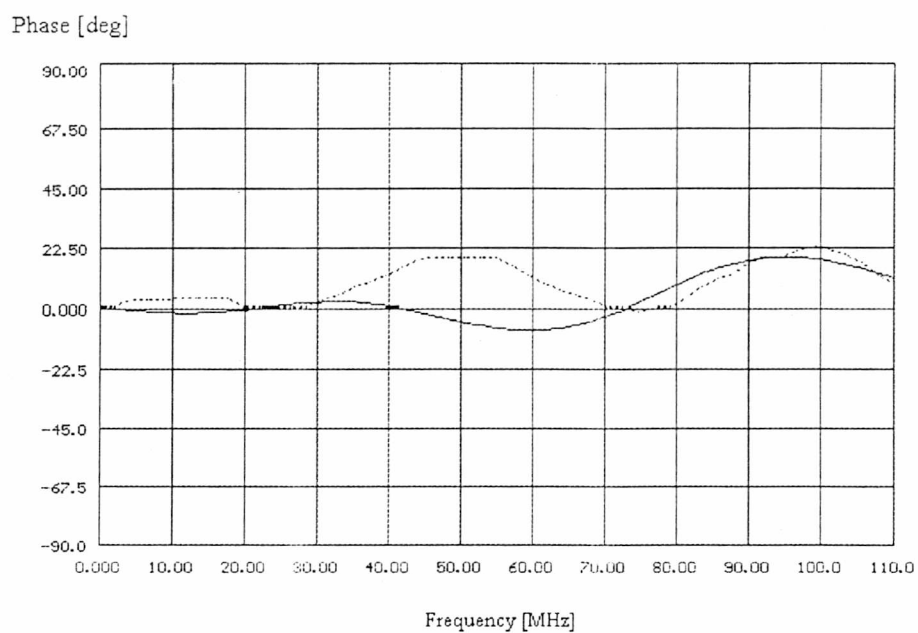


Fig.3.12b. Phase of the input impedance of the ITA cell
(dotted line - values measured using TD reflectometer)

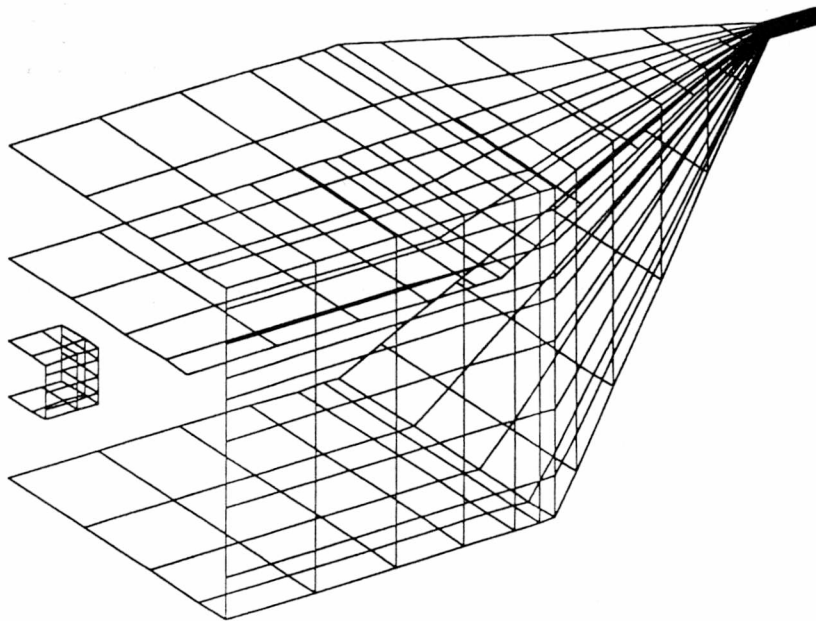


Fig.3.13. Segmentation of the loaded ITA asymmetric cell (1/4 of the volume)

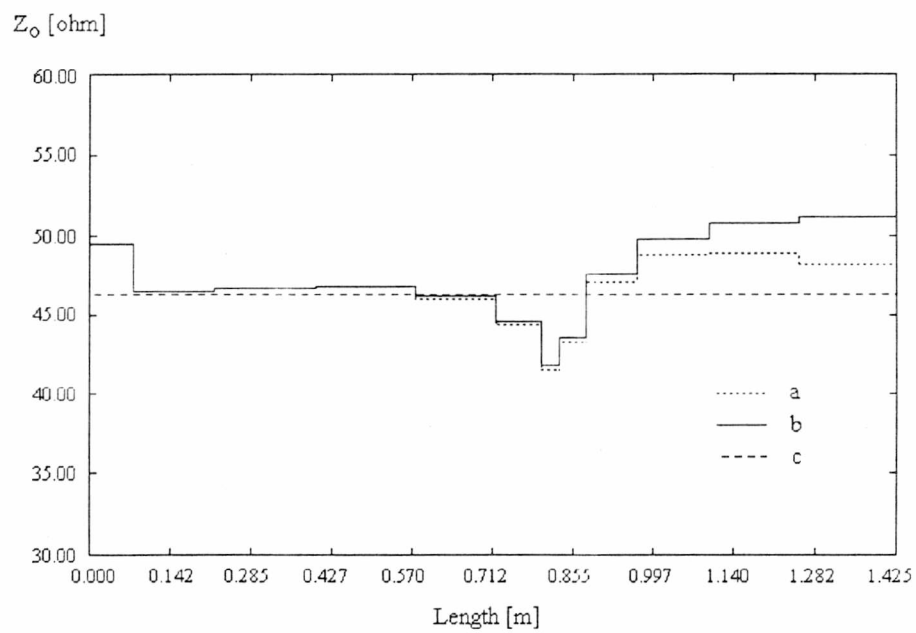


Fig.3.14. Characteristic impedance distribution along the ITA cell (half of the cell)
a - loaded cell, calculated using three-dimensional model,
b - empty cell, calculated using three-dimensional model,
c - loaded cell, calculated using two-dimensional model

3.3.2. Loaded cell

Next, the EUT's influence on characteristic impedance distribution along the cell has been investigated. As an object rectangular perfectly conducting cylinder has been taken with dimensions the same as in two-dimensional model. The EUT has been placed in the middle of the lower part of the cell. The object potential has been assumed to be $0V$.

The segmentation of the loaded ITA cell (1/4 of the cell volume) is shown in Fig. 3.13. The number of divisions is the same like for the empty cell.

Results are presented in Fig. 3.14. The comparison was made between (a) loaded cell (3D model), (b) empty cell (3D model) and (c) loaded cell (2D model) characteristic impedance distribution. It is easy to notice, that the object presence affects the characteristic impedance only in the region close to the EUT. It is also seen, that two-dimensional model gives too pessimistic estimation (the reasons are discussed in detail in the section 3.1).

3.4. GTEM cell project

Techniques of modelling described in Chapters 3.1 and 3.2 have been applied to construct the new TEM cell model (GTEM cell) in ITA. The following assumptions have been specified:

- characteristic impedance - 50Ω ,
- inner conductor (septum) at $3/4$ of the cell height,
- height to width ratio - $2/3$,
- angle between the septum and the floor - 15° ,
- angle between the septum and the ceiling - 5° ,
- working volume dimensions - minimum $0,8m \times 1,0m \times 1,0m$.

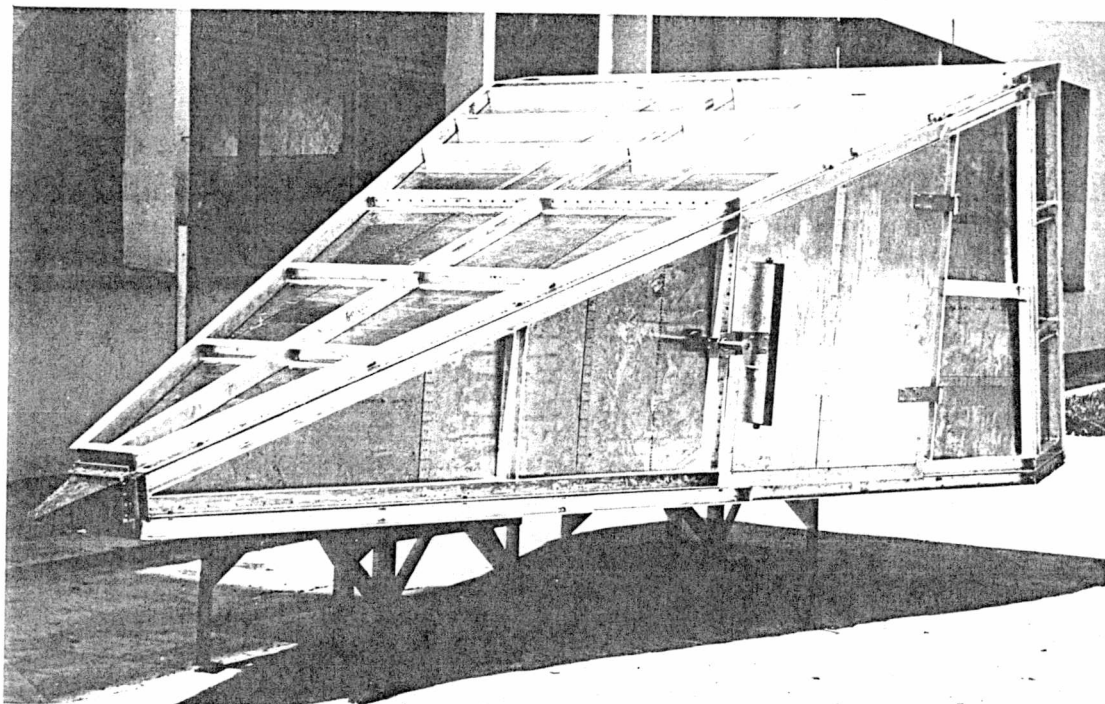


Fig. 3.15. GTEM cell constructed in ITA

The dimension to calculate was the septum width. To do this work the two-dimensional model (p.3.1) was used. The process of calculations was iterative. The calculations showed, that in order to obtain the characteristic impedance equal to 50Ω the width of the septum must be equal to 0.65 of the width of the cell. This result was verified using special curves (see p.2).

The dimensions of the cell assuring desired work volume are about $2 \times 3 \times 5$ m (see Fig. 3.15).

The separate problem was the distributed resistive load. Following factors were taken into account:

- the load must not disturb the current distribution across the cell - the situation "seen" from the cell should be that like in the case of infinite transmission line,
- the resistance distribution along the cell should enable the soft transition between the 50 ohm impedance of the cell and the impedance of the short-cut at the end of the cell.

On the bases of above assumptions the method of the load calculation was developed.

It was assumed, that the resistance distribution in the cross section of the cell should reflect the current distribution in the inner electrode, in the plane of the cross section. It means, that the resistance distribution function should be a reciprocal of the current distribution function. Of course the impedance of each row of resistors should assure the resultant resistance equal to 50 ohms. The current distribution in the cross section of the cell was calculated using the method from p.3.1. The segmentation of the septum used in calculations was adequate to the physical distribution of the resistors on the load plates. It should be noticed, that because of the constructional reasons this distribution was not regular - it was taken into account in the model and related in the fact that the resistance distribution function has characteristic notches in the points of linking the load plates (see Fig. 3.16).

As far as the longitudinal resistance was concerned the exponential model was taken, which is commonly used in the case of waveguides and microwave circuits.

The resultant function of the resistance distribution is presented in Fig. 3.16.

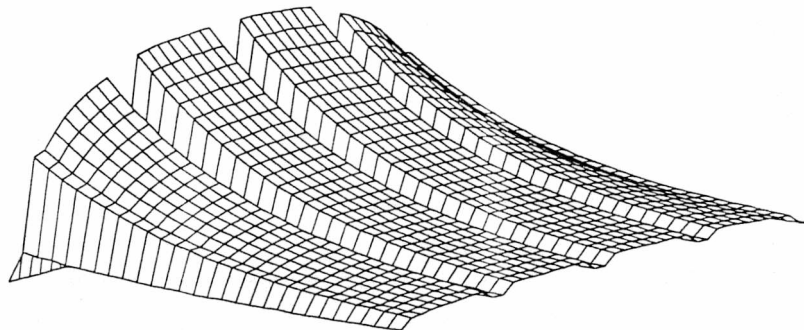


Fig. 3.16. Resistance distribution in the load of GTEM cell
- septum on the left side of the picture.

Theoretical accuracy of obtained resistive load calculation was 0.2%. Because of the necessity of applying resistors from the E12 series and because of the tolerance errors the practical (measured for DC) accuracy obtained was 0.1% (which means the difference 0.05Ω while characteristic impedance equal to 50Ω).

Thanks to the method applied it was possible to obtain optimal resistive load of the GTEM cell for the medium range of frequency, i.e. below the range of absorbing materials influence.

The resistor plates are shown in Fig. 3.17.

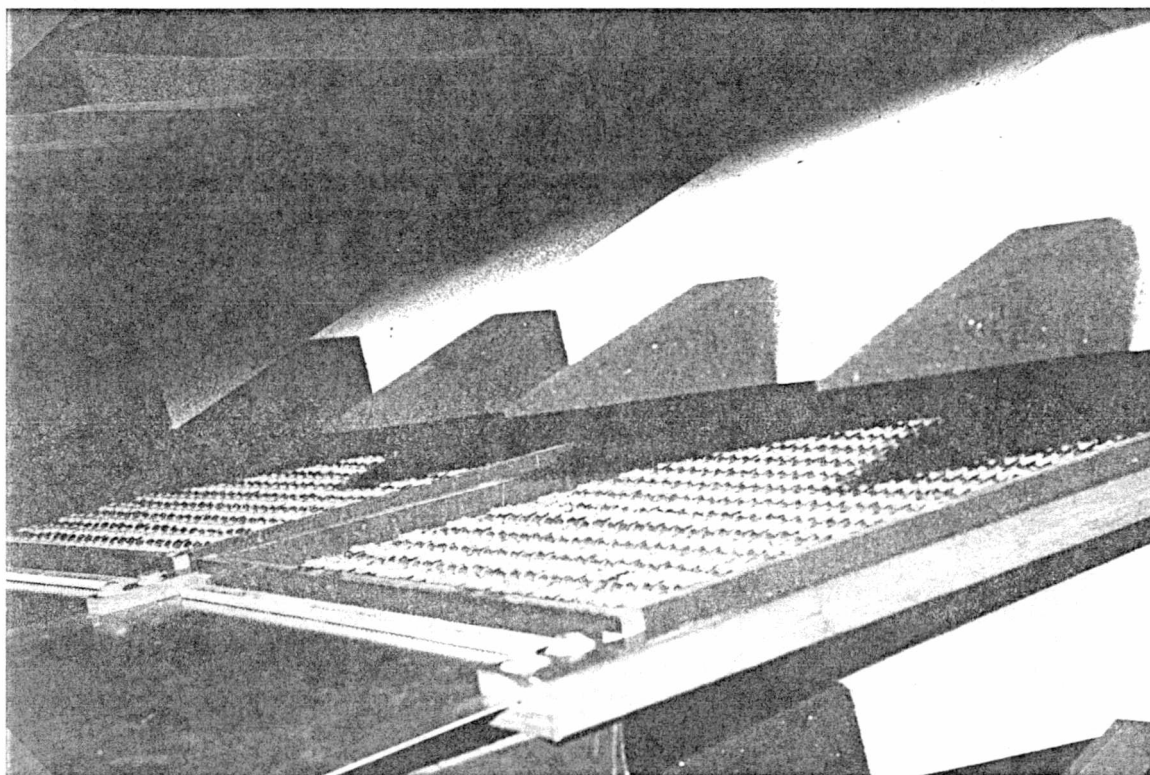


Fig. 3.17. Resistor plates used in the load of GTEM cell

The verification of the project was the measurements of the real GTEM cell build in ITA. The standing wave ratio (SWR) of the cell was measured in the frequency range 300 kHz to 3 GHz. The mean value of the SWR was below 1.5. Such a result allows to state that the GTEM cell designed using numerical techniques presented in Chapter 3.1 and build in ITA represents an international standard in this branch. Let the substantiation of this statement be the fact, that the GTEM cell produced by the world leader, the Electro-Mechanics Company (EMCO), USA, has similar parameters.

In this way the efficiency of the modelling techniques was practically tested.

4. Transmission-Line Matrix method

In Chapter 3 the electromagnetic field model is limited to TEM mode only. In order to obtain full information about fields in a TEM cell in the whole frequency range (i.e. also in the range where higher order modes occur) it is necessary to develop more sophisticated method of electromagnetic field modelling. Up to now a number of different numerical techniques for solving electromagnetic problems are available.

The methods can be divided into two categories: methods that operate in the frequency domain - Finite Element Method, Moment Method, Finite Differences Frequency Domain Method, and methods that operate in the time domain: Finite Differences Time Domain Method and Transmission-Line Matrix Method. The short comparison of above modelling techniques can be found in [24].

In this work the author decided to use a time domain method, because of the following advantages:

- 1) there is a possibility to calculate all cutoff or resonant frequencies in one computational cycle - frequency domain methods usually demand the full computational cycle for all considered frequencies,

- 2) more precise calculations can be done with the use of results of calculations done before,

- 3) using time domain method allows user to find out, which parts of a TEM cell are responsible for electromagnetic waves reflections and which are, in turn, sources of unwanted resonances.

From two mentioned methods the Transmission-Line Matrix method has been chosen. The Transmission-Line Matrix (TLM) method is one of the newest numerical electromagnetic modelling techniques [21]. In terms of its capabilities it is similar to the Finite-Differences-Time-Domain (FDTD) method, but its approach is unique. Like FDTD, analysis is performed in the time domain and the entire region of the analysis is girded. Instead of interleaving E-field and H-field grids, however, a single grid is established and the nodes of the grid are interconnected by virtual transmission lines. Excitations at the source node (or nodes) propagate to adjacent nodes through these transmission lines at each time step.

The advantages of using TLM method are similar to those of the FDTD method. Complex, non-linear materials and objects with complicated shapes are readily modelled. Impulse responses and the time-domain behaviour are determined explicitly. Also, like FDTD, this technique is suitable for implementation on massively parallel machines (the technique of diakoptics [44]).

Of course also typical disadvantages of FDTD method are shared by TLM technique. The main disadvantage is, that voluminous problems that require a fine grid (complicated geometries) need excessive amounts of computation.

In comparison with FDTD method TLM method has one great advantage: the electric and magnetic field components are determined in the same points of space, which is extremely important in resonant problems (determining cutoff and resonant frequencies) as well as in modelling complex boundary geometries. Also infinitely thin conductor plates (common case in theoretical investigations) can be easily modelled using TLM technique, while it is rather difficult with FDTD method.

A disadvantage is, in turn, more complicated treating of complex materials, which leads to greater requirements of computational memory per node.

The bases of the TLM method are in detail described below.

4.1. Discretization of the Huygens's principle

The TLM method bases on the Huygens's principle (it is different attitude than in FDTD method, where Maxwell's field equations are used directly [53]) which states, that a wavefront consists of a number of secondary radiators, which give rise to spherical wavelets. The envelope of these wavelets forms a new wavefront that, in turn, gives a new generation of spherical wavelets, and so on (see Fig. 4.1).

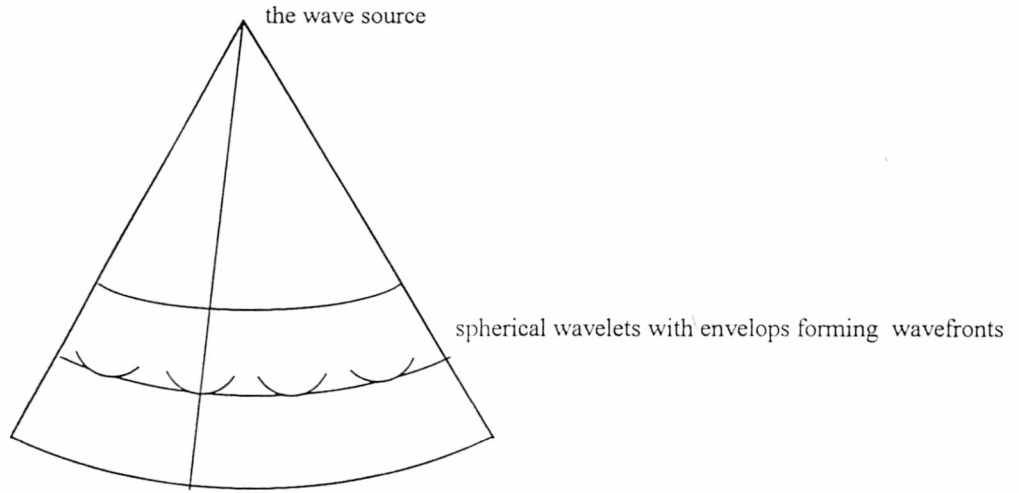


Fig. 4.1. Illustration of the Huygens's principle

In spite of certain difficulties in the mathematical formulation of this mechanism, its application nevertheless leads to an accurate description of wave propagation and scattering.

In order to implement the Huygens's model in a computer program, one must formulate it in a discretized form. This leads to dividing both space and time into finite elementary units Δl and Δt , which are related by the light velocity in vacuum :

$$\Delta t = \frac{\Delta l}{c} \quad (4.1)$$

In this way two- or three-dimensional space is modelled by a Cartesian matrix of points (nodes), separated by the mesh parameter Δl . The unit time Δt can be interpreted as the time required for an electromagnetic pulse to travel from one node to the next.

Let's consider two-dimensional case (the three-dimensional one can be treated as a generalisation of it). Assume that a delta function impulse is incident upon one of the nodes from the negative x -direction. The energy of the pulse is equal to unity. In

accordance the Huygens's principle this energy is scattered isotropically in all four directions, each radiated pulse carrying one fourth of the incident energy. The corresponding field quantities must then be $1/2$ in magnitude. Furthermore, the reflection coefficient "seen" by the incident pulse must be negative in order to satisfy the requirement of field continuity at the node. The situation is illustrated in Fig. 4.2.

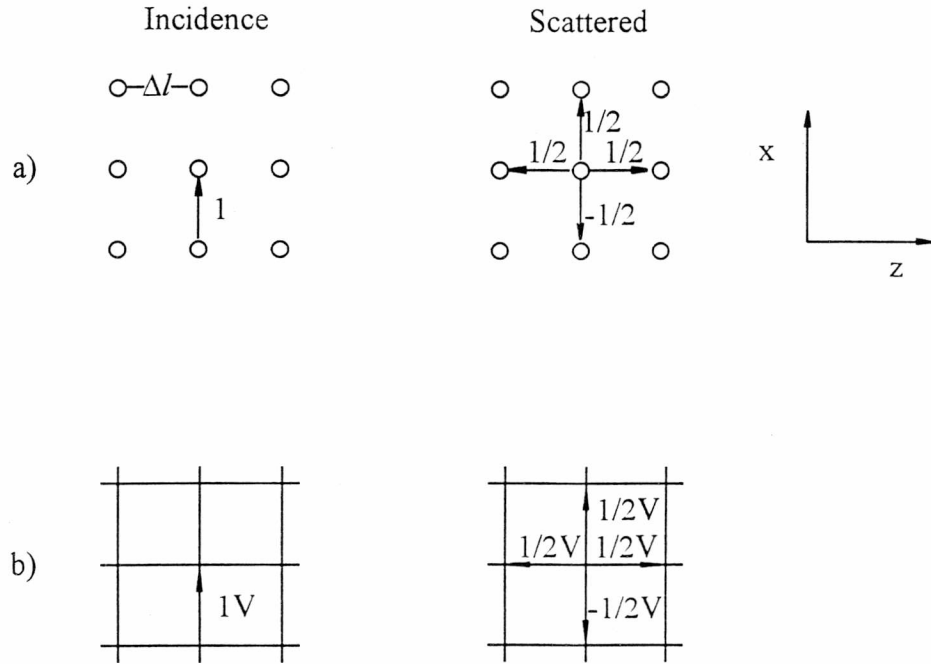


Fig. 4.2. Discretized Huygens's wave model (after Hoefer [21]):
a) in two-dimensional space,
b) in an equivalent Cartesian mesh of transmission lines

This model has a network analogy in the form of a mesh of orthogonal transmission lines, or transmission line matrix, forming a Cartesian array of shunt nodes that have the same scattering properties as the nodes in Fig. 4.2a.

As it will be seen later, there is a direct equivalence between the voltages and currents on the line mesh, and the electric and magnetic fields in Maxwell's equations.

From the mathematical point of view the elementary mesh in the method is represented at each node by a four-element vector (matrix) of numbers describing the magnitude of the incident voltages along the four coordinate directions. The matrix describing the whole TLM network consists of a number of such submatrices, corresponding to the number of nodes in the network. Thus, if the voltage impulses incident on a node at time k are represented by:

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_k^i \quad (4.2)$$

at the time $k+1$ they became reflected pulses:

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_{k+1}^r \quad (4.3)$$

In [21] it is shown, that above vectors are related to each other with the formula:

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_{k+1}^r = \frac{1}{2} \begin{bmatrix} -1 & 1 & 1 & 1 \\ 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \end{bmatrix} \cdot \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \end{bmatrix}_k^i \quad (4.4)$$

The reflected pulses from each node became the incident pulses on neighbouring nodes, which can be denote in the form (taking x, z system of coordinates):

$$\begin{aligned} {}_{k+1}V_1^i(z, x) &= {}_{k+1}V_3^r(z, x-1) \\ {}_{k+1}V_2^i(z, x) &= {}_{k+1}V_4^r(z-1, x) \\ {}_{k+1}V_3^i(z, x) &= {}_{k+1}V_1^r(z, x+1) \\ {}_{k+1}V_4^i(z, x) &= {}_{k+1}V_2^r(z+1, x) \end{aligned} \quad (4.5)$$

The process is repeated on iterative time steps. Each iteration corresponds to a unit of time that is required for pulses to travel from one node to its neighbour. Consequently, if the magnitudes, positions, and directions of all impulses are known at a time step $k\Delta t$, the corresponding values at the time $(k+1)\Delta t$ can be obtained by operating (4.4) and (4.5) at each node of the network. The impulse response of the network is then found by initially fixing the magnitudes, directions and positions of all pulses at $t=0$, and then calculating the state of the network at successive time intervals.

The scattering process described above forms the basic algorithm of the TLM method. Of course, such a network represents only free space situation. In order to take into account conducive or dielectric objects the method must be extended.

The whole algorithm can be described in the following form:

$$\mathbf{V}_{k+1}^i = \mathbf{C} \mathbf{S} \mathbf{V}_k^i + \mathbf{C} \mathbf{V}^s \quad (4.6)$$

where

$\mathbf{C}$ is a connection matrix with elements:

$$c_{ij} = \begin{cases} 1, & \text{if port "i" is connected to port "j"}; \\ 0, & \text{otherwise.} \end{cases}$$

$\mathbf{S}$ is the TLM impulse scattering matrix ,

$\mathbf{V}^s$ is a scattered impulse

$\mathbf{S}$ matrix in (4.6) is the 4 x 4 element matrix from (4.4).

A simple example shown in Fig. 4.3 visualises spreading of the injected energy across the two-dimensional network. The sequence of events resembles the disturbance of a pond due to falling drop of water. Unluckily there is a difference - the discrete nature of the method causes dispersion of the velocity of the wavefront, which means that the velocity of a signal component in the mesh depends on its direction of propagation and on its frequency.

4.2. Wave properties of a TLM network

As it was mentioned at the end of previous point, the discretization used in the method can be a source of some errors. Therefore the properties of the elementary mesh must be carefully investigated.

The basic building block of the two-dimensional TLM network is a node with four sections of transmission lines of length $\Delta/2$. Such a configuration can be approximated by the lumped-element model (Fig. 4.4).

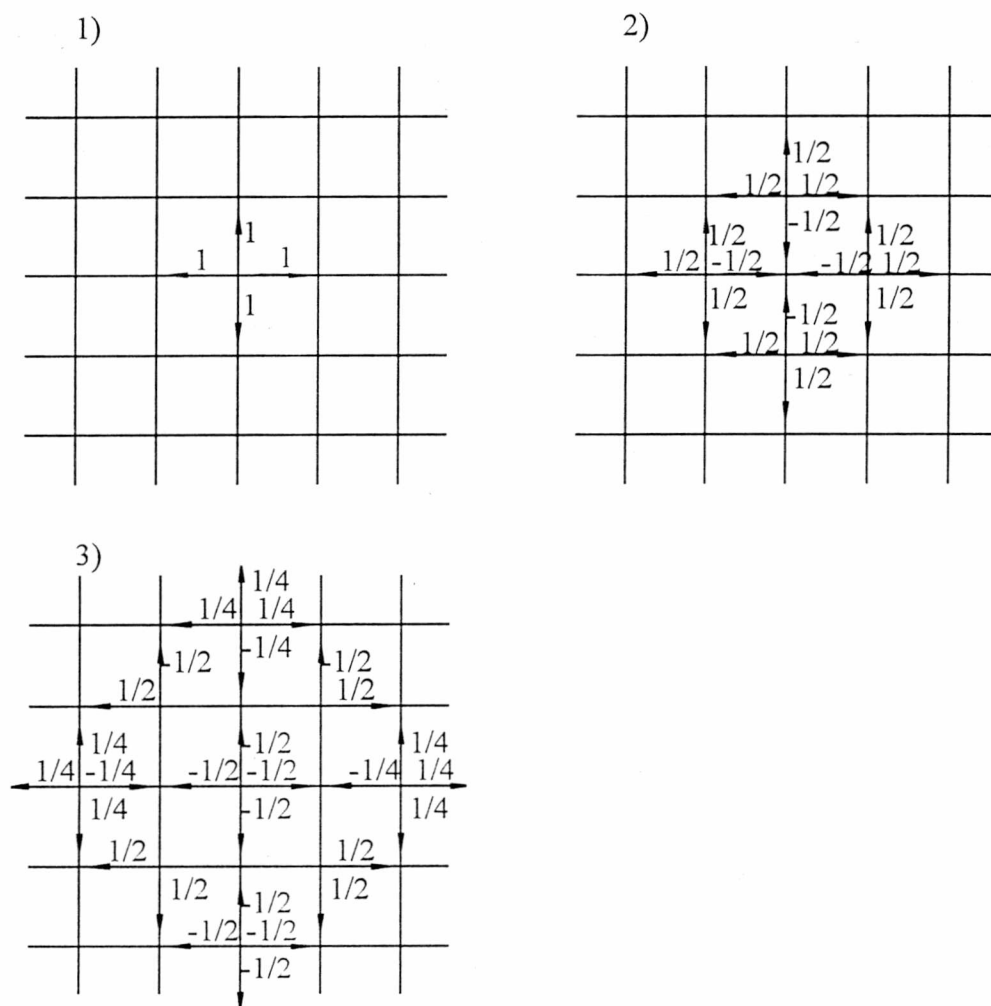


Fig. 4.3. Three first iterations in a TLM network

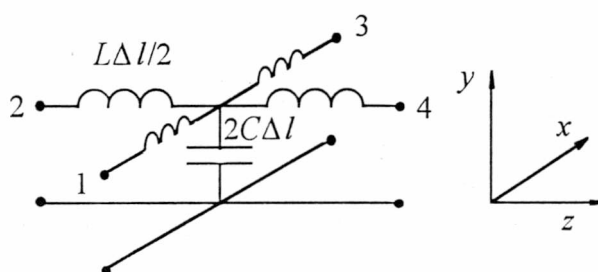


Fig. 4.4. Elementary block in two-dimensional TLM method

Then formulas describing the voltages and currents in such a block can be written as:

$$\begin{aligned}
 \frac{\partial V_y}{\partial x} &= -L \frac{\partial I_x}{\partial t} \\
 \frac{\partial V_y}{\partial z} &= -L \frac{\partial I_z}{\partial t} \\
 \frac{\partial I_x}{\partial x} + \frac{\partial I_z}{\partial z} &= -2C \frac{\partial V_y}{\partial t}
 \end{aligned} \tag{4.7}$$

The Maxwell's equations for $\frac{\partial}{\partial y} = 0$ take form (one of two possible sets):

$$\begin{aligned}
 \frac{\partial E_y}{\partial x} &= -\mu \frac{\partial H_z}{\partial t} \\
 \frac{\partial E_y}{\partial z} &= -\mu \frac{\partial H_x}{\partial t} \\
 \frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} &= \varepsilon \frac{\partial E_y}{\partial t}
 \end{aligned} \tag{4.8}$$

One can notice that we can take equivalence:

$$\begin{aligned}
 \varepsilon &\equiv 2C \\
 \mu &\equiv L \\
 E_y &\equiv V_y \\
 -H_z &\equiv \Delta I_x = I_3 - I_1 \\
 -H_x &\equiv \Delta I_z = I_2 - I_4
 \end{aligned} \tag{4.9}$$

For elementary transmission lines in the TLM network and for $\mu_r = \varepsilon_r = 1$, the inductance and capacitance per unit length are related by:

$$\frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\epsilon_0 \mu_0}} = c \quad (4.10)$$

Hence, assuming that voltage and current waves of each component of the transmission line travel at the speed of light, the complete network of intersecting transmission lines represents a medium of relative permittivity twice that of free space. That means, that as long as the equivalent circuit in Fig. 4.4 is valid, the propagation velocity in the TLM mesh is $1/\sqrt{2}$ the velocity of light.

At this point it should be noticed, that the (4.9) equivalence will be also valid when E and H fields (ϵ_r and μ_r) will be exchanged. It is related to the fact that electric and magnetic fields are dual in the Maxwell's equations. In fact, choosing proper set of field quantities in TLM method depends on choosing boundary conditions.

Considering the mesh as a periodic structure Johns [28] obtained the following dispersion relation for propagation along the main mesh axes:

$$\sin\left(\beta_n \frac{\Delta l}{2}\right) = \sqrt{2} \sin\left(\omega \frac{\Delta l}{2c}\right) \quad (4.11)$$

where β_n is the propagation constant in the network.

The resulting ratio of velocities on the matrix and in free space $v_n / c = \omega / (\beta_n c)$ is shown in Fig. 4.5.

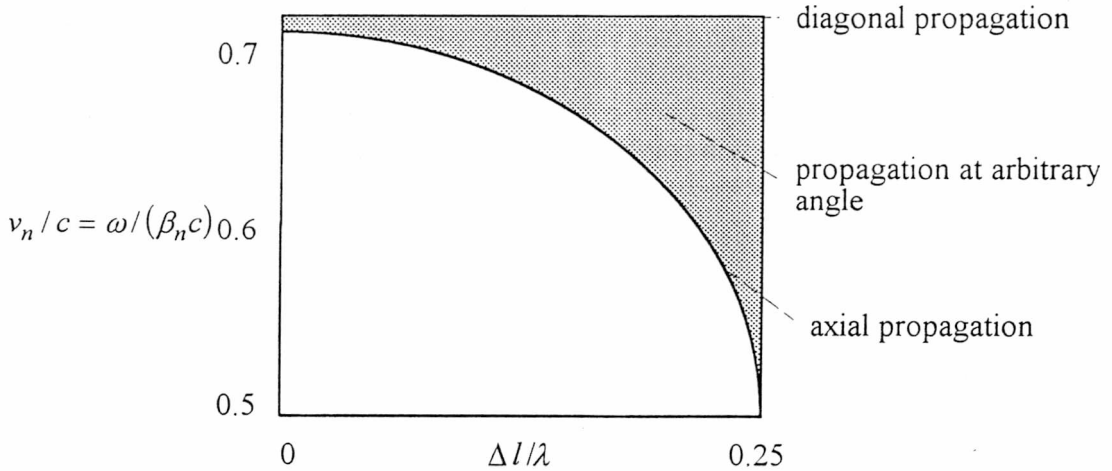


Fig. 4.5. Ratio of propagation velocities in TLM network and in free space

It appears, that a first cutoff occurs for $\Delta l / \lambda = 0.25$ (λ is the free space wavelength). However no cutoff occurs in the diagonal direction, where velocity is

frequency independent. In intermediate directions the velocity ratio lies somewhere between the two curves (shaded area).

In conclusion, the TLM network simulates an isotropic propagation medium only as long as all frequencies are well below the network cutoff frequency, in which case the network propagation velocity may be considered constant and equal to $c/\sqrt{2}$.

Thus, the Δl mesh parameter must be chosen not only to allow modelling of small elements of the object geometry, but also to make it possible to make simulations in the desired frequency range.

Another type of error that occurs in the TLM network is caused by the fact that the pulse response of the structure must be truncated in time. Of course this error decreases while increasing the number of iterations. Also some windowing techniques can be useful in minimising the error level (Hanning profile - see Chapter 5).

4.3. Representation of lossless and lossy boundaries

Electric and magnetic walls are represented by short and open circuits, respectively, put at appropriate positions in the TLM mesh. In order to ensure synchronism such circuits must be placed halfway between proper nodes. It means that electric and magnetic walls cannot be placed at every position. In practice the position of boundaries relatively to TLM network is achieved by setting Δl parameter as a proper structure dimension divided by an integer number. In the computation open or short circuit simulation is achieved by returning the pulse after one Δt period with equal or opposite sign. The new works published by Hoefer et al. [4,11,48] point out that there is also a possibility to place boundaries directly in the nodes of TLM network.

Curved walls must be of course represented by a piecewise straight boundaries, which is the source of the third type of error in TLM network (sometimes called "coarseness error"). To avoid problems with this type of error the Δl parameter must still be kept very small. This may be rather impractical because of the increasing rapidly size of matrices, another solution is to apply nonuniform mesh (see p.4.6).

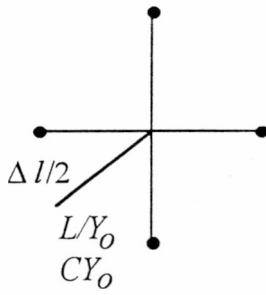
Lossy boundaries can be modelled in the same way as lossless ones, but with applying proper reflection coefficient. For good but imperfect conductor of conductivity σ reflection coefficient can be written as [21]:

$$\rho \cong -1 + 2\sqrt{\frac{\varepsilon_0 \omega}{2\sigma}} \quad (4.12)$$

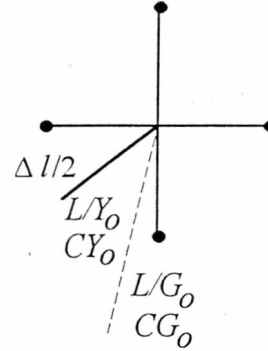
Because the coefficient is a function of frequency, the value of it is accurate only for frequency that has been selected while determining ρ .

4.4. Representation of dielectric and magnetic materials

Dielectric and magnetic materials can be taken into account in TLM method by loading inside nodes with reactive stubs of appropriate characteristic impedance and a length equal to $\Delta l/2$ (see Fig. 4.6).



Permittivity stub



Permittivity stub and loss stub

Fig. 4.6. TLM nodes with stubs modelling dielectric or magnetic material

From the analysis presented in [21] it is known, that the characteristic admittance of stubs (normalised to the admittance of network lines) is related to ϵ_r with the formula:

$$\epsilon_r = 2 \left(1 + \frac{Y_0}{4} \right) \quad (4.13)$$

It is true under assumption, that the voltage in the network represents the electric field. When the voltage represents the magnetic field, open shunt stubs describe a permeability. Unluckily the latter situation causes some troubles, because the interface conditions are not satisfied and a special correction must be introduced in the form of local reflection and transmission coefficients at the interface between the different media.

Dielectric losses can be modelled in the TLM method in to different ways:

- by applying lossy transmission lines (reduction of the magnitude of each pulse travelling through the "lossy node",

- by loading nodes of a lossless mesh with "loss stubs" - each node is loaded with a matched (no reflected impulse) transmission line of characteristic admittance G_0 extracting energy from each node at every iteration.

The second way seems to be better because interface conditions have not to be changed. For conductivity σ :

$$G_0 = \sigma \Delta l \sqrt{\frac{\mu_0}{\epsilon_0}} \quad (4.14)$$

The generalised set of equations (4.4) takes a form:

$$\begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \\ V_5 \end{bmatrix}_{k+1} = \frac{1}{(Y_0 + G_0 + 4)} \begin{bmatrix} -(Y_0 + G_0 + 2) & 2 & 2 & 2 & 2Y_0 \\ 2 & -(Y_0 + G_0 + 2) & 2 & 2 & 2Y_0 \\ 2 & 2 & -(Y_0 + G_0 + 2) & 2 & 2Y_0 \\ 2 & 2 & 2 & -(Y_0 + G_0 + 2) & 2Y_0 \\ 2 & 2 & 2 & 2 & (Y_0 - G_0 - 4) \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ V_3 \\ V_4 \\ V_5 \end{bmatrix}_k \quad (4.15)$$

V_5 voltage is a voltage of reactive stub (Fig. 4.6). The voltage of "loss stub" does not have to be taken into account, because, as it was stated before, there is no reflection pulse from this stub. In order to make the algorithm finished the (4.5) set of equations must be completed with equation concerning V_5 :

$${}_{k+1}V_5^i(x, z) = {}_{k+1}V_5^r(x, z) \quad (4.16)$$

4.5. Representation of perfectly absorbing walls

The local TLM network reflection coefficient equal to zero is not equivalent to simulation of a perfectly absorbing (non-reflective) wall. As it was said in [11,48] better approximation is to model situation, when transmission lines of the network are truncated at the absorbing boundary with impedance equal to characteristic impedance of the line divided by $\sqrt{2}$.

In fact, the local reflection coefficients in the situation when no reflection occurs are dependent on the time step. In the case of pulse excitation, the series of reflection voltages forms a pulse response. When the "margin" of nodes on the other side of the wall is big enough (it means that the number of additional node slices is equal at least to the length of the pulse response divided by two) the wall can be treated as perfectly absorbing. The length of the pulse response is the number of time steps for which the level of pulses is higher than some reference level. Taking lower reference level results

in better modelling of the absorbing wall. Number of slices equal to half the length of the pulse response results from the fact, that the unwanted reflected wave must travel through the margin twice: from the simulated wall to the end of the margin and back.

On the bases of the above idea Hoefer proposed in [48] another solution - the "margin" can be replaced by computing a convolution of real voltages values with obtained before pulse response of the non-reflective wall written in a matrix form (Johns matrix [11,48]). Of course the convolution must be done repetitively for each step of the calculations taking into account proper number of travelling pulses (the same as the length of the pulse response). In this solution also windowing techniques can be used to obtain "soft" cut of the pulse response.

The Hoefer method is the only method to obtain full perfect absorbing wall simulation, but requires much time- and memory consuming computation, especially when absorbing wall are big, and have complicated shapes. Therefore in the simple cases also taking simple imprecise methods can be used as a first approximation.

4.6. Nonuniform TLM grid

The serious lack of the standard TLM method is that the elementary block size must be small enough to model the smallest part of the structure and to model curved walls well. Sometimes it leads to a great number of blocks in regions, where so subtle modelling is not necessary, and of course makes computation much longer.

The solution is applying a nonuniform grid [21]: in the regions with small details the grid dimension is small, and in the rest of the structure - bigger. The situation is similar to that known for example from FEM method. In TLM method however the changes in the grid dimensions are strictly limited - the ratio of the grid dimensions must be an integer number to ensure synchronism of pulses in the whole network.

Let's denote as p the ratio of grid dimensions. To keep the velocity of travelling impulses the same in all branches, the inductivity per unit length of the longer mesh lines must be increased by a factor p while their capacity per unit length must be reduced by $1/p$. This in turn increases their characteristic impedance by a factor p , and the local scattering matrix S of nodes connecting cells with different size must be modified accordingly.

To preserve synchronism, impulses travelling in longer branches are kept in store for p iterations before being reinjected at the next node.

The experiment with the nonuniform network applied to TEM cell cross section pointed out, that in this case better results are obtained while using uniform mesh with much greater number of iterations (see Chapter 5).

4.7. Three-dimensional TLM method

The only interesting method of extending the standard two-dimensional TLM method for the third dimension has been proposed by Johns [29]. Other solutions [21] do not preserve one of the main advantages of the TLM method over the other methods (like FDTD) - the ability to obtain all six field quantities at one point of the space.

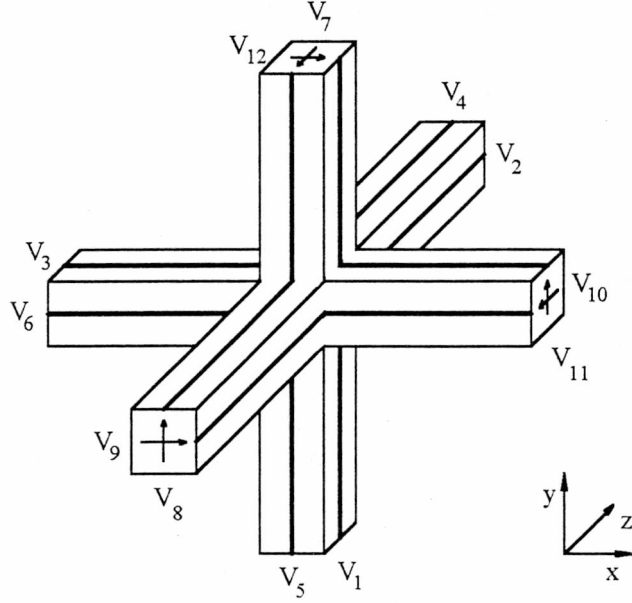


Fig. 4.7. Symmetrical TLM node

The elementary TLM node described by Johns is presented in Fig. 4.7. It consists of the set of twelve transmission lines. The propagation of pulses is described by node matrix $\mathbf{S}$ (as before), which in this case has a dimension 12 x 12 elements:

$$\mathbf{S} = \frac{1}{2} \begin{bmatrix} 0 & 1 & 1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & -1 & 0 \\ 1 & 0 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & -1 & 0 & 1 \\ 1 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & -1 \\ 0 & 0 & 1 & 0 & 1 & 0 & -1 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 & 0 & 1 & 0 & -1 & 0 & 1 & 0 & 0 \\ 0 & 1 & 0 & 0 & 1 & 0 & 1 & 0 & -1 & 0 & 0 & 0 \\ 0 & 0 & 0 & -1 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 & -1 & 0 & 1 & 0 & 0 & 0 & 1 & 0 \\ 1 & 0 & 0 & 0 & 0 & -1 & 0 & 0 & 0 & 1 & 0 & 1 \\ 0 & -1 & 0 & 0 & 1 & 0 & 1 & 0 & 1 & 0 & 0 & 0 \\ -1 & 0 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 & 0 & 1 \\ 0 & 1 & -1 & 0 & 0 & 0 & 0 & 0 & 1 & 0 & 1 & 0 \end{bmatrix} \quad (4.17)$$

The rules of modelling are the same as for the two-dimensional TLM method. The propagation velocity in the TLM network is now $c/2$ (assuming that pulses in the network travel with the speed of light c). The analysis of propagation properties of the three-dimensional TLM network is till now unavailable. The only rule for the users is to keep the Δl parameter small enough to avoid problems with dispersion for desired frequency range.

The relations between voltages at the node and the electric and magnetic fields are as follows:

$$\begin{aligned}
 E_x &= \frac{1}{2\Delta l} (V_1^i + V_2^i + V_9^i + V_{12}^i) \\
 E_y &= \frac{1}{2\Delta l} (V_3^i + V_4^i + V_8^i + V_{11}^i) \\
 E_z &= \frac{1}{2\Delta l} (V_5^i + V_6^i + V_7^i + V_{10}^i) \\
 H_x &= \frac{1}{2Z_0\Delta l} (V_4^i - V_5^i + V_7^i - V_8^i) \\
 H_y &= \frac{1}{2Z_0\Delta l} (-V_2^i + V_6^i + V_9^i - V_{10}^i) \\
 H_z &= \frac{1}{2Z_0\Delta l} (-V_3^i + V_1^i + V_{11}^i - V_{12}^i)
 \end{aligned} \tag{4.18}$$

For excitation points voltages for given fields values are given in the form:

$$\begin{aligned}
 V_1^i &= \Delta l (E_x + Z_0 H_z) / 2 \\
 V_2^i &= \Delta l (E_x - Z_0 H_y) / 2 \\
 V_3^i &= \Delta l (E_y - Z_0 H_z) / 2 \\
 V_4^i &= \Delta l (E_y + Z_0 H_x) / 2 \\
 V_5^i &= \Delta l (E_z - Z_0 H_x) / 2 \\
 V_6^i &= \Delta l (E_z + Z_0 H_y) / 2
 \end{aligned} \tag{4.19}$$

$$V_7^i = \Delta l (E_z + Z_0 H_x) / 2$$

$$V_8^i = \Delta l (E_y - Z_0 H_x) / 2$$

$$V_9^i = \Delta l (E_x + Z_0 H_y) / 2$$

$$V_{10}^i = \Delta l (E_z - Z_0 H_y) / 2$$

$$V_{11}^i = \Delta l (E_y + Z_0 H_z) / 2$$

$$V_{12}^i = \Delta l (E_x - Z_0 H_z) / 2$$

Z_0 is the free space wave impedance.

Johns has proposed also a solution for the case of modelling dielectric and magnetic materials. The TLM node is in this case completed with six stubs [29]. Corresponding **S** matrix has in this case 18 x 18 parameters, and the elementary cell in the space is described using 18 voltages. Details concerning this case are described in [29]. In this work only nodes without stubs (12 voltages per node) have been used because of computer memory limitations.

5. Higher order modes in TEM cells

As it has been said in Chapter 1, one of the limitations of the TEM cells is the presence of higher order modes. Those higher order modes destroy the regularity of the field distribution inside the cell and cause the unwanted resonances which leads to limitation of the useful frequency range. It is well known from the theory of microwave waveguides, that the presence of higher order modes, and their magnitude depends on the way of exciting fields inside, as well as of the irregularities of the transmission line (or waveguide) structure.

The problem arises when the line (in this case the TEM cell) is not empty. From this point of view any object (EUT) inside the cell can be treated as the irregularity of the line and therefore be the reason of exciting higher order modes.

Of course problems connected to it are much greater when using standard (Crawford) TEM cells, but in the case of the GTEM cells are also important.

In the case of Crawford cells it is a common habit to treat the cutoff frequency of the cell as the frequency limiting cell bandwidth [6,7]. From the other hand, the knowledge about cutoff frequencies of higher modes is also important. Some authors suggest using TEM cells between resonances making use of the fact, that resonances in Crawford cells are rather sharp.

Therefore the ability of calculation of cutoff frequencies of the higher order modes as well as the field distribution for these modes is of a great importance. In this work the Transmission-Line Matrix method has been applied to obtain both cutoff frequencies and the field distribution of higher order modes TEM cells.

5.1. Cutoff frequencies

It is known, that the cutoff frequency of a structure depend on its cross section. In this case the structure is treated as infinitely long. Under such condition the two-dimensional analysis is allowed. Of course assuming two-dimensional analysis leads to taking one of two possible sets of field quantities: E_x, E_y, H_z or H_x, H_y, E_z (when the structure infinite dimension lies along z -axis). In fact it is the result of putting into the set of Maxwell equations the z derivative equal to zero. So, the first of the above sets is taken when TE modes are investigated, the second one - for TM modes. From computational point of view both situations are dual. The difference occurs when putting the boundary conditions. For TE modes simulating a perfectly conducting wall is equivalent to open-circuiting the transmission lines on metal surfaces, for TM modes - to short-circuiting them. Numerically open- and short-circuiting is obtained by the simulation of a voltage reflection between nodes. For short-circuit the reflection coefficient is assumed to be equal -1, for open +1. Of course such attitude means, that all metal surfaces must lie in the middle between nodes, which demands a subtle segmentation in the case when the shape of the structure is complicated. In the case of the empty TEM cell, or the TEM cell loaded with a rectangular object, the segmentation is rather easy (see Fig. 5.1). If the shape of the EUT is more complicated it may be necessary to apply TLM method with non-uniform grid. In order to check the method both possibilities were tested.

It can be proved [27], that cutoff frequencies of the structure are the resonant frequencies of the two-dimensional model applied to structure cross section. The procedure enabling obtaining those resonant frequencies is with the TLM method very

simple. It concerns finding the pulse response of the structure, and then, after the Fourier transform, finding resonant frequencies.

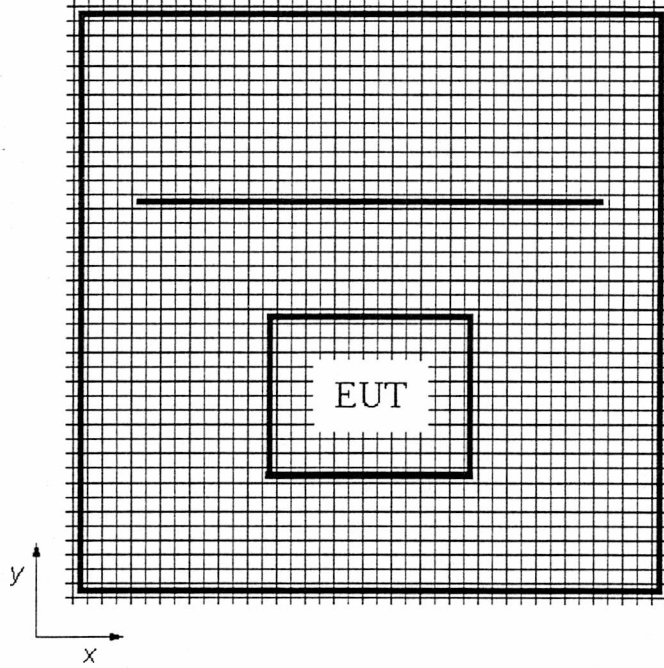


Fig. 5.1. Cross section of a loaded TEM cell over the TLM grid

The numerical procedure starts with the pulse excitation of a selected node inside the structure. Next, the propagation of pulses is calculated in discrete time steps, following the rules of the TLM method (see Chapter 4). Then the time-domain pulse response is obtained in another node of the grid. The choice of the excitation and response nodes should be done with much care - inadequate location results with the lack of information about some events inside the structure. Unluckily there is no simple way to determine those locations. In this work the nodes were chosen experimentally, applying several trials with different locations. The obtained pulse response is then transformed into frequency domain using the fast Fourier transform (FFT). Problems arise with the aliasing effects which are the result of cutting the pulse response after a number of steps, while real pulse is infinitely long. Therefore some windowing techniques must be used in order to clear the situation. It was proved [5,21] that using the Hanning profile is the best solution for electromagnetic scattering problems. Then the pulse transformed with the FFT takes the form:

$$p'(i) = \left(1 + \cos\left(\frac{i \cdot \pi}{N}\right) \right) \cdot p(i), \quad i = 1, \dots, N \quad (5.1)$$

where: $p(i)$ - original pulse response obtained from the model,
 $p'(i)$ - recalculated pulse response (after applying the Hanning profile),
 N - number of iterations.

In all calculations in this section the number of iterations equal to 4096 was used. Such number allows to obtain good accuracy of results, and minimise numerical errors (see Chapter 4).

First, in order to prove usefulness of the method, calculations were done for rectangular waveguides, for which simple analytical solutions are available. Results of modelling for 1.2m x 0.9m waveguide are presented in table 5.1.

Table 5.1. Cutoff frequencies of a rectangular waveguide

TE mode	Cutoff frequency [MHz]		Error [%]
	analytical	numerical	
1 0	125	124.3	0.6
0 1	167	165.7	0.8
1 1	208	207.2	0.4
2 0	250	248.6	0.6
2 1	300	303.8	1.3
0 2	333	331.5	0.5
1 2	356	352.2	1.1
3 0	375	372.9	0.6
2 2	417	414.3	0.9

As one can see, maximum error is 1.3%. The accuracy of calculations may be better when longer pulse response is taken into account.

Next, the cutoff calculations have been done for the TEM cell constructed in the Institute of Telecommunications and Acoustics (ITA) of the Technical University of Wroclaw. The results for TE modes have been obtained in two ways: using TLM method with uniform grid, and using approximate equations given by P.F. Wilson [63]. The use of the non-uniform grid is in this case not adequate, because of the simple geometry of the problem. The author, however, has made an experiment with non-uniform grid, too. The experiment has shown, that computation using non-uniform TLM mesh requires much longer time, than in the case of uniform grid, without making results more precise. The conclusion is, that the non-uniform TLM grid should be applied in cases, when the main limitation is computer memory.

Results of calculations for first three modes are presented in Table 5.2.

Table 5.2. Cutoff frequencies of the empty cell

TE mode	Cutoff frequency [MHz]		Error [%]
	TLM	[63]	
01	75.51	74.3	1.4
10	124.30	125.0	0.6
11	161.86	160.1	1.1

It is worth to notice that for the other TE modes and for TM modes approximate analytical formulas for asymmetric TEM cells have not been published yet, so numerical methods are the only way to obtain their cutoff frequencies.

5.2. Field distribution

The distribution of the fields of higher modes in the cross section of the cell can be obtained by exciting the structure with sinusoidal excitation, and then after achieving

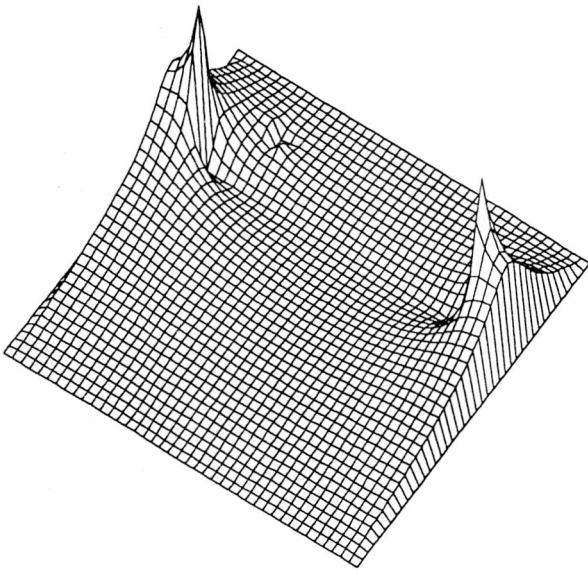


Fig. 5.2a. Field distribution at 74.3 MHz
- E_x component

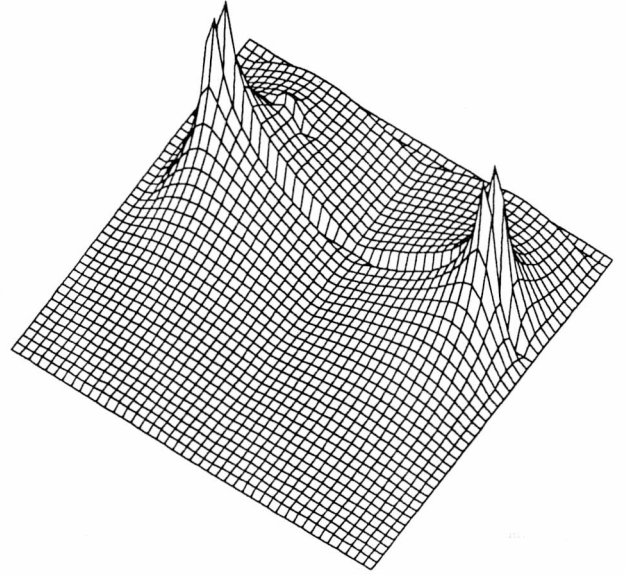


Fig. 5.2b. Field distribution at 74.3 MHz
- E_y component

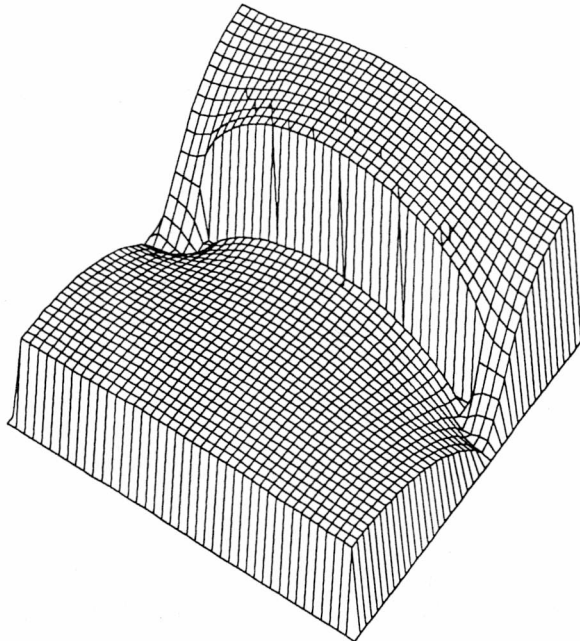


Fig. 5.2c. Field distribution at 74.3 MHz
- H_z component

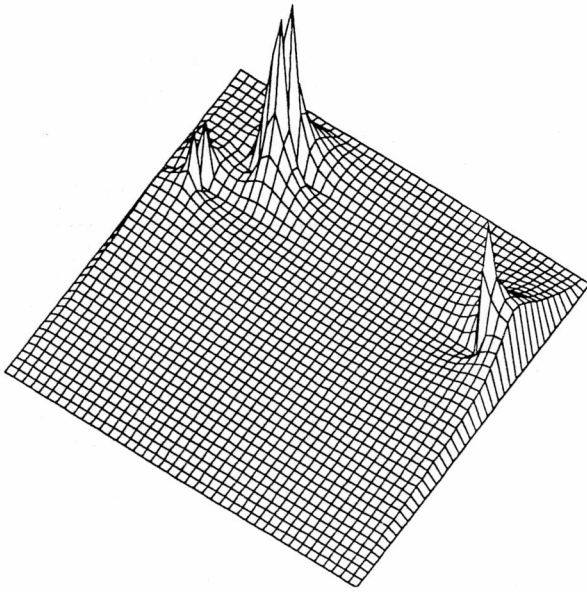


Fig. 5.3a. Field distribution at 125 MHz
- E_x component

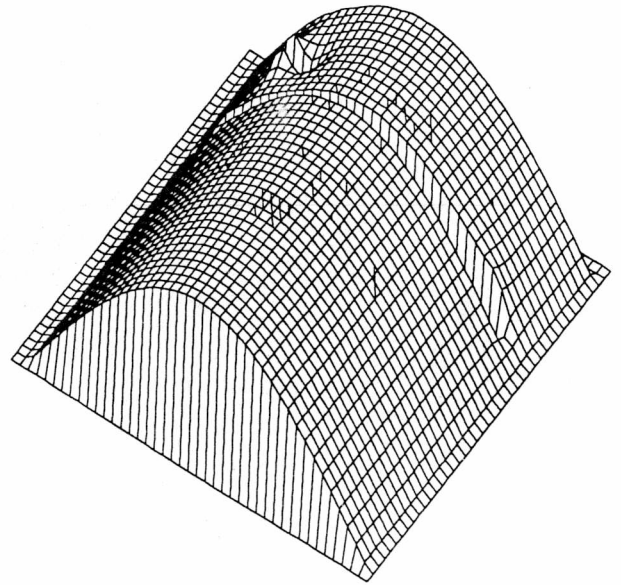


Fig. 5.3b. Field distribution at 125 MHz
- E_y component

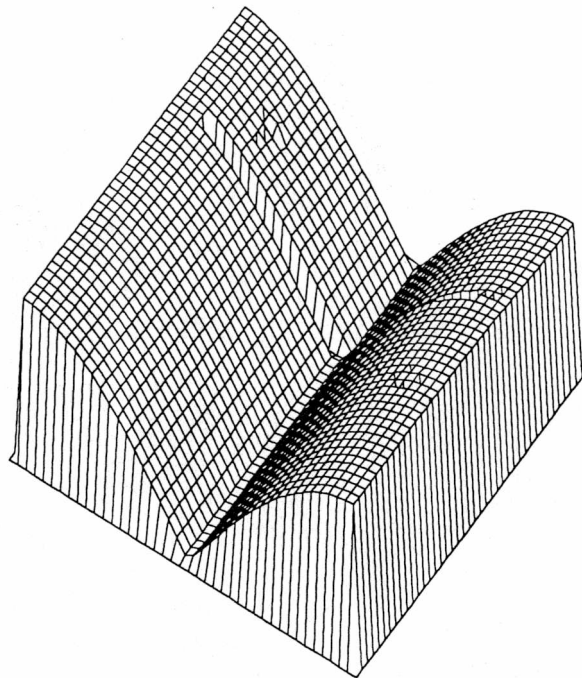


Fig. 5.3c. Field distribution at 125 MHz
- H_z component

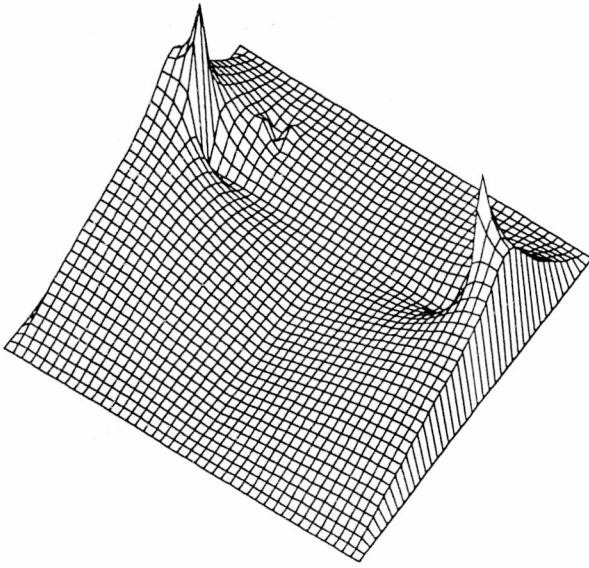


Fig. 5.4a. Field distribution at 160 MHz
- E_x component

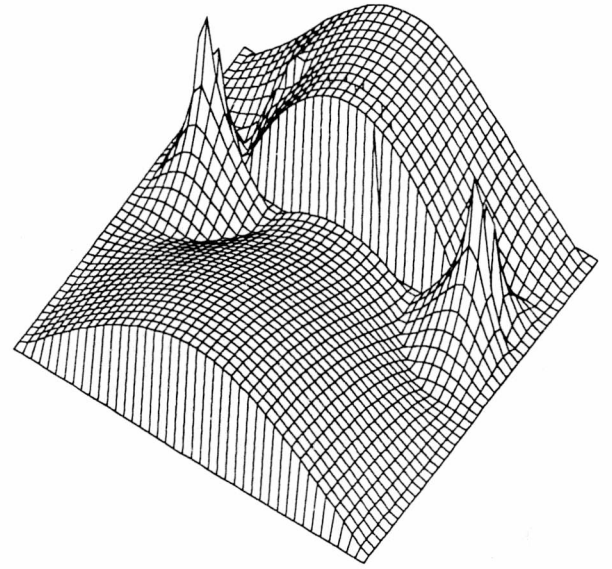


Fig. 5.4b. Field distribution at 160 MHz
- E_y component

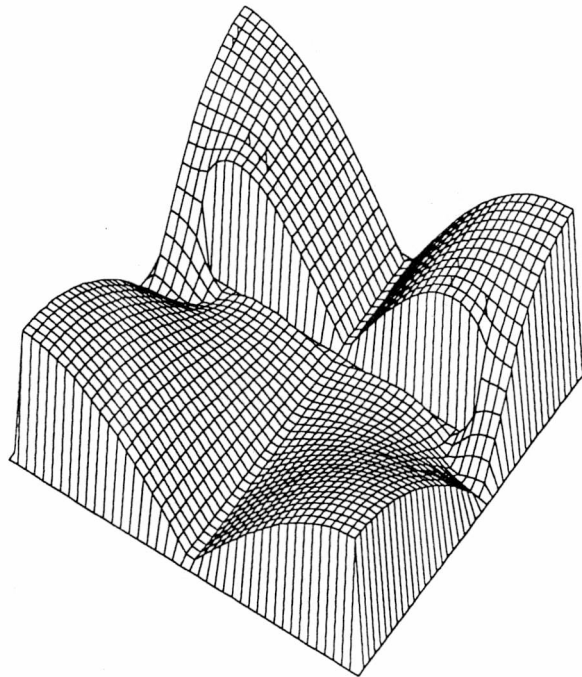


Fig. 5.4c. Field distribution at 160 MHz
- H_z component

a steady state situation, taking the maximum of the field level at each node, during one full cycle of the sinusoid. Of course the steady state is possible to obtain only for cutoff frequencies which are in fact the resonances of the two-dimensional structure. Applying frequency that is not a cutoff frequency results with waves "travelling" around the septum (in the case of TEM cell).

The cross section field distribution was calculated for frequencies presented in the Table 5.2 (TE modes). Results of modelling are presented in Figures 5.2-5.4 (E_x , E_y and H_z fields - standing waves, xy -plane corresponding to cross section of the cell). The comparison between those figures and results from [64] points out, that it is a good agreement of both methods.

Presented fields are not scaled along the z -axis. That is because the amplitude of each higher order mode depends on the way of the field excitation in the real structure. One can check, that also in analytical solution for (for instance) rectangular waveguides, the field quantities are given with an amplitude as a parameter [47,69]. Farther more - the most important to understand the mechanism of the higher order modes generation is the field structure. Therefore in all cases (for each mode) the excitation was the same, which enabled comparing results with each other.

The fact, that from the figures conclusions may be drawn concerning the mechanism of generating higher order modes on certain frequencies is especially important. For example one can see, that gaps between the centre conductor and walls of the cell are responsible for the TE_{01} mode, or that the centre conductor has no influence on the TE_{10} mode, because the field distribution is the same as for rectangular waveguides (Fig. 5.3).

Of course, in the same way calculations for TM modes can be made, but because TM modes cutoff frequencies are much higher then those of TE modes, author's interest was focused on the latter.

5.3. Influence of the EUT on higher order modes

The results of calculations presented in the previous point have proved usefulness of the TLM method at the field of cutoff frequencies and field distribution investigations. Next, the method has been applied to model the situation, in which EUT is placed into work volume of the TEM cell. As the EUT a rectangular perfectly conducting box has been taken. Its size has been assumed to be one third of linear dimensions of the cell volume.

Cutoff frequencies of the loaded cell are presented in the Table 5.3. For comparison frequencies for empty cell are given.

Table 5.3. Cutoff frequencies of the loaded cell

Mode	Cutoff frequency [MHz]	
	empty cell	loaded cell
TE_{01}	75.51	75.96
TE_{10}	124.30	105.07
TE_{11}	161.86	145.01
TE_{02}	193.42	156.06
TE_{20}	246.43	235.29

From above results one can see, that putting an object into TEM cell changes cutoff frequencies of higher order modes and, therefore, frequencies of resonances

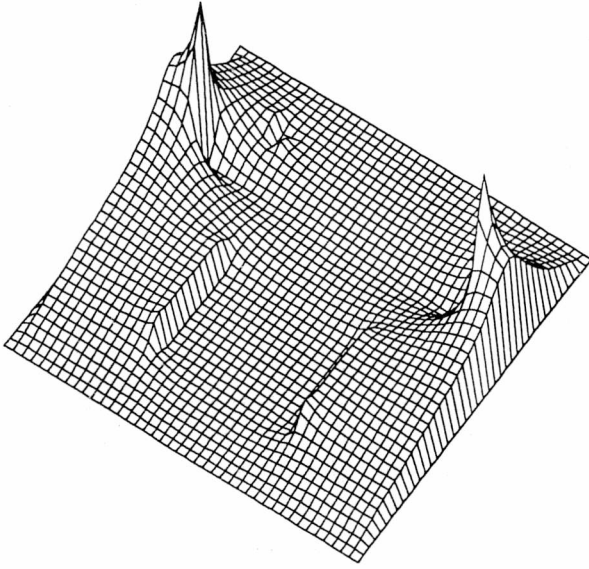


Fig. 5.5a. Field distribution in a loaded cell
at 74.3 MHz - E_x component

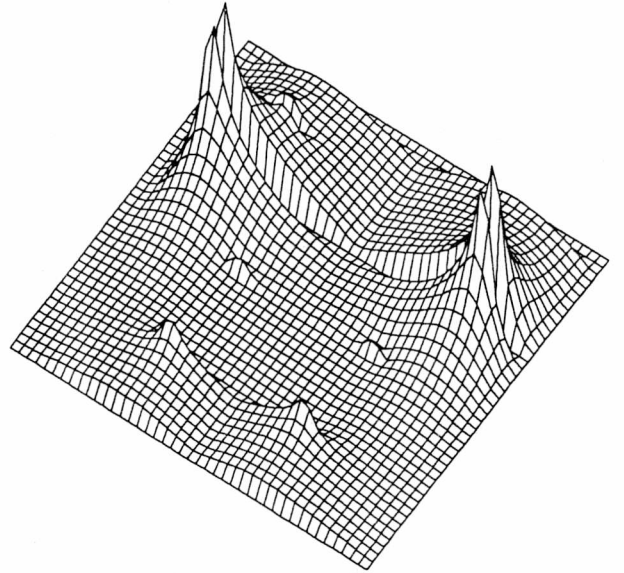


Fig. 5.5b. Field distribution in a loaded cell
at 74.3 MHz - E_y component

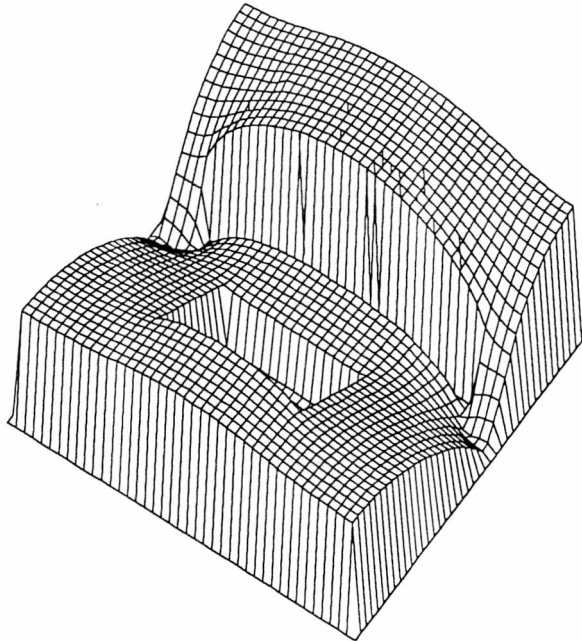


Fig. 5.5c. Field distribution in a loaded cell
at 74.3 MHz - H_z component

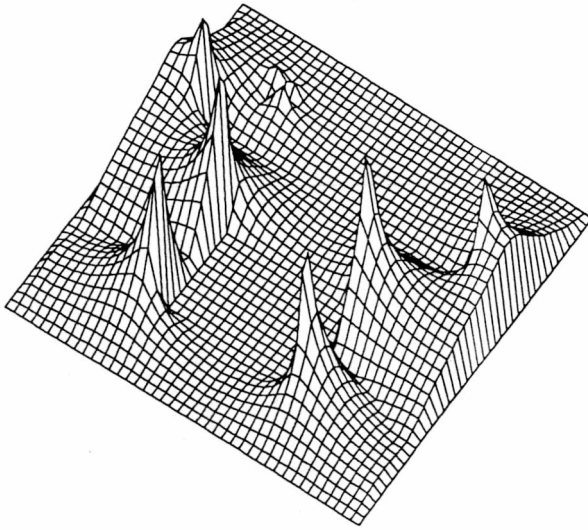


Fig. 5.6a. Field distribution in a loaded cell
at 105 MHz - E_x component

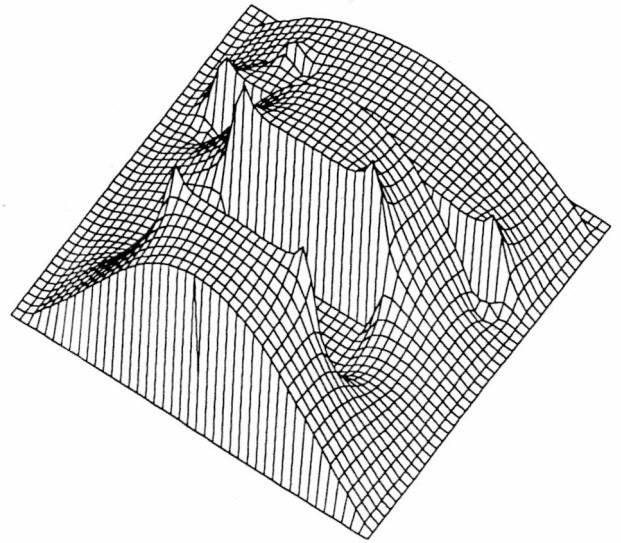


Fig. 5.6b. Field distribution in a loaded cell
at 105 MHz - E_y component

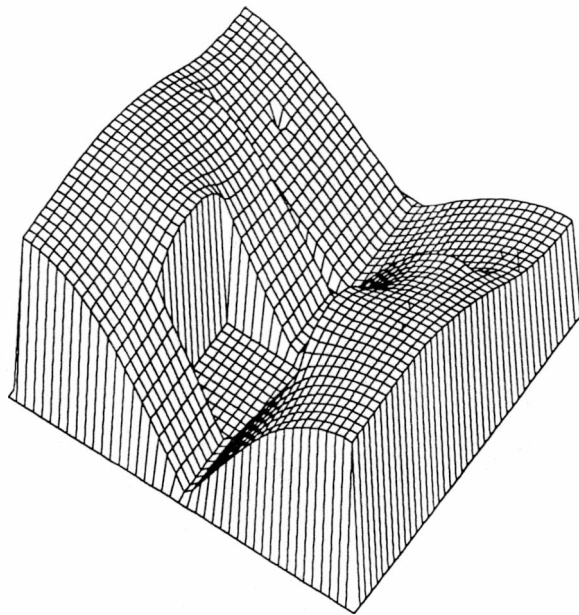


Fig. 5.6c. Field distribution in a loaded cell
at 105 MHz - H_z component

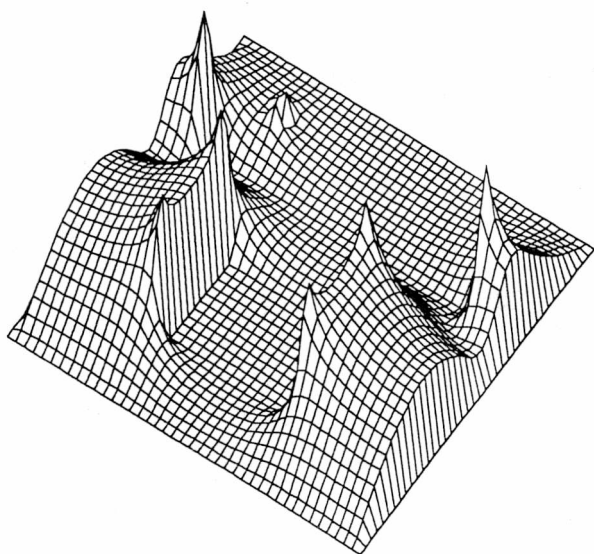


Fig. 5.7a. Field distribution in a loaded cell
at 145 MHz - E_x component

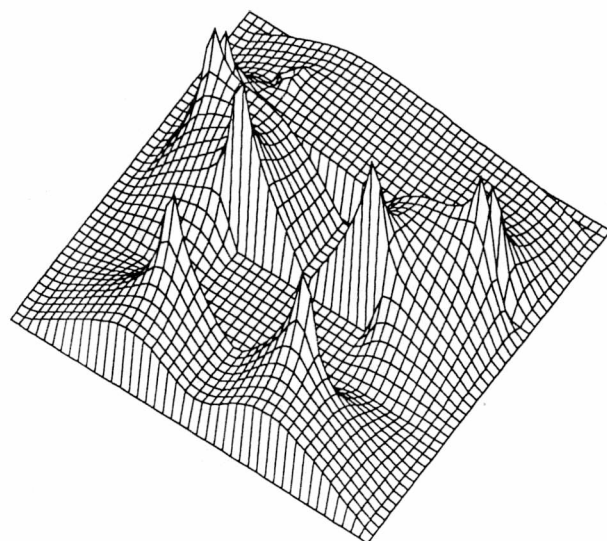


Fig. 5.7b. Field distribution in a loaded cell
at 145 MHz - E_y component

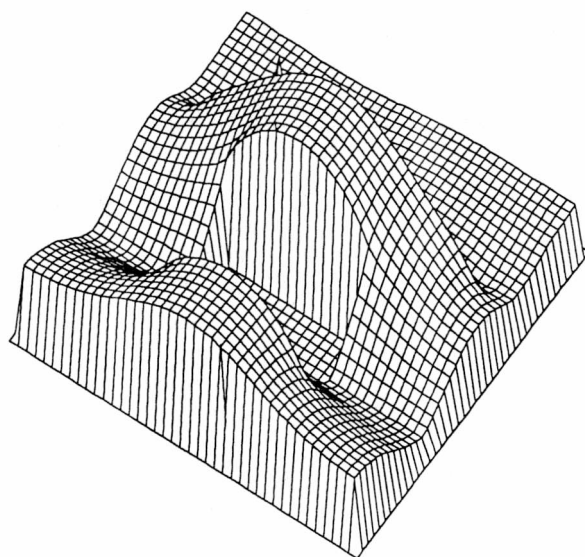


Fig. 5.7c. Field distribution in a loaded cell
at 145 MHz - H_z component

corresponding to those cutoff frequencies. Of course, as it was discussed earlier, two-dimensional model applied to calculations assumes, that the EUT is infinite along the z -axis that is not true, but the calculations were done rather to investigate some tendencies than to obtain precisely computed resonant frequencies of loaded cells. For this second purpose three-dimensional modelling is necessary.

The field distribution for first three modes in a loaded cell is presented in Figures 5.5-5.7.

It is easy to notice that the smallest changes in cutoff frequencies are related to the situation, when the EUT (a metal box) is put in the node of the standing wave (for instance TE_{01} mode) and the biggest when the field distribution is affected by putting the object in the area at which the maximum of the standing wave occurs (for example TE_{10} mode).

Above results prove, that putting an EUT into cell volume not only changes considerably a field distribution inside the TEM cell, but also changes higher order modes cutoff frequencies. It can be observed, that cutoff frequencies in the object presence are decreased, which can also decrease the resonant frequencies related to corresponding modes. This can cause decreasing the usable bandwidth of the cell!

TEM cells (especially Crawford cells) are sometimes used above their first resonance, taking advantage of the fact that resonances are sharp. From presented investigation it is clear, that determining resonant and cutoff frequencies for empty cell is not enough to make successful measurements when the EUT is present.

Full, reliable analysis of resonant frequencies in the TEM cell demands however taking three-dimensional model. The tool enabling such investigation is described in the next chapter.

6. TEM cell modelling in the whole frequency range

The calculations presented in Chapter 3 make it possible to find out the electromagnetic field distribution as far as only TEM mode is propagated inside the TEM cell (especially the field distribution in TEM cell cross section perpendicular to the direction of the wave propagation). However such a model (even three-dimensional) does not give reliable information about resonant problems in the cell, and about field distribution when higher order modes are present. In particular it is impossible to detect, which parts of the cell construction are responsible for higher order modes fields generation, and what, in fact, is the electrical length (see Chapter 2) of the cell.

In order to investigate those problems the TLM method has been applied to investigate fields distribution in Crawford cell and, partly, in GTEM cell.

6.1. Two-dimensional TLM model of a TEM cell

First, the two-dimensional TLM method has been applied to the longitudinal cross sections of the TEM cells. "Longitudinal cross section" means a vertical section from the input of the cell to the cell's load, as it is shown in Fig. 6.1.

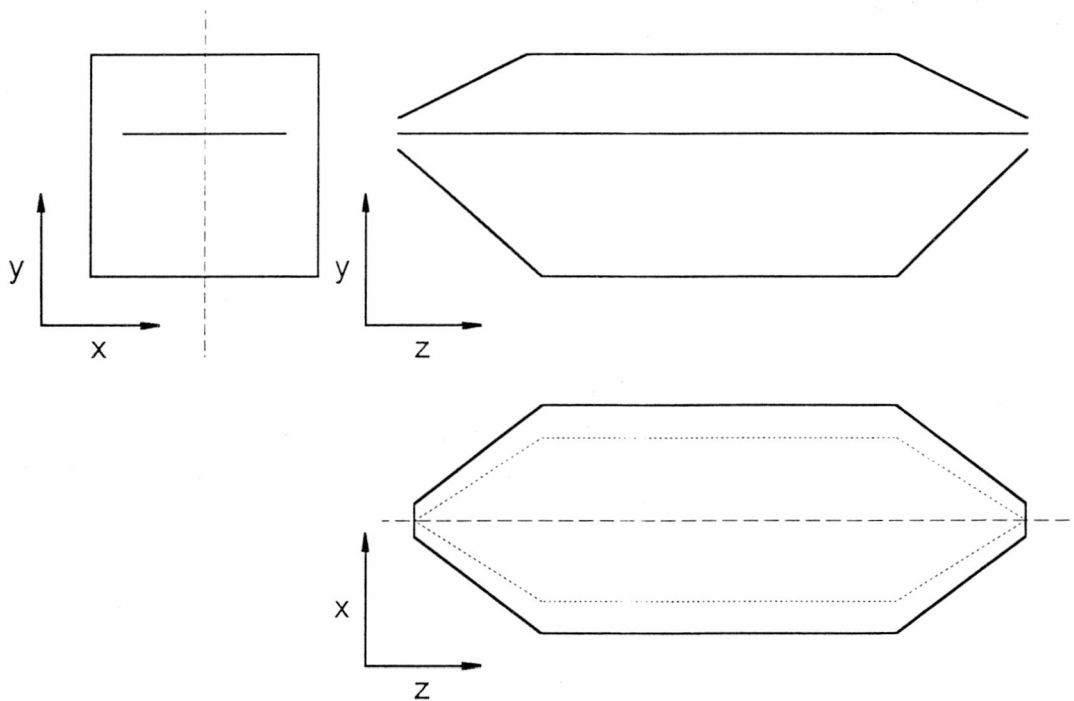


Fig. 6.1. Longitudinal cross section of the Crawford TEM cell

Two types of excitations have been used, both applied to the input of the cell: the sinusoidal excitation to investigate standing waves distribution at different frequencies, and the pulse excitation to investigate reflections from elements of the cell. In the latter

case the pulse with the limited bandwidth has been used in order to eliminate problems related to the limited bandwidth of the TLM method.

While using sinusoidal excitation two frequencies have been used: 30 MHz that is for sure the frequency below all possible resonances, and 300 MHz that is in the range of resonances (the ITA Crawford cell and the ITA GTEM cell). No particular resonance frequencies have been used, because two-dimensional model is not good enough to obtain reliable values. The purpose of applying such a model is to show mechanisms of field propagation rather than to calculate concrete resonant frequencies values and the field distribution.

6.1.1. Modelling of an empty cell

The results of the ITA Crawford cell modelling for the sinusoidal excitation are presented in figures 6.2 (30 MHz) and 6.3 (300 MHz).

As it was expected for 30 MHz the electric field distribution is flat, which means no standing waves in the cell's volume exists. In fact the 30 MHz frequency is far below any resonances (first measured resonance occurred at about 60 MHz), and below the cutoff frequencies of higher order modes of the field (see table 5.2). For such low frequencies the cell behaves exactly in the way expected by constructors - only TEM wave is propagated, which enables comparisons with free space.

At 300 MHz the field distribution inside the Crawford cell is very irregular (especially in the lower part - which is the main working volume of the cell). In this way, when the user puts any equipment into the cell, the field level to which the EUT is submitted cannot be determined without field mapping techniques. Furthermore, at such high frequencies each part of EUT is submitted to different electromagnetic field (concerning field levels and directions).

In order to check what is the reason of the standing waves generation the pulse travelling along Crawford cell has been investigated. The Fig. 6.4 shows four situations in different time steps. Because the TEM wave is under investigation, the E_y component is shown. Picture shows the pulse in the input tapered section (6.4a), after reaching the main part of the cell (6.4b), in the middle of the cell (6.4c) and after reaching the load tapered section (6.4d). From Fig. 6.4. it is obvious, that the field distortion is caused by the transitions between sections of the cell. It is especially well seen in Fig. 6.4c, where we have in fact two waves - direct TEM wave, and the wave reflected from the floor. It can also be noticed, that in the upper part of the cell, where the angle between tapered section and the ceiling is more obtuse, the problem of reflections is much smaller. It means that making the tapered sections longer should improve the situation in the cell. However making them too long is rather impractical. This way of thinking leads directly to GTEM cell concept!

In the same way the GTEM cell has been investigated. The results are shown in Fig. 6.5. Because in this case cell behaviour at high frequencies is interesting, the calculations have been done for 300 MHz excitation only. As one can see the field distribution surfaces are flat - the peaks around the E_z picture are the result of inability to model flat surface lying aslant, without "steps", and should not be taken into account. A slight standing wave that can be seen in Fig. 6.5a is the result of simplified modelling of the absorber - if the absorber wall was perfect it would be not standing wave at all. This statement can be proved by following the pulse propagation in GTEM cell. Fig. 6.6 shows the pulse propagation at four different moments. It can be noticed that there are not reflections in the cell. The "tail" of the pulse is the result of applying the pulse with limited bandwidth.

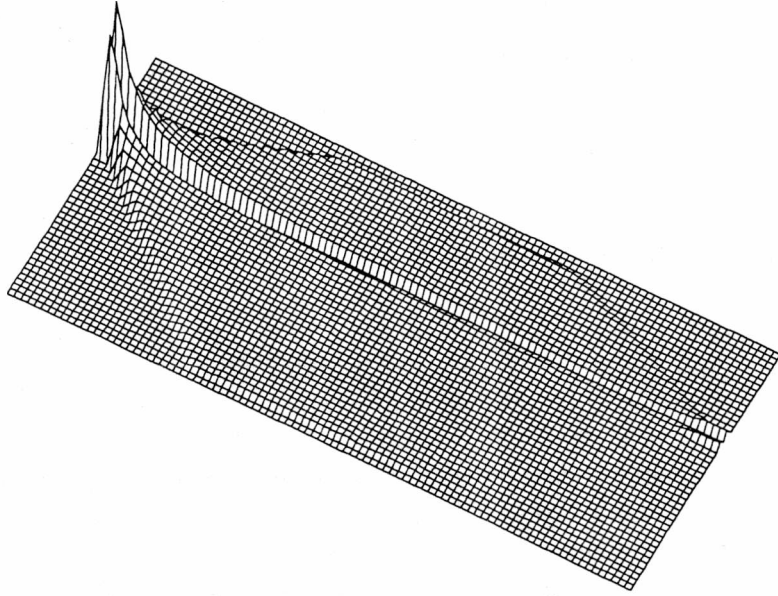


Fig. 6.2a. Field distribution in an empty Crawford cell at 30 MHz - E_y component

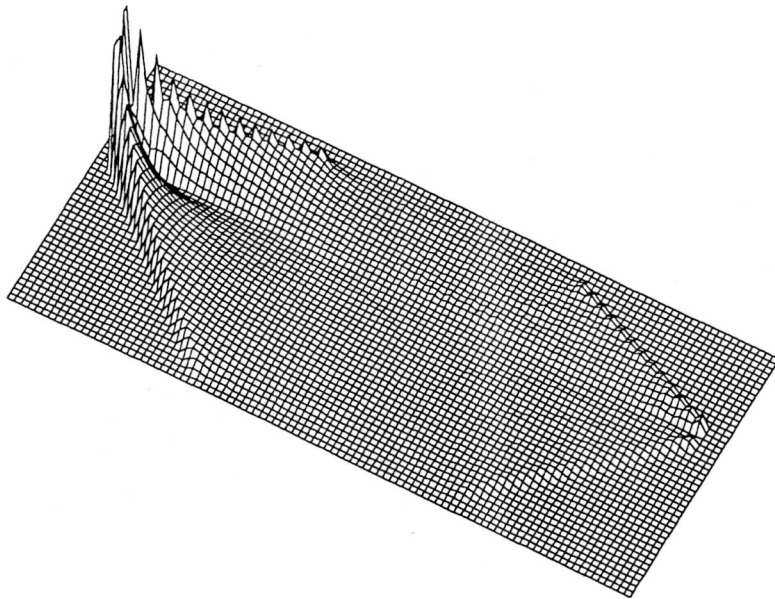


Fig. 6.2b. Field distribution in an empty Crawford cell at 30 MHz - E_z component

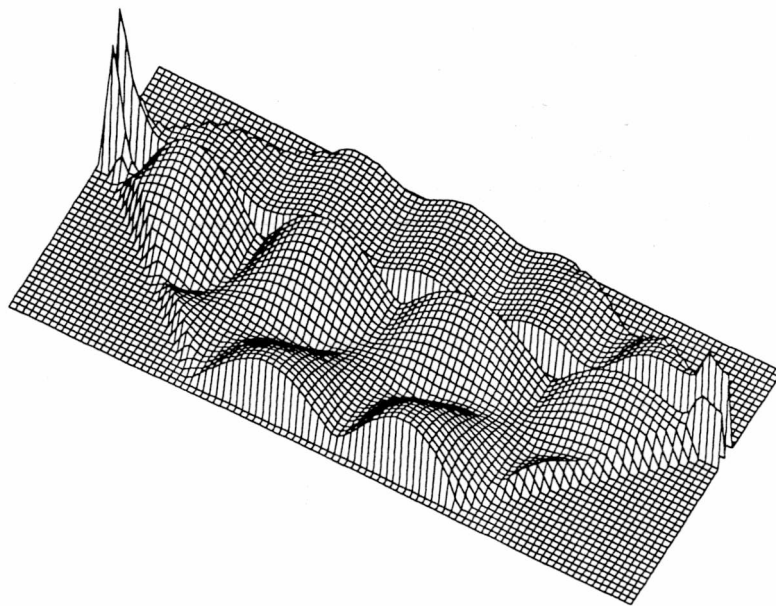


Fig. 6.3a. Field distribution in an empty Crawford cell at 300 MHz - E_y component

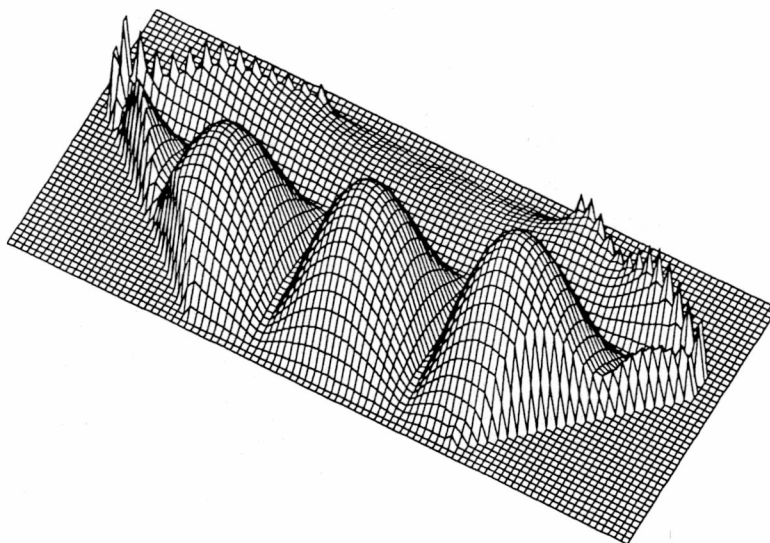


Fig. 6.3b. Field distribution in an empty Crawford cell at 300 MHz - E_z component

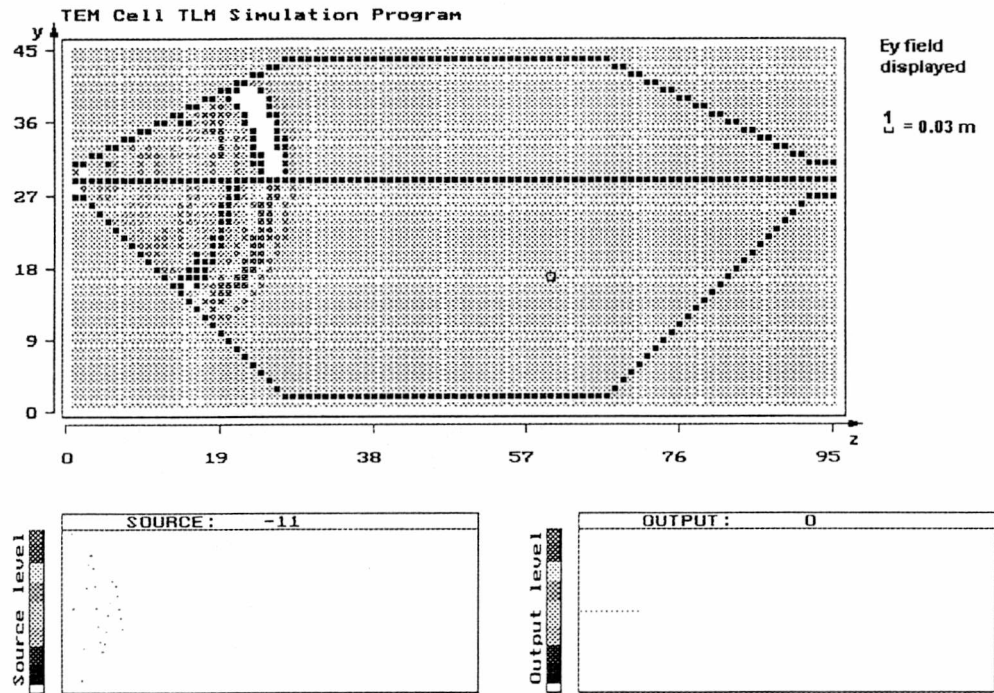


Fig. 6.4a. Pulse propagation inside the TEM Crawford cell after approx. 2,7 nsec

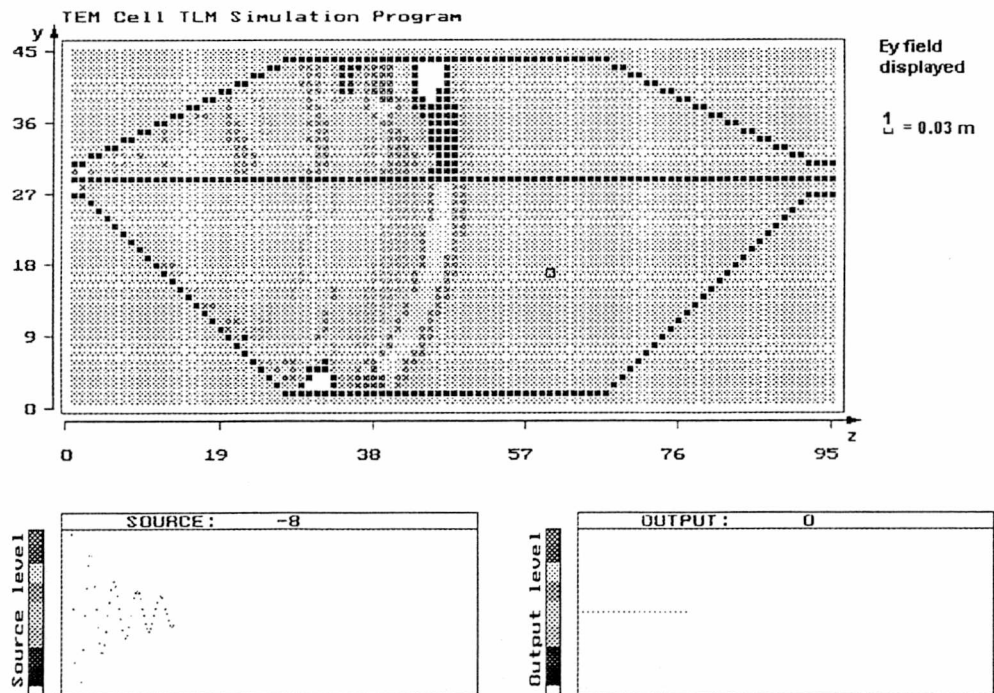


Fig. 6.4b. Pulse propagation inside the TEM Crawford cell after approx. 4,8 nsec

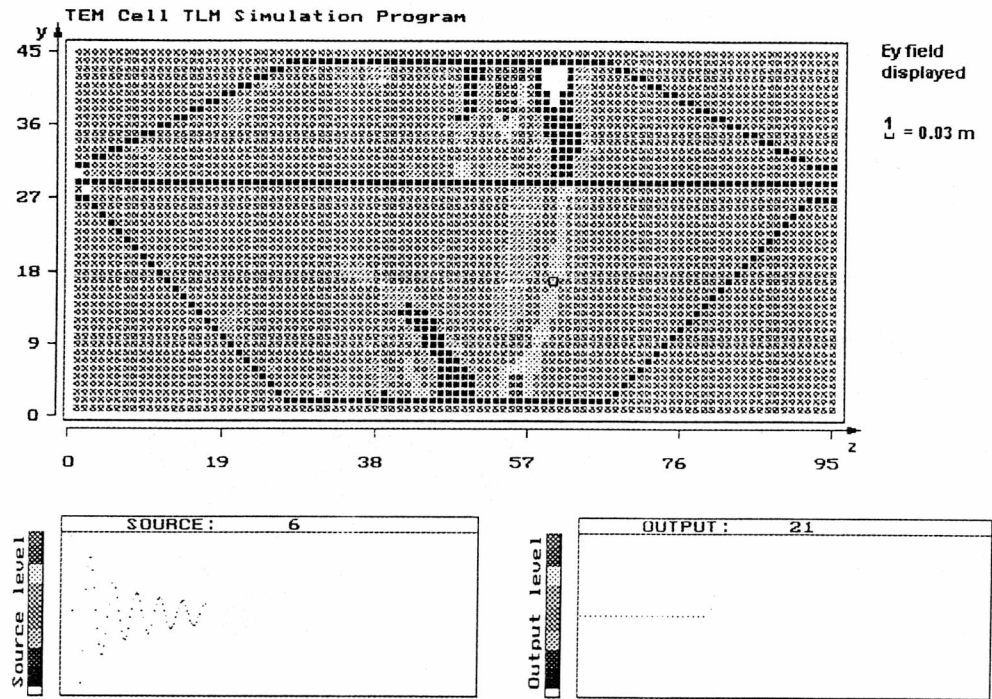


Fig. 6.4c. Pulse propagation inside the TEM Crawford cell after approx. 6,3 nsec

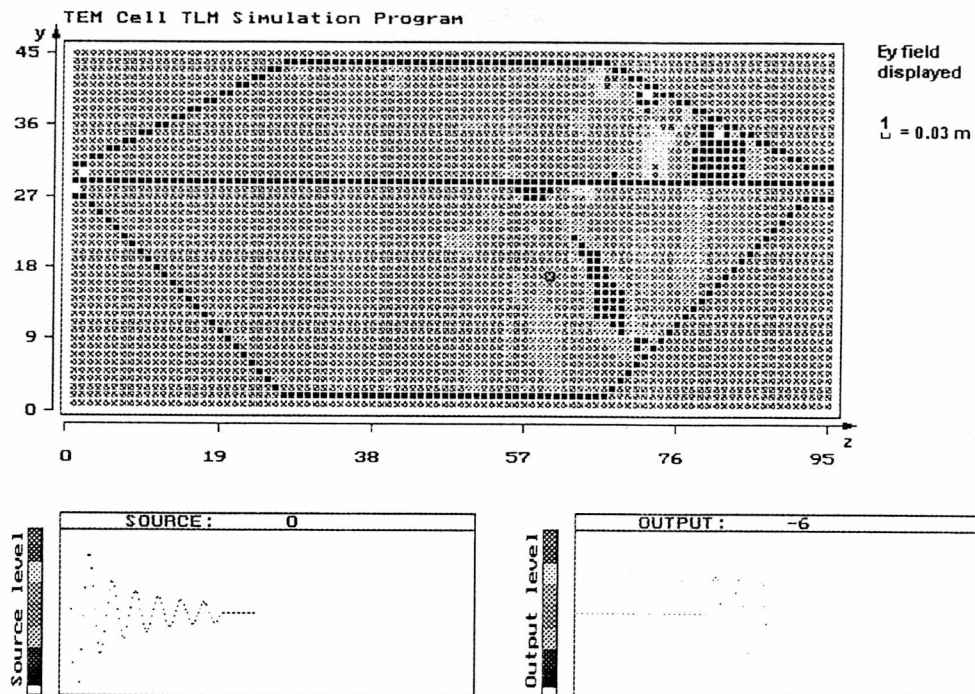


Fig. 6.4d. Pulse propagation inside the TEM Crawford cell after approx. 8,5 nsec

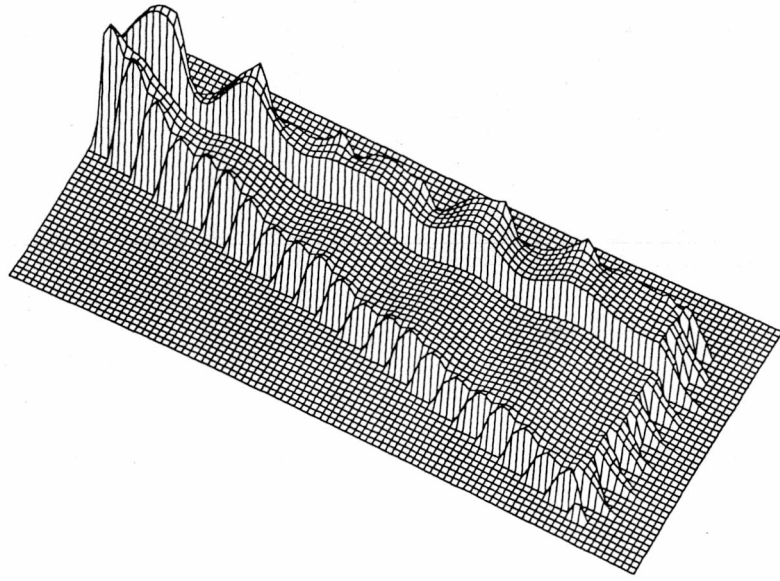


Fig. 6.5a. Field distribution in an empty GTEM cell at 300 MHz - E_y component

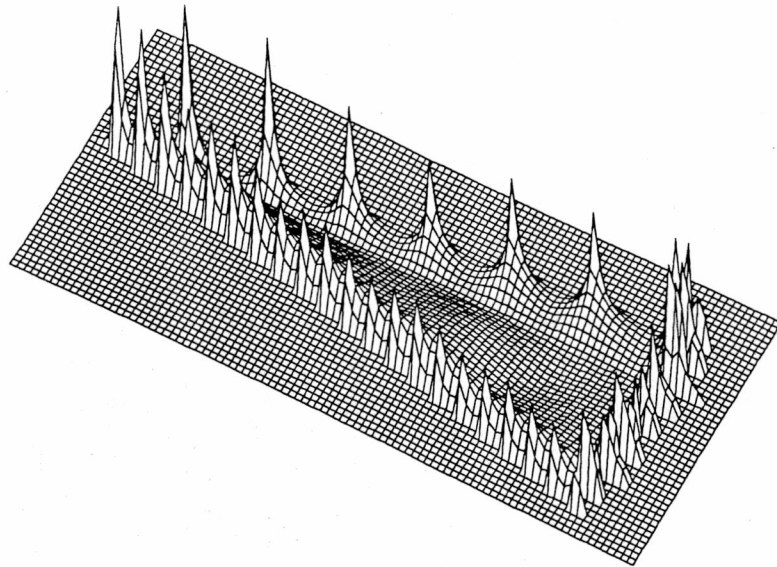


Fig. 6.5b. Field distribution in an empty GTEM cell at 300 MHz - E_z component

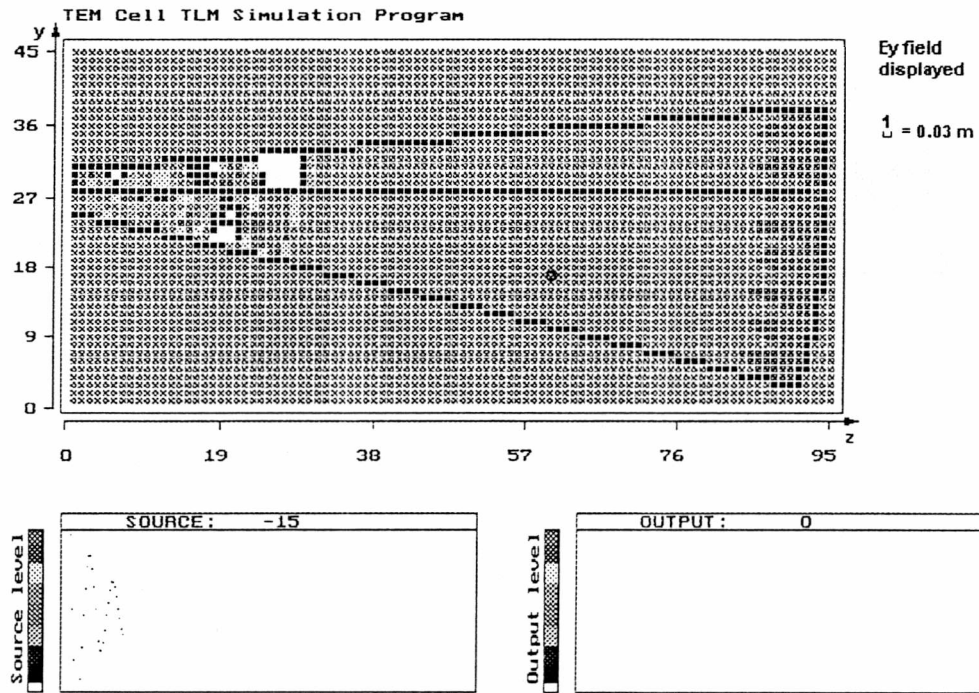


Fig. 6.6a. Pulse propagation inside the GTEM cell
after approx. 2,7 nsec

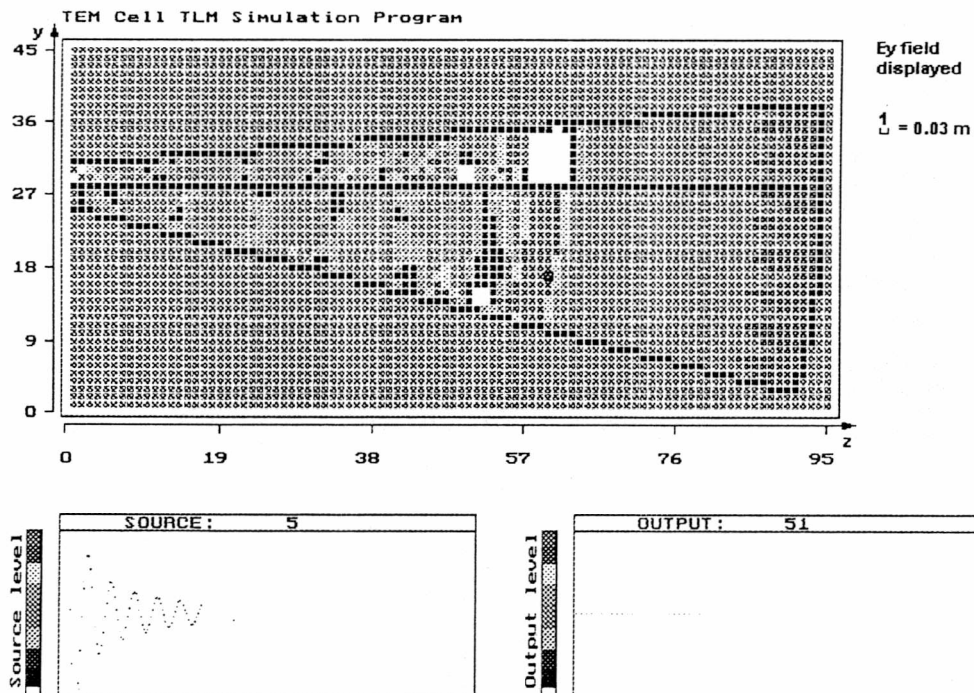


Fig. 6.6b. Pulse propagation inside the GTEM cell
after approx. 6,2 nsec

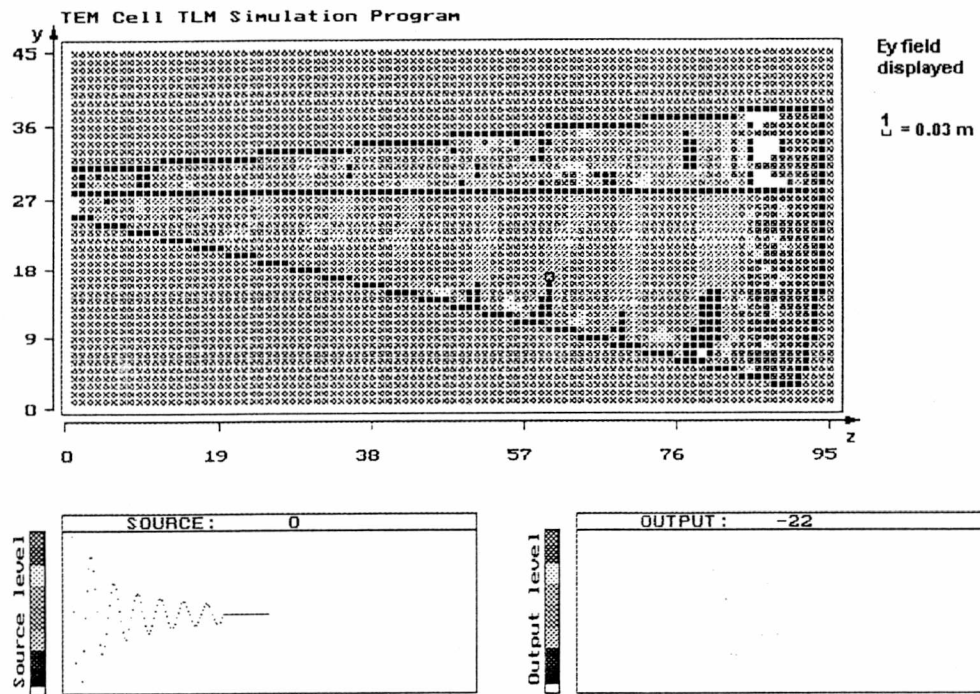


Fig. 6.6c. Pulse propagation inside the GTEM cell after approx. 9,0 nsec

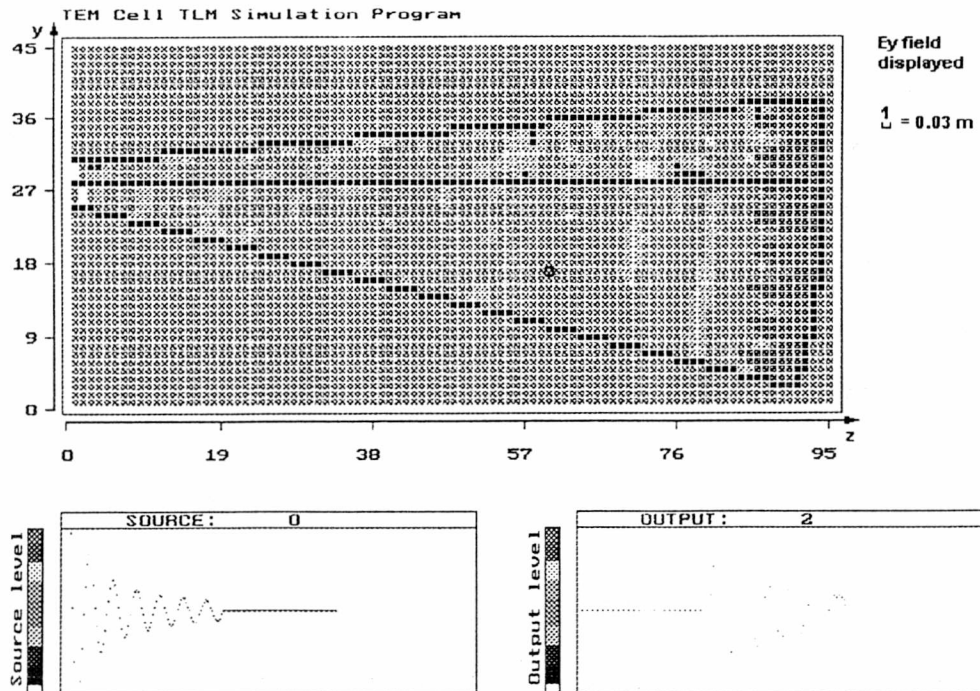


Fig. 6.6d. Pulse propagation inside the GTEM cell after approx. 12,0 nsec

6.1.2. Object inside a TEM cell

In the same way cells with EUT inside have been investigated. From Fig. 6.7 it can be seen, that as long as TEM cell operates below resonances it acts exactly in the way predicted by constructors - the metal object causes only increasing the field between in the neighbourhood, which can be interpreted in the way presented in Chapter 3. When operating at high frequencies (Fig. 6.8 - at 300 MHz) the situation is quite different. The field distribution seen in Fig. 6.8. is different from that in Fig. 6.3. Also it can be seen that there is no one field level that EUT is submitted. The field magnitude changes depending on the position on the EUT walls. Of course, such result was expected, and therefore Crawford TEM cells should not be used at frequencies above first resonances. The mechanism of reflection generation is shown in Fig. 6.9.

More interesting is the influence of EUT on the field distribution in GTEM cells. The situation is presented in Fig. 6.10 and 6.11. The size of the object is big enough to bring into relief field changes. It can be seen that the SWR is much bigger in this case than in the case of empty cell. The reason of it is the resonance between the input part of the cell (which cannot be eliminated by good matching - it is the result of tapering the cell structure rather) and the EUT. The backward reflection from the EUT cannot be eliminated - therefore large cell bandwidth reported by GTEM cell inventors may be a point of discussion when the loaded cell is investigated. It should be considered if applying absorbing materials near the input of the cell would be a solution.

6.2. Three-dimensional TLM model of a Crawford cell

Finally, full three-dimensional TLM method has been applied to investigate electromagnetic fields distribution in a Crawford ITA cell.

The model uses $21 \times 21 \times 49$ elementary blocks to build the TLM network. It means that the rectangular piece of space in which TEM cell is placed is subdivided into 21609 parts, for which electric and magnetic fields components are calculated. Applying more subtle segmentation on *DECstation 5000* machine is possible, but leads to very long computation times, because the operational memory must be temporally stored into computer hard disk (the technique called *swapping*). Of course for each elementary block 12 numbers must be stored (not taking into account any additional information), because of the TLM method rules (see Chapter 4).

From above limitations it is obvious, that results of modelling must be treated as approximate. The accurate calculations require more subtle segmentation.

Four series of calculations have been done: three with sinusoidal excitations for empty and loaded cell for different frequencies, and one with impulse excitation in order to find out resonant frequencies of the cell.

In all cases the excitation point has been placed near the input of the cell, in the lower (bigger) part. The radiator is an isotropic source of the electric field - it is the reason of relatively high peaks in some points of fields distribution, which can be seen on plots.

6.2.1. Calculations for an empty cell

Results of calculations for 30 and 300 MHz frequencies are presented in figures 6.12 and 6.13 respectively, in the form of standing waves distributions. Because the illustrated quantities are vectors in three-dimensional space, the author has decided to

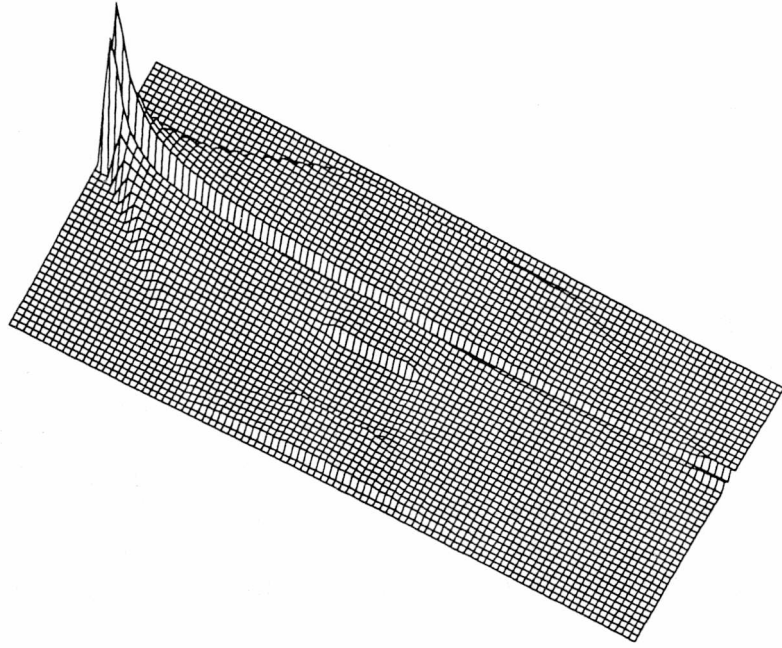


Fig. 6.7a. Field distribution in a loaded Crawford cell at 30 MHz - E_y component

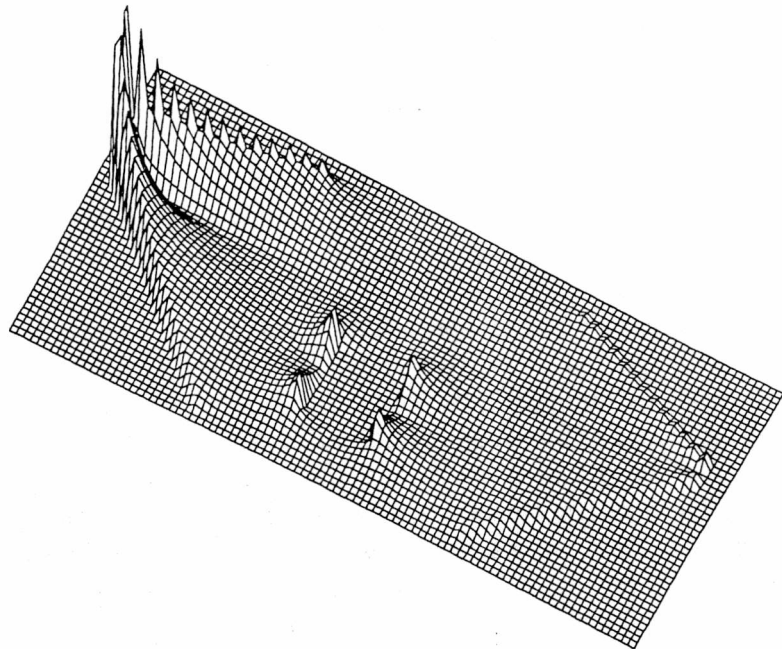


Fig. 6.7b. Field distribution in a loaded Crawford cell at 30 MHz - E_z component

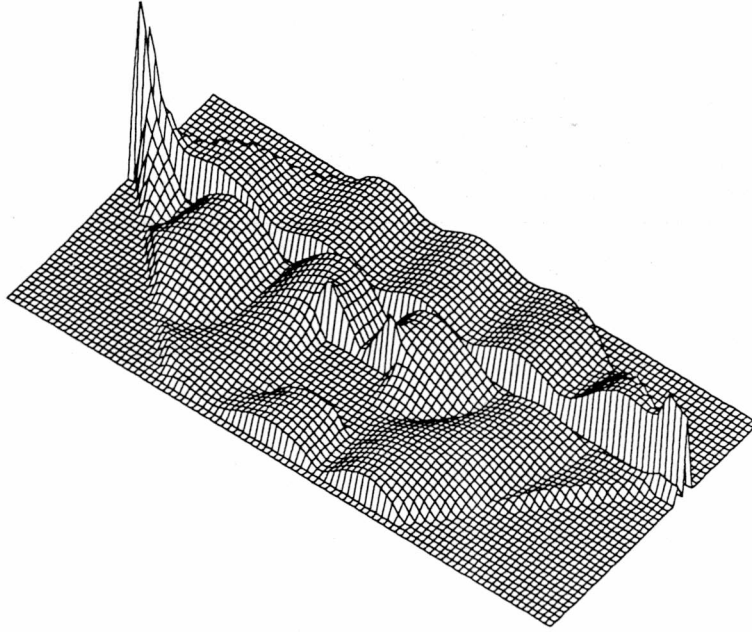


Fig. 6.8a. Field distribution in a loaded Crawford cell at 300 MHz - E_y component

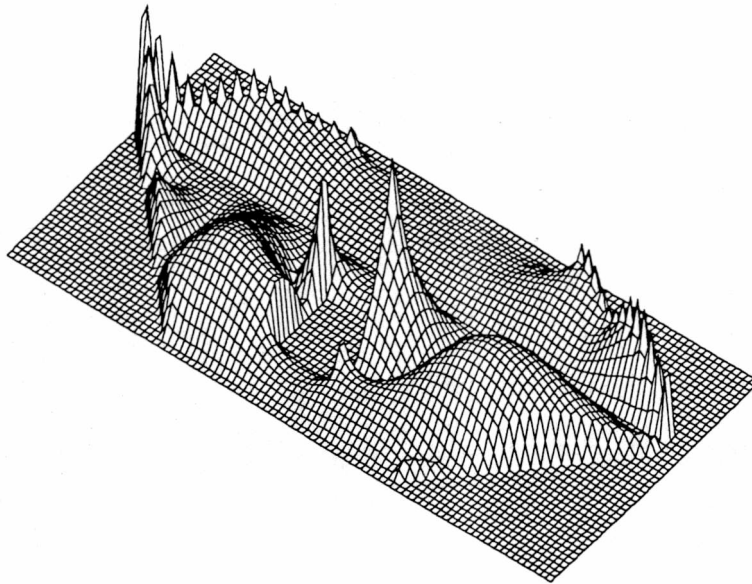


Fig. 6.8b. Field distribution in a loaded Crawford cell at 300 MHz - E_z component

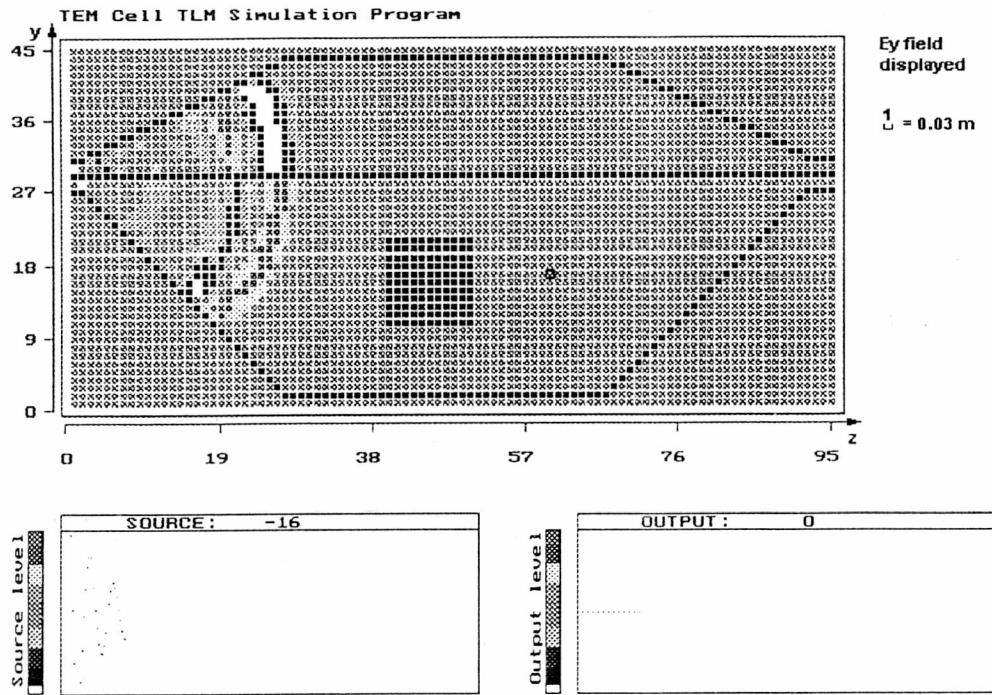


Fig. 6.9a. Pulse propagation inside the loaded TEM Crawford cell after approx. 2,7 nsec

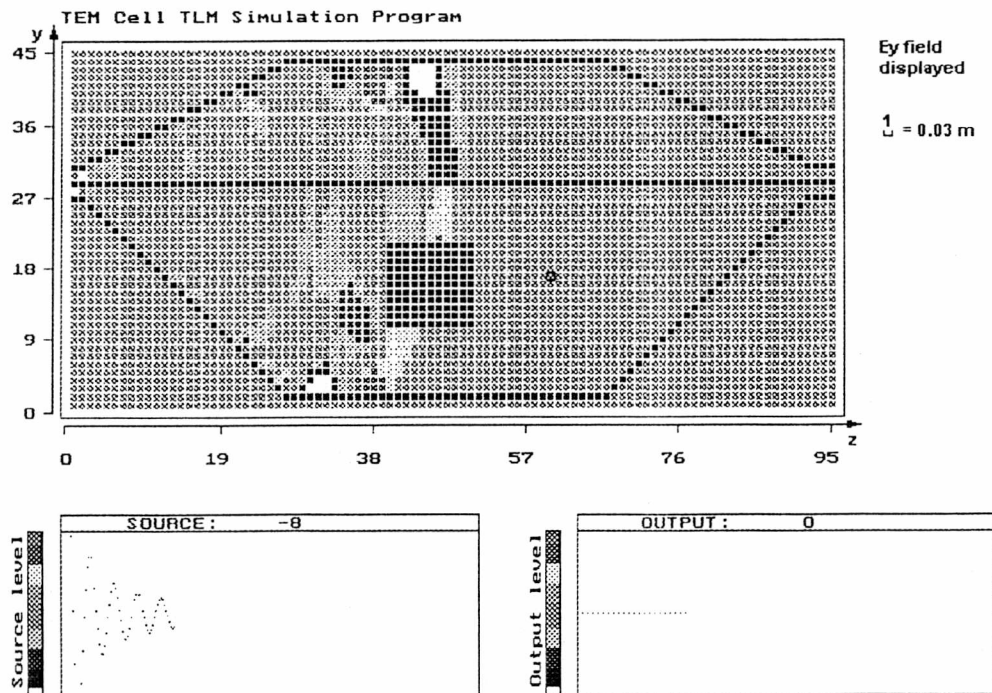


Fig. 6.9b. Pulse propagation inside the loaded TEM Crawford cell after approx. 4,8 nsec

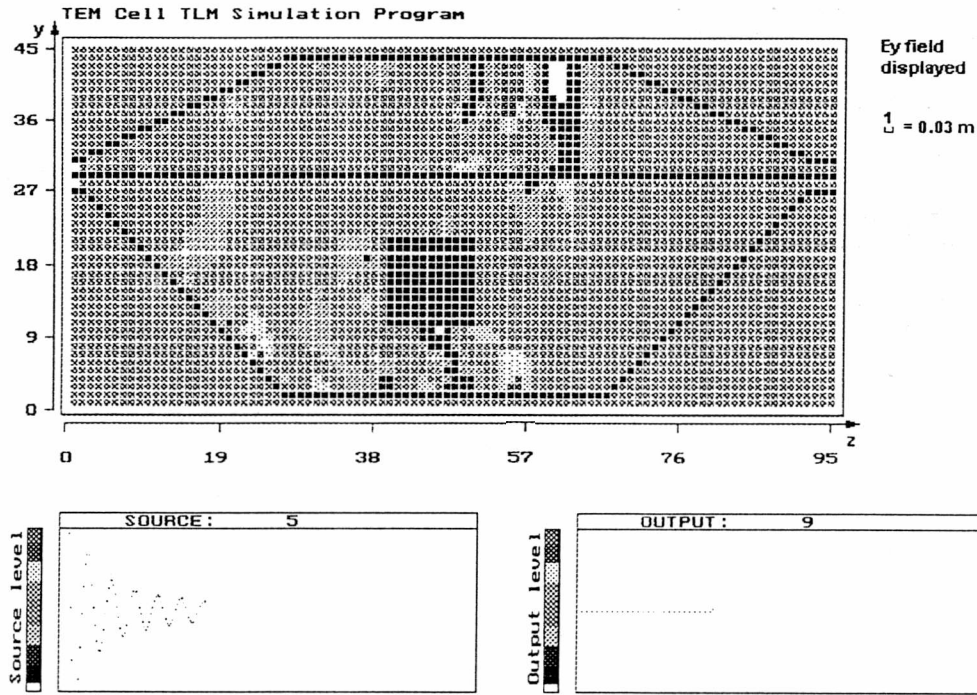


Fig. 6.9c. Pulse propagation inside the loaded TEM Crawford cell after approx. 6,3 nsec

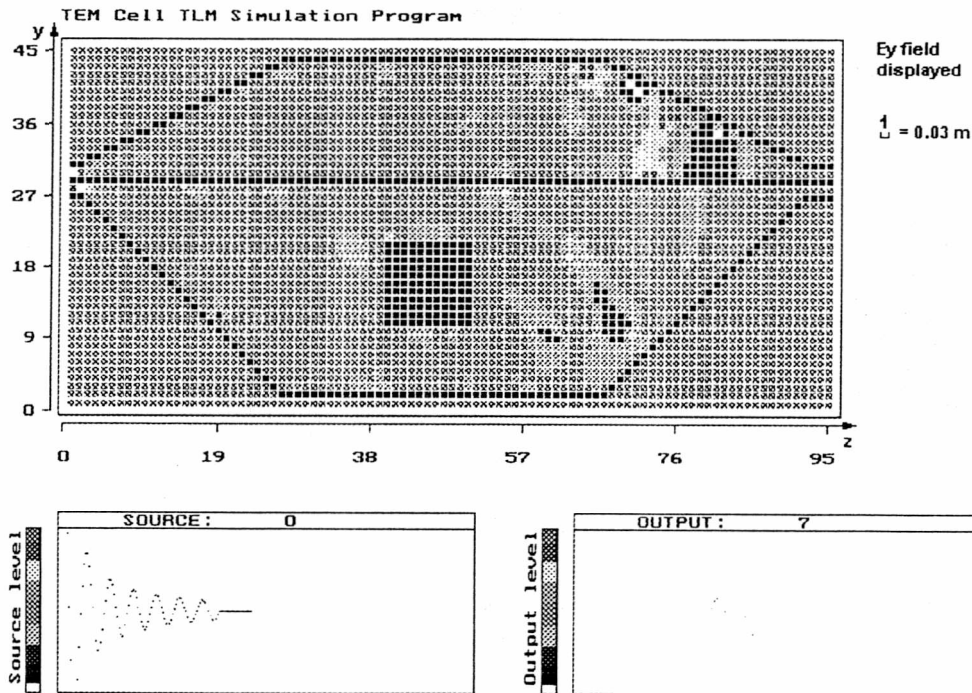


Fig. 6.9d. Pulse propagation inside the loaded TEM Crawford cell after approx. 8,5 nsec

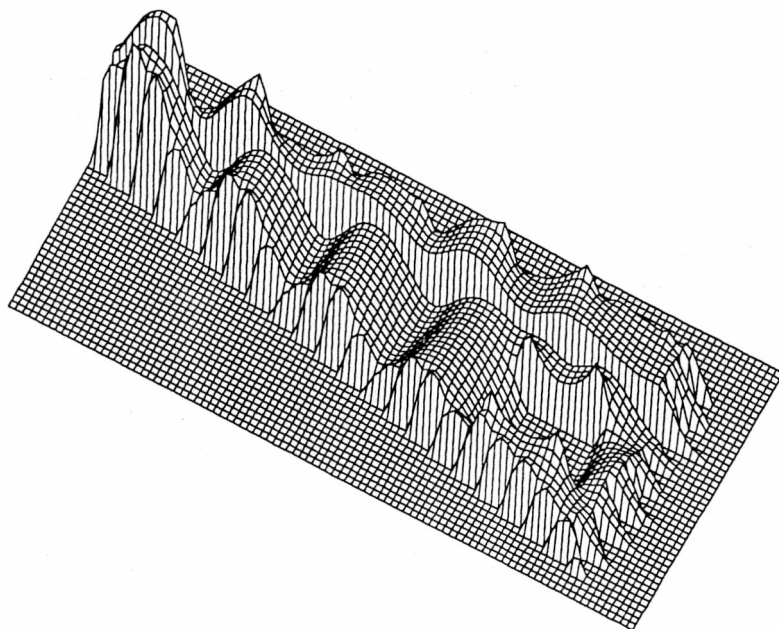


Fig. 6.10a. Field distribution in a loaded GTEM cell at 300 MHz - E_y component

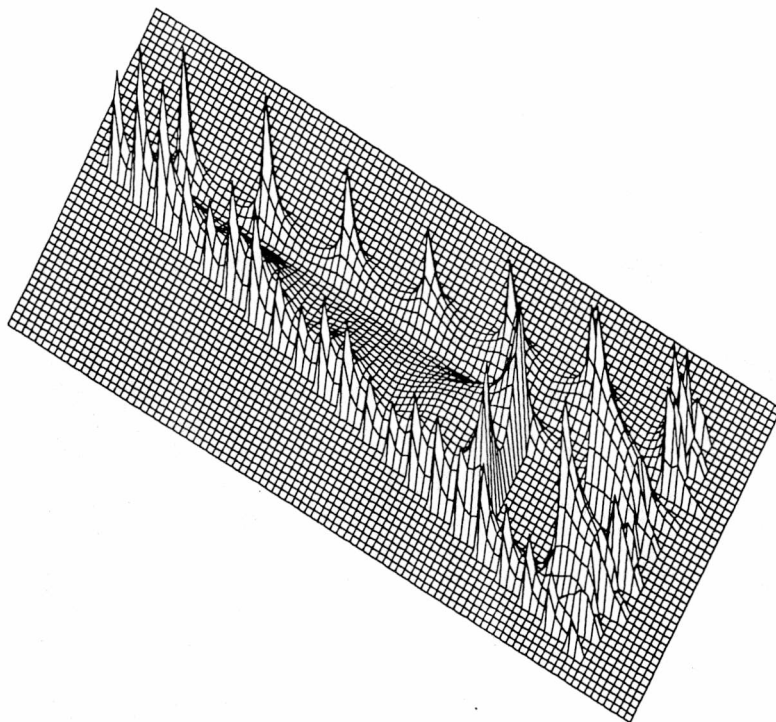


Fig. 6.10b. Field distribution in a loaded GTEM cell at 300 MHz - E_z component

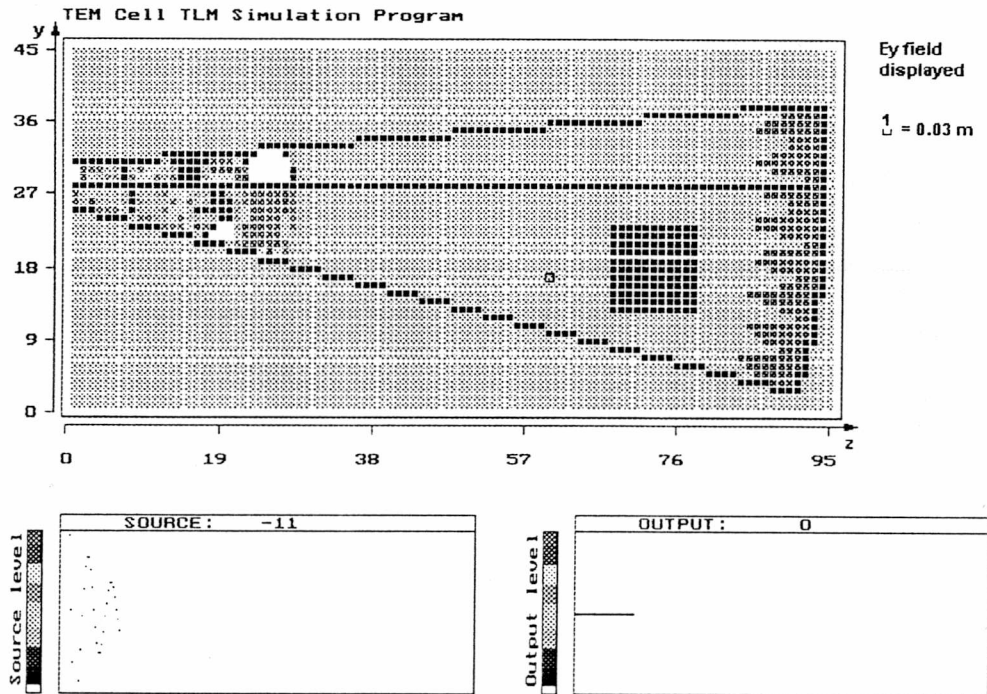


Fig. 6.11a. Pulse propagation inside the loaded GTEM cell after approx. 2,7 nsec

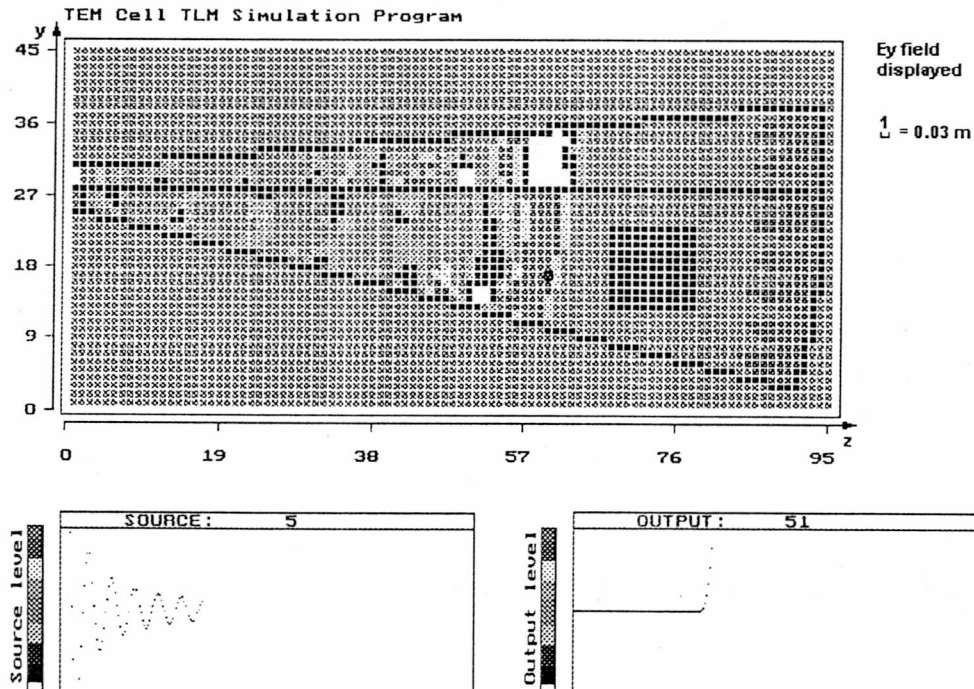


Fig. 6.11b. Pulse propagation inside the loaded GTEM cell after approx. 6,2 nsec

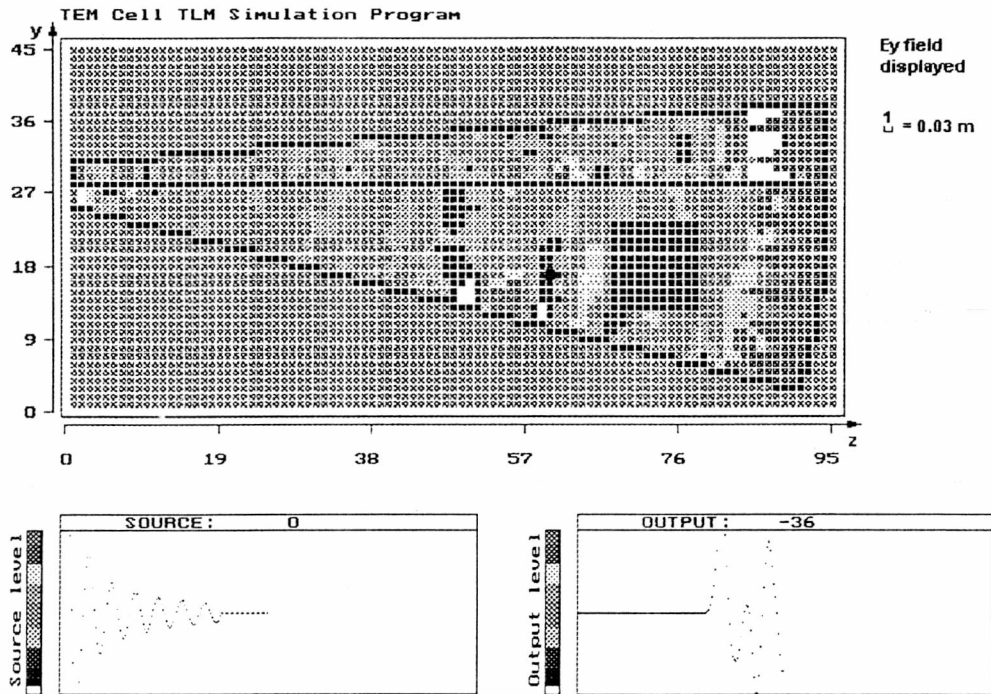


Fig. 6.11c. Pulse propagation inside the loaded GTEM cell
after approx. 9,0 nsec

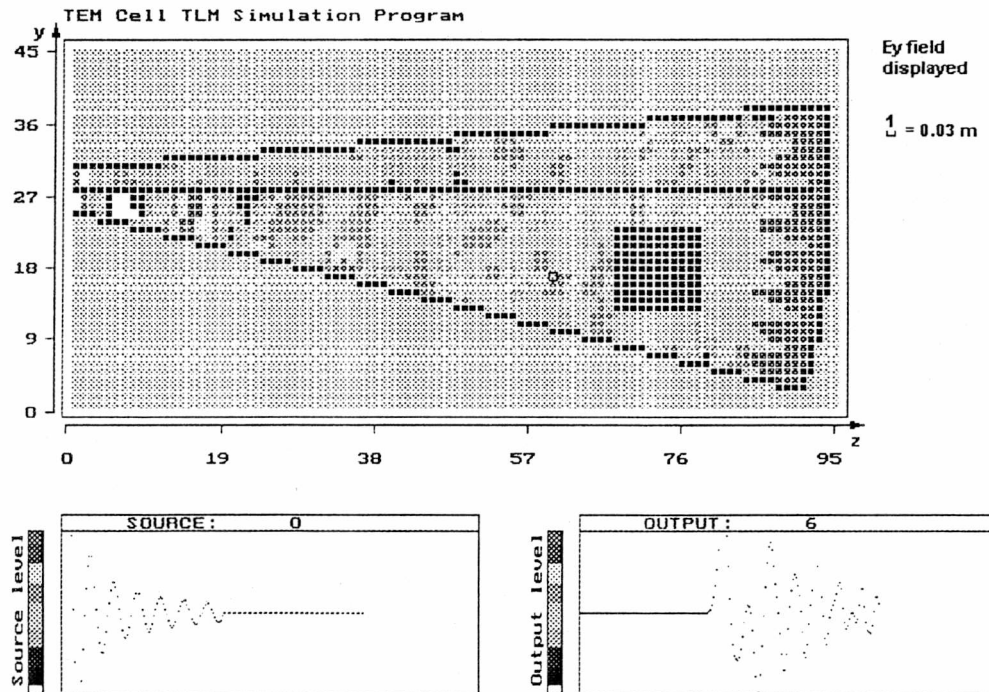


Fig. 6.11d. Pulse propagation inside the loaded GTEM cell
after approx. 12,0 nsec

present results in the form of all six field components in different cross sections of the TEM cell. So, in *a* and *d* pictures vertical cross sections perpendicular to the direction of propagation are shown. In *b* and *e* pictures vertical cross sections lying along the direction of propagation are presented, and finally in *c* and *f* pictures one can see horizontal cross sections of the TEM cell. The sections for which field distributions are presented are schematically shown on small pictures of the TEM cell on the top of each page.

From the Fig. 6.12 it is obvious, that at 30 MHz only TEM is propagated inside the cell. There are no E_z and H_z fields in the work volume. The E_x and E_y fields distributions near the middle part of the cell have exactly the same shape as calculated in Chapter 3 using the method of moments (see Fig. 3.2). The fact that both (entirely different) methods give the same results while describing the TEM mode inside TEM cells is the base to state that obtained results are correct. It should also be said that similar results were obtained by Hansen et al. from a model using FEM [34], and by Crawford [7] from measurements using the method of field mapping. The author could not do measurements himself, because of the lack of proper field probes. Such probes for both E- and H-field measurements in the near zone are under construction in the Radio Department of ITA, in the team under direction of the author.

The same calculations have been repeated for the frequency equal to 300 MHz, which was expected to lie in the range of resonances and higher order modes of electromagnetic field (Fig. 6.13). The reader can compare results of three-dimensional analysis with those of two-dimensional one (Fig. 6.3). It can be noticed that the shape of electric field components distribution is similar, especially why taking into account the middle part of the cell (compare Fig. 6.3a and b with Fig. 6.13b, Chapter 3). It means that as long as exact field levels are not so important as the field distribution (shape), the two-dimensional model can be used - for example to find out the mechanisms of the field reflections. From Fig. 6.13 it is also seen that at 300 MHz both E_z and H_z field components are not equal to zero in the middle part of the cell. It means, that Both TE and TM modes are present at this frequency. In fact the first TE cutoff frequency is 74.3 MHz (TE_{01} mode) and first TM cutoff frequency - 223.3 MHz (TM_{11} mode), as it was said in Chapter 5.

Careful reader can also check, that in all cases boundary conditions are satisfied, i.e. the electric field is always perpendicular to the walls near metal surface, and the magnetic field is always tangential.

6.2.2. Influence of the EUT on the field distribution

Calculations have been done to test the EUT influence on the electric and magnetic field distribution. As an object a rectangular metal box was assumed, like in pervious chapters. The EUT has been placed in the middle of the work space in the Crawford cell. The calculations have been done for 30 MHz frequency, in order to investigate field properties in the TEM mode range.

Results are presented in Fig. 6.14a-f. The comparison can be made with results of moment method modelling (Fig. 3.5, because the object is not grounded) and with two-dimensional TLM method modelling (Fig. 6.7). Once again the agreement of results is excellent.

The calculations have proved, as it was said before, that putting an object into work volume of the TEM cell changes the field distribution (but not very much while taking into account the TEM mode) and, what is more important, the field levels,

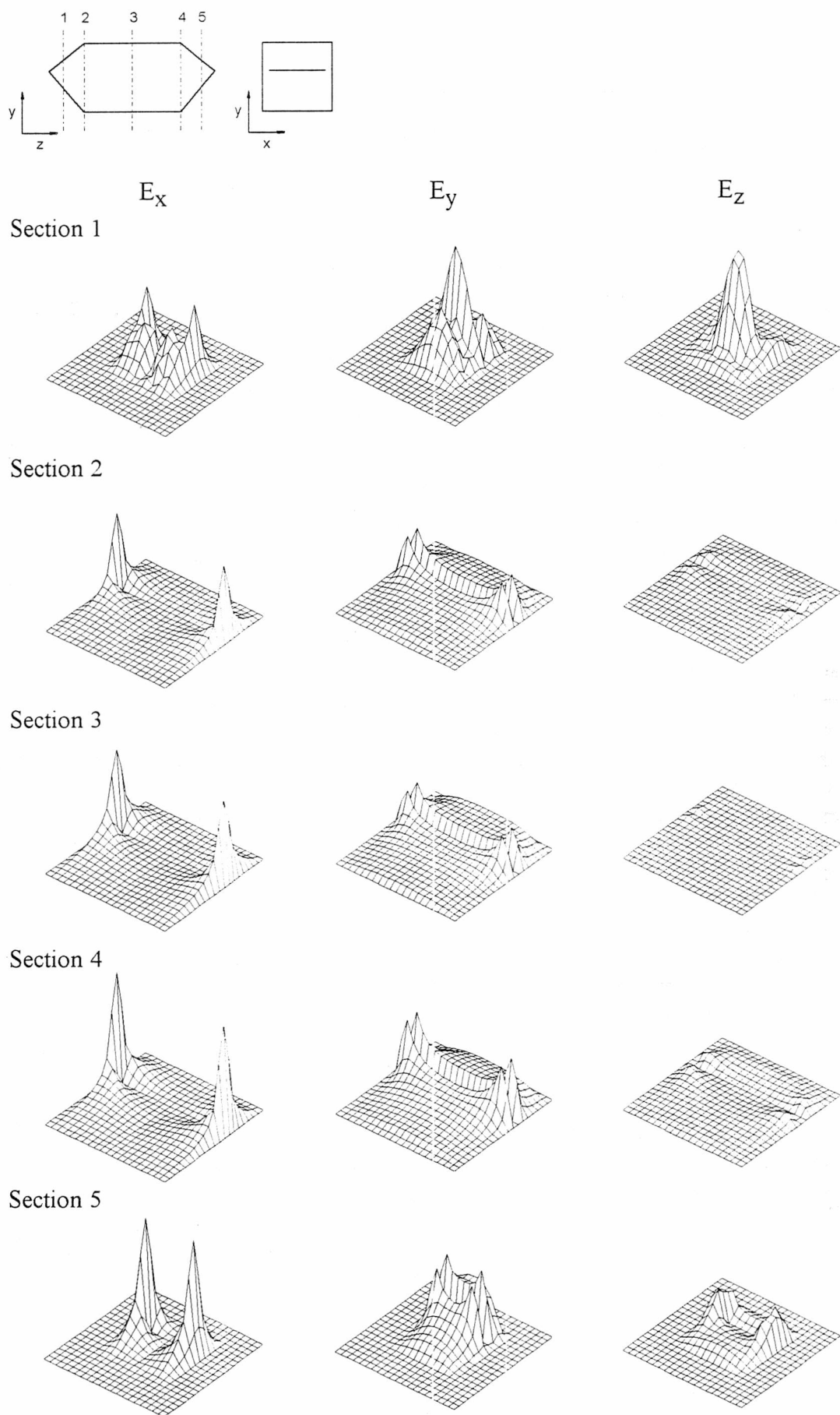
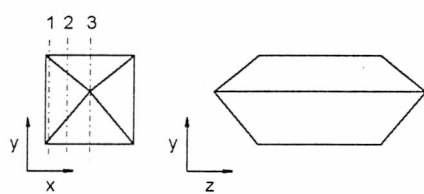


Fig. 6.12a. XY-electric-field distribution at 30 MHz (empty Crawford cell)

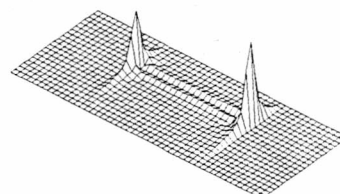
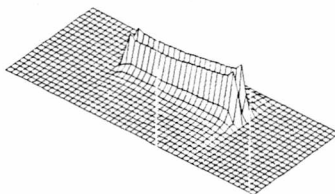
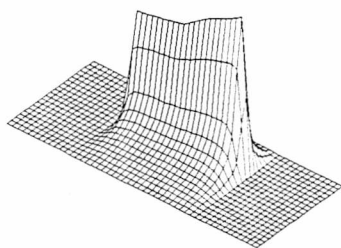


E_x

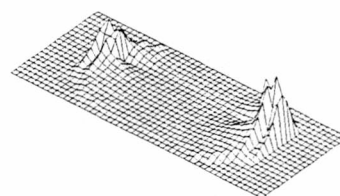
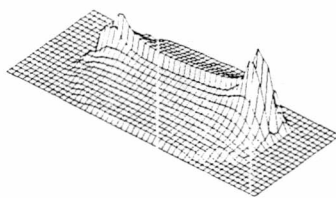
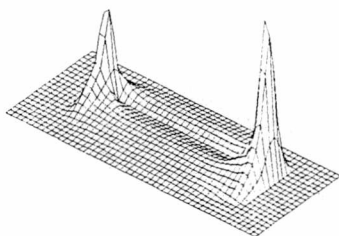
E_y

E_z

Section 1



Section 2



Section 3

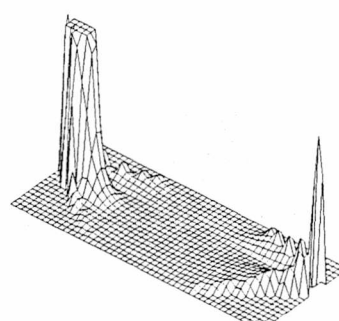
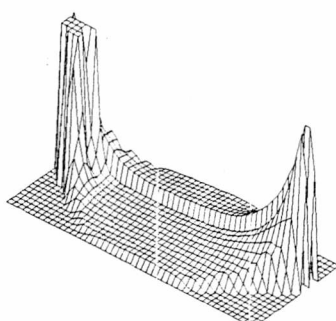
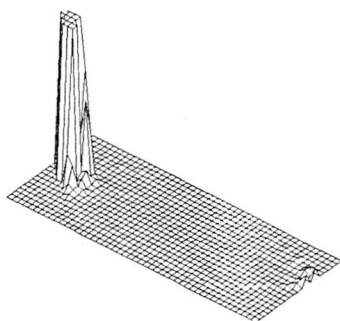
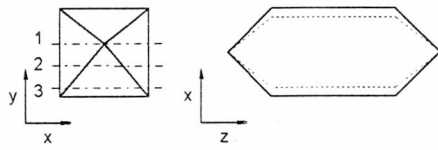


Fig. 6.12b. YZ-electric-field distribution at 30 MHz (empty Crawford cell)

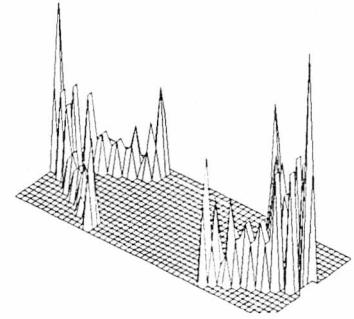
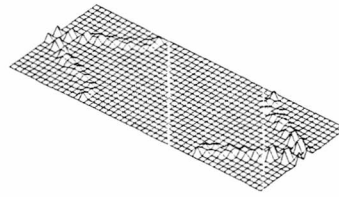
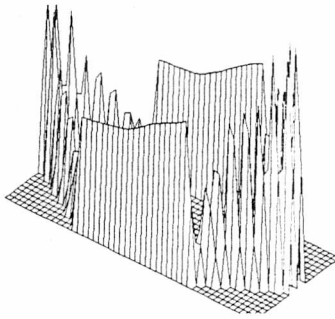


E_x

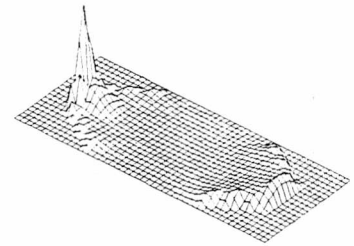
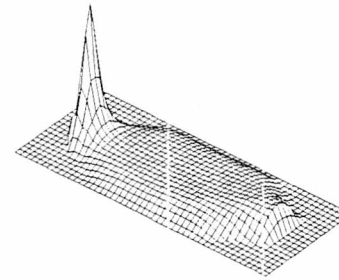
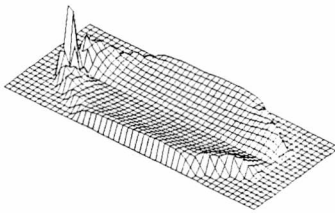
E_y

E_z

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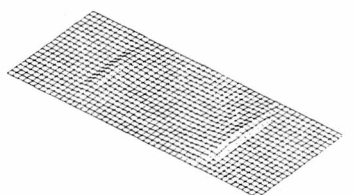
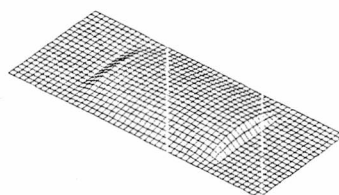
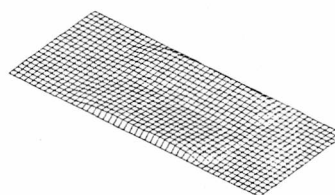


Fig. 6.12c. XZ-electric-field distribution at 30 MHz (empty Crawford cell)

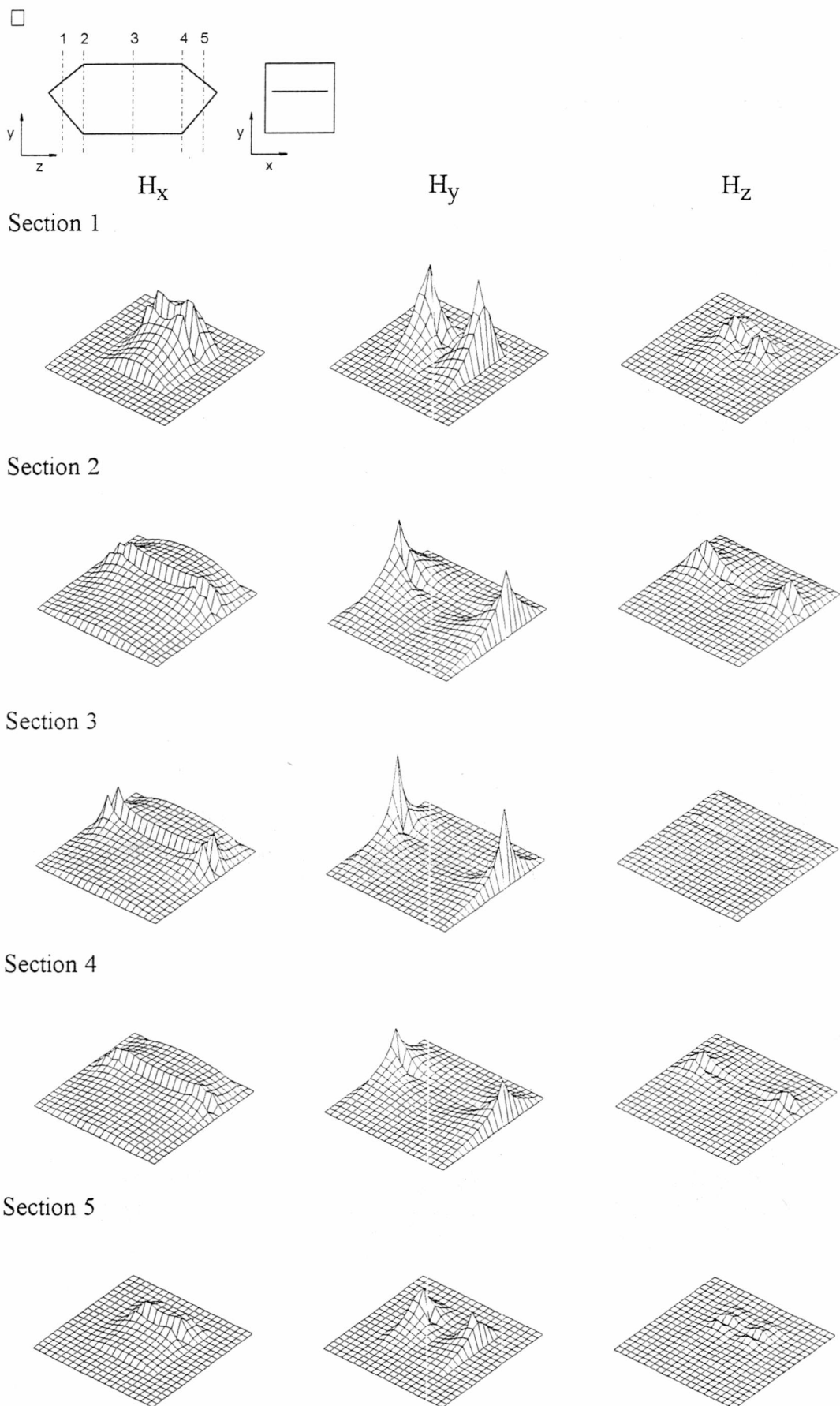
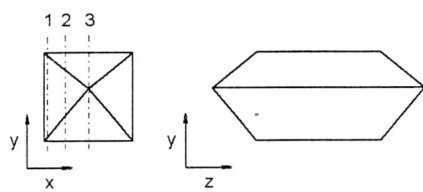


Fig. 6.12d. XY-magnetic-field distribution at 30 MHz (empty Crawford cell)

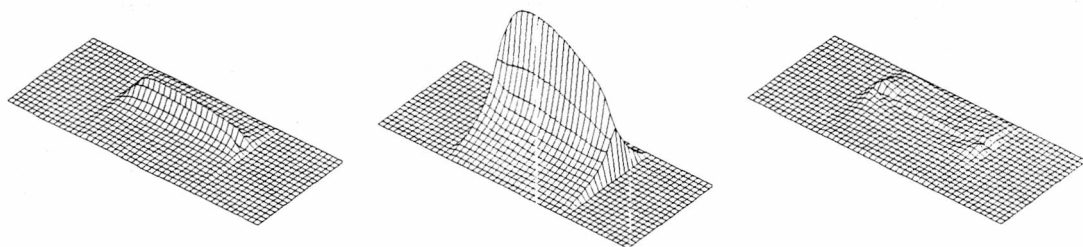


H_x

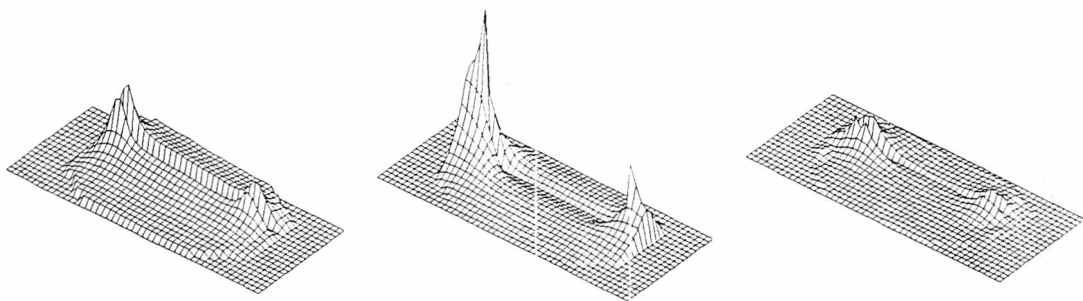
H_y

H_z

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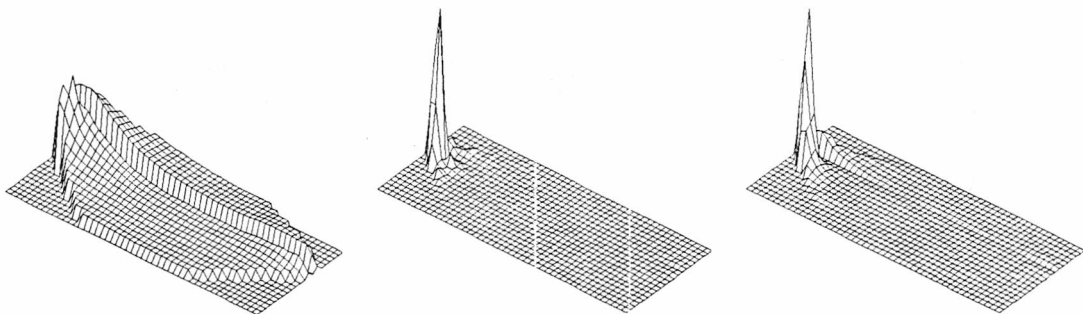
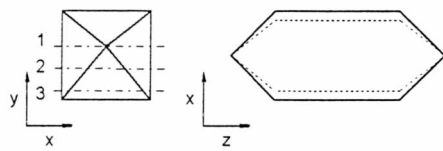


Fig. 6.12e. YZ-magnetic-field distribution at 30 MHz (empty Crawford cell)

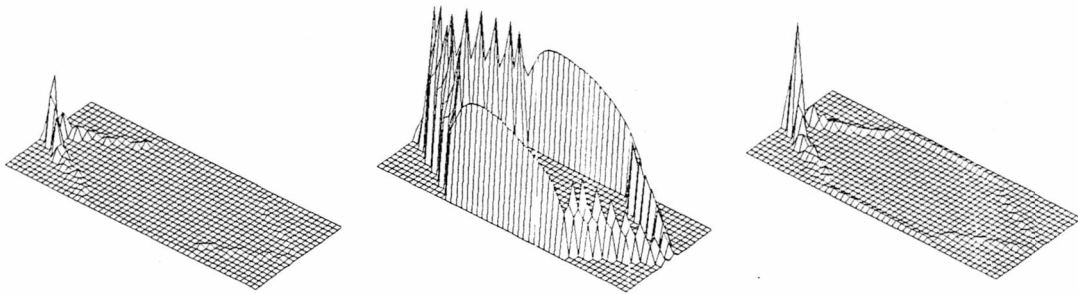


H_x

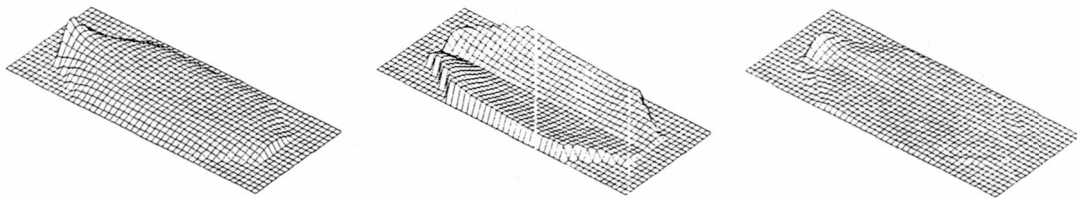
H_y

H_z

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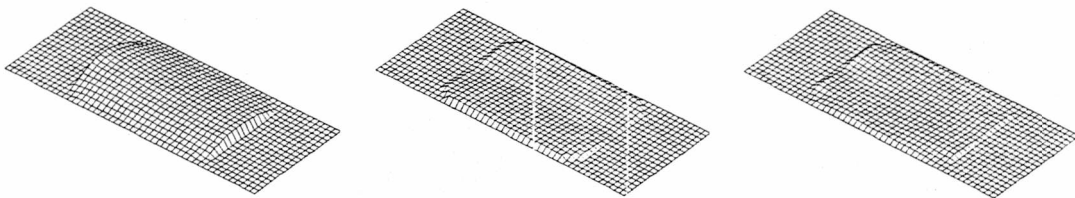


Fig. 6.12f. XZ-magnetic-field distribution at 30 MHz (empty Crawford cell)

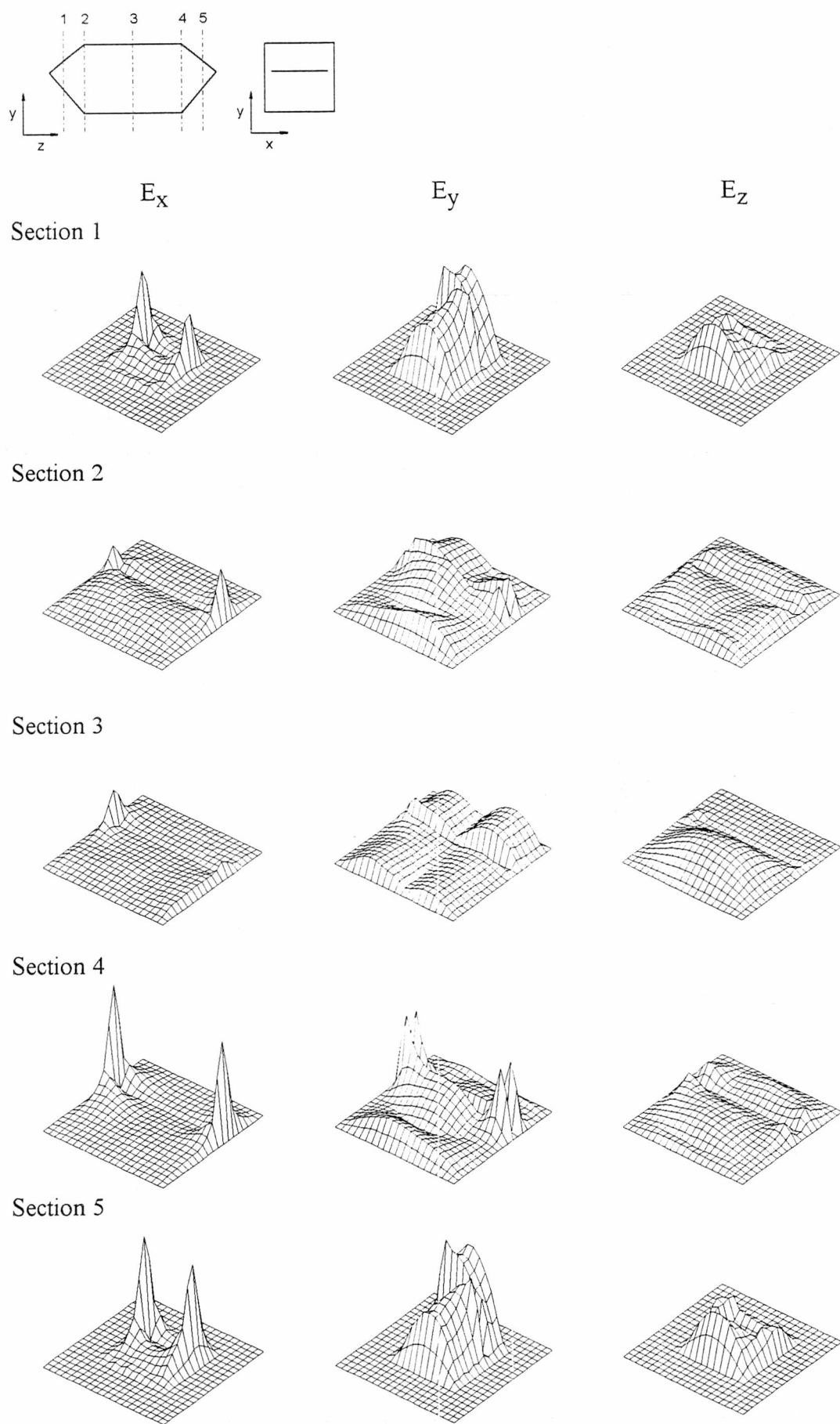


Fig. 6.13a. XY-electric-field distribution at 300 MHz (empty Crawford cell)

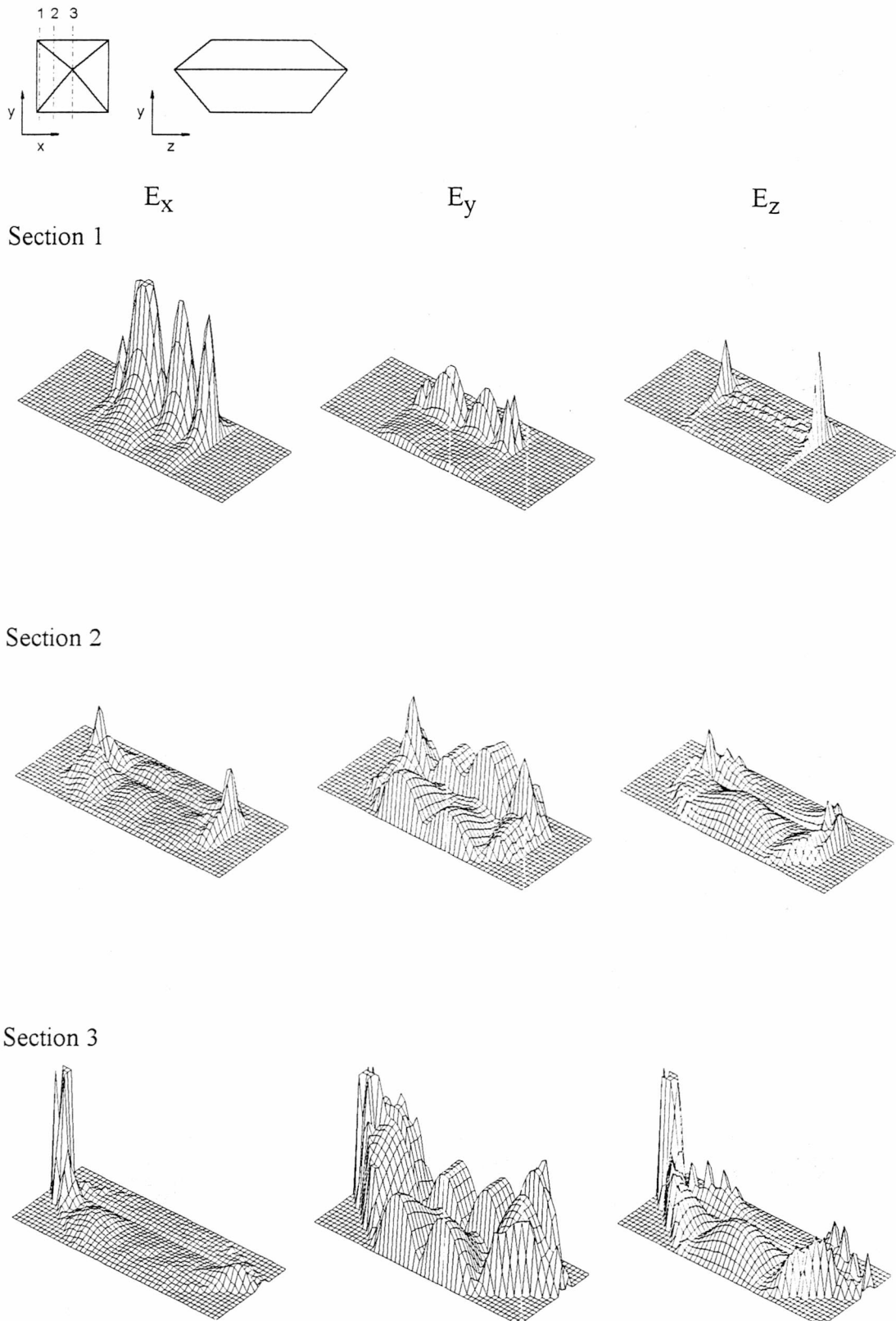


Fig. 6.13b. YZ-electric-field distribution at 300 MHz (empty Crawford cell)

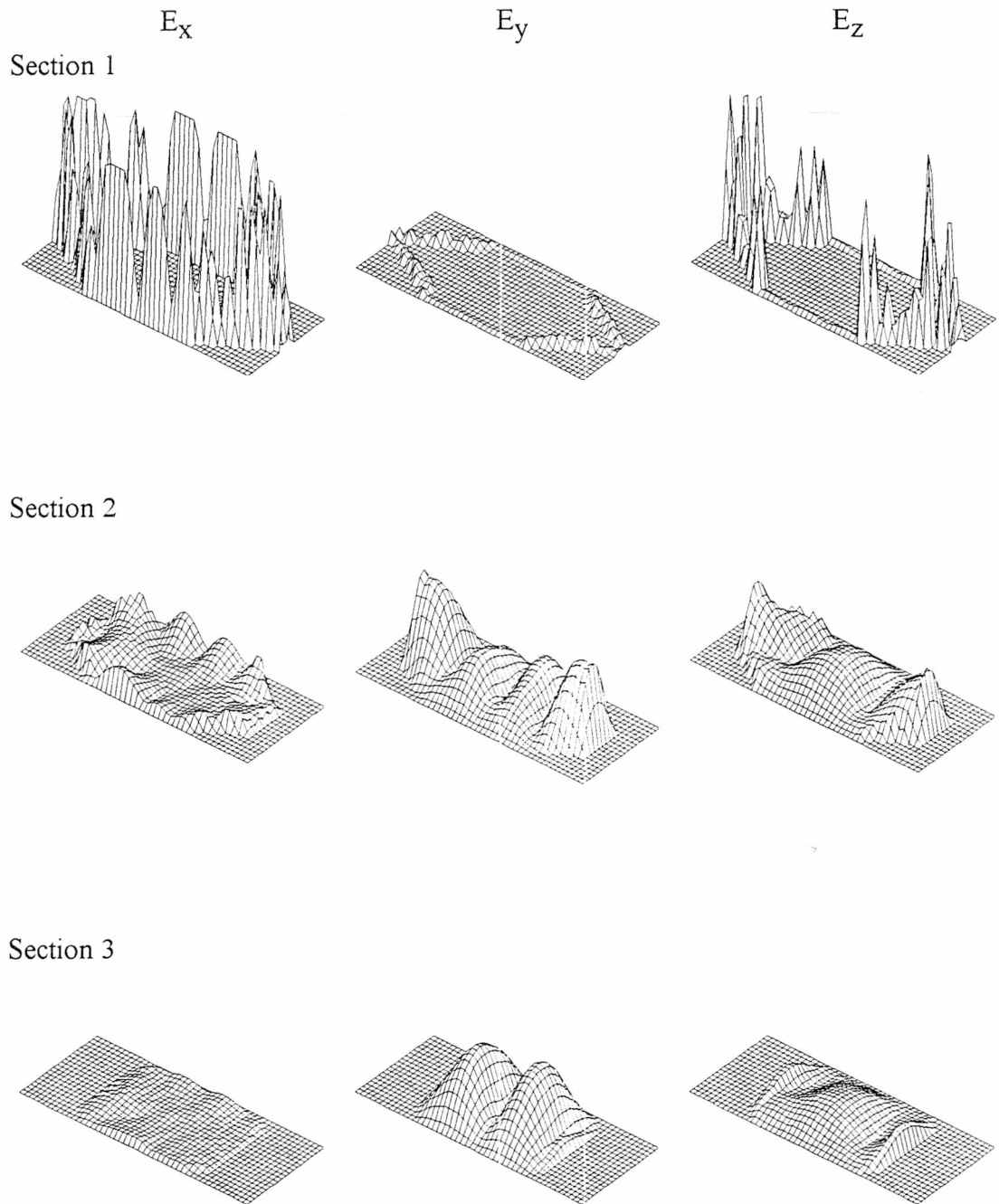
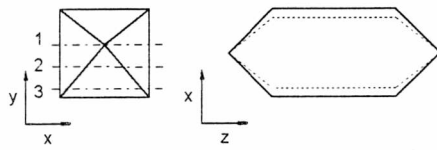


Fig. 6.13c. XZ-electric-field distribution at 300 MHz (empty Crawford cell)

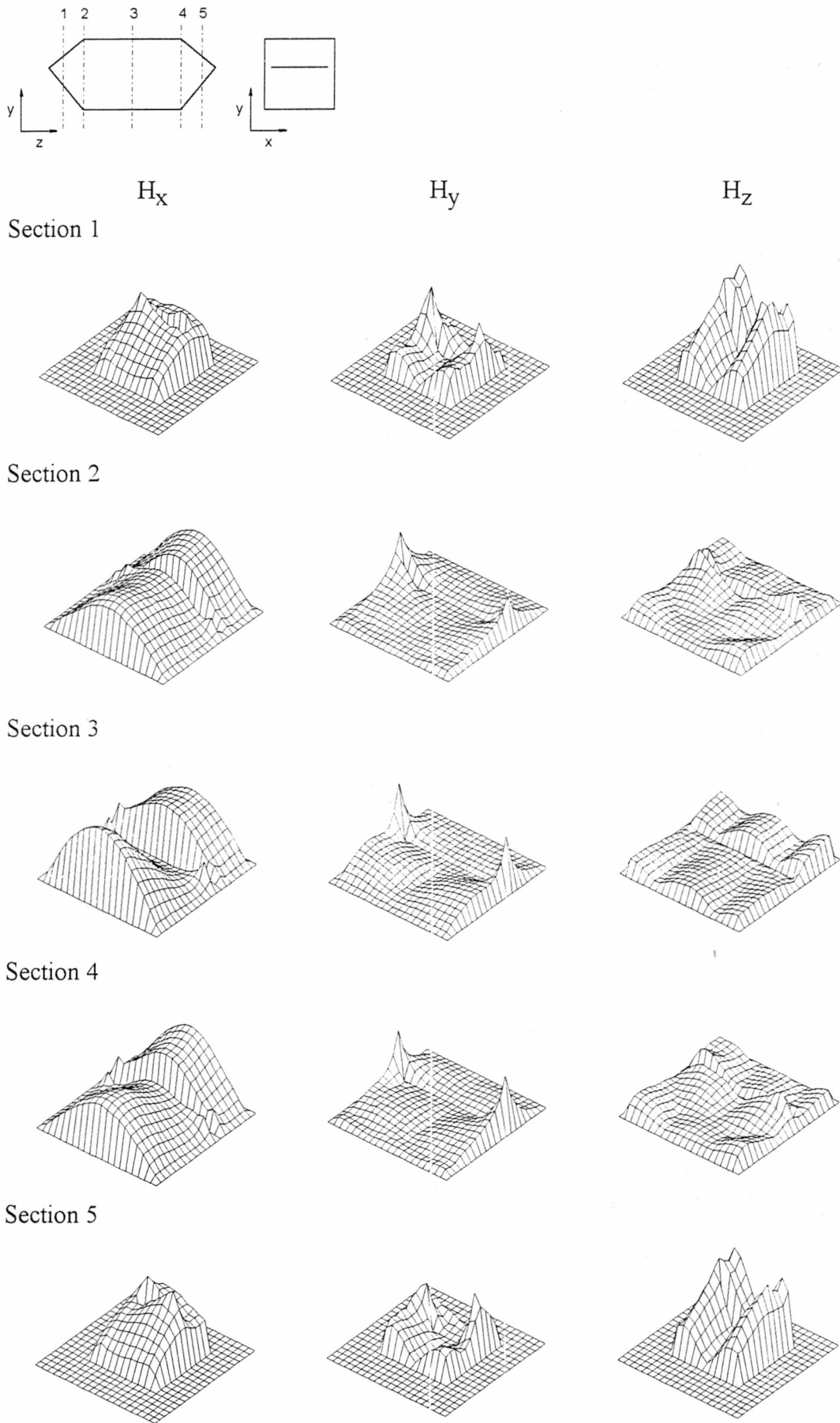
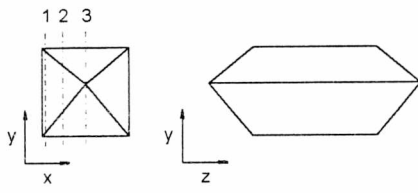


Fig. 6.13d. XY-magnetic-field distribution at 300 MHz (empty Crawford cell)

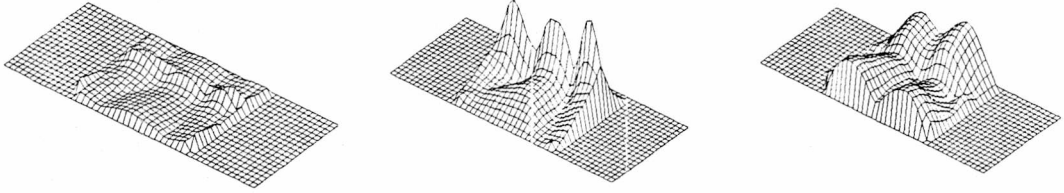


H_x

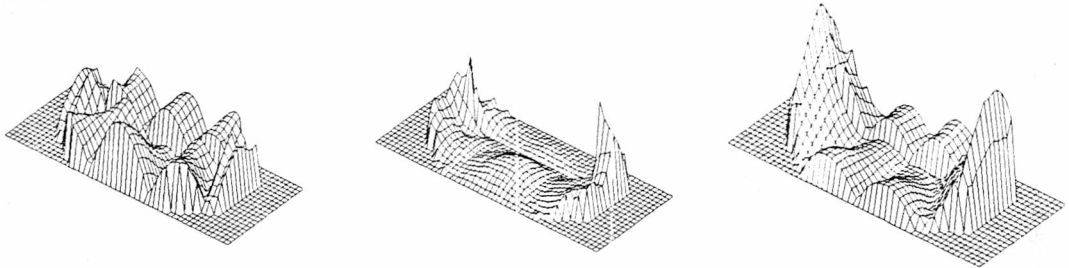
H_y

H_z

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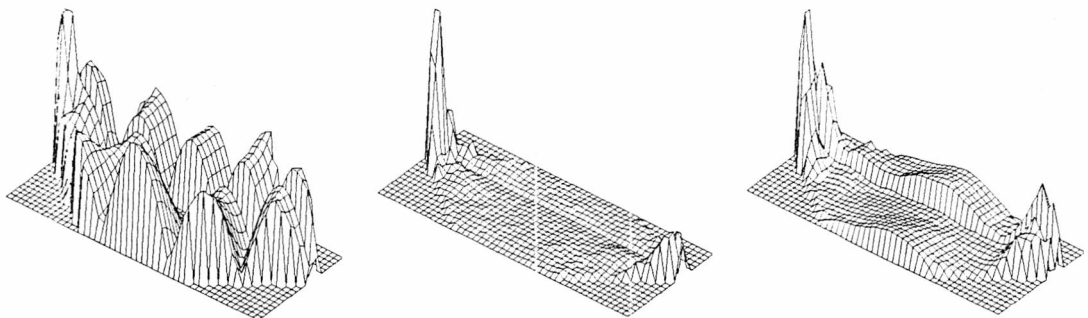
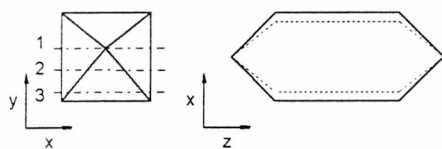


Fig. 6.12e. YZ-magnetic-field distribution at 30 MHz (empty Crawford cell)

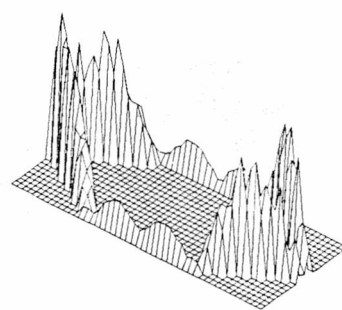
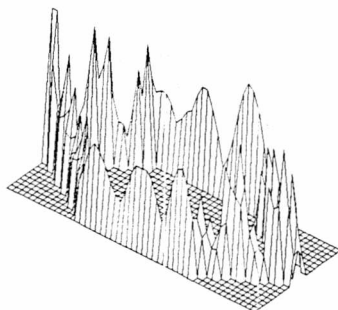
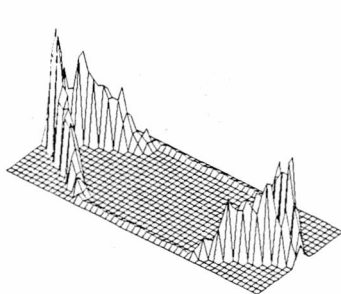


H_x

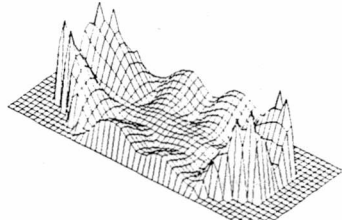
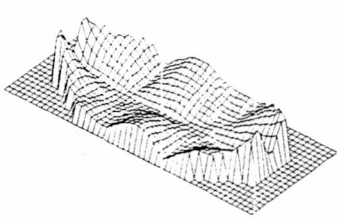
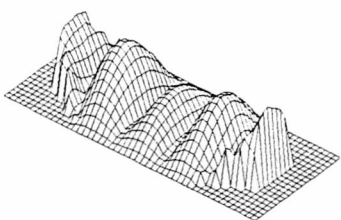
H_y

H_z

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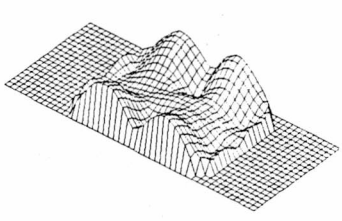
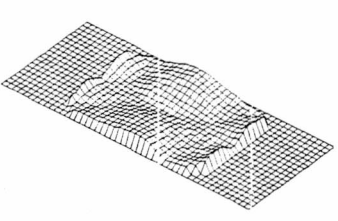
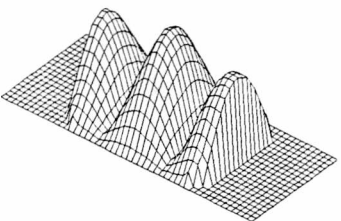


Fig. 6.13f. XZ-magnetic-field distribution at 300 MHz (empty Crawford cell)

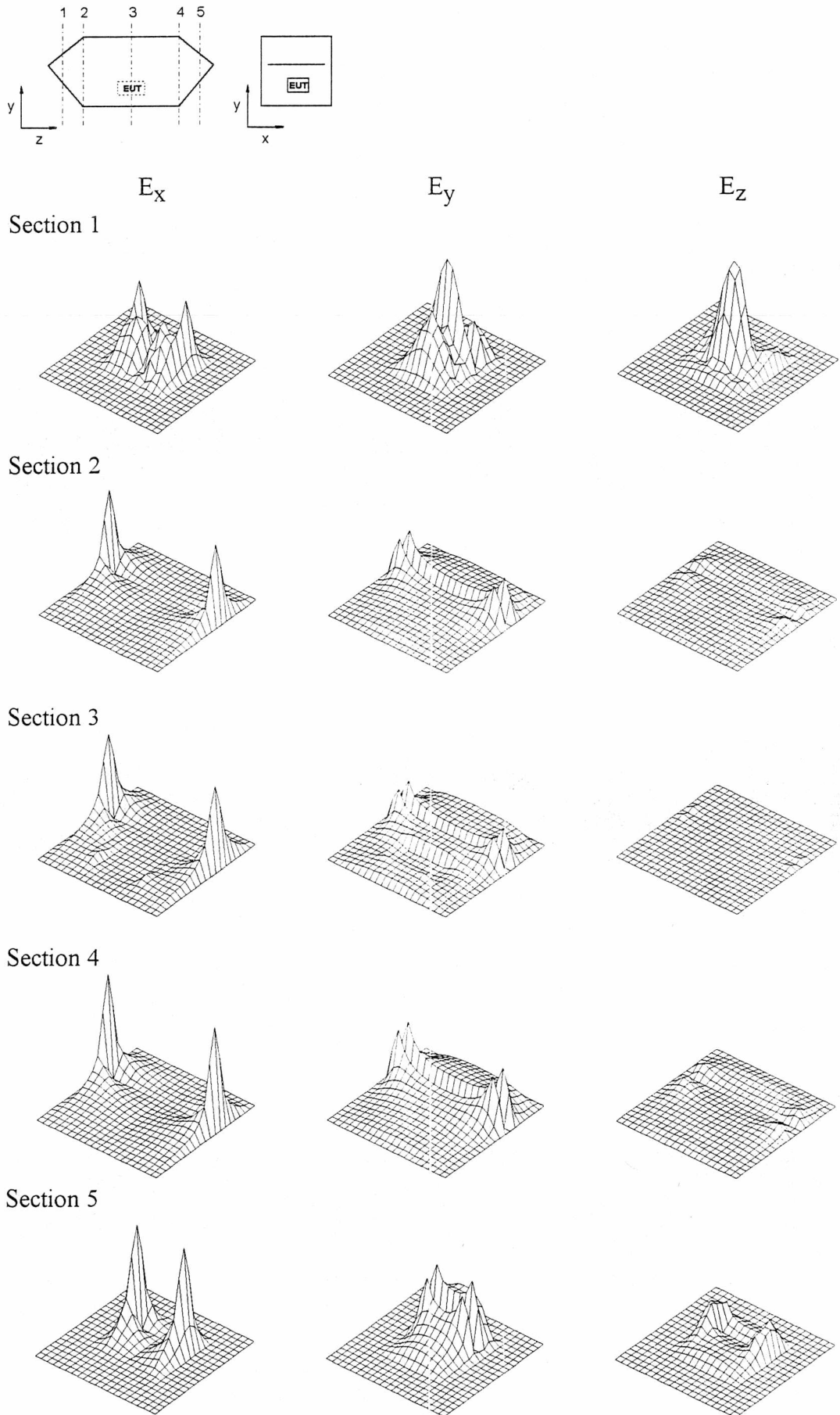
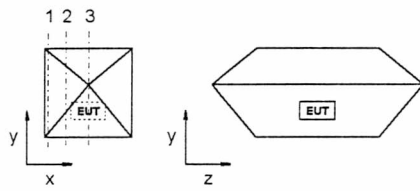


Fig. 6.14a. XY-electric-field distribution at 30 MHz (loaded Crawford cell)

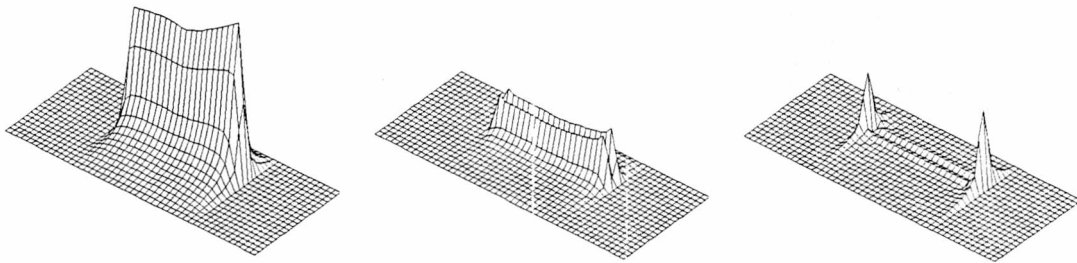


E_x

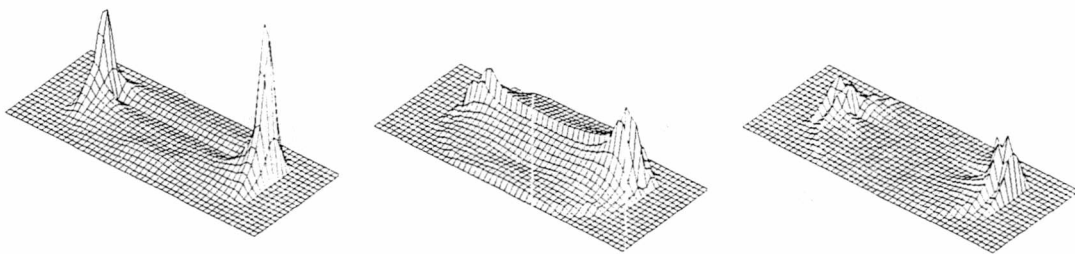
E_y

E_z

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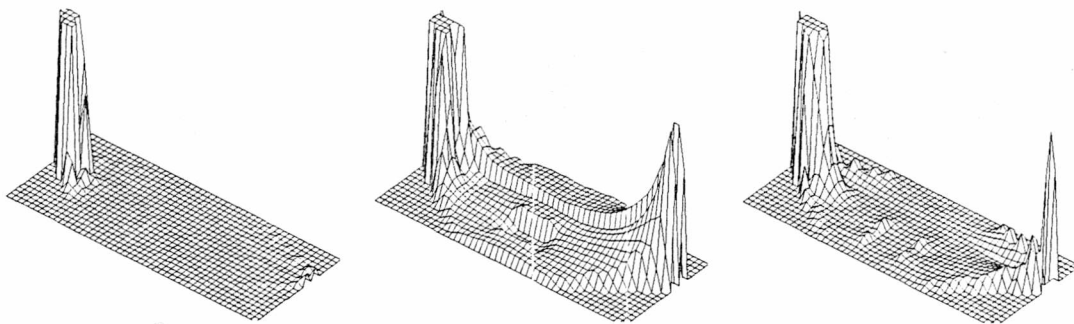


Fig. 6.14b. YZ-electric-field distribution at 30 MHz (loaded Crawford cell)

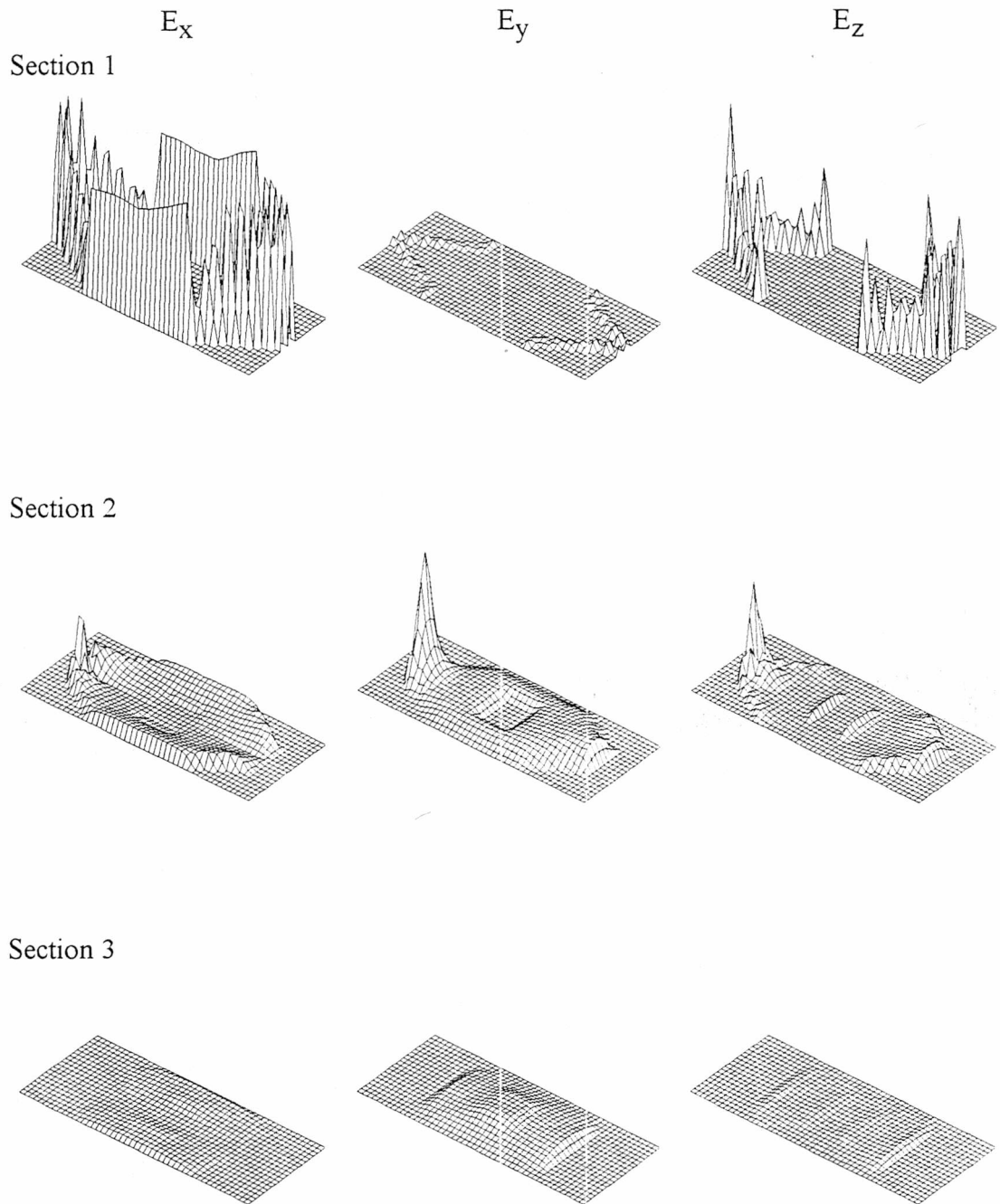
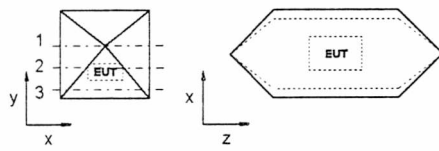


Fig. 6.14c. XZ-electric-field distribution at 30 MHz (loaded Crawford cell)

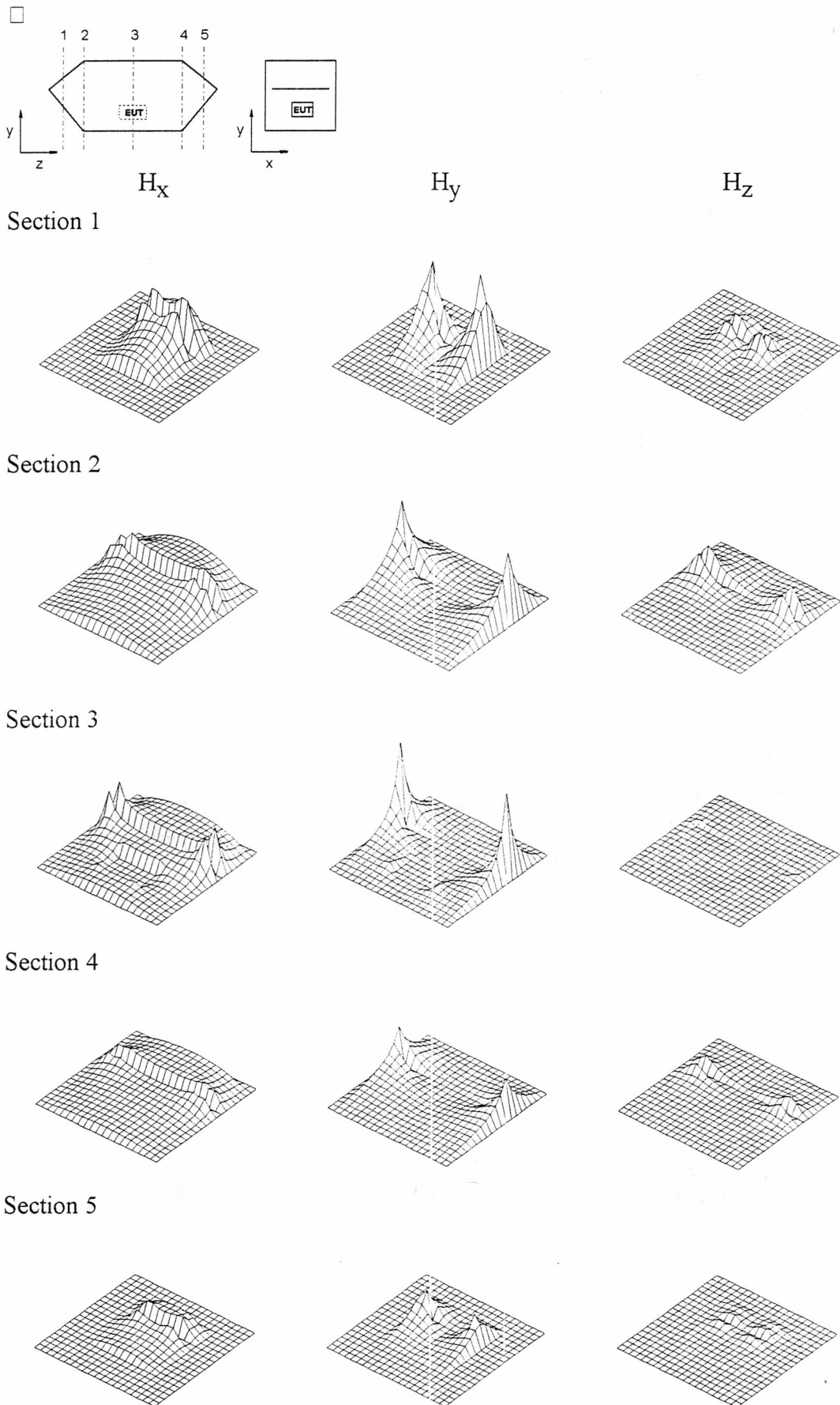
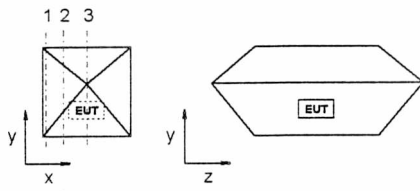


Fig. 6.14d. XY-magnetic-field distribution at 30 MHz (loaded Crawford cell)

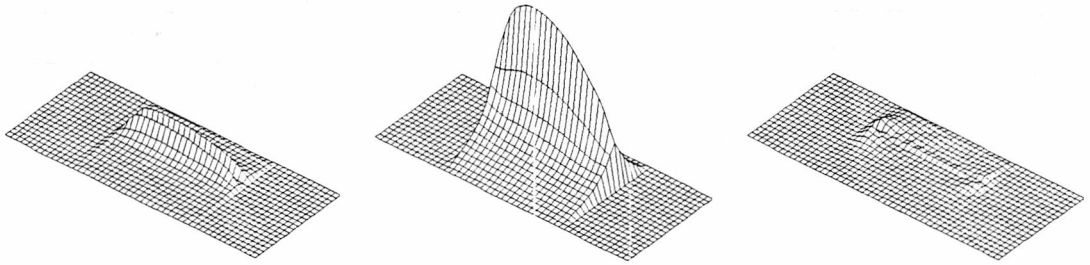


H_x

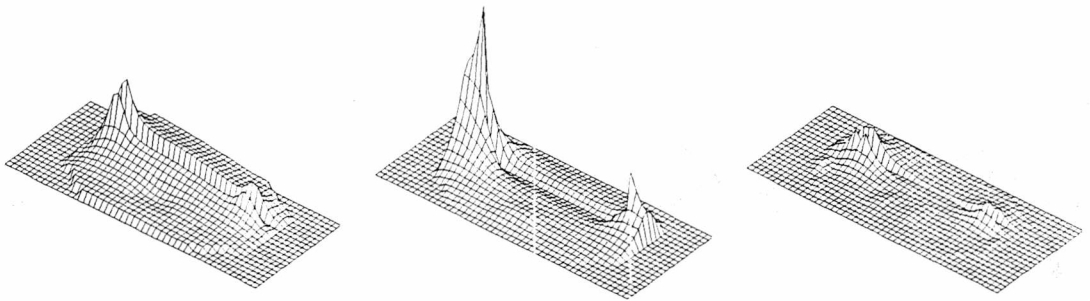
H_y

H_z

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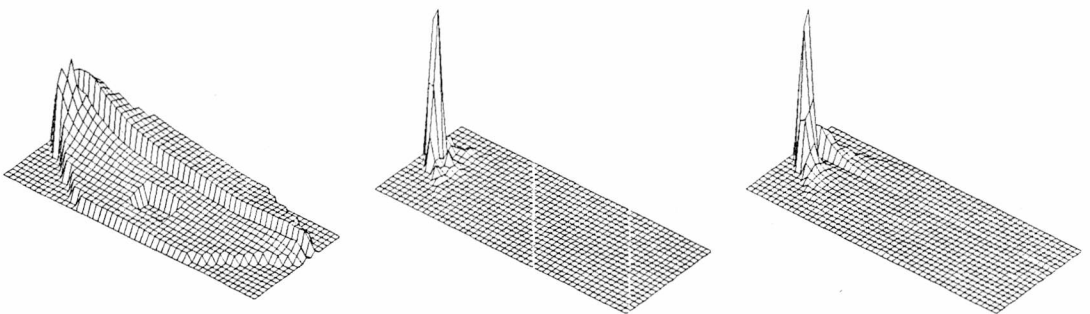
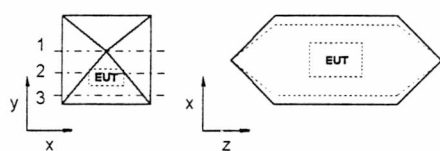


Fig. 6.14e. YZ-magnetic-field distribution at 30 MHz (loaded Crawford cell)

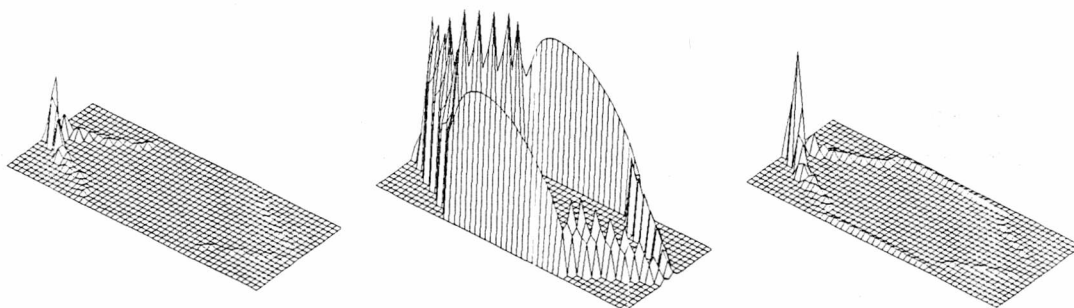


H_x

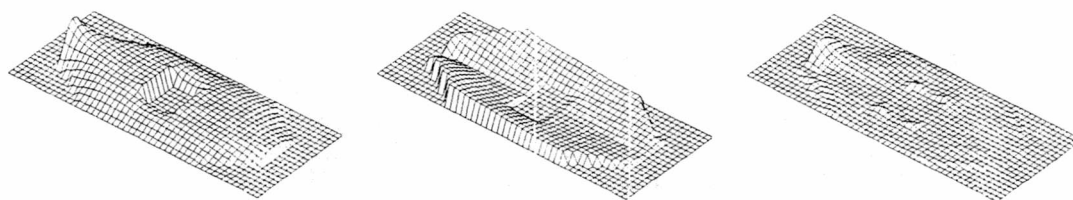
H_y

H_z

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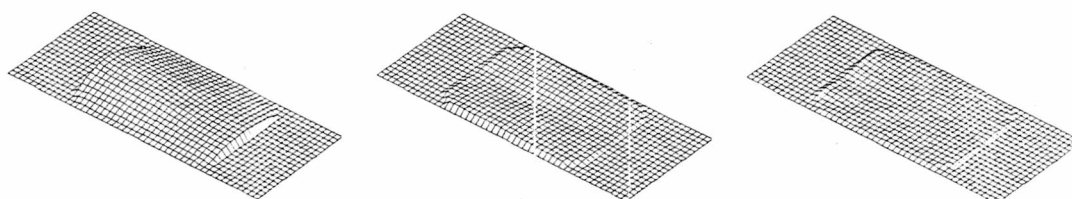


Fig. 6.14f. XZ-magnetic-field distribution at 30 MHz (loaded Crawford cell)

which makes it rather risky to compare measurements results with those obtained in free space. It is expected, that presented methods, after measurement tests, will allow to predict field levels in loaded TEM cells, making measurements with those devices more reliable and comparable to other techniques.

6.2.3. Resonant frequencies of a Crawford cell

Finally, the simulation of impulse propagation inside the TEM cell has been done. The excitation source has again been placed near the input of the cell, and the output (receiving) point - in the middle of the work volume. The pulse response has been transformed into frequency domain using FFT (with windowing technique - Hanning profile). Resulting resonant frequencies are shown in the table 6.1. At this point some measurement verification has been done. The resonant frequencies of the ITA Crawford cell have been measured using Hewlett-Packard HP 8752 A Spectrum Analyser.

The first measured resonant frequency which is approximately 58 MHz has been assumed to be the result of TEM wave reflection (it is well below all cutoff frequencies of higher order modes), has been used to calculate an approximate resonant length of the cell l_{res} which is equal to 2.6 m. This resonant length has been used to calculate resonant frequencies corresponding to various field modes cutoff frequencies, which are presented in Table 6.1 in the column "2D model".

Table 6.1. Resonant frequencies of the ITA Crawford cell

Resonant frequency [MHz]			
measured	calculated with 2D model	calculated with 3D model	resonance type
58	58	54	001
104	94	93	011
134	137	134	012
148	-	-	unknown
153	159	161	102
182	188	185	013
198	198	189	112
205	214	210	103
233	236	-	113
243	243	242	014
262	263	256	104
268	282	269	114
289	290	292	005
303	299	303	015
309	315	307	105
330	331	333	115
337	-	337	unknown
353	355	353	016

The resonance identification in Table 6.1 bases on two-dimensional calculations and on similar work done by Hill for the case of a large TEM cell [20]. Two-dimensional analysis has been done by taking into account first three TE modes (i.e. 01, 10 and 11) cutoff frequencies (Table 5.2) and the calculated above resonant length of the cell. Resonances denoted as $00p$ are the resonances of the TEM wave (standing wave within unmatched TEM line). In many cases the identification is not obvious, because two resonances of different types lie too close to each other. Full, reliable resonance identification requires more sophisticated analysis (especially of the electromagnetic field structure) and is not subject of present work. The same is true about measured resonances described as "unknown". The resonance at 148 MHz does not appear in the theoretical model of the cell. Probably it is caused by elements of the cell not taken into account in the model - for example dielectric support of the septum. Similarly, the mechanism of generation of resonance at 337 MHz is not obvious - therefore calculation with 2D method is difficult - it requires some *a priori* knowledge about the resonance.

The purpose of this chapter is to point out, that such analysis is possible with the use of both two- and three-dimensional TLM methods.

The reader could notice, that three-dimensional analysis of GTEM cell has not been performed. It is because of problems with exact modelling of absorbing walls. The modelling of radio-frequency absorbers using TLM method is possible, but requires machines with great computational power. Also, what is even more important, the composition of absorbing materials is kept secret by producers. The problem of modelling RF absorbers is the subject for separate study.

7. Conclusions

The relatively rapid progress concerning use of TEM cells in electromagnetic compatibility measurements caused the demand of the field analysis inside those devices. Such analysis is so important, because up to now, the only standardised and well-described measurement technique is the measurement in free space. Because neither TEM cells parameters nor measurements setups are described in international regulations, it is extremely important to be able to predict electromagnetic field distribution inside and next to precisely control it while doing measurements. Of course this second task is only possible when the fields structure is known, and phenomena concerning generation of such a field structure are well understood.

The present work is the result of applying the newest techniques of numerical analysis of electromagnetic fields to solve tasks mentioned above. The main purpose of the author was to work out reliable methods of such analysis, which could next be helpful in the works concerning constructing new types of TEM cell, improving existing cells parameters and, in the further future, enabling making comparisons between TEM cells and free space measurements results. The methods should have made it possible to investigate the following factors:

- the characteristic impedance of the cell,
- the cutoff and resonant frequencies of the cell,
- the electromagnetic field distribution both when only TEM wave is propagated, and when higher order field modes occur,
- the influence of cell construction details on cell's parameters,
- the object influence on all cell parameters.

The author has worked out the number of calculation techniques and computer programs that enable above investigations. After careful analysis of numerical techniques used by other authors, who had partially solve above problems (chapter 2), the author had chosen the best (in his opinion) method, which he had next improved and generalised in order to obtain algorithms of the analysis of electromagnetic fields inside TEM cells. In this way the analysis of characteristic impedance both using two- and three-dimensional models has been done, with special attention put on the object (EUT) influence on this parameter. Also the analysis of the field distribution when only basic, TEM mode is propagated inside TEM cell has been successfully done (Chapter 3). The computer programs written by the author were used during ITA GTEM cell construction. It is the pleasure of the author to state, that the GTEM cell built in ITA on the bases of his calculations has excellent electrical parameters comparable to those of the leading EMC centres of the world.

Next, the author has investigated higher order modes of electromagnetic field and the field distribution in the wide frequency range. At this point it has been found out, that no general method of calculating those parameters has been worked out up till now. After review of methods commonly used while solving electromagnetic field problems the author has decided to make use of the relatively new method - the

Transmission-Line Matrix method, which enables both cutoff and resonant frequencies investigation together with determining electromagnetic field distributions. Because there is no one full report concerning the TLM method features, the author decided to shortly describe the TLM method, taking into account the latest tendencies and published improvements (Chapter 4).

The two-dimensional TLM method has been applied to investigate cutoff frequencies and higher order modes distributions in TEM cells (Chapter 5). The results of the analysis have been compared with those obtained by other authors in those special cases when such works are available. An excellent agreement of the results has been achieved. It must be said, that however some methods of analysis were known before [27,64], the method used in this work gives much bigger possibilities, especially while investigating fields distribution and the object presence.

Finally, the full three-dimensional TLM model of the Crawford ITA TEM cell has been presented (Chapter 6). Because up to now it is the only full analysis of the electromagnetic field inside the TEM cell, there is no material to make any comparison. In order to check the validity of the solution some special cases have been taken into account. First, calculations have been done for relatively low frequency at which only TEM mode is propagated. The results have been compared to those obtained for TEM mode in Chapter 3. A good agreement of results has been achieved. Additionally the two-dimensional TLM algorithm tested while determining cutoff frequencies has been used to calculate approximate field distribution along the cell. Still an excellent agreement of results has been obtained, which allows to believe, that the three-dimensional algorithm is correct. Of course, full confirmation of the validity of the results is possible at the measurements way only. However up till now in ITA there is no possibility to make reliable field measurements inside TEM cell, because of the lack of adequate field probes. Such probes are under construction and it is believed that in the future such measurements will be possible. From the above reasons only simple measurements have been done first concerning the characteristic impedance (Chapter 3) and resonance frequencies (Chapter 6) of the cell. In both cases results of measurements agree well with results of calculations.

All computer programs used in the calculations have been written by the author himself on the bases of presented methods. The programming languages used are PASCAL and C. The programs have been used either on IBM PC AT compatible machine or on *DECstation 5000* (under control of the Ultrix operating system).

As his original achievement the author reckons the following:

- working out the first full description of electromagnetic field inside a TEM cell, and their parameters like characteristic impedance, and cutoff and resonant frequencies,
- improving of the methods of numerical analysis of TEM cells in the low frequency range with special attention put on generalisations that enable investigating the presence of an object inside the cell,
- making use of the two-dimensional TLM method to investigate the electromagnetic field distribution of the higher order modes,
- applying full three-dimensional TLM method to model electromagnetic phenomena inside TEM cells.

The valuable attainment of the author is also the working out the set of computer programs that enable testing described methods and making comparisons of the results.

The results of the presented work can be formulated in the form of the following thesis:

It is possible to use computer simulation to investigate electromagnetic field distribution and determine cell parameters both in the case of an empty cell, and in the case of a cell with an EUT inside it.

The present work is a contribution into the subject of modelling electromagnetic fields for the purpose of electromagnetic compatibility. The author thinks, that the work should be continued especially in two fields:

- improving the methods of analysis to make possible investigating more complicated situations, with the special attention put on modelling RF absorbers used in GTEM cell as the load, and in Crawford cells as a way of expanding the usable frequency range,
- verifying results of simulations doing measurements of the fields inside TEM cell, which requires constructing of the special field probes.

The author hopes, that presented methods and results of modelling will help both TEM constructors and users to understand and model the electromagnetic phenomena, that take place in TEM cells.

8. References

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Appendix A - Surface integration of the $1/r$ function

It is assumed that all pieces of surfaces, that are the result of the segmentation process (Chapter 3.3) are small enough to approximate them with the piece of flat plane (in the case of TEM cells, where all electrodes are made of flat metal plates there is no approximation error, but the EUT can be of any shape). Such a flat piece, which is always a polygon, can be subdivided into rectangles or rectangular triangles. Because the integration is a kind of summation, the integral (3.11) can be written as a sum of elementary integrals (omitting $1/4\pi\epsilon_0$):

$$I = \int_{\Delta L} \frac{1}{|\mathbf{r}_m - \mathbf{r}'|} d\mathbf{l}' \quad (\text{A.1})$$

where ΔL denotes elementary rectangles or rectangular triangles.

Because the coordinate system can be easy transformed in order to place the origin in any point of space, and to make the element lie in the xy -plane, the problem can be reformulated in order to make the situation as on one of the following pictures:

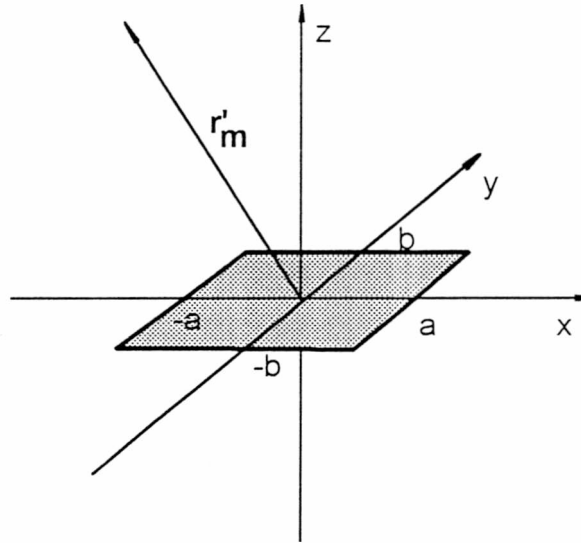


Fig. A.1a. Rectangular piece of surface after coordinate transformation

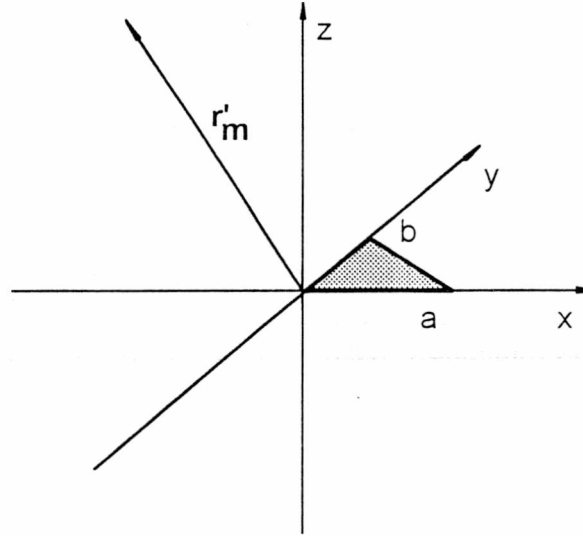


Fig. A.1b. Triangular piece of surface after coordinate transformation

Of course, one can see that each rectangle can be easily subdivided into two triangles, but as it will be seen the rectangle integration is much less complicated, and therefore much faster than triangle integration. Therefore it is more optimal to develop two different integration schemes for those two cases.

A.1. Rectangle integration

Let I be a desired integral. We can write:

$$I = \int_{-a}^a \int_{-b}^b \frac{1}{\sqrt{(x-x_0)^2 + (y-y_0)^2 + z_0^2}} dy dx \quad (\text{A.2})$$

x_0, y_0, z_0 - denote the end of the $\mathbf{r}'_m$ vector.

After internal integration we have:

$$I = \int_{-a}^a \ln \left(b - y_0 + \sqrt{x^2 + 2xx_0 + x_0^2 + z_0^2 + (b - y_0)^2} \right) dx \quad (\text{A.3})$$

$$- \int_{-a}^a \ln \left(-b - y_0 + \sqrt{x^2 + 2xx_0 + x_0^2 + z_0^2 + (b + y_0)^2} \right) dx$$

Both integrals in (A.3) require calculation of the integral of the type

$$\int \ln \left(m + \sqrt{x^2 + kx + l} \right) dx.$$

$$\begin{aligned} \int \ln(m + \sqrt{x^2 + kx + l}) dx &= x \ln(m + \sqrt{x^2 + kx + l}) - \int \frac{x}{m + \sqrt{x^2 + kx + l}} \cdot \frac{2x + k}{2\sqrt{x^2 + kx + l}} dx \\ &= x \ln(m + \sqrt{x^2 + kx + l}) - R \end{aligned} \quad (\text{A.4})$$

The right integral R in (A.4) can be found using substitution:

$$\sqrt{x^2 + kx + l} = t - x \quad (\text{A.5})$$

Then, after recalculation:

$$R = \int \frac{(t^2 - l)(2t^2 + 2kt + k^2 - 2l)}{(t^2 + t(k + 2m) + km + l)(2t + k)^2} dt \quad (\text{A.6})$$

R can be decomposed into the sum of integrals:

$$R = \frac{1}{2} \left(\int 1 dt + \int \frac{A_1}{2t + k} dt + \int \frac{A_2}{(2t + k)^2} dt + \int \frac{Bt + C}{t^2 + t(k + 2m) + km + l} dt \right) \quad (\text{A.7})$$

All four integrals in (A.7) are elementary and can be easily found using standard formulas:

$$R = \frac{1}{2} \left(t + \frac{1}{2} A_1 \ln \left| t + \frac{k}{2} \right| - \frac{1}{4} A_2 \frac{1}{t + \frac{k}{2}} + S \right) \quad (\text{A.8})$$

Calculating S depends on the case:

$$\text{if } z_0 = 0 \quad S = B \ln \left| t + m + \frac{k}{2} \right| - \frac{C - B \left(m + \frac{k}{2} \right)}{t + m + \frac{k}{2}} \quad (\text{A.9a})$$

else

$$S = \frac{1}{2} B \ln |t^2 + t(k+2m) + km + l| + \frac{\left(C - \frac{1}{2} B(k+2m) \right) \arctan \left(\frac{t + \frac{1}{2} k + m}{\sqrt{km + l - \frac{(k+2m)^2}{4}}} \right)}{\sqrt{km + l - \frac{(k+2m)^2}{4}}} \quad (\text{A.9b})$$

Constants A_1, A_2, B, C can be obtained from the set of equations:

$$\begin{bmatrix} 2 & 0 & 4 & 0 \\ 3k+4m & 1 & 4k & 4 \\ 4km+2l+k^2 & k+2m & k^2 & 4k \\ k^2m+kl & km+l & 0 & k^2 \end{bmatrix} \begin{bmatrix} A_1 \\ A_2 \\ B \\ C \end{bmatrix} = \begin{bmatrix} -4k-8m \\ -3k^2-12km-12l \\ -8kl-6k^2m-k^3 \\ 4l^2-3k^2l-k^3m \end{bmatrix} \quad (\text{A10})$$

The above way of calculating the integral I is not correct for the case when the (x_0, y_0, z_0) is placed in the origin of the coordinate system. In this case:

$$I = 2a \ln \left(\frac{\sqrt{a^2 + b^2} + b}{\sqrt{a^2 + b^2} - b} \right) + 2b \ln \left(\frac{\sqrt{a^2 + b^2} + a}{\sqrt{a^2 + b^2} - a} \right) \quad (\text{A.11})$$

A.2. Triangle integration

From the Fig. A.2 we have:

$$I = \int_0^a \int_0^{-\frac{b}{a}x+b} \frac{1}{\sqrt{(x-x_0)^2 + (y-y_0)^2 + z_0^2}} dy dx \quad (\text{A.12})$$

After internal integration:

$$I = \int_0^a \ln \left(-\frac{b}{a}x + b - y_0 + \sqrt{x^2 \left(1 + \frac{b^2}{a^2} \right) - 2x \left(x_0 + \frac{b}{a}(b - y_0) \right) + x_0^2 + z_0^2 + (b - y_0)^2} \right) dx \quad (\text{A.13})$$

$$- \int_{-a}^a \ln \left(-y_0 + \sqrt{x^2 - 2xx_0 + x_0^2 + y_0^2 + z_0^2} \right) dx$$

The second integral in (A.13) is of the type described in the section A.1 and can be solved with the part of the method developed for rectangle integration (of course with different values of m, k, l).

Solving the first integral needs calculating the integral of the type

$$\int \ln \left(mx + n + \sqrt{px^2 + qx + l} \right) dx.$$

$$\int \ln \left(mx + n + \sqrt{px^2 + qx + l} \right) dx = x \ln \left(mx + n + \sqrt{px^2 + qx + l} \right) -$$

$$- \int \frac{x}{mx + n + \sqrt{px^2 + qx + l}} \left(m + \frac{2px + q}{2\sqrt{px^2 + qx + l}} \right) dx \quad (\text{A.14})$$

$$= x \ln \left(mx + n + \sqrt{px^2 + qx + l} \right) - R$$

In order to find the solution the following substitution was used:

$$\sqrt{px^2 + qx + l} = \sqrt{p}(t - x) \quad (\text{A.15})$$

After substitution the right integral R in (A.14) can be written in the form of the sum of two integrals:

$$R_1 = \int \frac{2mp(pt^2 - l)(pt^2 + qt + l)}{(2pt + q)^2(t^2p(m + \sqrt{p}) + t(2pn + q\sqrt{p}) + nq - ml + l\sqrt{p})} dt \quad (\text{A.16a})$$

$$R_2 = \int \frac{\sqrt{p}(pt^2 - l)(2p^2t^2 + 2pqt + q^2 - 2pl)}{(2pt + q)^2(t^2p(m + \sqrt{p}) + t(2pn + q\sqrt{p}) + nq - ml + l\sqrt{p})} dt \quad (\text{A.16b})$$

After decomposition:

$$\begin{aligned} R_1 = & \int \frac{m}{2(m + \sqrt{p})} dt + \frac{1}{2(m + \sqrt{p})} \cdot \left(A_{11} \int \frac{1}{2pt + q} dt + A_{12} \int \frac{1}{(2pt + q)} dt \right. \\ & \left. \int \frac{B_1 t + C_1}{t^2 p(m + \sqrt{p}) + t(2pn + q\sqrt{p}) + nq - ml + l\sqrt{p}} dt \right) = \quad (\text{A.17}) \\ = & \frac{m}{2(m + \sqrt{p})} t + \frac{1}{2(m + \sqrt{p})} (i11(t) + i12(t) + i13(t)) \end{aligned}$$

where:

$$i11(t) = \frac{A_{11}}{2p} \ln \left| t + \frac{q}{2p} \right| \quad (\text{A.18a})$$

$$i12(t) = -\frac{A_{12}}{4p^2 \left(t + \frac{q}{2p} \right)} \quad (\text{A.18b})$$

If $z_0 = 0$

$$i13(t) = \frac{1}{p(m + \sqrt{p})} \left(B_1 \ln |t + E| - \frac{C_1 - B_1 E}{t + E} \right) \quad (\text{A.18c})$$

where

$$E = \frac{2pn + q\sqrt{p}}{2p(m + \sqrt{p})}$$

else

$$i13(t) = \frac{B_1}{2p(m+\sqrt{p})} \ln \left| t^2 p(m+\sqrt{p}) + t(2pn+q\sqrt{p}) + nq - ml + l\sqrt{p} \right| +$$

$$+ \frac{C_1 - B_1 E}{p(m+\sqrt{p}) \sqrt{\frac{nq - ml + l\sqrt{p}}{p(m+\sqrt{p})} - E^2}} \cdot \arctan \left(\frac{t + E}{\sqrt{\frac{nq - ml + l\sqrt{p}}{p(m+\sqrt{p})} - E^2}} \right) \quad (\text{A.18d})$$

where E the same as above.

Constants A_{11} , A_{12} , B_1 , C_1 can be obtained from the set of equations:

$$\begin{bmatrix} 2p^2(m+\sqrt{p}) & 0 & 4p^2 & 0 \\ 2p(2pn+q\sqrt{p}) + pq(m+\sqrt{p}) & p(m+\sqrt{p}) & 4pq & 4p^2 \\ 2p(nq - ml + l\sqrt{p}) + q(2pn+q\sqrt{p}) & 2pn+q\sqrt{p} & q^2 & 4pq \\ q(nq - ml + l\sqrt{p}) & nq - ml + l\sqrt{p} & 0 & q^2 \end{bmatrix} \cdot \begin{bmatrix} A_{11} \\ A_{12} \\ B_1 \\ C_1 \end{bmatrix} =$$

$$= \begin{bmatrix} -8mnp^3 - 4mp^2q\sqrt{p} \\ -mp(12pq n + 4pl\sqrt{p} + 5q^2\sqrt{p} - 4pml + q^2m) \\ -8pqml\sqrt{p} - 6pq^2mn - mq^3\sqrt{p} \\ -4pml^2(m+\sqrt{p}) - mq^2(nq - ml + l\sqrt{p}) \end{bmatrix} \quad (\text{A.19})$$

Similarly:

$$\begin{aligned}
 R_2 &= \int \frac{\sqrt{p}}{2(m+\sqrt{p})} dt + \frac{1}{2(m+\sqrt{p})} \cdot \left(A_{21} \int \frac{1}{2pt+q} dt + A_{22} \int \frac{1}{(2pt+q)} dt \right. \\
 &\quad \left. \int \frac{B_2 t + C_2}{t^2 p(m+\sqrt{p}) + t(2pn+q\sqrt{p}) + nq - ml + l\sqrt{p}} dt \right) = \quad (A.20) \\
 &= \frac{\sqrt{p}}{2(m+\sqrt{p})} t + \frac{1}{2(m+\sqrt{p})} (i21(t) + i22(t) + i23(t))
 \end{aligned}$$

$$i21(t) = \frac{A_{21}}{2p} \ln \left| t + \frac{q}{2p} \right| \quad (A.21a)$$

$$i22(t) = -\frac{A_{22}}{4p^2 \left(t + \frac{q}{2p} \right)} \quad (A.21b)$$

If $z_0 = 0$

$$i23(t) = \frac{1}{p(m+\sqrt{p})} \left(B_2 \ln |t+E| - \frac{C_2 - B_2 E}{t+E} \right) \quad (A.21c)$$

where

$$E = \frac{2pn+q\sqrt{p}}{2p(m+\sqrt{p})}$$

else

$$i23(t) = \frac{B_2}{2p(m+\sqrt{p})} \ln \left| t^2 p(m+\sqrt{p}) + t(2pn+q\sqrt{p}) + nq - ml + l\sqrt{p} \right| +$$

$$+ \frac{C_2 - B_2 E}{p(m+\sqrt{p}) \sqrt{\frac{nq - ml + l\sqrt{p}}{p(m+\sqrt{p})} - E^2}} \cdot \arctan \left(\frac{t + E}{\sqrt{\frac{nq - ml + l\sqrt{p}}{p(m+\sqrt{p})} - E^2}} \right) \quad (\text{A.21d})$$

where E the same as above.

Constants A_{21} , A_{22} , B_2 , C_2 can be obtained from the set of equations:

$$\begin{bmatrix} 2p^2(m+\sqrt{p}) & 0 & 4p^2 & 0 \\ 2p(2pn+q\sqrt{p}) + pq(m+\sqrt{p}) & p(m+\sqrt{p}) & 4pq & 4p^2 \\ 2p(nq - ml + l\sqrt{p}) + q(2pn+q\sqrt{p}) & 2pn+q\sqrt{p} & q^2 & 4pq \\ q(nq - ml + l\sqrt{p}) & nq - ml + l\sqrt{p} & 0 & q^2 \end{bmatrix} \begin{bmatrix} A_{21} \\ A_{22} \\ B_2 \\ C_2 \end{bmatrix} =$$

$$= \begin{bmatrix} -4p^3(q+2n\sqrt{p}) \\ pq^2m\sqrt{p} - 3p^2q^2 - 4mp^2l\sqrt{p} - 12p^3l - 12p^2qn\sqrt{p} \\ -8p^2ql - 6p^2ql - 6pq^2n\sqrt{p} - pq^3 \\ 4mpl^2\sqrt{p} + 4p^2l^2 - mlq^2\sqrt{p} - 3pq^2l - q^3n\sqrt{p} \end{bmatrix} \quad (\text{A.22})$$

Above integration schemes were applied in computer program used to calculate the S matrix elements (Chapter 3). The comparison of results obtained from above equations and from purely numerical integration shown the full compatibility of both methods, however numerical calculations with enough level of accuracy were about six times longer.

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POLITECHNIKA WROCLAWSKA
INSTYTUT TELEKOMUNIKACJI I AKUSTYKI

Raport nr I-28/PRE~~00~~794

NUMERYCZNE MODELOWANIE KOMÓR TEM
Dodatek do pracy doktorskiej

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Promotor: Prof. dr hab. inż. Daniel J. Bem

Słowa kluczowe:

kompatybilność elektromagnetyczna,
pole elektromagnetyczne,
metody numeryczne,
modelowanie komputerowe,
komora TEM

WROCLAW 1994

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1. Wstęp

Celem niniejszego "Dodatku" jest przedstawienie dodatkowych wyników obliczeń uzasadniających dokładność i poprawność metod zastosowanych w pracy "Numerical modelling of TEM cells". Szerzej omówiono także dodatkowe możliwości modelu TLM, takie jak modelowanie obiektów stratnych, czy źródłowych. Przedstawiono opis pomiarów, których wyniki zamieszczono w rozprawie, a także zaprezentowano koncepcję sond pomiarowych, które umożliwią pełną eksperymentalną weryfikację wyników obliczeń.

2. Dokładność wyników uzyskiwanych metodą opisaną w rozdziale 3. rozprawy

W rozdziale 3 rozprawy doktorskiej zaproponowano stosowanie metody momentów do rozwiązywania dwu- i trójwymiarowego równania Laplace'a. Metodę tę zastosowano w odniesieniu do całki równania Laplace'a. Rozwiązanie to ma następujące zalety w stosunku do innych metod stosowanych w rozwiązywaniu podobnych zagadnień:

- 1) Znaczne zredukowanie liczby zmiennych w przypadku obliczeń trójwymiarowych. W metodach siatkowych liczba zmiennych jest proporcjonalna do objętości modelowanego obszaru; w przypadku zastosowanej techniki poszukuje się rozkładu źródeł na brzegu obszaru, dzięki czemu liczba zmiennych proporcjonalna jest do pola powierzchni metalicznych. Ponadto nie występuje tu konieczność segmentacji i zadawania warunków brzegowych w płaszczyznach symetrii - warunek Neumana można spełnić przez sumowanie odpowiednich elementów macierzy opisującej układ. Dzięki temu zastosowana metoda jest również bardziej optymalna niż metoda elementów granicznych (BEM) - szacunki liczbowe dotyczące liczby zmiennych przedstawiono w rozprawie na stronie 38.
- 2) Łatwe modelowanie obszarów otwartych. Metody siatkowe z zasady działania wymagają zdefiniowania skończonych granic modelowanego obszaru oraz zadania dla nich odpowiednich warunków brzegowych. W związku z tym modelowanie powierzchni idealnie bezodbiciowych jest związane z poważnymi trudnościami metodologicznymi, a istniejące rozwiązania opierają się na różnego rodzaju przybliżeniach. Oznacza to, że o ile za pomocą np. metody elementów skończonych (FEM) można stosunkowo dobrze zamodelować zamkniętą komorę TEM, to ewentualne modelowanie np. otworu w komorze TEM, lub symulatora równoległopłytowego wymaga użycia bardzo skomplikowanych (i pochłaniających czas i pamięć komputera) technik. W zastosowanej metodzie momentów problemy te nie występują, gdyż zamiast rozwiązywania równania Laplace'a dla potencjału w obszarze nieograniczonym, poszukuje się najpierw rozkładu źródeł na brzegu tego obszaru, a następnie według prostego wzoru całkowego oblicza się potencjał w dowolnym punkcie przestrzeni.
- 3) Pomimo przyjęcia aproksymacji rozkładu źródeł na powierzchniach przewodzących w postaci kombinacji liniowych impulsowych funkcji prostokątnych, rozwiązania uzyskiwane metodą momentów umożliwiają dokładniejszą analizę pola w obszarach położonych daleko od źródeł (takich jak przestrzeń robocza komory TEM) - są one bardziej "fizyczne", gdyż rozkłady pola nie są aproksymowane zestawem funkcji bazowych (zazwyczaj będących funkcjami liniowymi, co oznacza, że rozkłady pola są rysowane jako linie łamane).

- 4) Nierównomierny podział powierzchni umożliwia znaczną poprawę dokładności obliczania rozkładu źródeł w rejonach krytycznych - np. na ostrych krawędziach.
- 5) Można modelować zarówno przewodniki o skończonej grubości, jak i nieskończenie cienkie, bez trudności związanych z niestabilnością rozwiązań. Oczywiście zastosowanie powierzchni nieskończenie cienkich oznacza dopuszczenie niefizycznych rozwiązań, w których występują punkty osobliwe. Metoda momentów uśrednia jednakże te osobliwości, dzięki czemu problemy związane np. z nieskończonymi gęstościami ładunku nie występują. Trudności takie występują natomiast w metodach siatkowych - np. autorzy programu High-Frequency Structure Simulator ostrzegają ([1], str. 5-13), że program ten nie będzie działał poprawnie w przypadku obiektów o znacznych różnicach rozmiarów w poszczególnych kierunkach i zalecają stosowanie struktur nieskończenie cienkich bez możliwości przeanalizowania popełnianych w ten sposób błędów.

W zaawansowanych metodach numerycznych dotyczących analizy pól elektromagnetycznych istnieją cztery metody weryfikacji poprawności uzyskiwanych wyników:

- porównanie z rozwiązaniami analitycznymi (jeżeli takie istnieją),
- porównanie z wynikami uzyskanymi za pomocą innych metod numerycznych,
- sprawdzenie, czy otrzymane rozwiązania w sposób satysfakcjonujący spełniają założenia wejściowe (np. zadane warunki brzegowe),
- weryfikacja pomiarowa.

Należy zaznaczyć, że wykonywane analizy pól dotyczą często zjawisk niemierzalnych, dlatego weryfikację pomiarową można zastosować stosunkowo rzadko.

W niniejszym punkcie przedstawiono weryfikację za pomocą sposobów pierwszego i trzeciego. Porównanie z wynikami uzyskiwanymi za pomocą w pełni ogólnej zastosowanej w pracy metody TLM oraz profesjonalnego pakietu High-Frequency Structure Simulator przedstawiono w punkcie 3 "Dodatku".

2.1. Rozkład ładunków na powierzchni doskonale przewodzącego przewodu taśmowego

W celu sprawdzenia dokładności metody momentów, zastosowano ją do obliczenia rozkładu źródeł dla nieskończenie cienkiego, odosobnionego przewodu taśmowego o szerokości 1 m, naładowanego do potencjału 1 V.

Rozwiązanie analityczne dla rozpatrywanego przypadku ma postać [Janiszewski, Karwowski, 26]:

$$\frac{q(x)}{\varepsilon_0} = \frac{2}{w \cdot \ln(2/w)} \frac{V_0}{\sqrt{1-(x/w)^2}}, \quad (2.1)$$

przy czym: w - połowa szerokości przewodu,
 V_0 - potencjał przewodu.

Obliczeń dokonano z zastosowaniem różnej liczby segmentów. Oszacowano:

- niedokładność metody w punktach, dla których obliczana jest gęstość ładunków;
- błąd związany z uśrednieniem osobliwości rozwiązania na krawędziach przewodu.

Tablica 2.1
 Porównanie rozkładu gęstości ładunku (liczba segmentów $N=10$)

Numer segmentu	Unormowana gęstość ładunku $\frac{q(x)}{\varepsilon_0}$		Błąd [%]
	wzór (2.1)	metoda momentów	
1	10.242	13.841	35.15
2	5.337	5.522	3.46
3	3.818	3.873	1.45
4	3.172	3.187	0.46
5	2.915	2.919	0.13

Tablica 2.2
 Porównanie rozkładu gęstości ładunku (liczba segmentów $N=20$)

Numer segmentu	Unormowana gęstość ładunku $\frac{q(x)}{\varepsilon_0}$		Błąd [%]
	wzór (2.1)	metoda momentów	
1	19.36	26.36	36.16
2	9.789	10.209	4.29
3	6.650	6.795	2.18
4	5.122	5.178	1.08
5	4.242	4.268	0.59
6	3.691	3.703	0.34
7	3.332	3.338	0.19
8	3.100	3.103	0.10
9	2.960	2.961	0.05
10	2.893	2.894	0.03

Tablica 2.3
Porównanie rozkładu gęstości ładunku (liczba segmentów N=50)

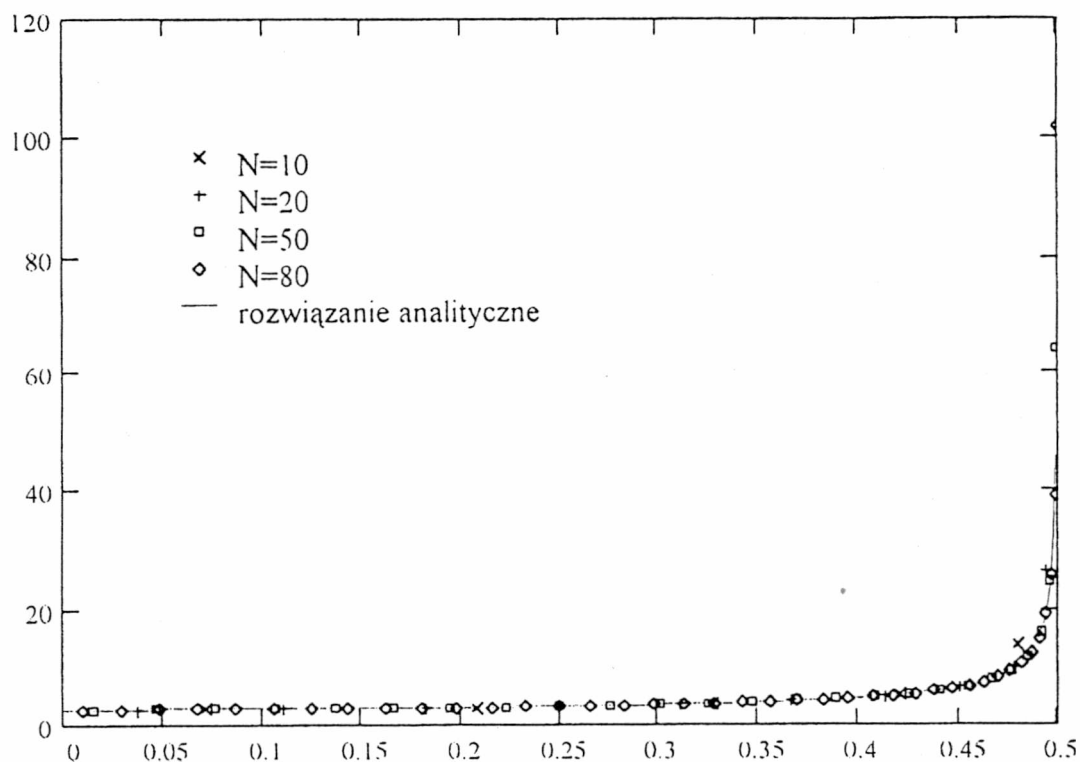
Numer segmentu	Unormowana gęstość ładunku $\frac{q(x)}{\varepsilon_0}$		Błąd [%]
	wzór (2.1)	metoda momentów	
1	46.870	63.973	36.49
2	23.480	24.553	4.57
3	15.703	16.085	2.44
4	11.830	11.987	1.33
5	9.518	9.598	0.84
6	7.987	8.033	0.57
7	6.904	6.931	0.40
8	6.099	6.117	0.29
9	5.481	5.493	0.23
10	4.994	5.003	0.17
11	4.602	4.609	0.14
12	4.283	4.287	0.11
13	4.019	4.022	0.08
14	3.800	3.802	0.07
15	3.616	3.618	0.05
16	3.461	3.463	0.04
17	3.332	3.333	0.03
18	3.223	3.224	0.03
19	3.133	3.134	0.02
20	3.059	3.060	0.01
21	3.000	3.000	0.01
22	2.954	2.954	0.01
23	2.920	2.920	0.01
24	2.898	2.898	0.00
25	2.887	2.887	0.00

Tablica 2.4

Porównanie rozkładu gęstości ładunku (liczba segmentów N=80)

Numer segmentu	Unormowana gęstość ładunku $\frac{q(x)}{\varepsilon_0}$		Błąd [%]
	wzór (2.1)	metoda momentów	
1	74.413	101.596	36.53
2	37.234	38.949	4.61
3	24.854	25.467	2.47
4	18.673	18.928	1.36
5	14.973	15.103	0.87
6	12.512	12.587	0.59
7	10.759	10.806	0.43
8	9.450	9.481	0.33
9	8.436	8.458	0.26
10	7.629	7.645	0.20
11	6.973	6.984	0.16
12	6.429	6.438	0.14
13	5.973	5.979	0.11
14	5.584	5.590	0.09
15	5.251	5.255	0.08
16	4.962	4.965	0.07
17	4.710	4.713	0.06
18	4.489	4.491	0.05
19	4.294	4.296	0.04
20	4.121	4.122	0.04
21	3.967	3.968	0.03
22	3.830	3.831	0.03
23	3.707	3.708	0.02
24	3.597	3.598	0.02
25	3.499	3.499	0.02
26	3.411	3.411	0.02
27	3.332	3.332	0.01
28	3.261	3.262	0.01
29	3.198	3.199	0.01
30	3.142	3.143	0.01
31	3.093	3.093	0.01
32	3.050	3.050	0.01
33	3.012	3.012	0.00
34	2.980	2.980	0.00
35	2.952	2.952	0.00
36	2.930	2.930	0.00
37	2.912	2.912	0.00
38	2.899	2.899	0.00
39	2.890	2.890	0.00
40	2.886	2.886	0.00

Przedstawione wyniki obliczeń zilustrowano wykresem pokazanym na rysunku 2.1. Zarówno w tablicach 2.1 - 2.4, jak i na rysunku 2.1 pokazano wyniki dla połowy szerokości przewodu taśmowego (ze względu na symetrię zagadnienia)



Rys. 2.1. Rozkład ładunku na powierzchni przewodu taśmowego

Jak widać, przy zwiększaniu liczby segmentów (podział Czebyszewa) największy błąd, dotyczący segmentów końcowych (osobliwość), pozostaje w przybliżeniu stały. Biorąc pod uwagę, że błąd ten dotyczy coraz krótszych odcinków przewodnika, oznacza to, że całkowity błąd może być zminimalizowany do zadanej wartości przez zastosowanie odpowiednio subtelnej segmentacji. Efekt ten ilustrują tablice 2.5 oraz 2.6.

Tablica 2.5
Unormowany ładunek zgromadzony na powierzchni skrajnych segmentów

Liczba segmentów	Szerokość segmentu [m]	Ładunek obliczony analitycznie	Ładunek obliczony numerycznie	Błąd [%]
10	$4.05 \cdot 10^{-2}$	0.585	0.561	4.115
20	$1.12 \cdot 10^{-2}$	0.306	0.294	3.631
30	$5.13 \cdot 10^{-3}$	0.207	0.200	3.528
40	$2.93 \cdot 10^{-3}$	0.156	0.151	3.491
50	$1.69 \cdot 10^{-3}$	0.126	0.121	3.473
80	$7.52 \cdot 10^{-4}$	0.079	0.076	3.453

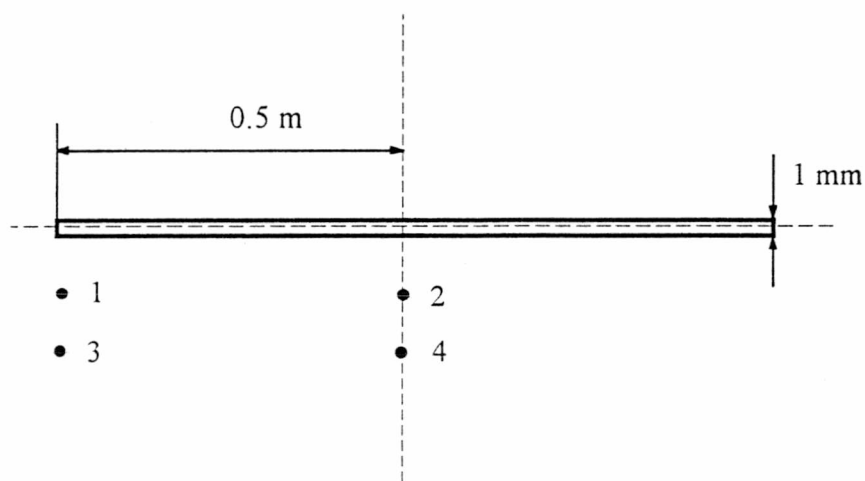
Tablica 2.6
Unormowany ładunek zgromadzony na powierzchni przewodu taśmowego

Liczba segmentów	Ładunek obliczony analitycznie	Ładunek obliczony numerycznie	Błąd [%]
10	4.532	4.486	1.028
20	4.532	4.519	0.286
30	4.532	4.526	0.133
40	4.532	4.529	0.077
50	4.532	4.530	0.049
80	4.532	4.531	0.020

Z danych przedstawionych w tablicy 2.5 wynika, że błąd obliczenia ładunku zgromadzonego na powierzchni skrajnych (najmniejszych) segmentów pozostaje w przybliżeniu stały i wynosi około 4 %. Przy coraz mniejszych szerokościach segmentów (kolumna 2 tablicy 2.5) powoduje to stałe, szybkie zmniejszanie się całkowitego błędu wraz ze wzrostem liczby segmentów. Jak pokazano w tablicy 2.6 błąd ten wynosi około 1 % dla $N=10$, ale już dla $N=20$ maleje do 0.3 %!. Oznacza to, że metoda daje dobre rezultaty już przy niewielkiej liczbie segmentów.

2.2. Błędy obliczania natężenia pola wynikające z zastąpienia przewodnika o skończonej grubości przewodnikiem nieskończenie cieniłym

W celu oszacowania wpływu przyjęcia do obliczeń nieskończenie cienkiego przewodnika dokonano porównania wyników obliczeń potencjału wytwarzanego przez przewód taśmowy o szerokości 1 m, naładowany do potencjału 1 V. Obliczono potencjały w odległości 0,1 m (punkty 1 i 2) i 0,2 m (punkty 3 i 4) od poziomej osi symetrii przewodu, w punktach odpowiadających środkowi oraz końcowi przewodu (rys. 2.2). Porównane zostały wyniki dla przewodnika nieskończenie cienkiego i dla przewodnika o grubości 1 mm.



Rys. 2.2. Punkty obliczeniowe w pobliżu przewodu taśmowego

Wyniki obliczeń przedstawiono w tablicach 2.7 i 2.8. Literą G oznaczono przewód o grubości 1 mm, literą C przewód nieskończenie cienki.

Tablica 2.7

Porównanie potencjałów obliczonych w odległości 0,1 m

N	Potencjał w punkcie 1			Potencjał w punkcie 2		
	G	C	Błąd [%]	G	C	Błąd [%]
10	0.67122	0.66209	0.75	0.85745	0.85633	0.13
20	0.67375	0.66947	0.64	0.85741	0.85654	0.10
50	0.67415	0.67180	0.35	0.85741	0.85665	0.09
100	0.67437	0.67216	0.33	0.85741	0.85666	0.09

Tablica 2.8

Porównanie potencjałów obliczonych w odległości 0,2 m

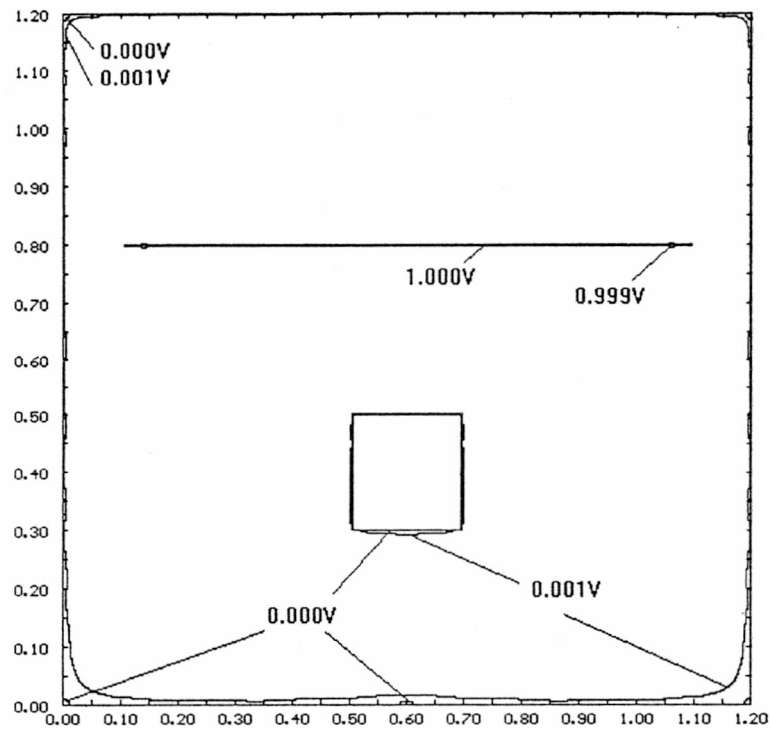
N	Potencjał w punkcie 3			Potencjał w punkcie 4		
	G	C	Błąd [%]	G	C	Błąd [%]
10	0.52961	0.52410	1.04	0.71922	0.71786	0.19
20	0.53106	0.52841	0.49	0.71933	0.71841	0.13
50	0.53138	0.52979	0.29	0.71936	0.71860	0.11
100	0.53143	0.53001	0.27	0.71936	0.71864	0.10

UWAGA: W przypadku przewodnika o skończonej grubości na N segmentów podzielono osobno płaszczyznę górną i dolną, tzn. dla $N=10$ liczba segmentów wynosi $2N+2=22$).

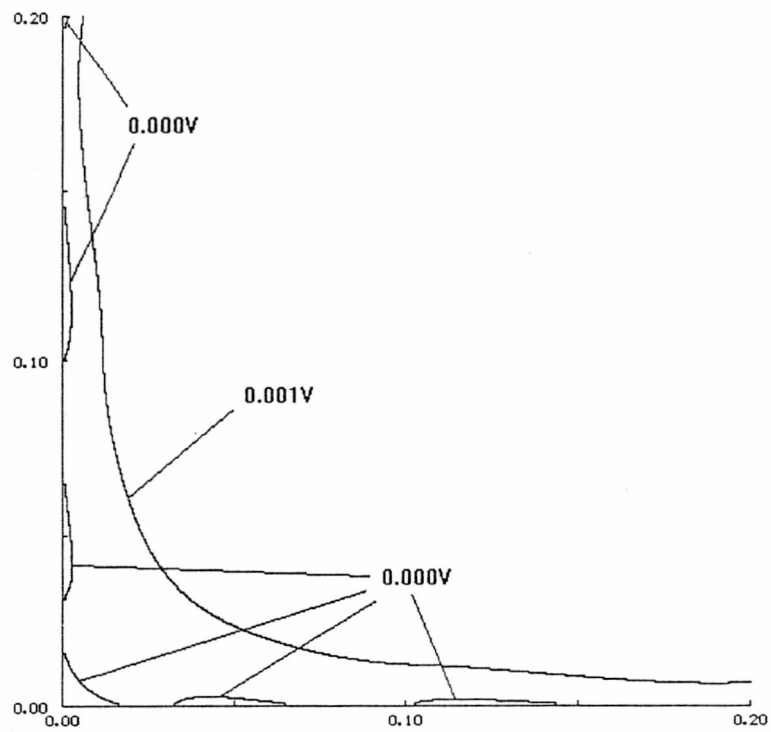
Przedstawione wyniki obliczeń wykazują, że już przy $N=10$ zastąpienie przewodnika o skończonej grubości przewodnikiem nieskończenie cienkim powoduje niewielkie (około 1 %) błędy w obliczaniu potencjału. Błąd obliczeń gwałtownie maleje ze wzrostem N . Należy zaznaczyć, że w przypadkach praktycznych interesujące są wyniki w pobliżu środka elektrody (tzw. przestrzeń robocza w komorze TEM), gdzie błędy przybliżenia są jeszcze mniejsze (około 0,2 %). Oczywiście wszystkie obliczenia wykonano dla punktów leżących w pewnej odległości od elektrody (1/10 szerokości) ponieważ zadaniem zastosowanej metody nie jest analiza pól w bezpośrednim sąsiedztwie krawędzi przewodnika. Poza tym dokładna analiza pola wokół krawędzi nieskończenie cienkiego przewodnika jest zadaniem akademickim i nie jest interesująca z punktu widzenia użytkownika komór TEM.

2.3. Weryfikacja spełnienia założeń początkowych (warunki brzegowe)

Założeniem wejściowym w obliczeniach za pomocą metody momentów jest przypisanie przewodzącym powierzchniom zadanych potencjałów. Potencjały te są warunkami brzegowymi dla rozwiązywanego równania Laplace'a. Jedną z metod weryfikacji uzyskanego rozwiązania jest sprawdzenie, czy uzyskany w wyniku obliczeń rozkład źródeł wytwarza rzeczywiście zadany potencjał na liniach tworzących, odpowiadających płaszczyznom przewodzącym. W celu dokonania takiej weryfikacji dla dwuwymiarowego modelu komory z obiektem o potencjale 0 V dokonano obliczenia rozkładu linii ekwipotencjalnych odpowiadających potencjałom 0 V, 0,001 V, 0,999 V i 1 V. Wyniki obliczeń przedstawiono na rysunku 2.3.



Rys. 2.3a. Odtworzenie wejściowego rozkładu potencjałów przez obliczony rozkład ładunków



Rys. 2.3b. Odtworzenie wejściowego rozkładu potencjałów przez obliczony rozkład ładunków - powiększenie narożnika komory

Jak pokazano na rysunku błędy odtworzenia wejściowych potencjałów (0 V i 1 V) nie przekraczają 0,5 %, a w rejonach najbardziej krytycznych (rogi komory) 1 %. Oznacza to, że zastosowana metoda jest w pełni poprawna. Dokładność obliczeń można poprawić stosując bardziej subtelną segmentację powierzchni.

2.4. Podsumowanie

W punktach 2.1 do 2.3 przedstawiono metody weryfikacji i oszacowano błędy obliczeń występujące w przypadku zastosowania metody momentów do modelowania komór TEM. Przedstawione wyniki stanowią podstawę stwierdzenia, że zastosowana metoda daje bardzo dobre rezultaty już przy niewielkiej liczbie zastosowanych segmentów.

Poprawność poczynionych założeń, w tym założenia, że do opisu rozkładu pola w komorze TEM dla modu TEM można zastosować trójwymiarowe równanie Laplace'a, została udowodniona w punkcie 3 niniejszego dodatku. W punkcie tym przedstawiono także porównanie wyników otrzymanych za pomocą modelu dwu- i trójwymiarowego.

3. Ogólny model komory TEM (metoda macierzy linii transmisyjnych - TLM) - uwagi uzupełniające

W celu analizy pola elektromagnetycznego w pełnym zakresie częstotliwości (a więc bez czynienia upraszczającego założenia o występowaniu w komorze wyłącznie modu TEM) wybrano i oprogramowano tzw. metodę macierzy linii transmisyjnych (TLM - ang. Transmission Line Matrix). Metoda ta (opisana szczegółowo w rozdziale 4 rozprawy) charakteryzuje się, w porównaniu z innymi metodami następującymi zaletami:

- 1) rozwiązania opierają się wprost na różniczkowych równaniach Maxwella (str. 54 rozprawy), a ponadto możliwe jest:
- 2) wprowadzanie dowolnych układów geometrycznych (z dokładnością do przyjętego skoku siatki),
- 3) modelowanie materiałów dielektrycznych oraz uwzględnianie strat w przewodnikach i dielektrykach,
- 4) modelowanie powierzchni absorpcyjnych,
- 5) modelowanie pola elektromagnetycznego w dziedzinie czasu lub w dziedzinie częstotliwości,
- 6) uzyskanie wartości częstotliwości rezonansowych komory w jednym przebiegu obliczeniowym,
- 7) umieszczanie dowolnej liczby pobudeń, w dowolnych miejscach przestrzeni,
- 8) modelowanie zjawisk otwartych,
- 9) obliczenie częstotliwości odcięcia wyższych modów pola.

Przykładowo program firmy Hewlett Packard "HP 85180 High-Frequency Structure Simulator", w którym zastosowano metodę elementów skończonych, nie ma możliwości 4 - 9. Szczególnie brak możliwości umieszczenia pobudzenia w obiekcie (źródła - porty muszą leżeć na zewnętrznych powierzchniach modelowanej struktury [1]) jest dyskwalifikujący z punktu widzenia modelowania komór TEM.

3.1 Porównanie wyników obliczeń z wynikami uzyskanymi za pomocą innych metod

Dokładność obliczeń za pomocą dwuwymiarowej metody TLM została wykazana w rozdziale 5 rozprawy. Dokonano tam porównania wyników obliczeń częstotliwości odcięcia wyższych modów pola za pomocą metody TLM dla przypadku falowodu prostokątnego (Tablica 5.1, str. 65) i dla komory TEM (Tablica 5.2, str. 65). Wyniki dla falowodu prostokątnego (proste wzory analityczne) wskazują, że błąd metody TLM w omawianym przypadku wynosi 0,5 - 1,0 %. Dla komory TEM dla pierwszych trzech modów istnieją [Wilson, Ma, 63] przybliżone wzory analityczne. Różnice między wynikami obliczeń na podstawie przybliżonych wzorów analitycznych i wyników uzyskanych za pomocą metody TLM nie przekraczają 1.4 %. Nie oznacza to jednak, że błędy leżą po stronie metody TLM (w [63] nie podano oszacowania błędów wprowadzanych przez rozwiązania przybliżone). Dokładność metody TLM może być dowolnie zwiększana poprzez przyjęcie mniejszego skoku siatki - ograniczeniem są tu raczej możliwości komputera (pamięć, czas), na którym dokonuje się obliczeń.

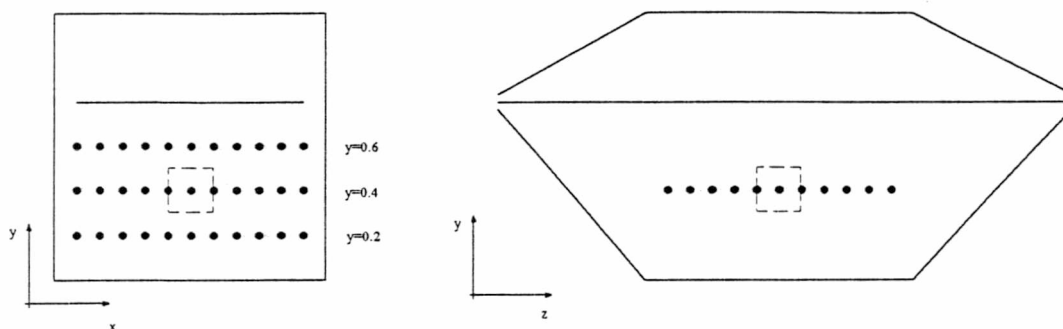
W celu weryfikacji poprawności rozwiązań uzyskanych za pomocą trójwymiarowej metody TLM (rozdział 6 rozprawy), a jednocześnie w celu wykazania zasadności stosowania metody uproszczonej dla modu TEM (rozdział 3), dokonano porównania wyników uzyskanych za pomocą czterech metod (w tym trzech zawartych w rozprawie):

- dwuwymiarowej metody momentów,
- trójwymiarowej metody momentów,
- trójwymiarowej metody TLM,
- trójwymiarowej metody elementów skończonych (profesjonalny program HP 85180A).

Na wykresach wyniki uzyskane za pomocą powyższych metod oznaczono odpowiednio: MM2, MM3, TLM, HFSS.

Obliczenia metodą TLM wykonano dwukrotnie: dla skoku siatki 3 cm i 6 cm. W drugim przypadku zastosowano dwukrotnie większą liczbę iteracji, co oznacza przyjęcie czterokrotnie dłuższego czasu stabilizacji przebiegów.

Obliczenia wykonano dla komory Crawforda pustej oraz dla komory Crawforda z obiektem (metalowy prostopadłościan), dla częstotliwości pobudzenia 30 MHz (mod TEM). Dokonano obliczeń natężenia pola elektrycznego, w punktach wskazanych na rysunku 3.1.



Rys. 3.1. Rozmieszczenie punktów obliczeniowych w przekrojach komory

W tablicach 3.1 i 3.2 zamieszczono porównanie wyników obliczeń dla przekroju poprzecznego komory pustej.

Tablica 3.1

Porównanie wyników obliczeń metodami MM2, MM3, TLM, HFSS,
moduł składowej E_y na wysokości 0,4 m

x [m]	MM2 [V/m]	MM3 [V/m]	TLM ¹ [V/m]	TLM ² [V/m]	HFSS [V/m]
0.1	0.431	0.437	0.530	0.380	0.391
0.2	0.776	0.775	0.842	0.807	0.695
0.3	1.006	0.999	1.080	1.063	0.999
0.4	1.124	1.122	1.139	1.175	1.246
0.5	1.181	1.179	1.189	1.222	1.208
0.6	1.197	1.196	1.194	1.234	1.210
0.7	1.181	1.179	1.189	1.221	1.153
0.8	1.124	1.122	1.154	1.181	1.049
0.9	1.006	0.999	1.074	1.067	1.032
1.0	0.776	0.775	0.865	0.806	0.795
1.1	0.431	0.437	0.508	0.378	0.435

Oznaczenia:

MM2 - dwuwymiarowa metoda momentów,

MM3 - trójwymiarowa metoda momentów,

TLM¹ - metoda TLM, skok siatki 3 cm,

TLM² - metoda TLM, skok siatki 6 cm,

HFSS - metoda elementów skończonych (program HP 85180A).

Tablica 3.2
Porównanie wyników obliczeń metodami MM2, MM3, TLM, HFSS,
moduł składowej E_x na wysokości 0,4 m

x [m]	MM2 [V/m]	MM3 [V/m]	TLM ¹ [V/m]	TLM ² [V/m]	HFSS [V/m]
0.1	1.072	1.036	1.076	1,169	1.037
0.2	0.875	0.848	0.862	0.941	0.842
0.3	0.632	0.611	0.644	0.650	0.648
0.4	0.397	0.384	0.438	0.398	0.368
0.5	0.189	0.183	0.234	0.198	0.138
0.6	0.000	0.000	0.124	0.049	0.059
0.7	0.189	0.183	0.256	0.193	0.269
0.8	0.397	0.384	0.475	0.391	0.331
0.9	0.632	0.611	0.680	0.641	0.580
1.0	0.875	0.848	0.897	0.933	0.837
1.1	1.072	1.036	1.119	1.163	1.080

Oznaczenia:

MM2 - dwuwymiarowa metoda momentów,

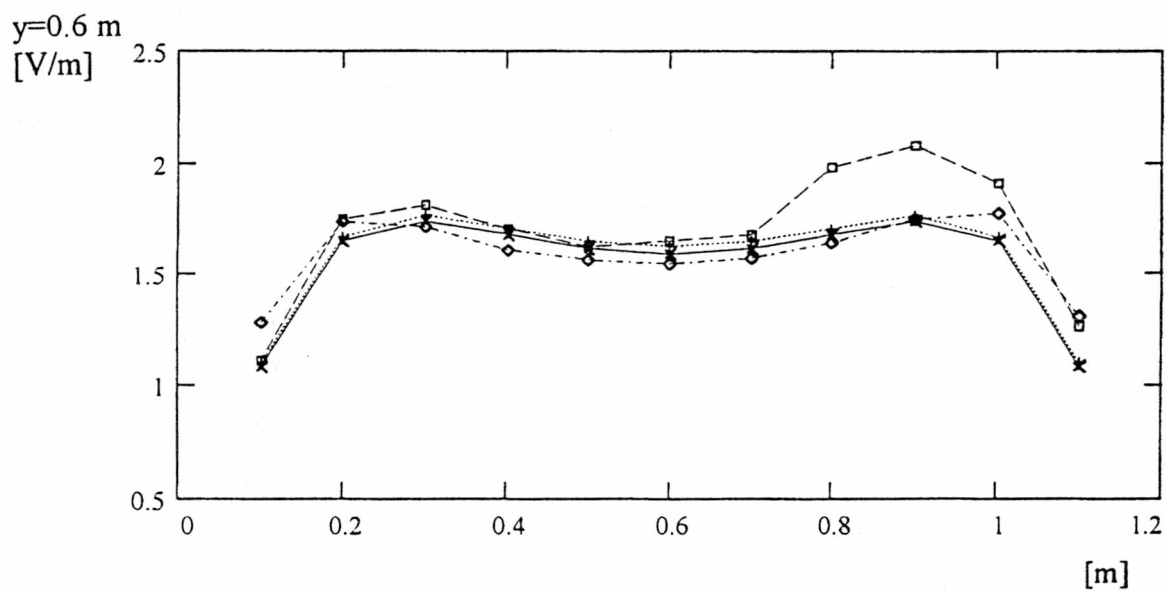
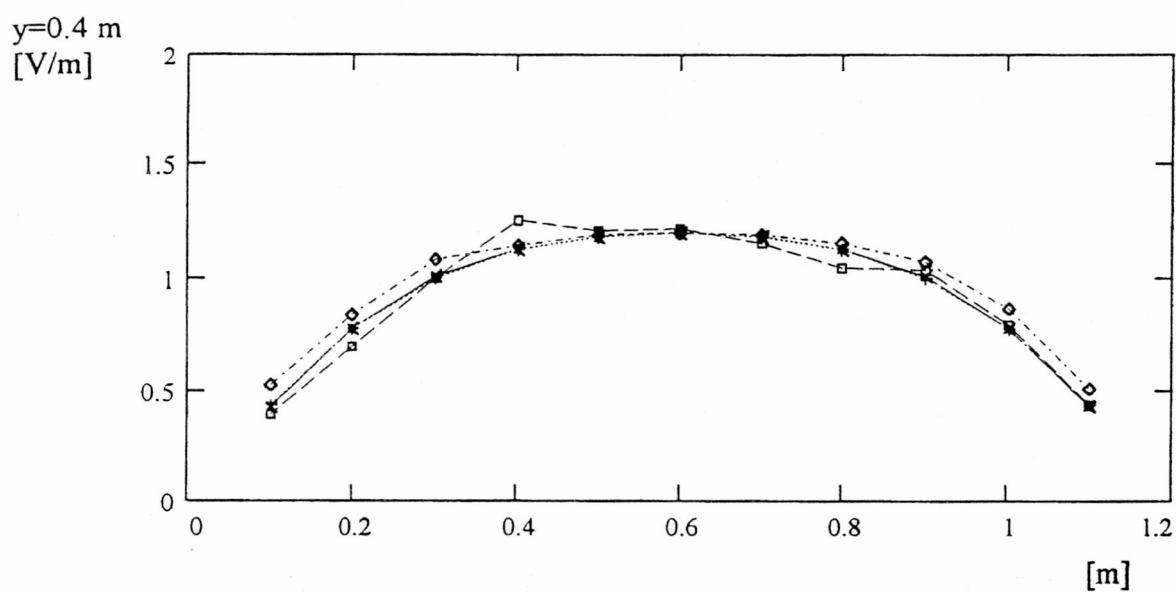
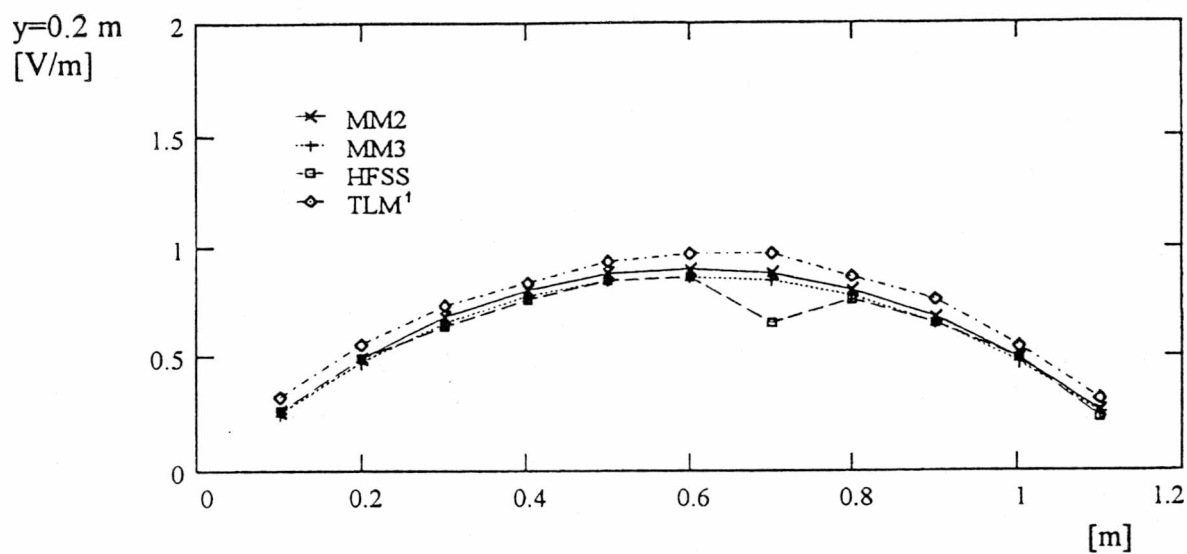
MM3 - trójwymiarowa metoda momentów,

TLM¹ - metoda TLM, skok siatki 3 cm,

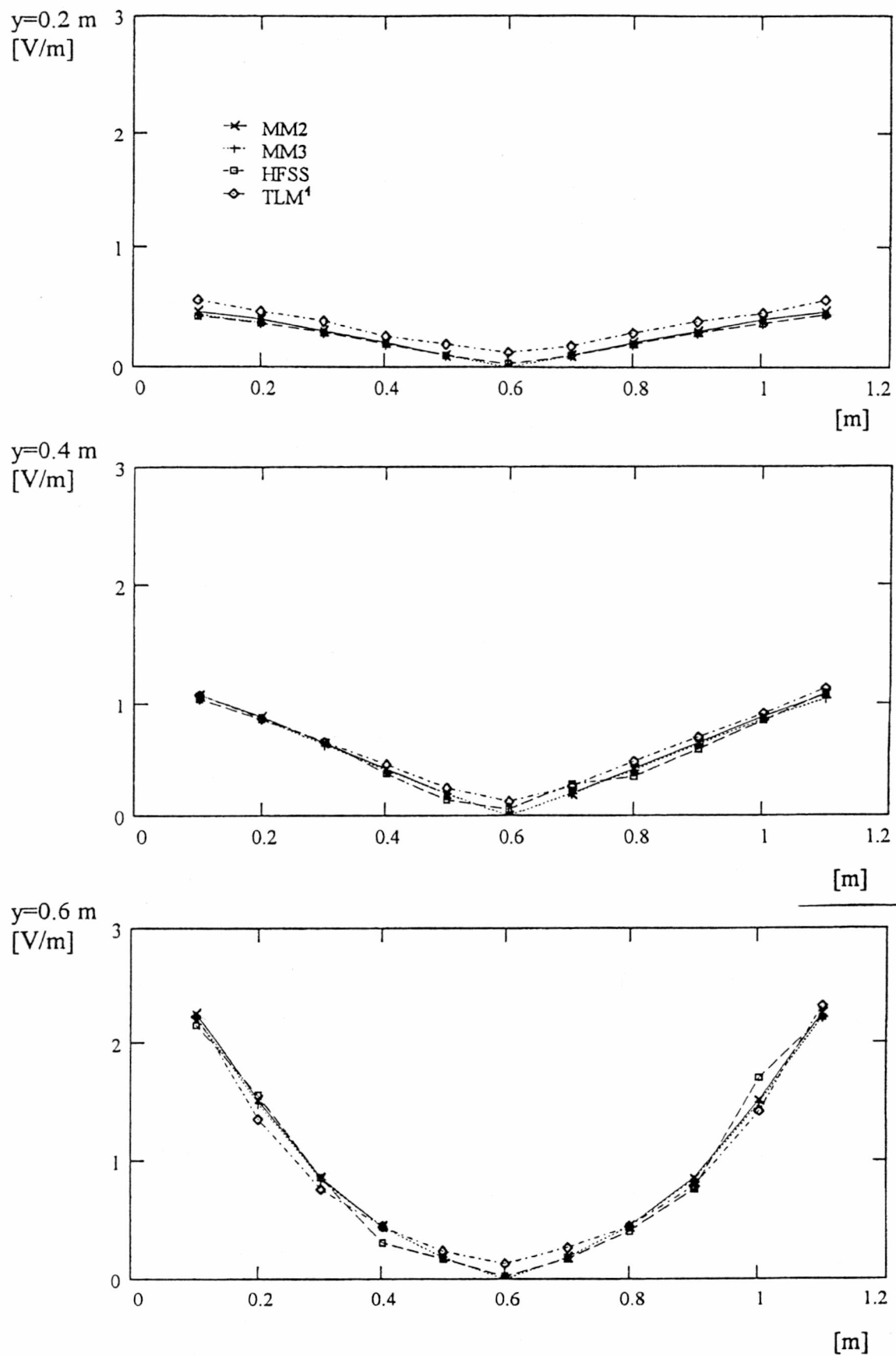
TLM² - metoda TLM, skok siatki 6 cm,

HFSS - metoda elementów skończonych (program HP 85180A).

Pełne wyniki obliczeń pokazano na rysunkach 3.2 - 3.7.



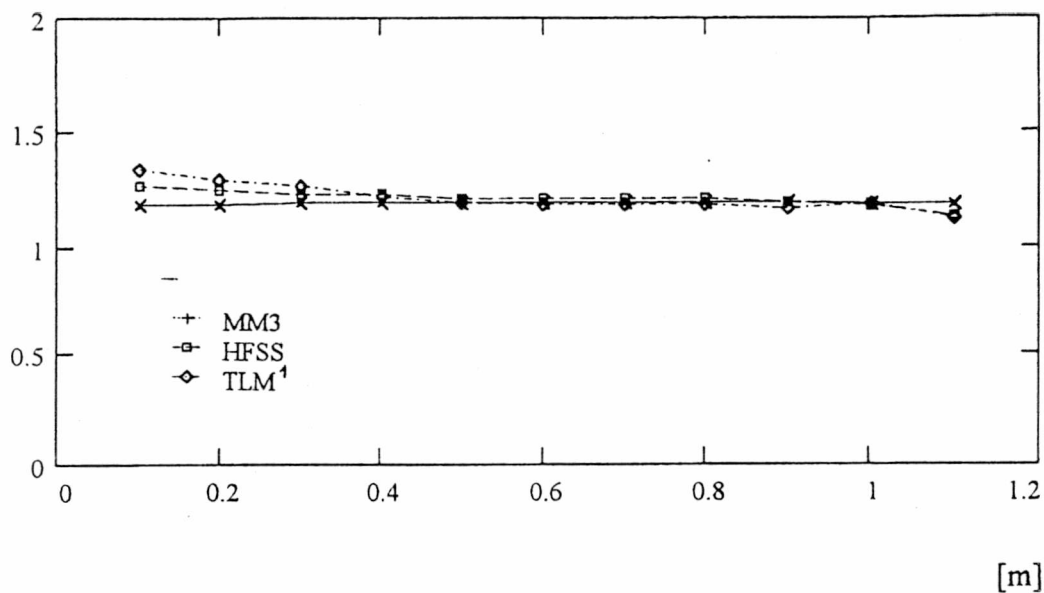
Rys. 3.2. Rozkłady pola w przekroju poprzecznym komory - moduł składowej E_y



Rys. 3.3. Rozkłady pola w przekroju poprzecznym komory - moduł składowej E_x

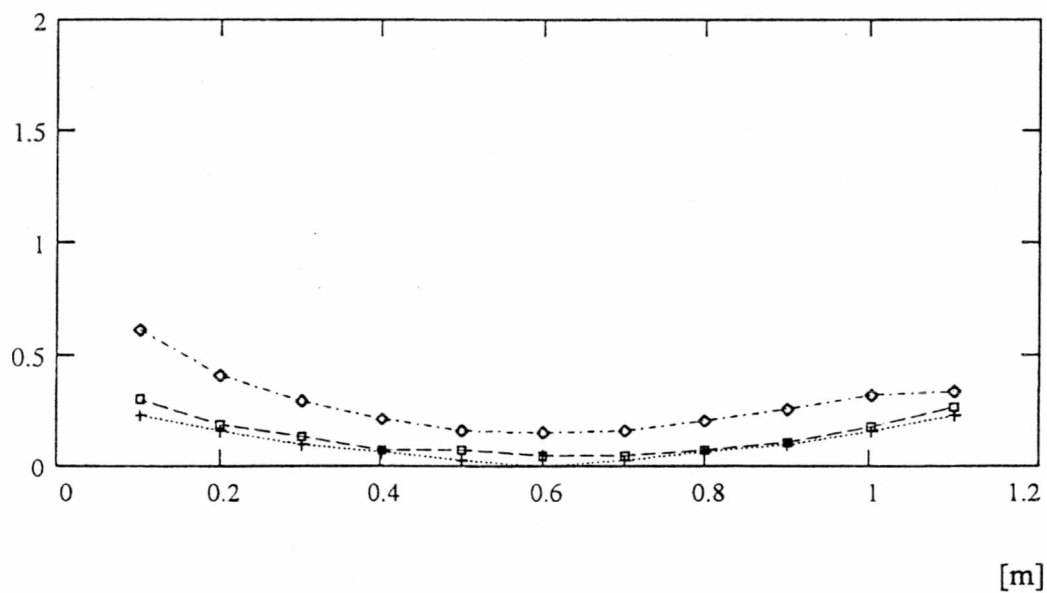
Moduł składowej E_y

[V/m]

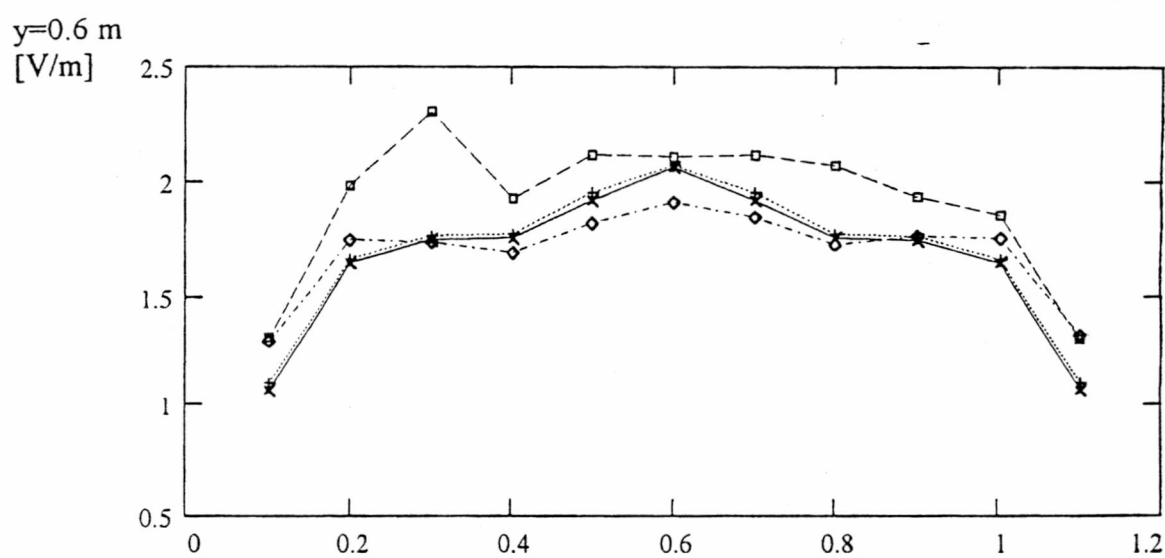
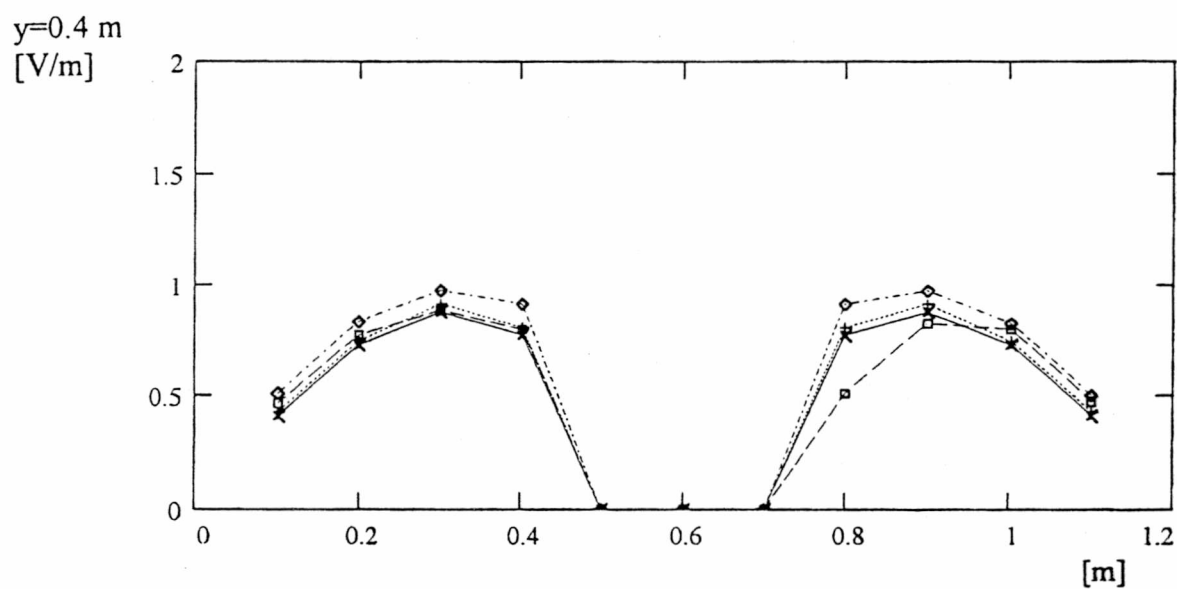
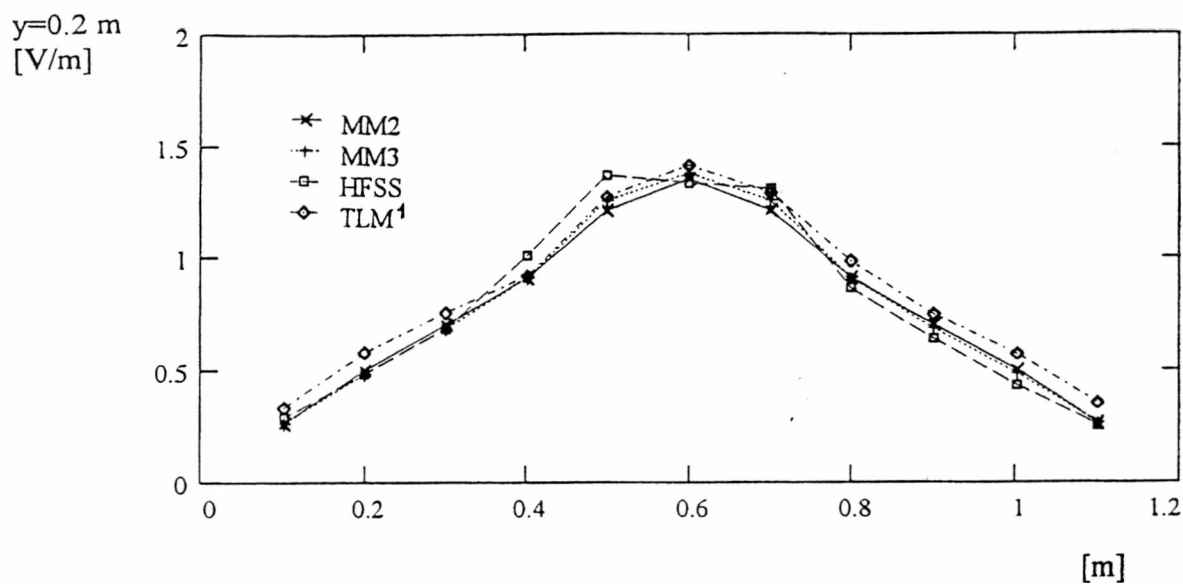


Moduł składowej E_z

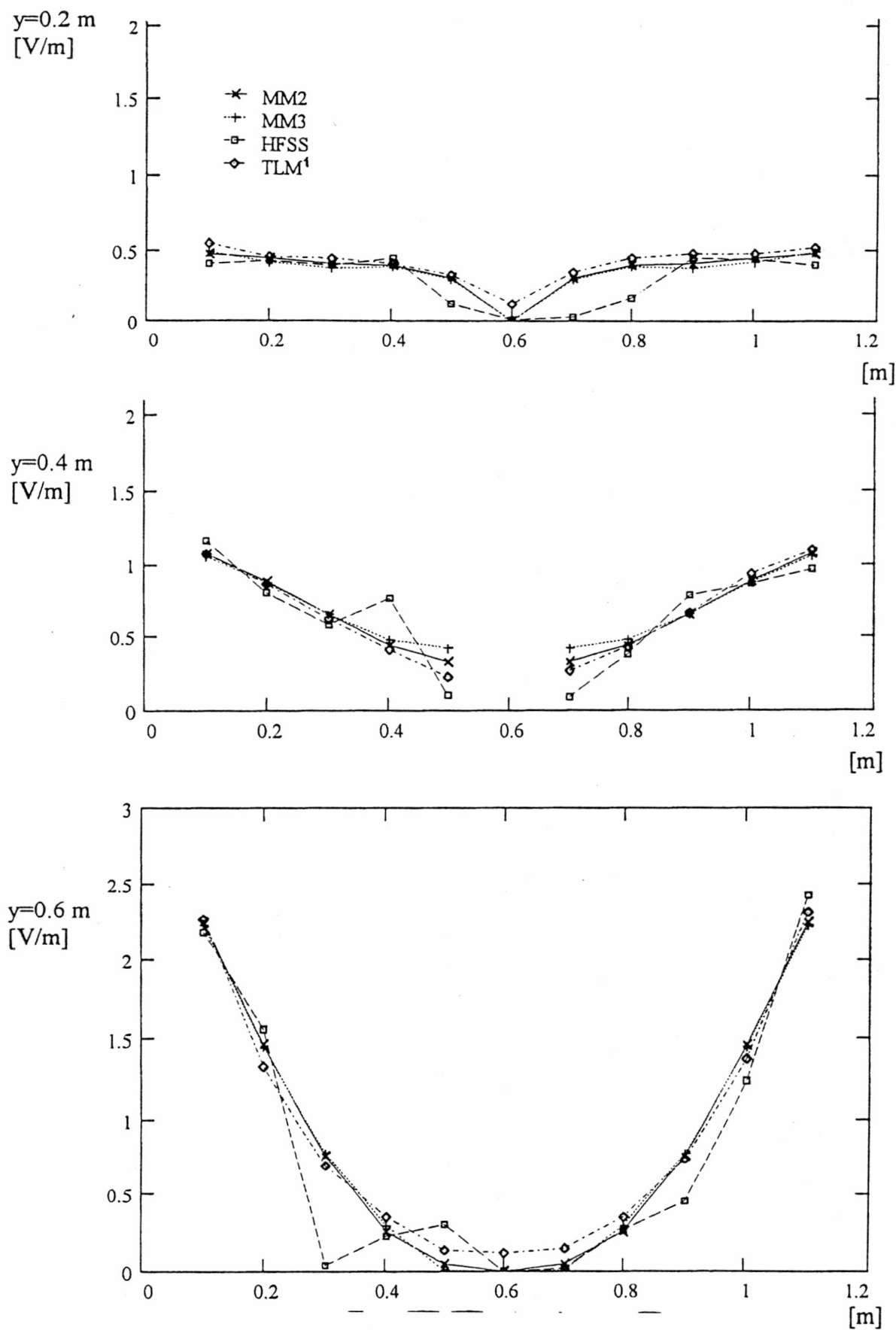
[V/m]



Rys 3.4. Rozkłady pola wzdłuż komory



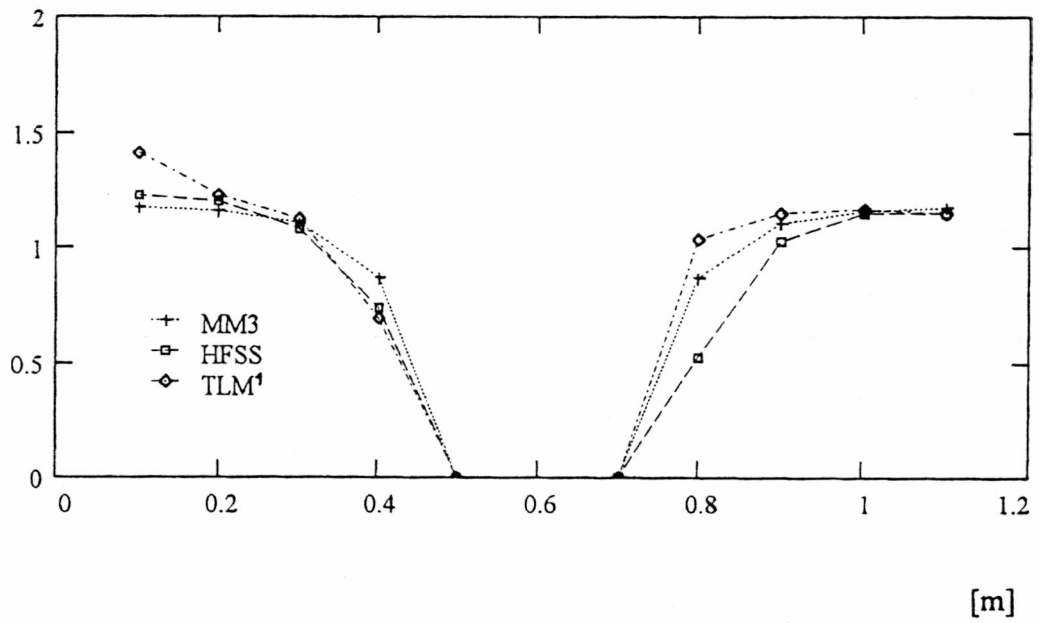
Rys. 3.5. Rozkłady pola w przekroju poprzecznym komory z obiektem
- moduł składowej E_y



Rys. 3.6. Rozkłady pola w przekroju poprzecznym komory z obiektem
- moduł składowej E_x

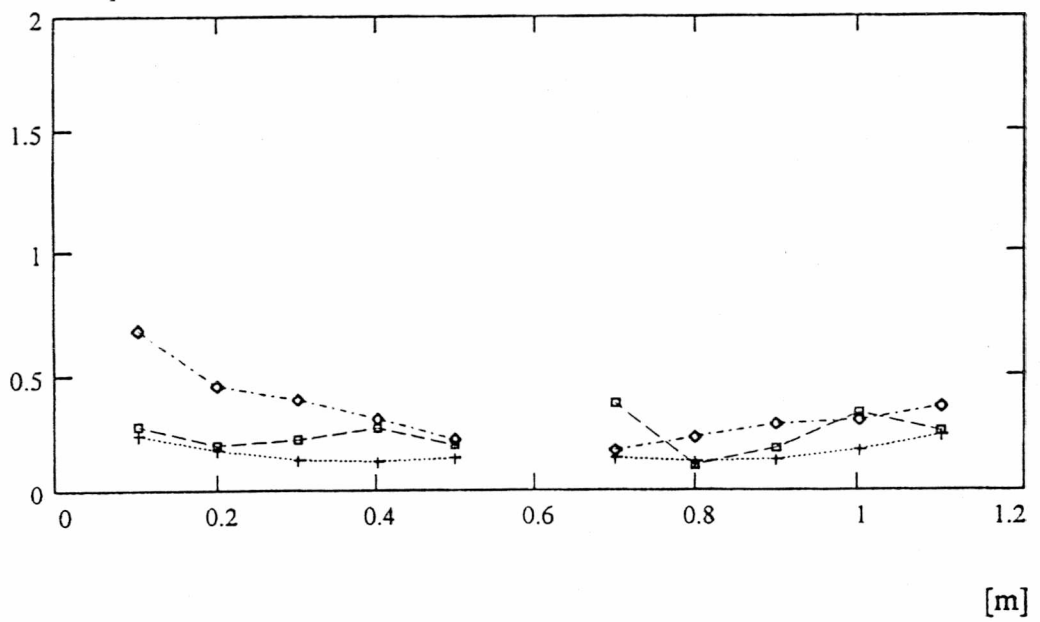
Moduł składowej E_y

[V/m]



Moduł składowej E_z

[V/m]



Rys. 3.7. Rozkłady pola wzdłuż komory z obiektem

Zaprezentowane wyniki obliczeń wykazują, że:

- wszystkie cztery zastosowane metody dają te same wyniki zarówno dla komory pustej, jak i komory z obiektem,
- różnice (błędy) są na poziomie założonej dokładności obliczeń; największe odchylenia w poszczególnych punktach daje metoda HFSS (rys 3.2, 3.5, 3.6), co jest związane ze stosowaniem w metodzie elementów skończonych dużych odległości pomiędzy węzłami dla roboczej części komory oraz interpolacji liniowej ([1], str. A-5),
- nie można jednoznacznie wskazać metody najdokładniejszej i najmniej dokładnej,
- największe różnice występują w przypadku szacowania rozkładu pola wzdłuż komory (wynika to z różnego rodzaju pobudzania pola w komorze w poszczególnych metodach),
- wykorzystanie uproszczonej metody (równanie Laplace'a) do poszukiwania rozkładów pola dla modu TEM daje bardzo dobre rezultaty zarówno dla komory pustej, jak i dla komory z przewodzącym obiektem. Jest to o tyle cenne, że program stosujący metodę momentów może być (i był) użyty na stosunkowo mało wymagającej platformie sprzętowej, jaką jest komputer osobisty klasy 386.

Reasumując, wyniki obliczeń potwierdziły w pełni poprawność zastosowanych do modelowania metod - zarówno zgodność wyników uzyskanych tymi metodami pomiędzy sobą, jak i w porównaniu z profesjonalnym pakietem firmy Hewlett Packard pozwala stwierdzić, że użyte metody są skutecznym narzędziem modelowania komór TEM.

3.2. Dodatkowe możliwości metody TLM

Z punktu widzenia użytkownika znajomość rozkładu pola w komorze z doskonale przewodzącym obiektem jest bardzo cenna - większość praktycznych obiektów można bowiem z dobrym przybliżeniem traktować jako obiekty o ścianach przewodzących (np. metalowe obudowy przyrządów pomiarowych i urządzeń telekomunikacyjnych). Jednakże w niektórych przypadkach może być pożądane rozważenie obiektów stratnych, a nawet źródłowych. Zastosowana w pracy metoda TLM umożliwia uwzględnienie w modelu obu tych przypadków. Możliwe jest także uwzględnienie obiektów o różnej od jedności przenikalności dielektrycznej i/lub magnetycznej (rozdział 4, str. 56-58).

W celu zilustrowania wymienionych możliwości metody TLM wykonano dodatkowe serie obliczeń dotyczące wymienionych przypadków.

Należy podkreślić, że użyte w pracy metody w żaden sposób nie ograniczają kształtu rozważanych obiektów - użycie do obliczeń obiektów prostopadłościennych jest jedynie przykładem (zresztą najbliższym praktyce).

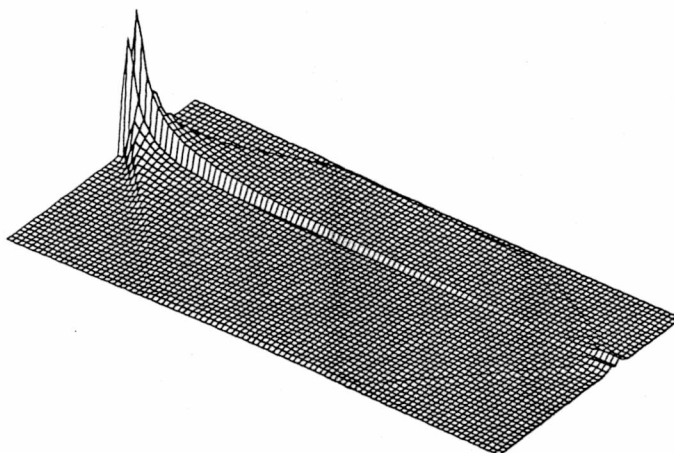
3.2.1. Obiekty stratne w komorach TEM

W obliczeniach przyjęto, że badany obiekt ma postać prostopadłościanu zbudowanego z materiału o skończonej konduktywności. Metoda TLM umożliwia także zadanie warunków brzegowych w postaci stratnej granicy obszaru (str. 56 pracy), jednakże w takim przypadku nie rozpatruje się pola wewnątrz badanego obiektu.

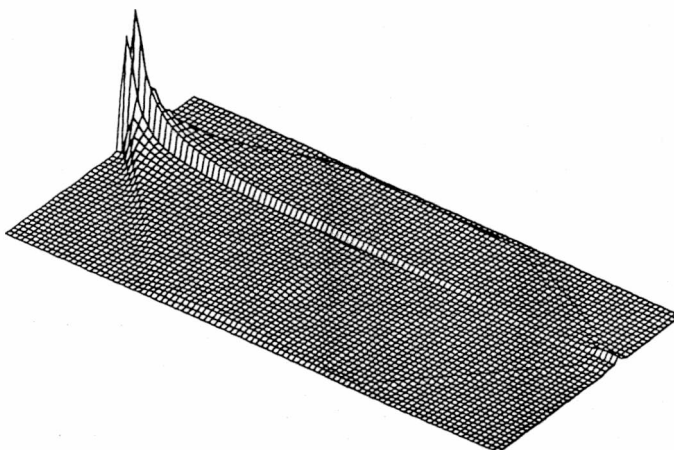
Obliczenia wykonano za pomocą dwuwymiarowej metody TLM, dla częstotliwości 30 i 300 MHz, dla trzech różnych wartości konduktywności obiektu. Wyniki przedstawiono na rysunkach 3.8 - 3.13

Wyniki są zgodne z oczekiwaniami - dla niewielkich wartości konduktywności obecność obiektu nie wpływa na rozkład pola elektromagnetycznego w komorze. Dla dużych wartości konduktywności rozkład pola zbliża się do rozkładu uzyskanego dla komory z obiektem idealnie przewodzącym (por. rysunki 6.7, 6.8 na str. 84, 85 rozprawy).

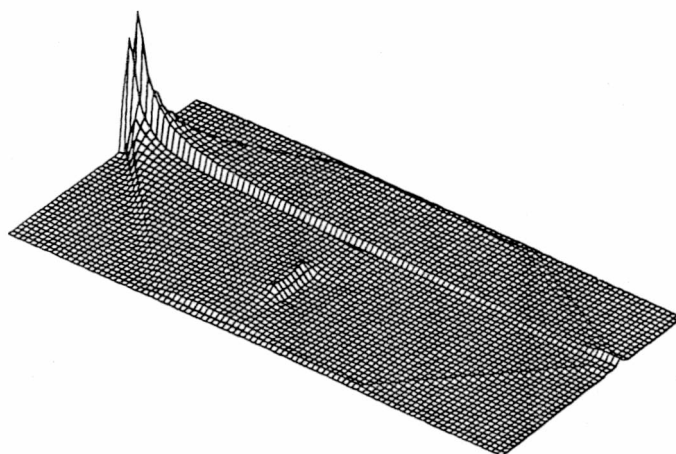
Na rysunkach można zaobserwować także rozkład pola wewnątrz badanego obiektu - świadczy to o możliwości szacowania natężeń pól oddziałujących na poszczególne części składowe urządzeń, a nie tylko pola wokół badanego obiektu.



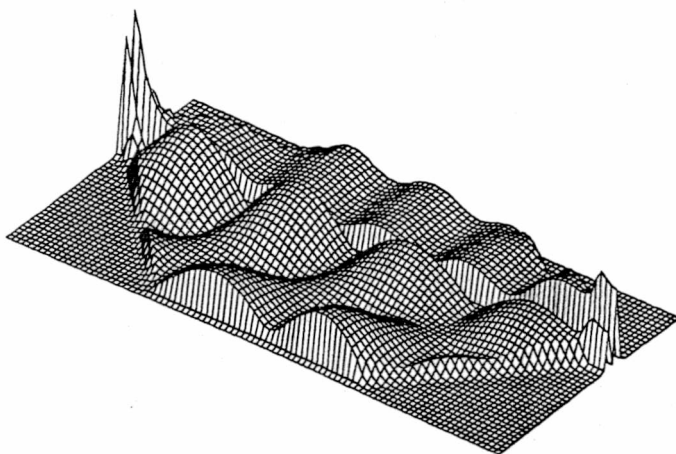
Rys. 3.8. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^{-2}$ S/m, przy częstotliwości 30 MHz



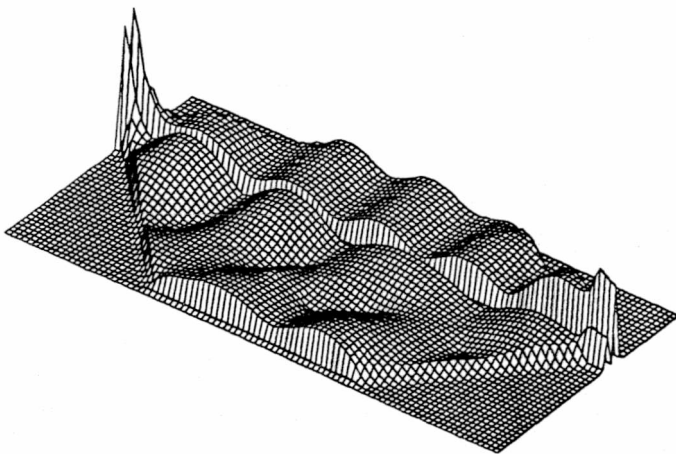
Rys. 3.9. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^{-1}$ S/m, przy częstotliwości 30 MHz



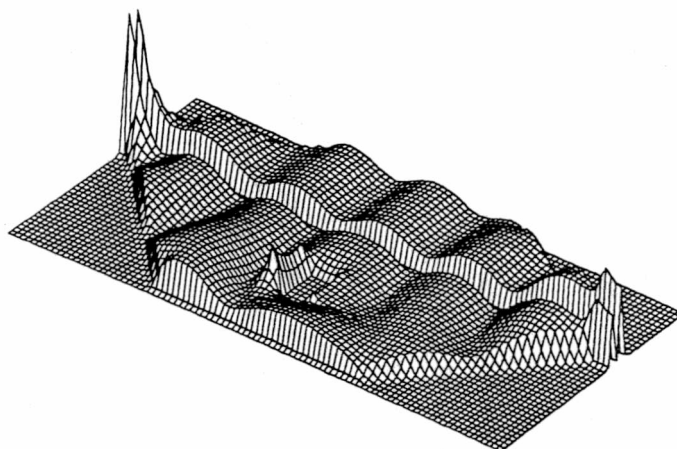
Rys. 3.10. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^5$ S/m, przy częstotliwości 30 MHz



Rys. 3.11. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^{-2}$ S/m, przy częstotliwości 300 MHz



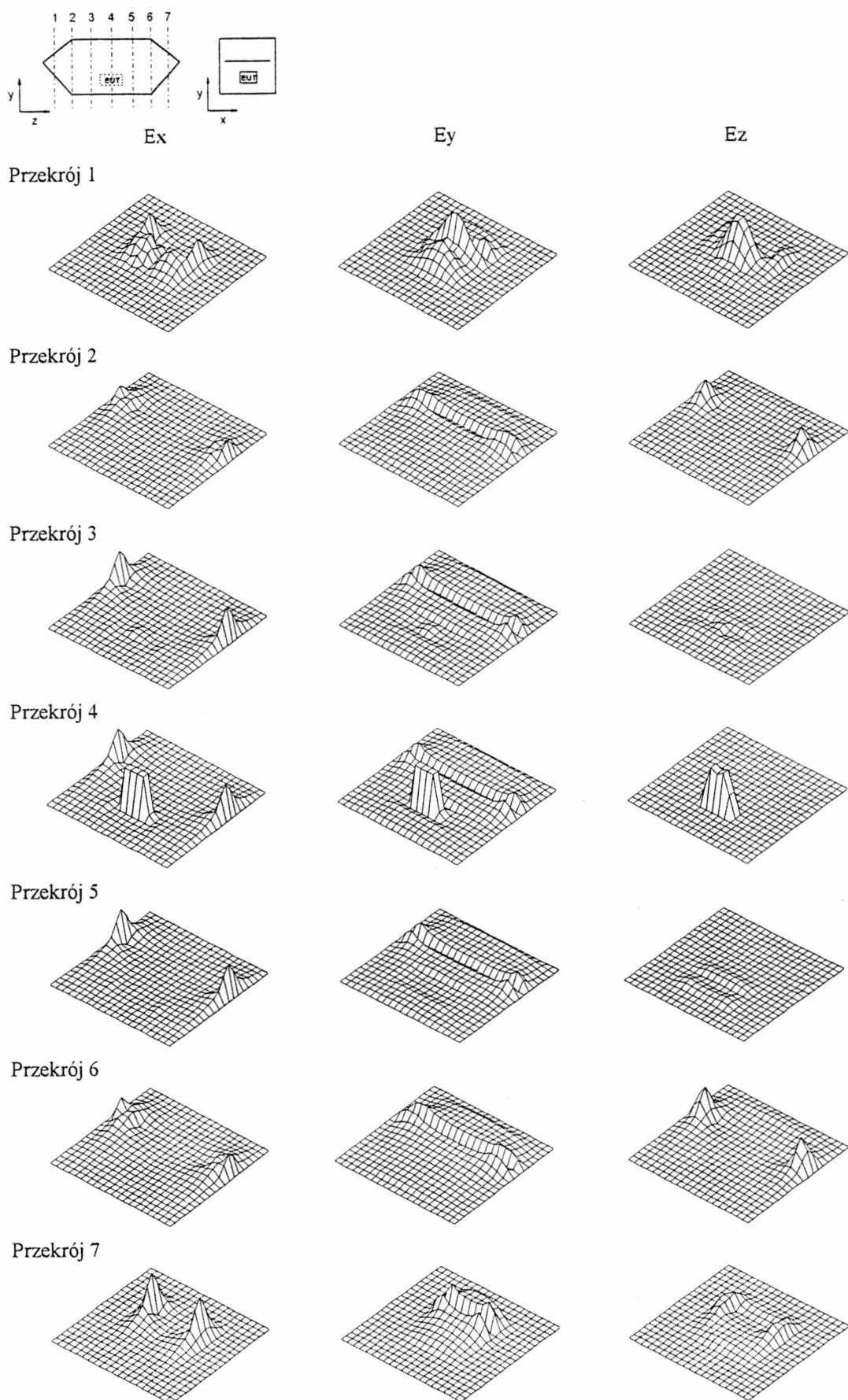
Rys. 3.12. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^{-1}$ S/m, przy częstotliwości 300 MHz



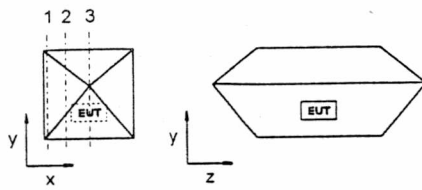
Rys. 3.13. Rozkład składowej E_y pola elektrycznego w komorze z obiektem o skończonej konduktywności, $\sigma = 10^5$ S/m, przy częstotliwości 300 MHz

3.2.2. Obiekt źródłowy w komorze TEM

W obliczeniach przyjęto, że obiektem jest metalowy prostopadłościan, w którego przedniej i tylnej ścianie (patrząc wzdłuż osi komory) znajdują się dwie poziome szczeliny. Wewnątrz obiektu umieszczono źródło izotropowe promieniujące pole elektryczne o częstotliwości 30 MHz i amplitudzie pięciokrotnie mniejszej niż amplituda źródła umieszczonego w porcie wejściowym komory. Wyniki przedstawiono zgodnie z konwencją przyjętą w rozprawie (str. 92-109). Obliczono rozkłady pola elektrycznego. Wyniki pokazano na rysunkach 3.14 - 3.16.



Rys. 3.14. Rozkład pola elektrycznego w płaszczyznach XY przy częstotliwości 30 MHz (komora Crawforda z obiektem źródłowym)

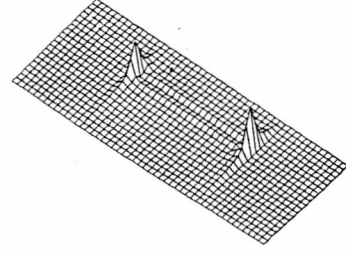
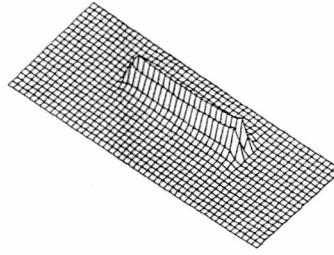
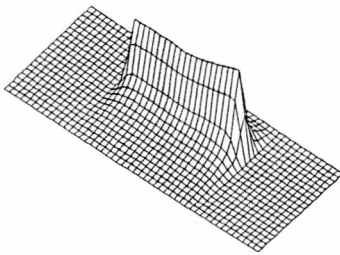


E_x

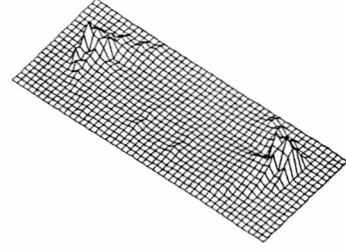
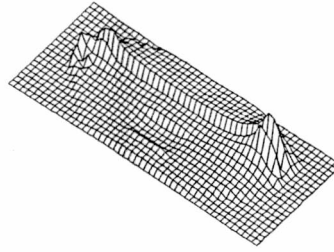
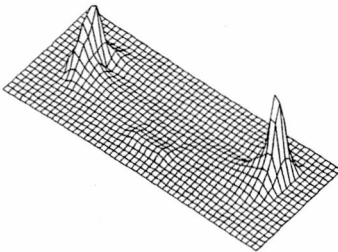
E_y

E_z

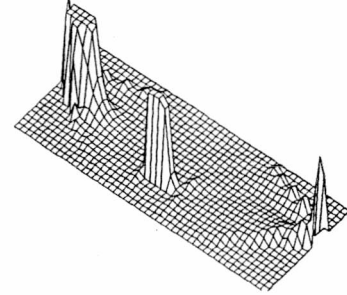
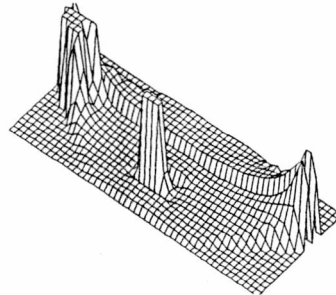
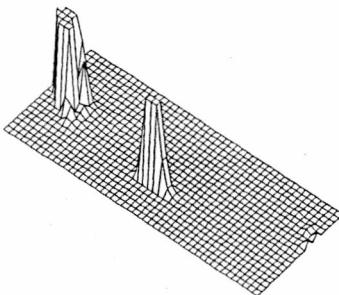
Przekrój 1



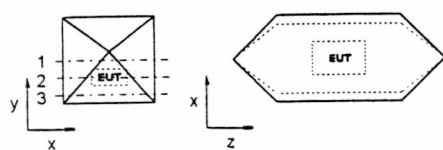
Przekrój 2



Przekrój 3



Rys. 3.15. Rozkład pola elektrycznego w płaszczyznach YZ przy częstotliwości 30 MHz (komora Crawforda z obiektem źródłowym)

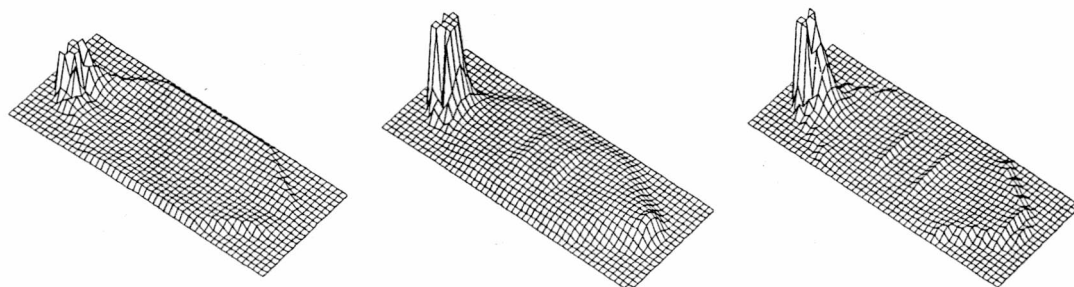


Ex

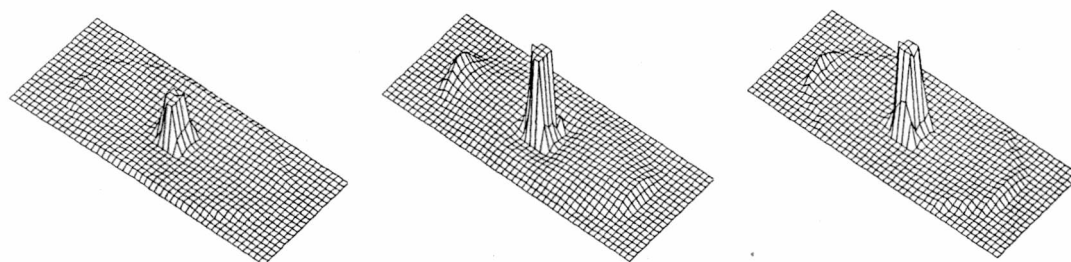
Ey

Ez

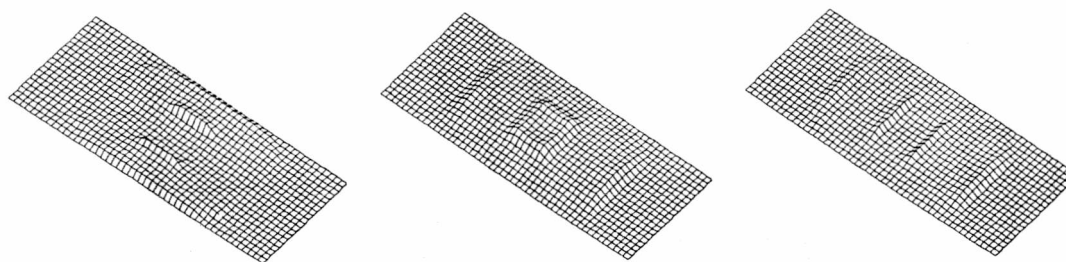
Przekrój 1



Przekrój 2



Przekrój 3



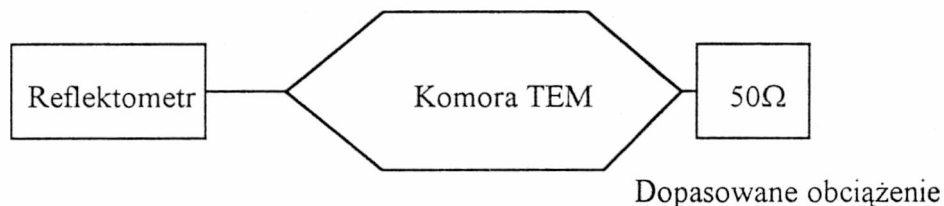
Rys. 3.16. Rozkład pola elektrycznego w płaszczyznach XZ przy częstotliwości 30 MHz (komora Crawforda z obiektem źródłowym)

4. Doświadczalna weryfikacja wyników symulacji komputerowych

Ze względu na brak w Instytucie Telekomunikacji i Akustyki PWr. sprzętu pomiarowego umożliwiającego wiarygodną weryfikację obliczeń rozkładów pól elektromagnetycznych, w ramach pracy zostały wykonane pomiary niektórych parametrów wtórnych, takich jak impedancja wejściowa w funkcji częstotliwości, częstotliwości rezonansowe komory Crawforda oraz współczynnik fali stojącej na wejściu komory GTEM. Wykorzystano także dostępne z literatury wyniki pomiarów rozkładu impedancji charakterystycznej komory Crawforda skonstruowanej w NBS, Colorado, USA. Poniżej przedstawiono zastosowane układy pomiarowe. Omówiono także koncepcję stanowiska pomiarowego umożliwiającego pełną weryfikację doświadczalną uzyskanych wyników.

4.1. Pomiar impedancji wejściowej komory Crawforda w funkcji częstotliwości

Pomiar został wykonany w układzie pokazanym na rysunku 4.1.

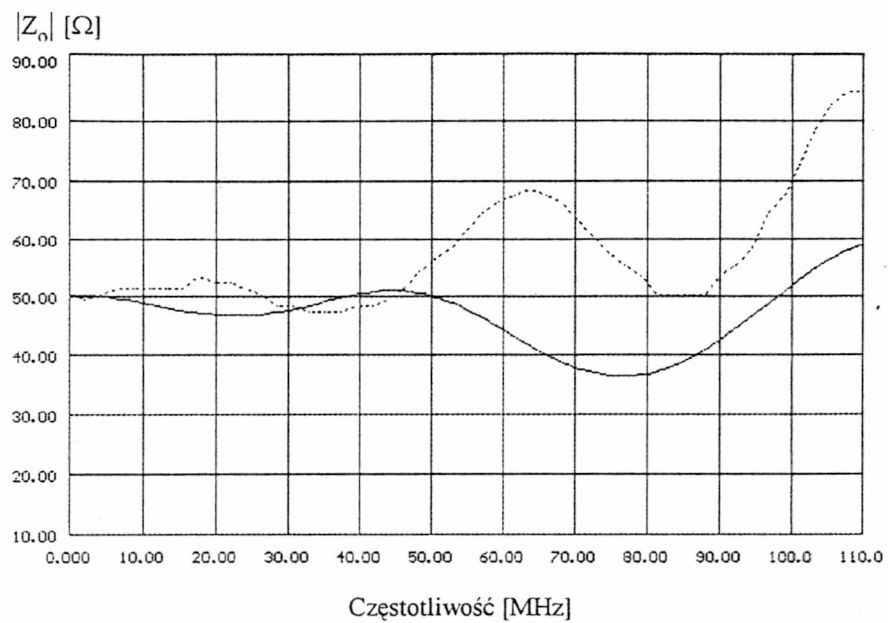


Rys. 4.1. Pomiar impedancji wejściowej komory w funkcji częstotliwości

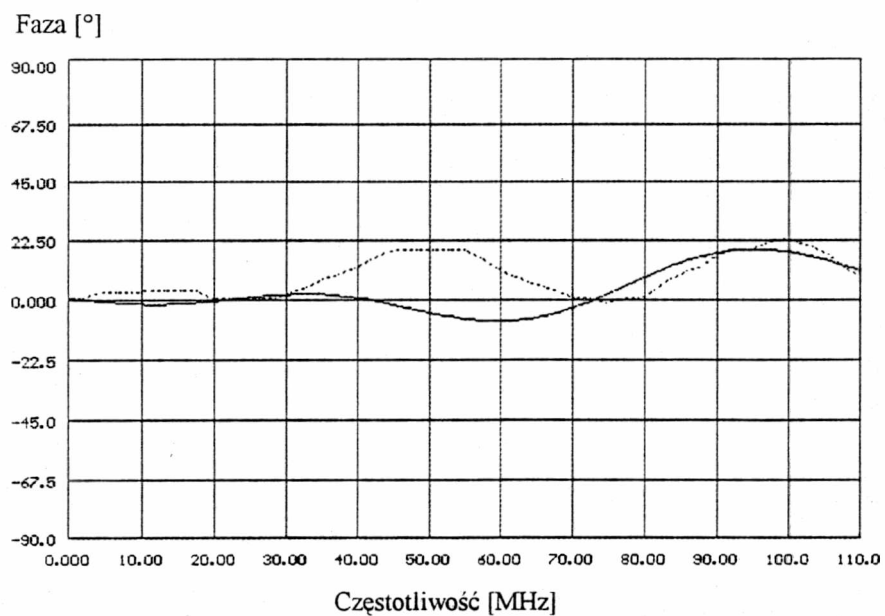
Wyniki pokazano na rysunku 4.2. Liniami przerywanymi oznaczono wartości zmierzone, liniami ciągłymi wartości obliczone za pomocą trójwymiarowej metody momentów (str. 42 - 43 rozprawy). Różnice impedancji obliczonej i zmierzonej są wynikiem niespełnienia założeń wejściowych metody obliczeniowej (równanie Laplace'a) w sekcjach zbieżnych komory. Metoda obliczeń została zresztą przewidziana do szacowania pola w przestrzeni roboczej komory Crawforda, a nie w sekcjach zbieżnych.

4.2. Pomiar impedancji charakterystycznej komory

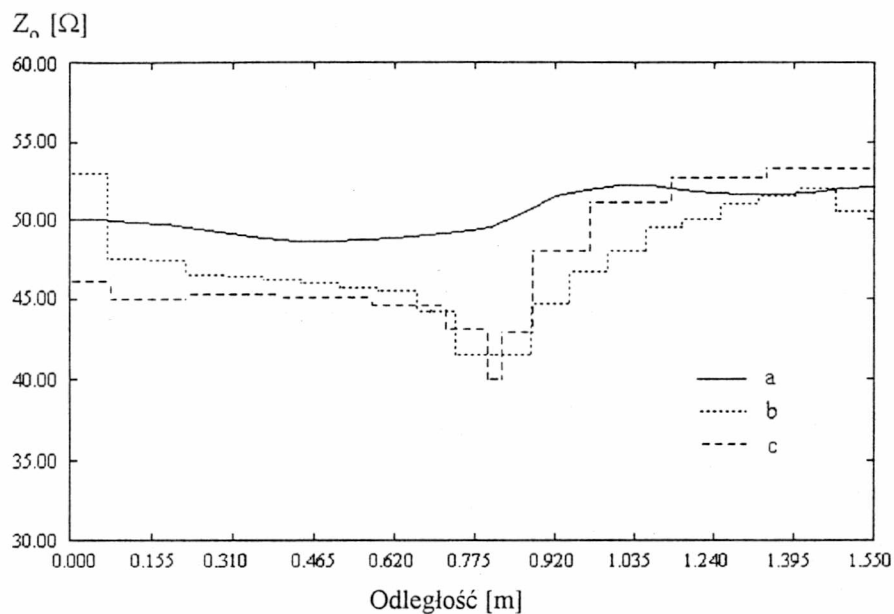
Wykorzystane zostały wyniki pomiarów dostępne w literaturze. Porównanie wyników przedstawiono na rysunku 4.3 (por. str. 40 rozprawy). Porównanie wykazuje (po raz kolejny), że zastosowana metoda obliczeń słuszna jest w obszarze roboczym komory Crawforda (błędy rzędu 3 %) nie daje natomiast poprawnych wyników w sekcjach zbieżnych (błędy około 10 %). Wynik ten potwierdza słuszność wniosku z punktu 4.1, w którym do obliczenia impedancji wejściowej użyte zostały obarczone stosunkowo dużym błędem wyniki obliczeń impedancji charakterystycznej dla części zbieżnych komory.



Rys. 4.2a. Moduł impedancji wejściowej komory Crawforda
(linia przerywana - wartości zmierzone)



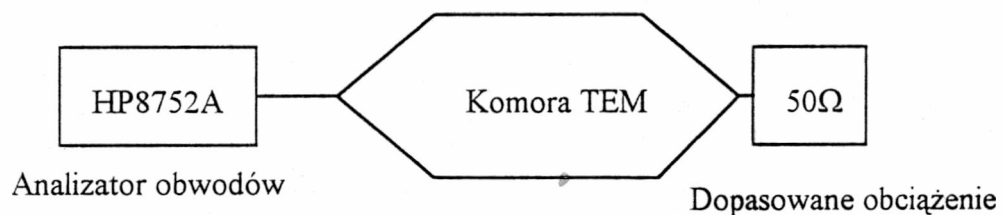
Rys. 4.2b. Faza impedancji wejściowej komory Crawforda
(linia przerywana - wartości zmierzone)



Rys. 4.3. Impedancja charakterystyczna komory Crawforda
a - wyniki pomiarów,
b - wyniki obliczeń za pomocą metody elementów granicznych,
c - wyniki obliczeń za pomocą metody momentów.

4.3. Pomiar częstotliwości rezonansowych komory Crawforda

Pomiar został wykonany w układzie przedstawionym na rysunku 4.4.



Rysunek 4.4. Pomiar częstotliwości rezonansowych komory Crawforda

Porównanie wyników przedstawiono w tablicy 4.1 (por. str. 110 w rozprawie)

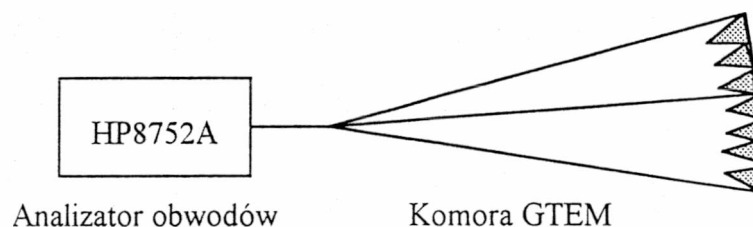
Tablica 4.1
Częstotliwości rezonansowe komory Crawforda

Częstotliwość rezonansowa [MHz]		
zmierzona	obliczona za pomocą dwuwymiarowej metody TLM	obliczona za pomocą trójwymiarowej metody TLM
58	58	54
104	94	93
134	137	134
148	-	-
153	159	161
182	188	185
198	198	189
205	214	210
233	236	-
243	243	242
262	263	256
268	282	269
289	290	292
303	299	303
309	315	307
330	331	333
337	-	337
353	355	353

Zagadnienie porównania i identyfikacji rezonansów zostało omówione w rozprawie na stronie 111.

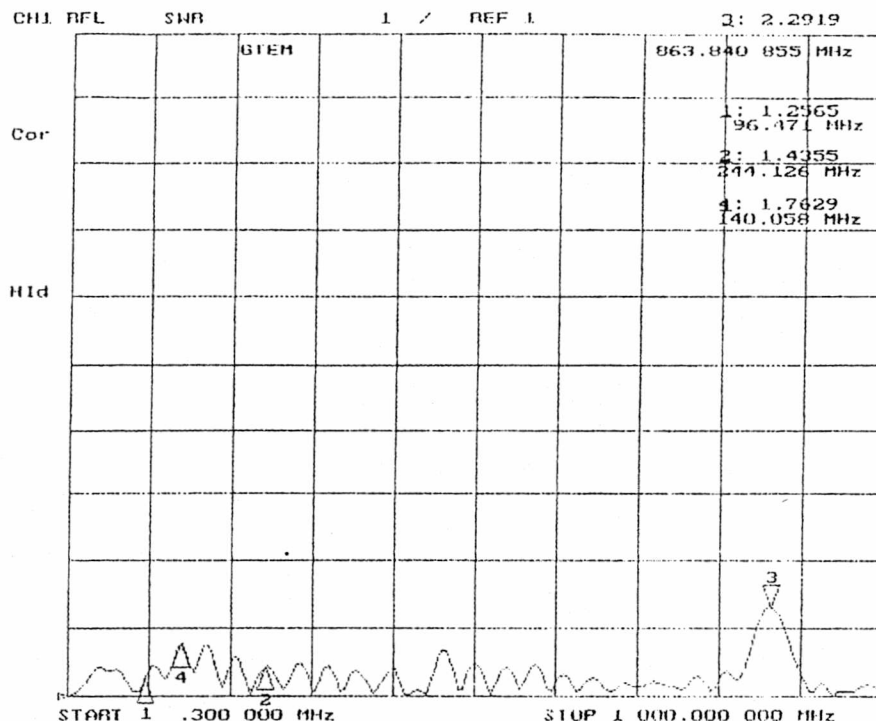
4.4. Pomiar współczynnika fali stojącej na wejściu komory GTEM

W podrozdziale 3.4 rozprawy opisano wykorzystanie opracowanego modelu symulacyjnego do zaprojektowania kształtu oraz obciążenia budowanej w ITA komory typu GTEM. Opisany na str. 47 pomiar współczynnika fali stojącej (SWR) na wejściu komory przeprowadzono w układzie przedstawionym na rysunku 4.3.



Rys. 4.5. Pomiar współczynnika fali stojącej na wejściu komory GTEM

Wyniki pomiarów przedstawia rysunek 4.6 (kopia ekranu analizatora HP8752A).



Rys. 4.6. Współczynnik fali stojącej na wejściu komory GTEM

Średnia wartość współczynnika fali stojącej w zakresie częstotliwości 300 kHz - 1 GHz wynosi poniżej 1,5. Uzyskany rezultat pozwala stwierdzić, że zaprojektowana w oparciu o przedstawione w rozprawie metody obliczeń i wykonana w ITA komora GTEM reprezentuje światowy poziom w tej dziedzinie. Uzasadnieniem tego stwierdzenia może być fakt, że komora tego typu produkowana przez lidera światowego - firmę Electro-Mechanics Company (EMCO) z USA ma WFS o takiej samej wartości.

4.5. Metodyka pomiarów pól elektromagnetycznych w komorach TEM - koncepcja sond pomiarowych

Przeprowadzenie precyzyjnych pomiarów natężenia pola w komorach TEM jest zagadnieniem trudnym i wymagającym specjalnych technik pomiarowych. Trudności wynikają z następujących czynników:

- 1) pole w komorach TEM w zakresie częstotliwości roboczych jest polem strefy bliskiej, konieczny jest więc niezależny pomiar poszczególnych składowych pola,
- 2) użyte w pomiarach anteny umieszczane są w pobliżu metalowych płaszczyzn (ściany komory), co powoduje istotne zmiany impedancji anten pomiarowych i wymaga osobnej analizy,
- 3) niemożliwe jest użycie standardowych kabli pomiarowych, ponieważ zakłócają one w sposób znaczny pierwotny rozkład pól w komorze [Kasyap, 31] i fałszują wyniki pomiarów - w celu doprowadzenia sygnału z sond do mierników pożądane jest zastosowanie łączny światłowodowych,

- 4) aby nie zakłócać rozkładu pola w komorze sondy pomiarowe muszą być małe elektrycznie - powoduje to problemy związane z niską czułością sond i wymaga stosowania silnych sygnałów pobudzających na wejściu komory.

Wymienione czynniki zostały uwzględnione w koncepcji sond pomiarowych do pomiarów pól elektromagnetycznych w strefie bliskiej opracowanej w Zakładzie Radiokomunikacji ITA PWi.

4.5.1. Podwójnie obciążona antena ramowa

Do przeprowadzenia równoczesnego pomiaru składowej elektrycznej i magnetycznej pola dobrze nadaje się podwójnie obciążona antena ramowa. Wiadomo [3,8], że natężenie prądu wzbudzanego w antenie ramowej przez zewnętrzne pole elektromagnetyczne zależy od obu składowych tego pola, tj. zarówno od składowej elektrycznej, jak i magnetycznej. O ile w przypadku anteny ramowej obciążonej pojedynczo nie jest możliwe rozdzielenie wpływu obu składowych (tzn. ich niezależny pomiar), to w przypadku anteny obciążonej podwójnie poprzez odpowiednie sumowanie i odejmowanie wartości napięcia na obu obciążeniach można uzyskać informację o składowej elektrycznej i magnetycznej pola w środku anteny. Podane twierdzenie zostało udowodnione dwiema metodami - analitycznie i numerycznie.

4.5.2. Prąd na zaciskach anteny

Rozważmy okrągłą antenę ramową zanurzoną w polu elektromagnetycznym o dowolnej strukturze. Załóżmy, że dwa identyczne obciążenia anteny o impedancji Z_1 są umieszczone w przeciwległych punktach $\varphi = \varphi_1$ i $\varphi = \varphi_2 = \varphi_1 + \pi$ (Rys. 4.5).

Równania opisujące prądy w obciążeniach mają postać [4]:

$$\begin{aligned} I(\varphi_1, \omega) &= \frac{-j\omega\mu\pi b^2 H_z^i(\omega) W(\varphi_1)}{1 + 2Y_0 Z_1} - \frac{2\pi b E_s^i(\omega) X(\varphi_1)}{1 + 2Y_1 Z_1} \\ I(\varphi_2, \omega) &= \frac{-j\omega\mu\pi b^2 H_z^i(\omega) W(\varphi_1)}{1 + 2Y_0 Z_1} + \frac{2\pi b E_s^i(\omega) X(\varphi_1)}{1 + 2Y_1 Z_1} \end{aligned} \quad (4.1)$$

Z równań tych widać, że suma prądów (lub napięć) na obciążeniach jest funkcją wyłącznie natężenia pola magnetycznego w środku anteny. Podobnie różnica prądów (napięć) zależy wyłącznie od natężenia pola elektrycznego.

W celu weryfikacji tego twierdzenia wykonano serię obliczeń numerycznych, wykorzystując program MOMIC autorstwa A. Karwowskiego [7]. Program ten, będący ulepszoną wersją znanego programu MININEC, umożliwia analizę struktur zbudowanych z segmentów cienkich przewodników. Niestety program nie umożliwia symulacji pobudzenia struktury zewnętrznym polem elektromagnetycznym. Możliwe jest natomiast "wbudowanie" w układ źródeł napięciowych.

W związku z wymienionymi ograniczeniami weryfikację przeprowadzono w oparciu o zasadę wzajemności. Dwa źródła pobudzające zostały umieszczone w miejscach obciążeń anteny. Zastosowano pobudzenie współfazowe oraz przeciwfazowe. Następnie obliczono pole elektryczne i magnetyczne w geometrycznym środku anteny.

Okrągłą antenę ramową o średnicy 20 cm podzielono na 20 segmentów. Na każdym segmencie rozpięto 4 prostokątne funkcje bazowe. Program rozwiązywał zatem układ 80 równań liniowych z 80 niewiadomymi. Obliczenia przeprowadzono dla częstotliwości 50 MHz, przyjmując pobudzenia o wartości 10 V. Obliczono wartości natężenia pola elektrycznego i magnetycznego w środku anteny (punkt o współrzędnych 0,0,0). W obliczeniach wykorzystano tzw. metodę krótkiego dipola - algorytm właściwy dla pól bliskich. Otrzymane wyniki przedstawiono w tablicach 4.2 - 4.5:

Tablica 4.2

Wartości natężenia pola elektrycznego, pobudzenia skierowane jednakowo

Składowa	Część rzeczywista [V/m]	Część urojona [V/m]	Moduł [V/m]	Faza [stopnie]
X	$-2.510 \cdot 10^1$	$-1.150 \cdot 10^{-2}$	$2.510 \cdot 10^1$	-180
Y	$-1.191 \cdot 10^{-4}$	$-2.464 \cdot 10^{-7}$	$1.191 \cdot 10^{-4}$	-180
Z	0	0	0	0

Tablica 4.3

Wartości natężenia pola magnetycznego, pobudzenia skierowane jednakowo

Składowa	Część rzeczywista [A/m]	Część urojona [A/m]	Moduł [A/m]	Faza [stopnie]
X	0	0	0	0
Y	0	0	0	0
Z	$-1.586 \cdot 10^{-8}$	$-9.369 \cdot 10^{-6}$	$9.369 \cdot 10^{-6}$	-90.1

Tablica 4.4

Wartości natężenia pola elektrycznego, pobudzenia skierowane przeciwnie

Składowa	Część rzeczywista [V/m]	Część urojona [V/m]	Moduł [V/m]	Faza [stopnie]
X	$-1.727 \cdot 10^{-2}$	$-1.038 \cdot 10^{-5}$	$1.727 \cdot 10^{-2}$	-180
Y	$1.797 \cdot 10^{-2}$	$1.105 \cdot 10^{-5}$	$1.797 \cdot 10^{-2}$	0
Z	0	0	0	0

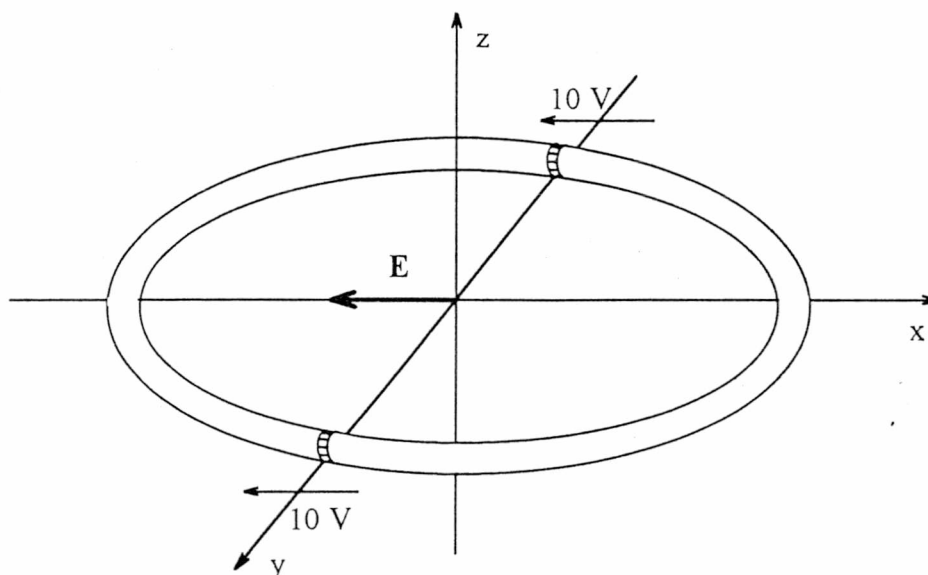
Tablica 4.5

Wartości natężenia pola magnetycznego, pobudzenia skierowane przeciwnie

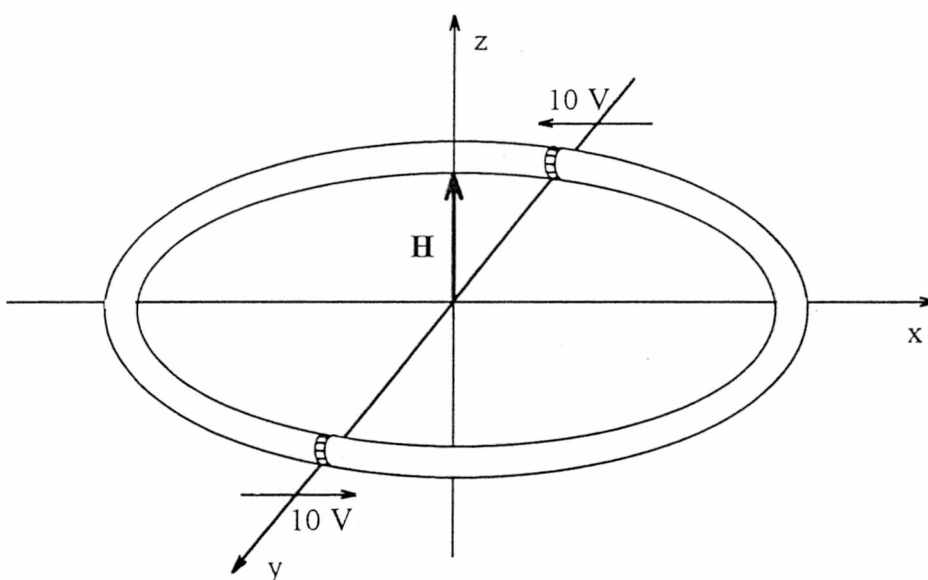
Składowa	Część rzeczywista [A/m]	Część urojona [A/m]	Moduł [A/m]	Faza [stopnie]
X	0	0	0	0
Y	0	0	0	0
Z	$1.459 \cdot 10^{-4}$	$6.395 \cdot 10^{-1}$	$6.395 \cdot 10^{-1}$	89.99

Z przedstawionych wyników obliczeń widać, że dla przypadku pobudzeń skierowanych jednakowo w środku anteny wytwarzane jest pole elektryczne (pozostałe niezerowe składowe mają poziom dużo niższy - ich pojawienie się jest związane z błędami obliczeń). Podobnie dla przypadku pobudzeń skierowanych przeciwnie wytwarzane jest pole magnetyczne.

Wytwarzane pola wraz z napięciami pobudzającymi przedstawiono na rysunku 4.7.



Rys. 4.7a. Podwójnie obciążona antena ramowa - pobudzenia skierowane jednakowo

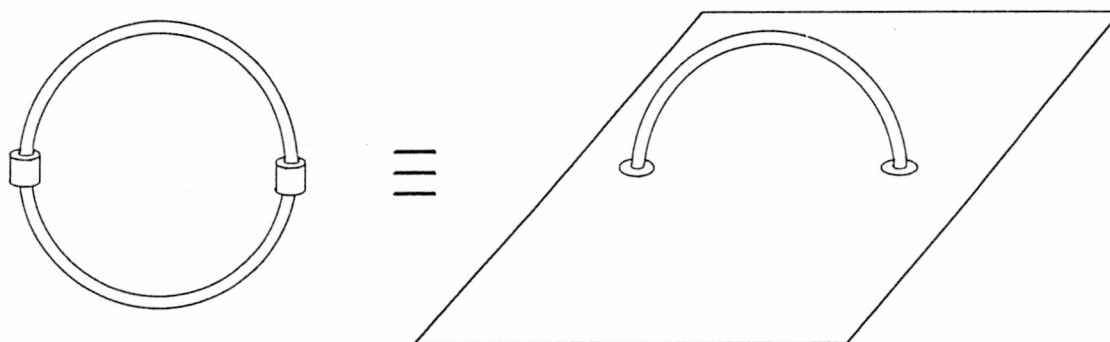


Rys. 4.7b. Podwójnie obciążona antena ramowa - pobudzenia skierowane przeciwnie

Przedstawione wyniki obliczeń numerycznych w pełni potwierdzają wyniki analizy teoretycznej. Zgodnie z zasadą wzajemności pole elektryczne w środku anteny spowoduje wyindukowanie się na jej obciążeniach napięć skierowanych jednakowo (suma jest różna od zera). Z kolei pole magnetyczne spowoduje wyindukowanie się napięć skierowanych przeciwnie (różna od zera jest ich różnica).

Należy zaznaczyć, że opisywane zjawiska nie są związane wyłącznie z antenami okrągłymi (które przyjęto do rozważań ze względu na istnienie opisu analitycznego). Kształt ramy jest w zasadzie dowolny. Dlatego też w praktycznych zastosowaniach można użyć anteny o innym, np. prostokątnym kształcie.

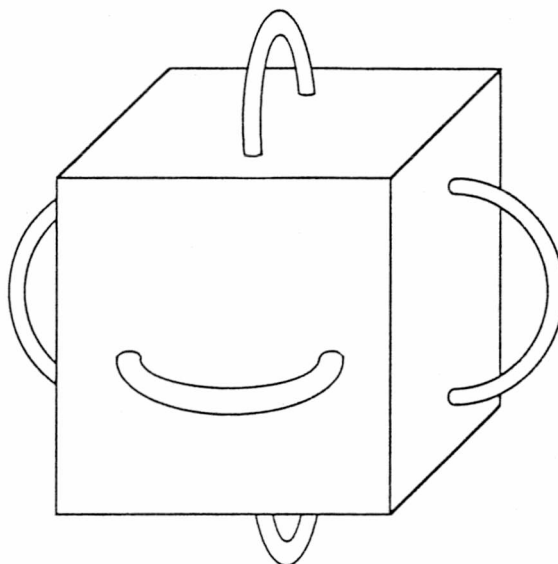
Dotychczasowe rozważania dotyczyły anten umieszczonych w wolnej przestrzeni. Analogiczne wyniki otrzymuje się w przypadku zastosowania połówki anteny ramowej umieszczonej nad doskonale przewodzącą płaszczyzną. Druga połówka jest symulowana przez zwierciadlane odbicie (Rys. 4.8).



Rys. 4.8. Antena ramowa uzyskana przez umieszczenie połówki pętli nad przewodzącą płaszczyzną

4.5.3. Koncepcja sondy pomiarowej

Opisane powyżej właściwości podwójnie obciążonych anten ramowych wykorzystano w projekcie izotropowej sondy pomiarowej do pomiaru pól elektromagnetycznych. Ponieważ pojedyncza antena ramowa reaguje tylko na jedną składową pola elektrycznego i jedną składową pola magnetycznego, to w celu otrzymania informacji o wszystkich sześciu składowych konieczne jest zastosowanie trzech anten umieszczonych ortogonalnie. Przyjęto koncepcję umieszczenia połówek anten na metalowym ekranie w kształcie sześciangu, co jest oryginalnym pomysłem opracowanym w ITA. Ekran ten stanowi zwierciadło "uzupełniające" anteny ramowe. Wielkość ekranu (który nie jest idealną płaszczyzną) będzie miała wpływ na kalibrację sondy. Ekran stanowi jednocześnie obudowę dla układów elektronicznych sondy (układy sumowania i odejmowania napięć, ewentualne wzmacniacze i przetworniki). Przykładową konstrukcję sondy przedstawia rysunek 4.9 (przypadek okrągłych anten ramowych). W celu komputerowego modelowania sond przygotowywany jest rozbudowany program analizy numerycznej (program MOMIC jest narzędziem zbyt słabym do takich analiz).

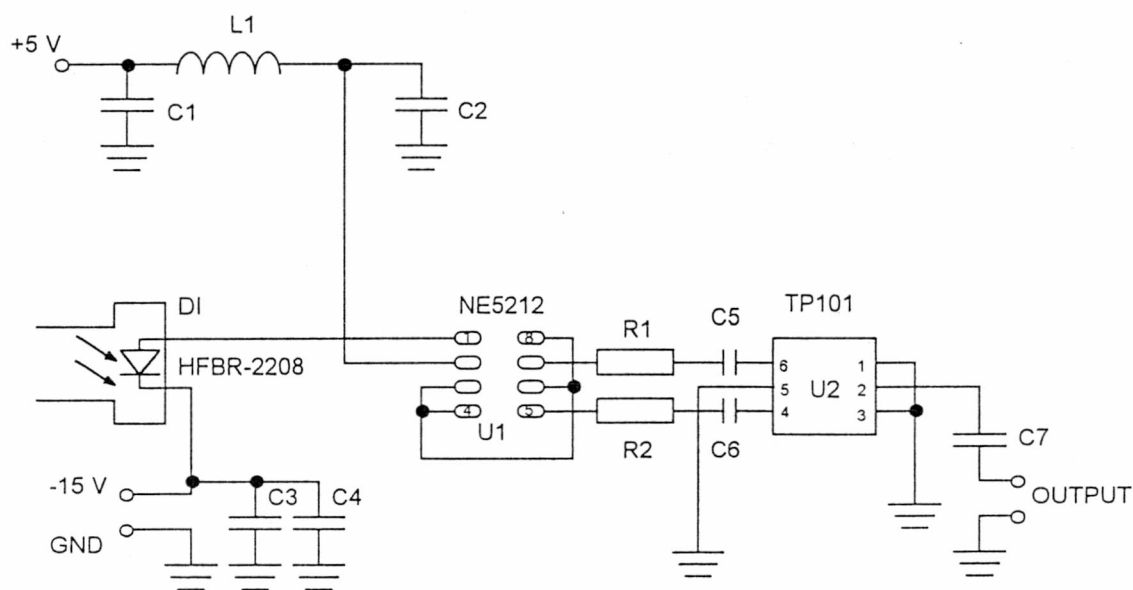


Rys. 4.9. Proponowana konstrukcja sondy

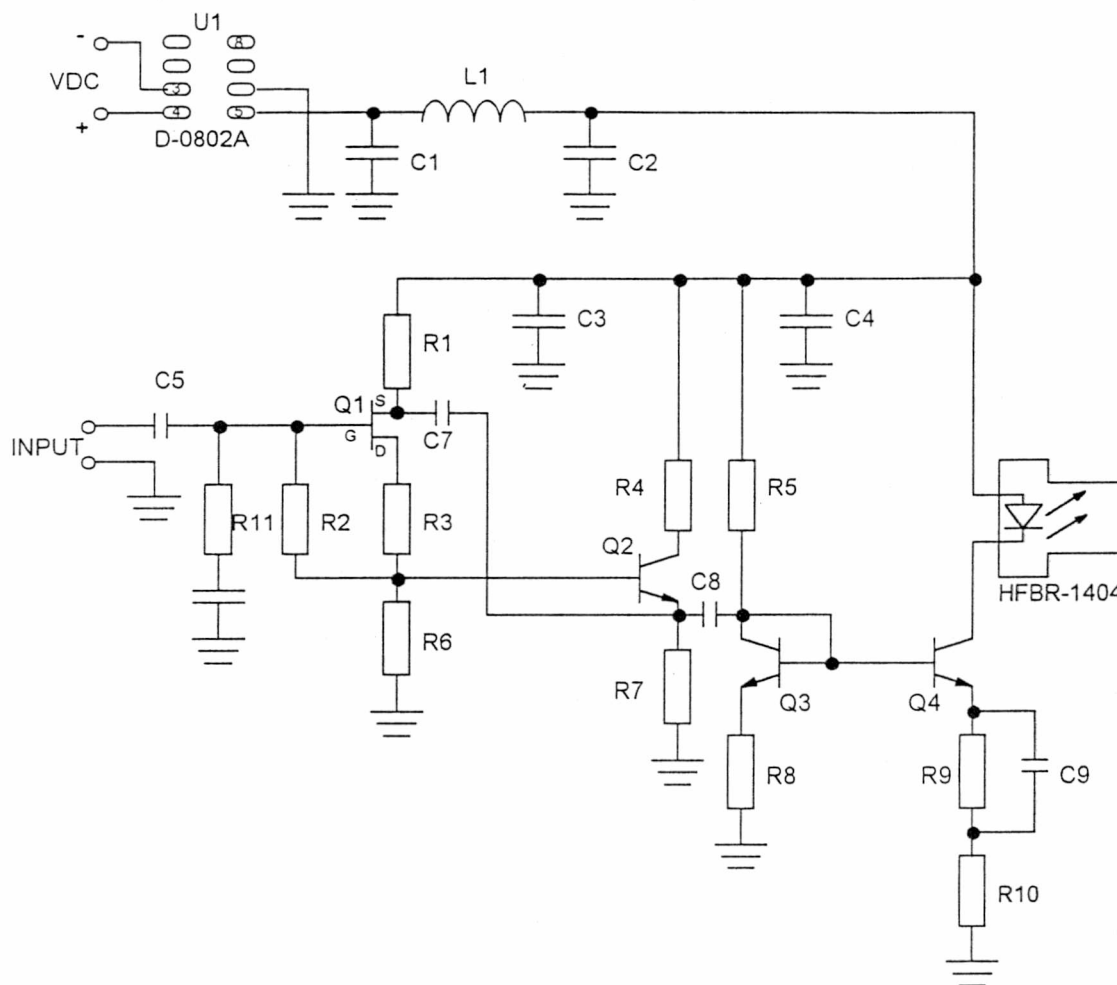
4.4.4. Łącze światłowodowe

W niniejszym podpunkcie zaprezentowano przykładowe rozwiązanie konstrukcji łącza optoelektronicznego, które można wykorzystać do przekazywania sygnałów pomiarowych z sondy do przyrządu pomiarowego. Rozwiązanie oparte jest na materiałach amerykańskiej firmy "The Electro-Mechanics Co."

Na rysunku 4.8 pokazano konstrukcję nadajnika optycznego, na rysunku 4.9 - odbiornika.



Rys. 4.10. Odbiornik łącza światłowodowego szerokopasmowej sondy pomiarowej



Rys. 4.11. Nadajnik łącza światłowodowego szerokopasmowej sondy pomiarowej

Prace nad sondą pomiarową są prowadzone w Zakładzie Radiokomunikacji ITA PWr w ramach Grantu Komitetu Badań Naukowych o numerze 8 S 504 025 04. Autor oczekuje, że realizacja opisanych sond umożliwi doświadczalną weryfikację modeli symulacyjnych przedstawionych w rozprawie.

5. Wnioski

Przeprowadzone w niniejszym dodatku analizy potwierdziły skuteczność zastosowania zaproponowanych w rozprawie metod do komputerowego modelowania komór TEM. Za swoje osiągnięcie autor uważa opracowanie metod i algorytmów umożliwiających:

- analizę pola w komorze pustej dla modu TEM (metoda momentów),
- analizę pola z obiektem przewodzącym dla modu TEM (metoda momentów),
- obliczenie częstotliwości odcięcia wyższych modów pola (dwuwymiarowa metoda TLM),

- obliczenie rozkładów pola odpowiadającym wyższym modom pola (dwuwymiarowa metoda TLM),
- analizę pola w komorze bez założeń upraszczających (w obecności wyższych modów, w komorze z dowolnym obiektem przewodzącym, dielektrycznym, stratnym, źródłowym) - trójwymiarowa metoda TLM,
- oszacowanie wartości częstotliwości rezonansowych komory - metoda TLM,
- symulację pobudzeń impulsowych w komorze - metoda TLM.

Rozprawa prezentuje pełną analizę struktury i właściwości pola elektromagnetycznego w komorach TEM, bez założeń upraszczających. Analiza taka była dotąd w literaturze niedostępna (patrz rozdział 2 rozprawy).

W niniejszym dodatku udowodniono dokładność zastosowanych metod poprzez porównanie wyników ilościowych uzyskanych przy ich pomocy pomiędzy sobą, a także z wynikami obliczeń przy pomocy profesjonalnego pakietu oprogramowania firmy Hewlett-Packard "HP 85180A High Frequency Structure Simulator". Należy tu podkreślić, że program ten został zakupiony do ITA PWr już po wydaniu rozprawy. Dlatego też porównanie metod było możliwe dopiero w niniejszym "Dodatku". Uzyskane wyniki w pełni potwierdziły zgodność wszystkich metod. Nie oznacza to jednakże, że zastosowane algorytmy są równoważne, w sensie ich wzajemnego zastępowania się przy realizacji poszczególnych zadań. Metoda TLM zastosowana przez autora ma możliwości niedostępne użytkownikowi programu HFSS (wymienione w punkcie 2 "Dodatku"). Z kolei program HFSS umożliwia łatwe niezależne śledzenie propagacji wyizolowanych wyższych modów pola, co w metodzie TLM jest trudne do implementacji (w metodzie TLM rozważane są jednocześnie wszystkie mody występujące w strukturze, dla zadanego pobudzenia). Należy więc raczej stwierdzić, że obie metody wzajemnie się uzupełniają.

Celowość stosowania metody uproszczonej dla modu TEM (rozdział 3 rozprawy) wynika natomiast z możliwości jej implementacji na maszynach o małej mocy obliczeniowej, klasy komputerów osobistych. Przeciętnego użytkownika komór TEM nie stać bowiem na sprzęt klasy stacji roboczych HP9000, a także na zakup profesjonalnego oprogramowania HP85180A HFSS (koszt programu rzędu kilkudziesięciu tysięcy dolarów).

W "Dodatku" przedstawiono także szczegółową koncepcję budowy sond pomiarowych, które w przyszłości powinny umożliwić eksperymentalną weryfikację zamieszczonych w rozprawie wyników obliczeń. Zaplanowanie i właściwe wykonanie takiego eksperymentu stanowi zagadnienie skomplikowane, wymagające zastosowania kosztownego oprzyrządowania i nie było przedmiotem przedstawionej rozprawy - stanowi natomiast jeden z punktów kontynuacji badań nad komorami TEM, wymienionych na str. 114 rozprawy.

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