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MODELLING ELECTROMAGNETIC RADIATION FROM DIGITAL ELECTRONIC SYSTEMS BY MEANS OF THE FINITE DIFFERENCE TIME DOMAIN METHOD

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ABSTRACT

The necessity for the control and minimisation of unintentional electromagnetic emissions from electrical systems has long been appreciated and much skilled effort is spent on EMI suppression. Due to the complexity of the problem, however, very little in the way of CAD tools is available to help the designer. In this contribution, a method is described, based on the Finite Difference Time Domain technique, whereby the radiation levels from digital circuits may be predicted. The predicted results are compared to measurements from some trial configurations.

INTRODUCTION

The necessity for the control and minimisation of unintentional electromagnetic emissions from electrical systems has long been appreciated. Nevertheless, many systems currently in production may fail to comply with the stringent EC Electromagnetic Compatibility (EMC) standards which will shortly become mandatory for all member states. There is no doubt that the EMC problem is assuming a prominent role within the design and manufacturing process.

Up to now, control of r.f. emission for EMC purposes in telecommunications systems has relied on the application of basic design principles, followed by careful measurement of prototypes. Of the few CAD tools available, none is capable of predicting the emissions expected from large, complex equipment such as a modern digital telephone exchange. If this predictive capability was improved, then development cost and time could be considerably reduced.

The approaches employed in the few available computer programs which can be applied to problems of this nature may be broadly divided into two philosophies. Some programs make many simplifying approximations in order to reduce the complexity of the problem and the required computer power. This approach has, however, a limited range of applications and cannot be satisfactorily used for large systems. The second approach is to solve Maxwell's equations directly, usually by applying the Method of Moments (MoM), the Transmission Line Matrix method (TLM) or Finite Elements (FE) as in (respectively) the public/commercial software packages NEC, STRIPES, and MSC/EMAS. Whilst this approach is, in principle, capable of giving accurate results, care has to be taken to ensure that the computational requirement is kept within bounds. Based on our experience with a wide variety of electromagnetic modelling techniques, we believe that a suitable CAD tool must include a rigorous solution of Maxwell's equations. Of the available formulations, we believe that the Method of Moments is not appropriate for this type of problem and that the Finite Difference Time Domain (FDTD) technique has advantages over both the TLM [1], with which it has similarities, and FE, and is most likely to give the best results (see later).

In this contribution, a technique is described, based on the Finite Difference Time Domain method, whereby accurate predictions of radiation may be obtained. This method is already proven for the prediction of crosstalk on printed circuit boards [2] and in this paper, calculations for the levels of radiation from a simple PCB are presented and compared with measured results. These results indicate that the technique has much potential for problems of realistic size and complexity.

POSSIBLE METHODS OF SOLUTION

The prediction of radiation levels from PCBs has received some attention in the literature. Many formulations use a TEM approximation, e.g. [3], but this becomes less satisfactory as faster signal rise-times need to be catered for. Of the published rigorous analyses, many use frequency domain methods such as the MoM, e.g. [4]. This method, however, requires computer storage and computation times of the order $(3N)^2$ and $(3N)^3$ respectively, where N is the number of unknowns. Although numerically efficient algorithms have been developed, the very long computation time and heavy storage requirements prohibit its application to problems involving complex geometries. The finite element technique, when used with efficient sparse matrix storage and inversion algorithms, offers a substantial improvement, with solution times and storage requirements proportional to $N^{1.5}$ and N respectively [5].

At Bristol University much work has been carried out over the past several years into the use of the FDTD technique for the analysis of electromagnetic problems involving complex geometries. In contrast to the FE method and the MoM, the FDTD method has storage and processing requirements proportional to N. Thus, some very complex problems that are too large to be handled by other techniques can be analysed by FDTD [6]. Moreover, a time domain method such as FDTD can be used with any desired excitation function, so enabling the effects of pulse shaping to be studied, and can provide information on the transient effects of crosstalk anywhere in or around the circuit without the need to use Fourier transforms.

Another important and useful advantage of using a time-domain technique is that non-linear components can be accommodated. Using frequency domain methods to analyse non-linear circuits creates major difficulties.

THE FINITE DIFFERENCE TIME DOMAIN METHOD

The basic FDTD technique is well described in the literature, so only the pertinent details are given here. The technique was first

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used by Yee [7] in 1966 who applied it to electromagnetic scattering problems. His method is to discretise Maxwell's curl equations in space and time to produce a simple time-stepping iterative procedure to calculate the field pattern as a function of time. No restrictions are imposed by the method on the geometry, materials or the incident fields and currents.

We start with Maxwell's two curl equations:

$$curl\ \mathbf{E} = \nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \qquad \ldots\ldots(1)$$

$$curl\ \mathbf{H} = \nabla \times \mathbf{H} = \mathbf{J} + \frac{\partial \mathbf{D}}{\partial t} \qquad \ldots(2)$$

These equations represent fundamental laws of electromagnetism and relate the field quantities E and H around a point in space.

They can be expanded in the rectangular co-ordinate system (x,y,z) to give six equations having the form shown in equation (3).

$$\frac{\partial E_x}{\partial t} = \frac{1}{\epsilon}\left(\frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} - \sigma E_x\right) \qquad \ldots(3)$$

Yee introduced a set of finite-difference equations analogous to the original Maxwell's equations. Following Yee's notation, a point in space is defined as

$$(i,j,k) = (i\delta, j\delta, k\delta) \qquad \ldots(4)$$

and any function F (x,y,z,t) of space and time is represented by its values at grid points:

$$F^n(i,j,k) = F(i\delta, j\delta, k\delta, n\delta t) \qquad \ldots(5)$$

where $\delta = \delta x = \delta y = \delta z$ is the space increment and δt is the time step.

Yee uses finite difference expressions for the space and time derivatives that are accurate to second order in both δ and δt.

$$\frac{\partial F^n(i,j,k)}{\partial x} = \frac{F^n(i+1/2,j,k) - F^n(i-1/2,j,k)}{\delta} + O(\delta^2) \qquad \ldots(6)$$

$$\frac{\partial F^n(i,j,k)}{\partial t} = \frac{F^{n+1/2}(i,j,k) - F^{n-1/2}(i,j,k)}{\delta t} + O(\delta t^2) \qquad \ldots(7)$$

To achieve the accuracy of (6), and to realise all of the space derivatives of (3), Yee positions the components of E and H about a unit cell. To achieve the accuracy of (7), E and H are evaluated at alternate half-time steps. The result of this procedure is a set of six finite-difference equations having the form shown in equation (8).

$$E_x^{n+1}(i+1/2,j,k) = \left[1 - \frac{\sigma(i+1/2,j,k)\delta t}{\epsilon(i+1/2,j,k)}\right] E_x^n(i+1/2,j,k) + \frac{\delta t}{\epsilon(i+1/2,j,k)\delta}\,[H_z^{n+1/2}(i+1/2,j+1/2,k) - H_z^{n+1/2}(i+1/2,j-1/2,k) + H_y^{n+1/2}(i+1/2,j,k-1/2) - H_y^{n+1/2}(i+1/2,j,k+1/2)] \qquad \ldots(8)$$

With this system the new value of a field vector component at any lattice point depends only on its previous value and on the previous values of the components of the other field vector at adjacent points.

The main difference between the FDTD method and the TLM method is that the FDTD method directly uses a finite difference form of Maxwell's equations, whereas the TLM method replaces the actual problem by an approximation based on a network of transmission lines. This approximate problem is then solved exactly. The direct approach of the FDTD means that computation time and storage is reduced for a given accuracy. In addition, it is easier to take account of dielectric materials, and to incorporate non-uniform grid sizes to optimise the computational efficiency.

It has recently been shown that the basic FDTD method and TLM are formally equivalent [8] so that the reduced computational requirements of the FDTD method can be exploited advantageously. At Bristol, several improvements have been made to the basic FDTD method which greatly improves computational efficiency. These are described below.

<u>Use of Variable Mesh Density</u>

The implementation of the Yee algorithm is necessarily based upon a finite number of cells. By using a non-uniform mesh, a greater number of nodes can be placed in regions of large field variations (e.g. track edges) to achieve good computational efficiency and accuracy. Nevertheless, most implementations of the FDTD method do not take advantage of it. For example in [9], no attempt has been made to use a non-uniform mesh or otherwise provide special treatment for the high field variations in the neighbourhood of the strip edge. Moreover in [10], the same authors report that an attempt to use a non-uniform mesh had a detrimental effect on the accuracy of their results. This necessitated their use of a large number of nodes (30x55x160) even for a single microstrip. In contrast to this, results have been obtained at Bristol University for a similar structure [11] using a highly non-uniform mesh without such a reduction in accuracy being observed. The ability to use a non-uniform mesh has enabled accurate results to be obtained using a much smaller number of nodes (12x20x36) for microstrip. More recent work at Bristol University [12] has confirmed that accurate results may also be obtained for radiating structures while still taking advantage of the benefits of a non-uniform mesh. For the problem of large system EMC, the use of such a technique is vital to ensure a realistically low computational requirement.

<u>Inclusion of a Priori Asymptotic Behaviour of the Fields</u>

By building on published work [13] on the behaviour of fields close to the edge of a metal surface, it has been possible to replace the standard finite difference equations with more accurate equations which take into account the non-linear field variation. We have shown [14] that an improvement of an order of magnitude in computational requirement can be obtained using this technique. To our knowledge, this enhancement has not been tried elsewhere.

APPLICATION TO A TRIAL PROBLEM

In order to prove the applicability of the technique to the problem of radiation from digital circuits, a set of simple trial problems was defined. Measurements were carried out and compared to predicted results obtained using the FDTD method. The structure of the problems which were analysed are shown in Figure 1. Although the geometries are simple, the range of feature sizes is realistic. It is the latter which poses the greatest challenge to the FDTD method. Here we have a PCB with a single microstrip track running in the x direction having a perfect voltage source at one end and a matched load at the other. Two different positions of the track were considered, firstly with the track centrally placed and secondly with the track placed 13mm from the edge of the PCB. The exact dimensions are shown in the figures. The excitation is a raised cosine pulse of unit peak field strength and a pulse width of 5 ns. The level of E field radiation was to be found at a distance of 3m from the track as a function of frequency over the range 5-600 MHz. In order to accommodate the large variation in feature size (1.6mm to 3m), highly non-uniform grids were set up. The effect of grid size on the accuracy of the model was checked by using two different grids as detailed in Table 1.

The grids are terminated by first order absorbing boundaries [15]. The time step in each case was chosen to be 1.7 ps which is safely less than the maximum value which guarantees stability.

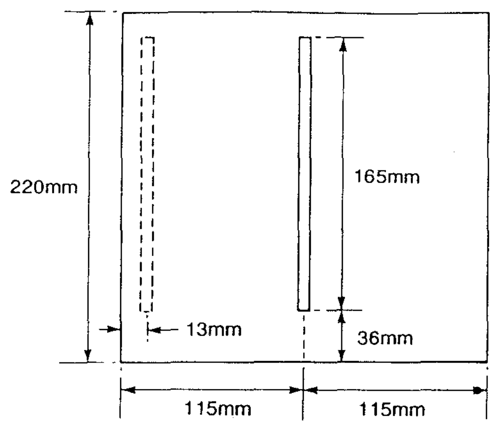

FIGURE 1a. PLAN OF PRINTED CIRCUIT BOARD. Both centre and off-centre microstrips are indicated.

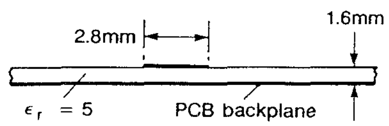

FIGURE 1b. ELEVATION OF PRINTED CIRCUIT BOARD.

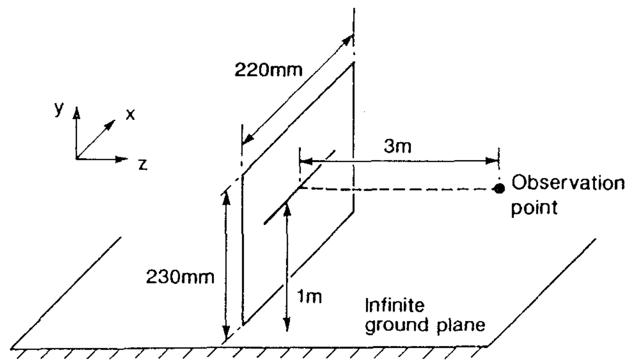

FIGURE 1c. MEASUREMENT GEOMETRY

RESULTS

For the theoretical model of the structure shown in Figure 1, observation points were chosen at the following co-ordinates: (0,1,0) i.e. between the strip and the PCB ground plane and (0,1,3), i.e. 3m in front of the PCB. Each of the three components of the E field were recorded. Note that for the purpose of this presentation, the co-ordinate origin lies on the ground plane directly below the strip mid-point.

Figure 2 shows the z component of the E field under the strip as a function of time.

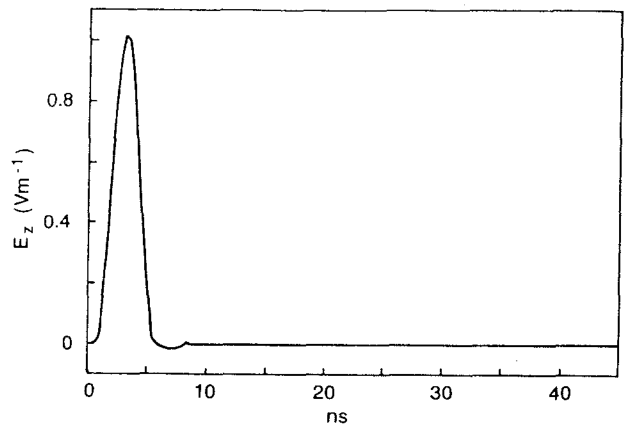

FIGURE 2. PREDICTED FIELD STRENGTH UNDER THE STRIP FOR THE STRUCTURE OF FIGURE 1.

Figure 3 shows the x component of the E field at the observation point 3 m in front of the PCB with the central track.

Measurements were carried out on the structures in the GPT 10m anechoic chamber at New Century Park, Coventry, England. The 50 ohm microstrip line was matched with a resistive load and excited by a screened battery-powered source consisting of a TTL crystal-controlled oscillator and type 74F37 buffer amplifier integrated circuit. The clock frequency of the source was 4.9152 MHz and its waveform was approximately square with a 10 ns rise time. The x component of the electric field strength at the 3m observation point was measured with the PCB and the antenna centre 1m above the chamber conducting ground plane using a biconical antenna below 200 MHz and a log-periodic antenna above that frequency. No attempt was made to correct the measured field values for coupling effects between the antenna and the chamber ground plane.

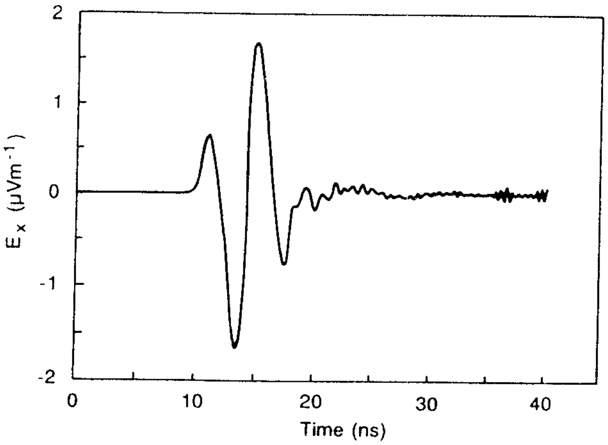

FIGURE 3. PREDICTED TRANSIENT RESPONSE FOR THE CENTRE STRIP PCB AT $\underset{\sim}{r}$ = (0,1,3)m.

In order to compare the measured and predicted results, a Field Transfer Loss was defined which was equal to the ratio of the field strength E_Z under the strip and the field strength E_X (or E_y) of the observation point. In other words.

$$FTL_x(f) = 20\log_{10}\left(\frac{|E_z(0,1,0;f)|}{|E_x(0,1,3;f)|}\right) \tag{2.1}$$

$$FTL_y(f) = 20\log_{10}\left(\frac{|E_z(0,1,0;f)|}{|E_y(0,1,3;f)|}\right)$$

In Figure 4 the measured Field Transfer Loss (FTL_X) for the PCB with the central track is compared with that predicted using the different sized grids. It can be seen that there is agreement between the predicted results using the different grids of better than 2 dB up to a frequency of approximately 250 MHz. This corresponds to a node spacing of approximately $\lambda/16$ for the finer grid and $\lambda/8$ for the coarser grid. Above this frequency the results obtained from the two grids diverge. Nevertheless the discrepancy between both sets of predicted results and the measurements is within the limits of experimental error up to about 400 MHz. This indicates that it may be possible to use grids coarser than the usually recommended node spacing ($\lambda/10$) for this type of problem.

The effect of the position of the PCB track on the radiation was investigated by considering the case where the strip is placed 13 mm from the edge of the board. The other details of the structure are the same as shown in Figure 1.

A comparison of the predicted field transfer loss FLT_X for the central and non-central strip showed no significant differences between the two geometries. This agrees with measured values (Figure 5) within likely experimental errors.

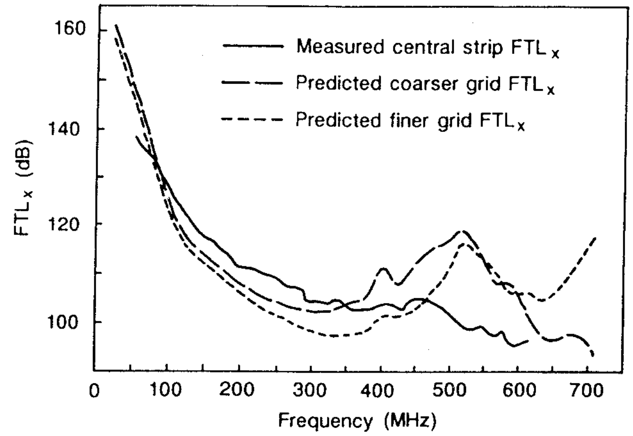

FIGURE 4. FTL_X FOR THE CENTRE STRIP PCB USING TWO DIFFERENT FDTD MESHES.

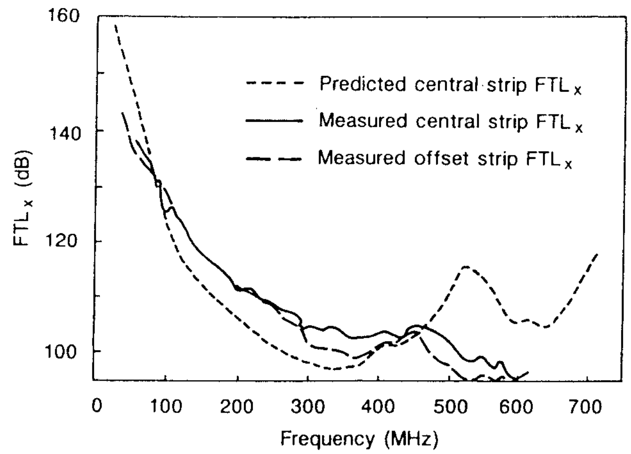

FIGURE 5. FTL_X FOR THE CENTRE AND OFF-CENTRE STRIP PCBs.

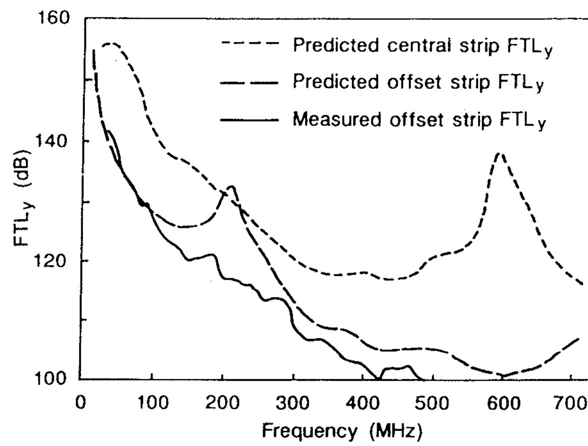

FIGURE 6. FTL_y FOR THE STRUCTURES OF FIGURE 1.

The magnitude of E_y (and therefore of FTL_y) for the central track could not be measured since it was below the noise levels of the measuring equipment. However, measured FTL_y values for the non-central track are compared with prediction in Figure 6.

It can be seen from Figure 6 that, apart from a feature at around 200 MHz, there is agreement between predictions and measurement to within 4dB. A substantially higher FTL_y for the PCB with the centrally placed strip is also correctly predicted by the model. The origin of the predicted 200 MHz feature is under investigation.

CONCLUSION

In this contribution we have shown that the Finite Difference Time Domain method is well suited to the problem of predicting radiation levels from printed circuit boards and offers advantages over alternative methods such as the MoM, TLM and FE. Results for trial problems which have simple structures but a realistic range of feature sizes have been presented and compared to measured results and, with a good choice of mesh configuration, yields good agreement, despite the fact that the model makes no allowance for the presence of the antenna. The technique lends itself well to being extended to cater for very complex systems by means of partitioning the problem into sets of "equivalent sources" and using field transformation techniques. Further work on radiation from open-circuit strip tracks and 3-layer strip-line geometries is under way and will be reported at the Conference.

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TABLE 1 GRIDS USED FOR THE STRUCTURE OF FIGURE 1

Region sizes in metres: Each region consists of 2 x 2 x 2 unit cells

Grid 1			Grid 2		
x	y	z	x	y	z
0.127	0.1475	0.1284	0.254	0.295	0.2568
0.127	0.1475	0.1284	0.254	0.295	0.2568
0.127	0.1475	0.1284	0.254	0.1475	0.2568
0.127	0.1475	0.1284	0.127	18.63e-3	0.2568
0.127	0.1475	0.1284	17.25e-3	18.63e-3	1.6e-3
0.127	0.1475	0.1284	17.25e-3	18.63e-3	1.6e-3
0.127	18.63e-3	0.1284	3e-3	18.63e-3	0.1284
17.25e-3	18.63e-3	0.1284	16.1e-3	18.63e-3	0.2568
17.25e-3	18.63e-3	1.6e-3	16.1e-3	18.63e-3	0.2568
3e-3	18.63e-3	1.6e-3	16.1e-3	2.6e-3	0.2568
16.1e-3	18.63e-3	0.1284	16.1e-3	2.6e-3	0.2568
16.1e-3	18.63e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	2.6e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	2.6e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	18.63e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	18.63e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	18.63e-3	0.1284	16.1e-3	18.63e-3	0.2568
16.1e-3	18.63e-3	0.1284	3e-3	0.295	0.2568
16.1e-3	18.63e-3	0.1284	0.127	0.295	0.2568
16.1e-3	18.63e-3	0.1284	0.254	0.295	0.2568
3e-3	0.1475	0.1284	0.254		0.2568
0.127	0.1475	0.1284	0.254		
0.127	0.1475	0.1284			
0.127	0.1475	0.1284			
0.127	0.1475	0.1284			
0.127	0.1475	0.1284			
0.127		0.1284			
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