Fast Simulation Methods for the Detection of Reflection- and Crosstalk Effects During the Design of Complex Printed Circuit Boards

E. Griese*, J. Schrage**, M. Vogt**

* Siemens Nixdorf Informationssysteme AG - Cadlab - Analog System Engineering
** University of Paderborn - Cadlab - Analog System Engineering
Bahnhofstrasse 32 · 33094 Paderborn · Germany

Abstract

In this paper two fast frequency domain methods are presented for the determination of the voltages at all terminations of coupled transmission line systems. Algorithms based on these methods can be used for a fast preanalysis of complex PCBs with respect to reflection- and crosstalk effects. The dynamical behaviour of a transmission line system is described by a frequency dependent transfer matrix which is calculated with two different methods. The first method is based on a Padé Approximation and the second requires the knowledge of the eigenvalues and the eigenvectors of the product of the inductance and capacitance matrix. Within both methods the symmetry of these matrices and a fundamental property of the matrix exponential were used to develop efficient algorithms.

1 Introduction

Increasing demands on circuits and printed circuit boards result in a higher complexity, increasing clock rates, and faster component technologies. Therefore, the probability of EMC-effects like reflection and crosstalk during the operation of modern electronic equipment is very high. To ensure an error-free operation of electronic systems, and to reduce the number of redesigns as well as the time to market it is necessary to detect and to handle these effects in an early design stage of PCBs. The reasons for reflections which may cause critical voltage levels and functional errors are for example transmission line discontinuities and mismatched terminations. Crosstalk problems may occur if two or more transmission lines have been routed parallel over a long distance. It is caused by electromagnetic coupling and may lead to functional errors. An intensive study of both effects has shown that they cannot be handled separately. This fact makes it more complicated and time consuming to analyse PCBs with regard to signal integrity effects.

The detection of these noise effects during the design of printed circuit boards and their necessary reduction to non-disturbing levels require special simulation methods.

2 Reflection- und Crosstalk Analysis

The simulation of reflection- and crosstalk effects is very time consuming because these effects can only be simulated together and the consideration of the nonlinear behaviour of the inputs and output gates of digital components requires time domain methods. To reduce the total effort and analysis time the complete analysis process can be subdivided into a fast and approximative preanalysis and an accurate simulation of those subnets which have been classified as critical within the preanalysis process.

The task of the preanalysis is to reduce the high number of nets to be simulated by an appropriate classification into probably critical nets and guaranteed not critical ones. In recent years different approaches for a fast preanalysis were made to estimate the voltage levels caused by reflection and crosstalk as described in [2] and [9]. In general most of them cannot be used within an automatic preanalysis process because their simplifying assumptions are very restrictive. The reflection behaviour of simple transmission line structures (e.g. two-point-nets) can be estimated by a rule driven approach presented in [7] and [9]. Also simply coupled transmission line areas can be handled by algorithmic rules [9]. As arbitrary structured nets have in general to be analysed a classification based on an approximative calculation of the voltages at all termination networks seems to be the most powerful approach.

The extraction of the transmission line structures from complex PCB designs is only efficient using ap-
appropriate CAD tools which are able to extract the relevant information automatically from the layout data [3]. As the presented method is based on the transmission line theory the transmission lines do not have to be described by their three-dimensional geometry. They can be handled by appropriate models which consist in the case of lossless arrangements solely of capacitive and inductive elements that can only be calculated by numerical methods in general.

### 3 Modelling

Precondition for the presented method is the validity of the transmission line theory. This means that the longitudinal components of the electric and magnetic field strength can be neglected in comparison to the transversal ones. This kind of wave propagation is called quasi-TEM wave propagation. Another precondition is that the influence of transmission line discontinuities like vias and edges can be neglected. Furthermore, all termination networks are assumed to have approximately a linear behaviour.

Using these assumptions and preconditions every transmission line structure on a printed circuit board can be described by a very simple model which consists solely of coupled and not coupled homogeneous areas. In general, these areas are modelled by a system of \( N \) coupled transmission lines which are described by their length \( l_i \) and their capacitance and inductance matrices \( \mathbf{C}_i ' \) and \( \mathbf{L}_i ' \) respectively (figure 1).

Transmission line structure:

![Transmission line structure](image1)

Structure model:

![Structure model](image2)

**Figure 1:** Example of a typical transmission line structure on digital printed circuit boards and its simplified structure model.

Equivalent to this model are the following equations which have to be satisfied by the voltages \( u_n \) and the currents \( i_n (n \in [1, N]) \) are equivalent to this model.

\[
\frac{\partial u_n (t, z)}{\partial z} = - \sum_{j=1}^{N} K'_{nj} \frac{\partial u_j (t, z)}{\partial t} + \sum_{j=1}^{N} L'_{nj} \frac{\partial i_j (t, z)}{\partial t}
\]  

Introducing voltage and current vectors which comprise the \( u_n \) and \( i_n \), and quadratic inductance and capacitance matrices:

\[
\mathbf{L}' = \begin{bmatrix} | L'_{ij} | \end{bmatrix}, \quad \mathbf{C}' = \begin{bmatrix} | C'_{ij} |, \quad C'_{ij} = \begin{cases} \sum_{n=1}^{N} K'_{in} & \text{for } i = j \\ - K'_{ij} & \text{for } i \neq j \end{cases} \end{bmatrix}
\]  

a vectorial partial differential equation of the order 2 is obtained:

\[
\frac{\partial}{\partial z} \left( \begin{bmatrix} u(t, z) \\ i(t, z) \end{bmatrix} \right) = - \left( \begin{bmatrix} 0 & \mathbf{L}' \end{bmatrix} \mathbf{C}' \begin{bmatrix} 0 \\ \mathbf{0} \end{bmatrix} \right) \frac{\partial}{\partial t} \left( \begin{bmatrix} u(t, z) \\ i(t, z) \end{bmatrix} \right).
\]  

The application of the Fourier Transformation yields

\[
\frac{d}{dz} \left( \begin{bmatrix} u(j \omega, z) \\ i(j \omega, z) \end{bmatrix} \right) = - j \omega \left( \begin{bmatrix} 0 & \mathbf{L}' \end{bmatrix} \mathbf{C}' \begin{bmatrix} 0 \\ \mathbf{0} \end{bmatrix} \right) \left( \begin{bmatrix} u(j \omega, z) \\ i(j \omega, z) \end{bmatrix} \right).
\]  

The solution of (6) can be described using the transition matrix \( \Psi(j \omega, z - z_0) \):

\[
\left( \begin{bmatrix} u(j \omega, z) \\ i(j \omega, z) \end{bmatrix} \right) = \Psi(j \omega, z - z_0) \cdot \left( \begin{bmatrix} u(j \omega, z_0) \\ i(j \omega, z_0) \end{bmatrix} \right),
\]  

which is given by the matrix exponential of the system matrix, multiplied by the length of the transmission line area:

\[
\Psi(j \omega, z - z_0) = \exp \left( - j \omega \left( \begin{bmatrix} 0 & \mathbf{L}' \end{bmatrix} \mathbf{C}' \begin{bmatrix} 0 \\ \mathbf{0} \end{bmatrix} \right) (z - z_0) \right).
\]  

As this method is working in the frequency domain, a linear behaviour of all components has to be assumed. Taking into account that short computation times are required, the input and output behaviour of digital components have to be modelled solely by a few components. The used models are shown in figure 2.

Output model of a driver:

![Output model of a driver](image3)

Input model of a receiver:

![Input model of a receiver](image4)

**Figure 2:** Linear input and output models.
The voltage source $u_q$ of the output model generates a (periodical) trapezoidal voltage which is characterized by the parameters rise time $t_r$, fall time $t_f$, low-level voltage $u_{q1}$, high-level voltage $u_{q2}$, the duration $T$, and the period $T_0$. The input model can be derived from the output model by setting $u_{q2}$ and $u_{q1}$ to zero. During the preanalysis process these models have to be parametrized by technology dependent values.

### 4 Computation of the Frequency Dependent Transition Matrix

The computation of the voltages at the inputs and outputs of the components requires the solution of (6) respective (8) for different frequencies. The necessary matrix exponential can be calculated exactly by the corresponding eigenvalues and eigenvectors. Furthermore, the matrix exponential can be computed by approximative methods by appropriate power series.

As the structure of the differential equation system (6) is very simple and the inductance and capacitance matrices are symmetric,

\[
\begin{align*}
L' &= L' \\
C' &= C'
\end{align*}
\]

the matrix exponential has a very simple structure too, which can be used to develop fast methods for its calculation. Using the corresponding Taylor series the required matrix exponential has the following structure:

\[
\Phi(j\omega, z - z_0) = \left( \begin{array}{c}
B_1 \\
C_1
\end{array} \right) + j \left( \begin{array}{c}
B_2 \\
C_2
\end{array} \right) + \cdots,
\]

where

\[
\begin{align*}
L_1' &= -\omega (z - z_0) L' \\
C_1' &= -\omega (z - z_0) C'.
\end{align*}
\]

In the following two methods are described which allow a fast and accurate calculation of the submatrices $B_1$ and $B_2$.

#### 4.1 Calculation of the Submatrices $B_1$ and $B_2$ by the Corresponding Eigenvalues and Eigenvectors

To compute $B_1$ and $B_2$ exactly, the system of differential equations given in (6) has to be transformed. Introducing two quadratic non singular matrices $M$ and $M$ with

\[
\begin{align*}
\bar{u}(j\omega, z) &= M \cdot \bar{u}(j\omega, z), \quad \bar{y}(j\omega, z) = M \cdot \bar{y}(j\omega, z),
\end{align*}
\]
equation (6) can be transformed into

\[
\frac{d}{dz} \left( \begin{array}{c}
\bar{u}(j\omega, z) \\
\bar{y}(j\omega, z)
\end{array} \right) = -j\omega \left( \begin{array}{c}
L' \bar{u}(j\omega, z) \\
C' \bar{y}(j\omega, z)
\end{array} \right),
\]

where $L'$ and $C'$ are diagonal matrices. This can be achieved by an appropriate choice of $M$ and $M$. For their determination let $\bar{u}^{(k)}$ be the $k$ th column vector of $M$, $\bar{u}^{(k)}$ be the $k$ th column vector of $M$, and $\bar{u}^{(k)}$ and $\bar{u}^{(k)}$ the $k$ th elements of the diagonal matrices $L'$ and $C'$. Then, these quantities have to satisfy the equations

\[
L' \cdot \bar{u}^{(k)} = \bar{u}^{(k)} \cdot \bar{u}^{(k)},
\]

\[
C' \cdot \bar{u}^{(k)} = \bar{u}^{(k)} \cdot \bar{u}^{(k)},
\]

which lead to the eigenvalue problem

\[
\left( L' \cdot C' - \frac{1}{v_k} \cdot I \right) \cdot \bar{u}^{(k)} = 0.
\]

Here $v_k$ ($v_k^2 = (i\bar{u}^{(k)} \cdot \bar{u}^{(k)})^{-1}$) is the propagation velocity of the $k$ th mode. $\bar{u}^{(k)}$ can be calculated from $\bar{u}^{(k)}$ with equation (15). It can be shown that $M$ and $M$ satisfy the equation

\[
M^T \cdot M = 1.
\]

If the product matrix $L' \cdot C'$ has a multiple eigenvalue $v_k^{-2}$ of the order $m$, a modification of Schmidt’s Orthonormalization Method can be used to avoid numerical problems. Furthermore, the application of this method preserves the property given in (17). Let $\bar{x}^{(kj)}$ ($j \in [1, m]$) be the corresponding $m$ linear independent eigenvectors of $L' \cdot C'$. Then, $\bar{u}^{(kj)}$, $\bar{u}^{(kj)}$, $\bar{u}^{(kj)}$, and $\bar{u}^{(kj)}$ can be calculated using the following formulas.

\[
\begin{align*}
\bar{x}^{(kj)} &= \bar{x}^{(kj)} - \sum_{i=1}^{j-1} \left( \bar{x}^{(kj)} \cdot \bar{u}^{(ki)} \right) \cdot \bar{u}^{(ki)}
\end{align*}
\]

\[
\bar{u}^{(kj)} = \frac{\bar{u}^{(kj)} \cdot |\bar{u}^{(kj)}|^{-1}}{\bar{u}^{(kj)} \cdot |\bar{u}^{(kj)}|^{-1}}
\]

\[
\bar{u}^{(kj)} = \frac{1}{\bar{u}^{(kj)} \cdot |\bar{u}^{(kj)}|^{-1}}
\]

As the matrices $C'$ and $L'$ are diagonal, the matrices $B_1$ and $B_2$ are given by

\[
B_1 = M \cdot \cos \left( \omega z \sqrt{L' \cdot C'} \right) \cdot M^T,
\]

\[
B_2 = M \cdot \sqrt{(L' \cdot C')} \cdot \sin \left( \omega z \sqrt{L' \cdot C'} \right) \cdot M^T,
\]

where $z := z - z_0$. The newly introduced matrices $\sin \left( \omega z \sqrt{L' \cdot C'} \right) \cos \left( \omega z \sqrt{L' \cdot C'} \right)$ are diagonal and therefore they can be computed very fast.
4.2 Calculation of $B_1$ and $B_2$ by a Padé Approximation

The required submatrices of the matrix exponential can also be computed fast and accurately by the aid of a Padé Approximation which is a fraction of polynomials $\exp A \approx D_n(A) \cdot N_n(A)$. The corresponding expressions for the matrix exponential of an arbitrary matrix $A$ are given in [1] and [6]. The Padé Approximation can be used if the norm of the argument matrix $A$ is not too large ($\|A\|_{\infty} = \max_{1 \leq i \leq m} \sum_{j=1}^{n} |a_{ij}|, A \in \mathbb{R}^{m \times n}$). Otherwise, roundoff errors and the computation time will not be acceptable. These difficulties can be controlled by exploiting a fundamental property of the matrix exponential

$$e^{m A} = (e^A)^m, \quad e^A = (e^{\frac{1}{m} A})^m.$$  \hspace{1cm} (25)

Then, the exponential of the normalized matrix $A/m$ can easily be computed. If $m$ is chosen to a power of two the matrix exponential will be obtained by repeated squaring. Using this method $B_1$ and $B_2$ of the normalized matrix are approximately given by

$$B_1 \approx (\hat{B}_1^2 - \phi \cdot \hat{B}_2^2) \cdot (\hat{B}_2^2 + \phi \cdot \hat{B}_1^2)^{-1},$$  \hspace{1cm} (26)

$$B_2 \approx 2 \hat{B}_1 \hat{B}_2 \cdot (\hat{B}_2^2 + \phi \cdot \hat{B}_1^2)^{-1},$$  \hspace{1cm} (27)

where $\phi$ is the product of the scaled matrices $L'$ and $C'$ ($\phi := L' \cdot C'/m^2$) and

$$\hat{B}_1 = \sum_{n=1}^{q \div 2} \frac{(2q - 2n)! q! (-\phi)^n}{(2q)! (2n)! (q - 2n)!},$$

$$\hat{B}_2 = \sum_{n=1}^{(q-1) \div 2} \frac{(2q - 2n - 1)! q! (-\phi)^n}{(2q - 2n)! (2n - 1)! (q - 2n - 1)!}. $$\hspace{1cm} (28)

In (28) and (29) the commutativity of $\phi$, $\hat{B}_1$, $\hat{B}_2$, and $(\hat{B}_2^2 + \phi \cdot \hat{B}_1^2)^{-1}$ has been exploited.

The quality of the approximation is determined by the parameter $q$ which is responsible for the upper summing indices in (28) and (29). Investigations on the accuracy of the Padé Approximation of the matrix exponential in [1] and [6] have shown that a relative error $\delta(q)$ and a quadratic error matrix $E$ can be defined so that the following correlations are valid:

$$\|A\|_{\infty} \leq 2^{-q(q+1)}$$

$$e^A = D_n(A) \cdot N_n(A)$$

$$\|E\|_{\infty} \leq \delta(q) \cdot \|A\|_{\infty}$$

$$\delta(q) = \frac{2^{q-2} q!^2}{2q!(q + 1)!}$$

Already for small values of the summing parameter $q$ an accuracy is obtained which is less than that of floating point representations of computers ($\delta(5) < 10^{-15}$).

If $q$ is chosen to be odd ($q = 2k + 1$) then the upper summing indices in (28) and (29) are equal and solely $k$ powers of $\phi$ have to be computed.

The necessary squaring of the scaled matrix exponential can be done computing the corresponding submatrices $B_1$ and $B_2$. It is obvious that the new submatrices $B_{1, n \omega_0}$ and $B_{2, n \omega_0}$ can be calculated from the old quantities as follows:

$$B_{1, n \omega_0} = B_{1, \omega_0} - \phi \cdot B_{2, \omega_0},$$

$$B_{2, n \omega_0} = 2 B_{1, \omega_0} B_{2, \omega_0}.$$ \hspace{1cm} (30)

$$B_{1, n \omega_0} = B_{1, \omega_0} - \phi \cdot B_{2, \omega_0},$$

$$B_{2, n \omega_0} = 2 B_{1, \omega_0} B_{2, \omega_0}.$$ \hspace{1cm} (31)

4.3 Calculation of $B_1$ and $B_2$ for Different Frequencies

The application of a Fourier analysis requires the knowledge of $B_1$ and $B_2$ for integer multiples of a basis frequency $\omega_0$. After $B_1$ and $B_2$ have been computed for $\omega_0$ as given in (23) and (24) or (30) and (31) respectively, the corresponding quantities for $n \omega_0$ ($n \in \mathbb{N}$) can be computed very fast by making use of the fundamental property of the matrix exponential given in (25). This means that the transition matrices and of course $B_1$ and $B_2$ can be computed recursively for different frequencies $(n + 1) \omega_0$ by

$$\Psi(j(n + 1) \omega_0) = \Psi(j \omega_0) \cdot \Psi(j n \omega_0),$$

$$B_{1, j(n + 1) \omega_0} = B_{1, j \omega_0} B_{1, j n \omega_0} + \hat{L} \hat{C} B_{2, j \omega_0} - B_{2, j \omega_0} B_{1, j n \omega_0},$$

$$B_{2, j(n + 1) \omega_0} = B_{1, j \omega_0} B_{2, j n \omega_0} + B_{2, j \omega_0} B_{2, j n \omega_0}.$$ \hspace{1cm} (32)

In case of a not coupled transmission line system $L'$ and $C'$ are diagonal. Therefore, the equations (23) and (24) can be used to calculate $B_1$ and $B_2$ taking into account that in this case $M_{\alpha}$ and $M_{\beta}$ are equal to the identity matrix. As the calculation of trigonometrical functions needs much more computation time than multiplications even in case of not coupled systems the above described method for computing the higher frequency matrices should be used.

4.4 Transfer Matrix of a System of Coupled Transmission Lines

The transfer matrix of the complete transmission line system is obtained by computing the product series of all transfer matrices of each homogeneous area. Again the block structure of the matrices can be used to simplify and fasten the calculations.

$$\Psi(j \omega) = \prod_{n=1}^{N} \Psi_n(j \omega, l_n) = \left( \begin{array}{c} \Psi_{11}(j \omega) \Psi_{12}(j \omega) \\ j \Psi_{21}(j \omega) \Psi_{22}(j \omega) \end{array} \right)$$

$$\Psi_n(j \omega, l_n) \in \mathbb{R}^{2 \times 2}$$ \hspace{1cm} (33)

Here $\Psi_n(j \omega, l_n)$ is the transfer matrix of the $n$th area with the length $l_n$. 

$$\Psi(j \omega) = \prod_{n=1}^{N} \Psi_n(j \omega, l_n) = \left( \begin{array}{c} \Psi_{11}(j \omega) \Psi_{12}(j \omega) \\ j \Psi_{21}(j \omega) \Psi_{22}(j \omega) \end{array} \right)$$

Here $\Psi_n(j \omega, l_n)$ is the transfer matrix of the $n$th area with the length $l_n$. 

$$\Psi(j \omega) = \prod_{n=1}^{N} \Psi_n(j \omega, l_n) = \left( \begin{array}{c} \Psi_{11}(j \omega) \Psi_{12}(j \omega) \\ j \Psi_{21}(j \omega) \Psi_{22}(j \omega) \end{array} \right)$$
In this section a system of two coupled transmission lines (figure 3) which has been extracted with a layout data extractor [3] from a real multilayer PCB design is analysed. The results are compared with those obtained by using the time domain transmission line simulator FREA CS [4].

The structure model of the transmission line system which consists of one coupled area (3) and four not coupled ones (1, 2, 4, 5) is shown in figure 4. The corresponding parameters are given in table 3. Transmission line 1 is active while line 2 is assumed to be passive (low pegel).

![Structure model of the transmission line system](image)

**Figure 4: Structure model of the transmission line system**

The simulation results shown in figure 5 and 6 have been obtained using the Padé Approximation of the matrix exponential. In figure 7 and 8 the results are illustrated which have been obtained calculating the eigenvalues and eigenvectors with the orthonormalization procedure. As this method is working in the frequency domain, the comparison with FREA CS-results was done by calculating five periods of the signal to ensure the steady state of the voltages.

![Voltage at the driver (u11) and the receiver (u12) of the active transmission line (FREA CS: ---, present method with a Padé Approximation: - - -)](image)

**Figure 5: Voltage at the driver (u11) and the receiver (u12) of the active transmission line (FREA CS: ---, present method with a Padé Approximation: - - -)**

![Voltages at the driver (u22) and the receiver (u12) of the passive transmission line (FREA CS: ---, present method with a Padé Approximation: - - -)](image)

**Figure 6: Voltages at the driver (u22) and the receiver (u12) of the passive transmission line (FREA CS: ---, present method with a Padé Approximation: - - -)**

As the coupled area of this example is located between the potential layers of the PCB both eigenvalues of the product matrix \(L' C'\) are identical. Therefore, the exact evaluation of the matrix exponential without the orthonormalization cannot be applied. The results are extremely sensitive against small variations.

![Table 1: Transmission line parameters](image)

**Table 1: Transmission line parameters**

<table>
<thead>
<tr>
<th>Coupled area (Area 3)</th>
<th>Area 1</th>
<th>Area 2</th>
<th>Area 4</th>
<th>Area 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>(l_{c} [\text{mm}])</td>
<td>77.254</td>
<td>2.885</td>
<td>101.008</td>
<td>2.591</td>
</tr>
<tr>
<td>(C'_{c} [\text{pF/m}])</td>
<td>96.689</td>
<td>95.354</td>
<td>95.354</td>
<td>96.689</td>
</tr>
<tr>
<td>(L'_{c} [\text{mH/m}])</td>
<td>329.390</td>
<td>498.252</td>
<td>498.252</td>
<td>329.390</td>
</tr>
</tbody>
</table>

![Table 2: Model parameters of all inputs and outputs](image)

**Table 2: Model parameters of all inputs and outputs**

<table>
<thead>
<tr>
<th>Inputs</th>
<th>Outputs</th>
</tr>
</thead>
<tbody>
<tr>
<td>(R_{in} = 1 \Omega)</td>
<td>(R_{out} = 26 \Omega)</td>
</tr>
<tr>
<td>(C_{in} = 5 \text{ pF})</td>
<td>(C_{out} = 12 \text{ pF})</td>
</tr>
<tr>
<td>(l_{f} = 0.5 \text{ ns})</td>
<td>(l_{f} = 0.2 \text{ ns})</td>
</tr>
<tr>
<td>(T = 20 \text{ ns})</td>
<td>(U_{ph_{1}} = 0 \text{ V})</td>
</tr>
<tr>
<td>(U_{ph_{2}} = 5 \text{ V})</td>
<td></td>
</tr>
</tbody>
</table>
of \( L' \) and \( C' \) which can in general be computed solely by numerical methods. If the numerical product of these matrices is proportional to the identity matrix the method without the orthonormalization fails.

Although, 25 Fourier coefficients have solely been considered, the results obtained using these methods agree very well with that obtained using FREACS. The maximum and minimum values of the voltages are approximated very well. Also the results of both presented methods agree very well. The simulation of this configuration requires less than 30% of the computation time which is necessary using FREACS for the analysis of one time period.

6 Conclusion

In this paper numerical methods have been presented which can be used within a fast reflection- and crosstalk preanalysis for the computation of the voltages at all terminations of the examined transmission line system. A submatrix formalism and the symmetry of the inductance and capacitance matrix have been exploited to develop and implement efficient algorithms which have been optimized with regard to the computation time and the required memory.

The comparison of obtained results with that of other methods shows a very good agreement while less computation time is necessary. The presented methods are applicable to each transmission line structure. Furthermore, there are no restrictions with respect to the transmission line parameters and the rise and fall time of the expected signals. Therefore, they can be used within an automatic preanalysis process of complex printed circuit boards.

Acknowledgements

This work is part of the JESSI project AC-5 Development of an EMC-Workbench for Microelectronic Application which is supported by the German Government, Department of Research and Technology under grant 01 M 2886 D and 01 M 2886 E. The responsibility for this publication is held by the authors only.

References