KONINKLIJKE VLAAMSE ACADEMIE VAN BELGIE
VOOR WETENSCHAPPEN EN KUNSTEN

# URSI FORUM 2003
# RADIO SCIENCE ON THE MOVE

## BRUSSELS, 18 DECEMBER 2003

**K.U.Leuven, afdeling ESAT-TELEMIC**
**Emmanuel Van Lil, Bart Nauwelaers, Dominique Schreurs, Antoine Van de Capelle, Guy Vandenbosch, eds.**

**Vlaams Kennis- en Cultuurforum**

Handelingen van het contactforum "URSI Forum 2003-Radio Science on the Move" (18 december 2003, hoofdaanvrager: Prof. Emmanuel Van Lil, Katholieke Universiteit Leuven) gesteund door het Vlaams Kennis- en Cultuurforum van de Koninklijke Vlaamse Academie van België voor Wetenschappen en Kunsten.

Afgezien van het afstemmen van het lettertype en de alinea's op de richtlijnen voor de publicatie van de handelingen heeft de Academie geen andere wijzigingen in de tekst aangebracht. De inhoud, de volgorde en de opbouw van de teksten zijn de verantwoordelijkheid van de hoofdaanvrager van het contactforum.

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URSI Forum 2003

Palace of the Academies
18th of December

PROGRAMME OF THE DAY

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| | | |
|---|---|---|
| **0930-0940** | **Welcome address** | Prof. dr. ir. Vloeberghs, Chairman URSI/BE |
| **0940-0945** | **Introduction to the Forum & Acknowledgements** | Prof. dr. ir. Van Lil, Chairman organisation |

**0945-1045** **ORAL SESSION A : TELECOMMUNICATIONS**
(chairman : Prof. dr. ir. E. Schweicher)

0945-1005 A1. *Validation of source parameter extraction with the sage algorithm using noise free data,*
J. Verhaevert, K.U.Leuven

1005-1025 A2. *Etude théorique et expérimentale de la distribution des ondes électromagnétiques ELF émises par une source dipolaire immergée en eau de mer peu profonde*,
O. Thongsamouth, ERM

1025-1045 A3. *Integral Function Method for the Analysis of Harmonic Distortion of SOI MOSFETs,*
B. Parvais, UCL

**1045-1125** **Coffee break (till 1115) & Poster session** (36 items)

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**1125-1225** A4. *Quelques Exemples de l'Évolution Récente des Technologies des Télécommunications Filaires et Hertziennes,*
Prof. P. Degauque, USTL, France

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1235-1350 **Lunch** (in University Foundation across the boulevard (Egmont street, 11))

**1400-1500 ORAL SESSION B : MICRO- AND NANOSTRUCTURES**
(chairman Prof dr. ir. B. Nauwelaers)

1400-1420 B1. *Ferromagnetic Nanowires for Microwave Filters and Nonreciprocal Devices*
A. Saib, UCL

1420-1440 B2. *Embedded Thin-Film HEMTs in Multi-Chip Modules*
R. Vandersmissen, IMEC

1440-1500 B3. *An efficient FMM-PML-MPIE Formalism for 2D Microstrips*
D. Vande Ginste, UGent

**1500-1615 Coffee break (till 1530) & Poster session**

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**1615-1630 Debriefing and comments**

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1630-1730 **Farewell cocktail**

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**Important notice for the authors of the POSTER presentations.**

The posters will be visible all day. So, the authors are kindly invited to post their presentations as early as possible. Special Velcro strips provided by the organisers will have to be used **(thumbnails are *strictly* forbidden).**

# 11th URSI Forum 2003
# in the Academy Palace on 18 December 2003

## INTRODUCTION AND ACKNOWLEDGEMENTS

**Professor dr. ir. Claude VLOEBERGHS**

Chairman of the URSI Belgian Committee

This year, the now traditional URSI forum is not held in a Belgian university, like in the past, but in the Academy Palace in the frame of the so-called "contact forums" organised by the Royal Academies for Science and the Arts of Belgium (RASAB). I would like to express my big gratitude to the Academies for the support we got and, I believe, we still get for the realisation of this Forum.

Even if this URSI Forum takes place for the first time in the beautiful Academy Palace, I have to emphasise that we are proud to celebrate this year the tenth anniversary of our Forum. Indeed, the first edition was held in the Royal Military Academy in 1993.

This anniversary gives me the opportunity to come back to the roots of these forums and also to rectify what I wrote last year in the introduction of the $10^{th}$ URSI Forum. If I correctly mentioned the name of Professor Schweicher, who was at that time the organiser of the first edition and the chairman of the Belgian URSI committee, I forgot to stress that the idea for the creation of these forums was coming from Professor Paul Delogne of the UCL. Indeed, when he was chairman of the committee, just before Professor Schweicher, he was very concerned by the fact that the only realisations of the Belgian committee at that moment consisted each year in four, let us say, "closed meetings" of a few URSI members. So, he formulated then the encouraging project to have a one-day special meeting where young researchers, preparing a PhD in the field of the domains covered by URSI, will get the opportunity to present their work and also to exchange ideas with their colleagues from other laboratories in Belgium. And, I think, this last point is very important for the organisers of each forum; they have to manage sufficient time during the posters sessions, during the breaks and the lunch to principally foster discussions between the young researchers themselves. So, I hope that it will be the case during this forum.

Since the first edition, the URSI Forum took place each year in different Belgian universities. It was organised at least one time by almost all the Faculties of Applied Sciences, alternatively from both sides of the language border, and also by the Scientific Institutes related with the of the Royal Observatory of Belgium in Uccle. Taking the remarks and the comments of the previous forums into account, the Belgian URSI committee and the organisers of the next forum always tried to enhance its setting and its programme. An important step was accomplished for the forum of the year 2000 when the committee decided to open the forum to researchers and speakers from European countries. Since then, we always had the great pleasure to note an increasing number of participants and, also, to welcome one or two guest speakers presenting an invited paper. For the present edition, Professor DEGAUQUE, from the University of LILLE in France, has accepted to address the forum with a special lecture about his research activities on the EM propagation. Professor, I thank you in advance for your presentation; I am sure you will raise much interest among the attendance.

Although the forum of this year is not held on the university site of its organising committee, it does not mean that the task of the forum's organisers became easier. The request for credit, the selection of the presentations, oral or by poster, the redaction of the proceedings, some logistic activities, briefly many different tasks have also to be accomplished like in the past. The Belgian URSI committee is very proud and pleased to can rely for the perfect achievement of the forum of this year on the team from the ESAT department of the KULeuven, led by its past-chairman, Professor Van Lil. The title he gave to this forum, "Radio Science on the Move", is broad enough to cover the different domains of URSI. But, moreover, this title emphasises the increasing role of wireless applications in our modern society. It is therefore an encouragement to our young scientists for the enhancement of our communication and information systems.

Before to give the floor to Professor Van Lil, who will inform us about the details of this forum, I would like that you congratulate him and his colleagues for the nice and interesting day they prepared for us. Thank you very much!

**KEYNOTE ADRRESS**
# QUELQUES EXEMPLES DE L'EVOLUTION RECENTE DES TECHNOLOGIES DES TELECOMMUNICATIONS FILAIRES ET HERTZIENNES

Pierre Degauque et Martine Liénard

Université des Sciences et Technologies de Lille

Avec l'émergence des nouvelles générations de téléphonie cellulaire, liée à celles des réseaux locaux sans fil, les premiers pas menant aux réseaux de communication mobile, universelle et personnelle (UMTS *Universal Mobile Telecommunications System*) ont été franchis. Si, dans les années 1970 – 1990, les liaisons mobiles pour le grand public étaient essentiellement axées sur la transmission de la voix, ne nécessitant donc que des débits de quelques kbits/s, l'émission et la réception de données à haut débit est devenue une nécessité quotidienne, la généralisation de l'Internet n'étant d'ailleurs pas étrangère à ce développement. Dans ce contexte, l'élaboration d'un nouveau système de communications passe d'abord par la caractérisation et l'optimisation de ce qui est communément appelé: la "couche physique". Pour un environnement donné, les paramètres caractéristiques, comme le bruit radioélectrique présent dans la bande d'émission et la fonction de transfert du canal, doivent d'abord être déterminés, soit expérimentalement, soit à partir de modèles théoriques de propagation ou d'une approche purement stochastique. A partir de cette connaissance, et en introduisant des contraintes liées au système de transmission, les techniques de modulation et de codage les plus appropriées pour assurer le meilleur compromis entre débit et taux d'erreurs, peuvent être précisées. Compte tenu de l'ampleur des efforts déployés actuellement dans ce domaine, nous nous proposons d'illustrer cette évolution des technologies en nous limitant, dans la plupart des exemples présentés, aux transmissions de données dans des structures confinées, comme l'intérieur des bâtiments, et habituellement désignées sous le vocable de transmissions *indoor*. Nous envisagerons également le cas d'une liaison filaire pour nous attarder sur une technique en émergence.

Un des défis majeurs dans le domaine des télécommunications mobiles, donc dans le domaine hertzien, est d'assurer une augmentation des débits de transmission tout en minimisant la puissance émise et la bande de fréquences nécessaires. De nombreuses recherches ont été menées ces dernières années sur l'utilisation de réseaux d'antennes à l'émission et à la réception. Ces techniques, connues sous l'appellation anglo-saxonne de MIMO, acronyme de *Multiple Input Multiple Ouput*, sont basées sur l'utilisation d'un codage spatio temporel. Cependant, leur efficacité dépend fortement de la décorrélation entre canaux et donc des propriétés mathématiques de la matrice de transfert liant les antennes d'émission et de réception. Après avoir décrit l'évolution des techniques de réception ou (et) d'émission en diversité, nous montrerons sur quelques exemples, les potentialités que l'on peut attendre d'un tel système implanté dans un environnement *indoor*. Mentionnons enfin que les perspectives ouvertes par les transmissions ultra large bande et la montée en fréquence vers le spectre des ondes millimétriques sont également des moyens de répondre à ce besoin de haut débit.

Si on envisage maintenant une liaison fixe, différentes technologies d'accès sont commercialisées et notamment l'ADSL (*Asymetric Digital Subscriber Line)* qui utilise, comme support physique, les paires téléphoniques cuivrées. D'autres techniques sont encore à l'étude et notamment les transmissions sur ligne d'énergie PLC (*Power Line Communication*). Dans ce cas, aucun câblage supplémentaire n'est nécessaire, puisque les infrastructures industrielles ou résidentielles possèdent un vaste réseau électrique. Cela procure également l'avantage d'une grande souplesse d'utilisation puisque la plupart des appareils électroniques sont, très souvent, branchés sur le « réseau secteur ». Néanmoins de nombreux points restent en suspens, liés notamment au problème de perturbation radioélectrique et à la prévision de portée de la liaison. Ces différents aspects seront évoqués ainsi que d'autres applications potentielles de cette technique comme, par exemple, dans le domaine des voitures 'tout électronique' du futur.

# EXTENDED ABSTRACTS

***Commission C: Radio-Communication systems & signal processing***

# VALIDATION OF SOURCE PARAMETER EXTRACTION WITH THE SAGE ALGORITHM USING NOISE FREE DATA

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## 1. INTRODUCTION

Because of the success of wireless communications, the understanding and the prediction of the radio-propagation characteristics in different indoor and outdoor configurations becomes more and more important. Of course, it is very valuable to have all parameters that model the environment accurately, without conducting expensive and time-consuming propagation measurements. The extraction techniques need to be verified, which is the topic of this paper.

In this work, the Direction of Arrival (DOA) estimation with the Space Alternating Generalised Expectation maximisation (SAGE) Algorithm will be tested with simulated data. Not only the time delay (or the distance) and the angle will be estimated, but also the complex amplitude (or the power) of every significant path. The simulation set-up is based on a real indoor measurement set-up at the University of Kassel [1]. The measured bandwidth is 600 MHz at a carrier frequency of 1.8 GHz. Instead of using the measurement data, the configuration set-up will be modelled with the Enhanced Propagation for Indoor Communications Systems (EPICS) program, a ray-tracing tool available at our university, and results in the simulated data.
The organisation of this paper is as follows. Section 2 handles the set-up of the simulations. Section 3 presents the simulation results. Conclusions will be drawn in Section 4.

## 2. SIMULATION SET-UP

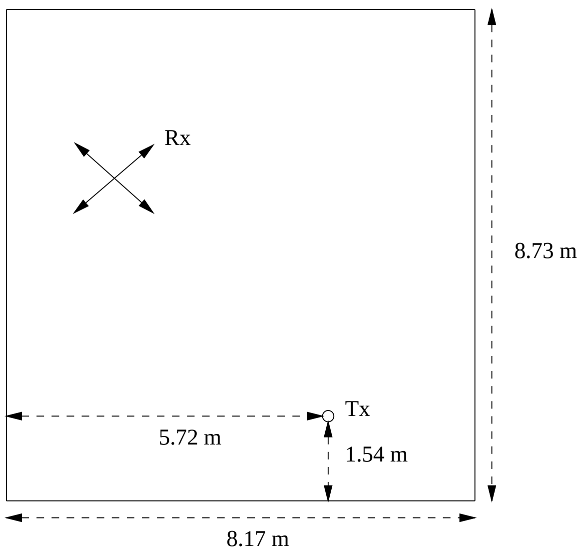


Figure 1: Topview of the configuration set-up

The simulations were done in a classic indoor configuration, which can be seen in Figure 1. The size is 8.17 m on 8.73 m. The origin is placed in the lower left corner of the Figure. The co-ordinates of the antennas are given as (x, y). The transmitting antenna Tx is positioned at (5.72 m, 1.54 m). The receiving antenna array Rx is a mill's cross antenna array, where one linear array is perpendicular to the Line of Sight (LOS). The beginning co-ordinates thereof are (1.71 m, 5.64 m) and the ending ones are (2.08 m, 5.99 m). For the other linear antenna array, parallel with the LOS, the co-ordinates are (1.75 m, 6.03 m) and (2.04 m, 5.60 m) respectively. The inter-antenna spacing is 0.012 m. Because the ground floor and the ceiling are not modelled, the simulation set-up can be seen as a two-dimensional configuration. Each antenna array consists of 45 antennas, hence resulting in the total number of the mill's cross antenna array of 90 antennas. The

advantages of this array are extensively described in [2].
This configuration was simulated with the Enhanced Propagation for Indoor Communications Systems (EPICS), which is a ray-tracing tool, available at our university ([3], [4]). With EPICS, the transfer function for the 90 different antennas as function of 511 frequencies in a bandwidth of 1.2 GHz can be calculated.
These data, generated in the frequency domain, are transformed to the time domain and a training sequence is included, because the available implementation of the Space Alternating Generalised Expectation maximisation (SAGE) Algorithm is programmed in the time-domain, using training sequences. This transformation is extensively explained in [5]. A theoretical study of the SAGE Algorithm can be found in [6], [7] and [8]. This optimisation algorithm is used to replace the high dimensional optimisation procedure necessary to compute the joint maximum likelihood estimate of all the parameters by several separate maximisation processes, which can be performed sequentially.
By applying this SAGE Algorithm on the simulated and processed data, time-domain values of the Direction of Arrival (DOA) parameters can be determined in a few seconds. The extracted parameters are the time delay (or the distance), the angle and the complex amplitude (or the power) for every significant path.

## 3. SIMULATION RESULTS

In Figure 2, the results of the extraction with the SAGE Algorithm (plotted in stars) are compared with the results of the EPICS program, which are plotted in circles. The power levels are given on the Figure. For every ray, the first number is the power calculated with EPICS, the second one is the power of the estimation with the SAGE Algorithm. The simulated and calculated rays are the direct ray and the single reflections; because we consider only the two-dimensional situation, the total number of rays is limited to 5.
The DOA parameter extraction with the SAGE Algorithm matches the ray-tracing results from EPICS very accurately. Even the estimation of the power (or the complex amplitude) is done in a precise way.

## 4. CONCLUSIONS

In this paper, the Direction of Arrival (DOA) estimation with the Space Alternating Generalised Expectation maximisation (SAGE) Algorithm has been shown to co-incidence with the results from the EPICS ray-tracing tool, validating both the algorithm and the tool. Not only the time delay and the angle are estimated accurately, but also the complex amplitude (and hence the power) has been shown to correspond with the simulations very precisely. The mill's cross antenna array has been proven to eliminate the angle ambiguity.

## 5. ACKNOWLEDGEMENTS

The authors gratefully acknowledge the support of Dave Trappeniers for adapting the output of the EPICS program.

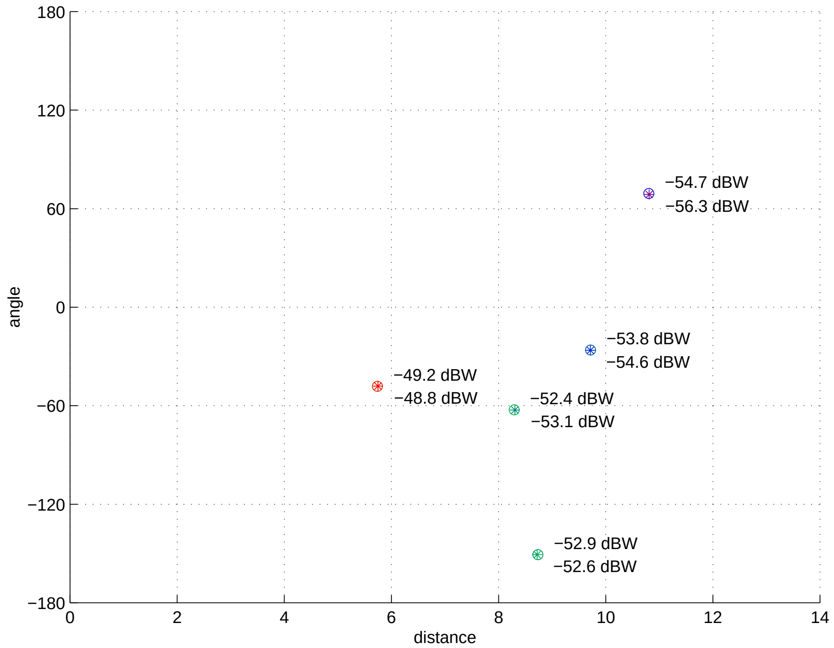


Figure 2: The results: comparison between SAGE and EPICS

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## ETUDE THEORIQUE ET EXPERIMENTALE DE LA DISTRIBUTION DES ONDES ELECTROMAGNETIQUES ELF EMISES PAR UNE SOURCE DIPOLAIRE IMMERGEE EN EAU DE MER PEU PROFONDE

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## 1. INTRODUCTION

Depuis la Deuxième Guerre mondiale, il est connu que le bruit et les champs électromagnétiques générés par les vaisseaux de surface peuvent être détectés par les mines. Ces mines intelligentes sont équipées de capteurs sismiques, acoustiques, magnétiques ou électriques qui leur permettent d'identifier un navire en particulier. La signature électromagnétique des navires est située dans la bande de fréquences des ELF (Extremely Low Frequency, $\nu < 3$kHz). La menace des mines peut être réduite par une maîtrise des signatures acoustique et électromagnétique du navire ou par un système de leurres.
Afin d'évaluer les risques encourus par les vaisseaux de surface et l'efficacité des systèmes de leurres, nous devons donc être capables de prédire les signaux émis dans l'eau de mer par une source connue.
Dans ce papier, on se propose d'étudier la distribution des champs électromagnétiques émis

- par des sources élémentaires horizontales,
- en régime harmonique,
- dans la bande de fréquence des ELF,
- en champ proche,
- et en eau de mer peu profonde.

Nous présentons tout d'abord un modèle théorique adéquat permettant de calculer les vecteurs champs électrique et magnétique. Des résultats expérimentaux obtenus lors d'une des campagnes de mesures en environnement réel sont ensuite analysés avant d'être finalement comparés aux résultats théoriques.

## 2. FORMULATION THEORIQUE DES CHAMPS EN MILIEU CONDUCTEUR A COUCHES PLANAIRES MULTIPLES

Les formulations analytiques des champs électrique et magnétique s'obtiennent en utilisant une décomposition de champs développée dans le passé pour les lignes de transmission dans les circuits intégrés [1]. Le principe consiste tout d'abord à transformer les champs et la source dans le domaine de Fourier spatial, dans les directions x et y et de séparer ensuite les champs en deux catégories comprenant chacune trois composantes : une catégorie sans composante Ez (mode TE) et une catégorie sans composante Hz (mode TM).

Nous considérons un milieu composé de quatre couches : l'air, la mer et deux types de sol. Nous utilisons un système de coordonnées cartésiennes dans lequel l'axe z est perpendiculaire aux couches et orienté positivement vers le haut, le plan (x,y) correspondant au plan des couches. Chaque couche est numérotée j de bas en haut, où j=1,...,n et n est le nombre total de couches (n=5 dans notre cas car une couche fictive a été ajoutée pour pouvoir introduire la source). La première et la dernière couche sont semi-infinies. Chaque couche j est homogène et isotrope, et est caractérisée par une permitivitté $\varepsilon_j$, une conductivité $\sigma_j$ et une perméabilité $\mu_j$. La coordonnée z des interfaces entre deux couches j-1 et j est notée $z_{(j-1,j)}$.

Nous calculons les champs électromagnétiques rayonnés par un dipôle électrique horizontal (HED) et un dipôle magnétique vertical (VMD). Dans le cadre de la menace des mines, nous sommes amené à nous intéresser aux ondes électromagnétiques émises par une source immergée en eau de mer peu profonde et en champ proche. Dans le cas du HED, les deux modes TE et TM interviennent tandis que dans le cas du VMD, seul le mode TE est excité. Les champs sont représentés dans chaque couche j par une équation du type

$$E_j^{TX} = A_j^{TX} e^{-\Gamma_j (z - z_{j-1,j})} + B_j^{TX} e^{\Gamma_j (z - z_{j-1,j})} \tag{1}$$

$$H_j^{TX} = \frac{1}{Z_j^{TX}} \left( A_j^{TX} e^{-\Gamma_j (z - z_{j-1,j})} - B_j^{TX} e^{\Gamma_j (z - z_{j-1,j})} \right) \tag{2}$$

où TX signifie TE ou TM; $\Gamma_j = \left(k_x^2 + k_y^2 - k_j^2\right)^{1/2}$, $\Re e(\Gamma) > 0$ ; $k_j^2 = -i\omega\mu_j(\sigma_j + i\omega\varepsilon_j)$ ; $Z_j^{TE} = -\dfrac{i\omega\mu_j}{\Gamma_j}$ et $Z_j^{TM} = \dfrac{\Gamma_j}{\sigma_j + i\omega\varepsilon_j}$ . La résolution d'un système d'équations considérant les conditions aux limites aux interfaces nous permet de trouver les valeurs des coefficients $A_j^{TX}$ et $B_j^{TX}$. Après transformation de Fourier inverse, les composantes cartésiennes des champs électrique et magnétique s'expriment par des intégrales de Sommerfeld. Ces intégrales sont calculées numériquement. Le calcul de ces intégrales est facilité par le fait de travailler
avec des ondes ELF dans l'eau de mer et en champ proche. En effet, les fonctions à intégrer ne présentent pas de pôle sur l'axe réel, oscillent faiblement et les intégrales convergent assez rapidement. Les résultats ont été validés théoriquement par comparaison avec d'autres modèles à deux et trois couches.

## 3. CAMPAGNE DE MESURES

Une campagne de mesures a été réalisée lors de tests OTAN de brouillage de mines sur la côte Ouest de la Floride à Panama City. Les mesures ont été effectuées en eau peu profonde (profondeur moyenne de 16m). Nous disposions d'un dipôle électrique horizontal et d'un système de mesures multi-influences, appelé SIGMA (SIGnatures and Mine Algorithms), appartenant à la Défense britannique (Qinetic).

Le dipôle horizontal de 2m de long est fixé, à l'aide d'un support en GRP (Glass Reinforced Plastic), en dessous de la coque non-métallique d'un bateau de 7m de long et alimenté par un générateur fournissant un courant alternatif (sinusoïdal) d'une intensité de 10A . Le capteur électrique est un capteur triaxial à gradient de potentiel composé d'électrodes à bruit faible. Le capteur magnétique est un magnétomètre fluxgate . Les données sont transmises en temps réel à la tour de contrôle via des câbles sous-marins. Le SIGMA est situé au fond de l'eau à une position DGPS connue (30° 08' 40.2 N, 85° 47' 14.5 W). L'expérience consiste à faire des allers et retours avec le dipôle au-dessus du SIGMA. Un DGPS embarqué à bord du bateau permet de connaître sa trajectoire exacte. Nous avons effectué trois séries de mesures à des fréquences distinctes, chaque série comprenant plusieurs passages du bateau au-dessus du SIGMA. Au début et en fin de chaque série de mesures, une mesure "à vide", c'est-à-dire avec le dipôle éteint, a été prise afin de pouvoir distinguer le signal du bruit. Les mesures étant relativement bruitées, nous avons eu recours au filtrage numérique.

## 4. COMPARAISON ENTRE LES RESULTATS THEORIQUES ET EXPERIMENTAUX

Les mesures filtrées sont ensuite comparées aux courbes théoriques. Les résultats théoriques présentent soit un retard, soit une avance sur toutes les composantes par rapport aux signaux expérimentaux. Nous expliquons ce décalage par l'incertitude de position du GPS. Les trajectoires ont ainsi été adapté en conséquence. Mis à part quelques composantes, les erreurs absolues en amplitude sont, après correction de trajectoire, en général inférieures à 3dB. La figure 1 est une comparaison entre les résultats théoriques et expérimentaux obtenus pour la composante x du champ électrique avec et sans correction de trajectoire. Les mesures ayant été réalisées en environnement réel, nous pouvons conclure que les résultats obtenus répondent à nos attentes.

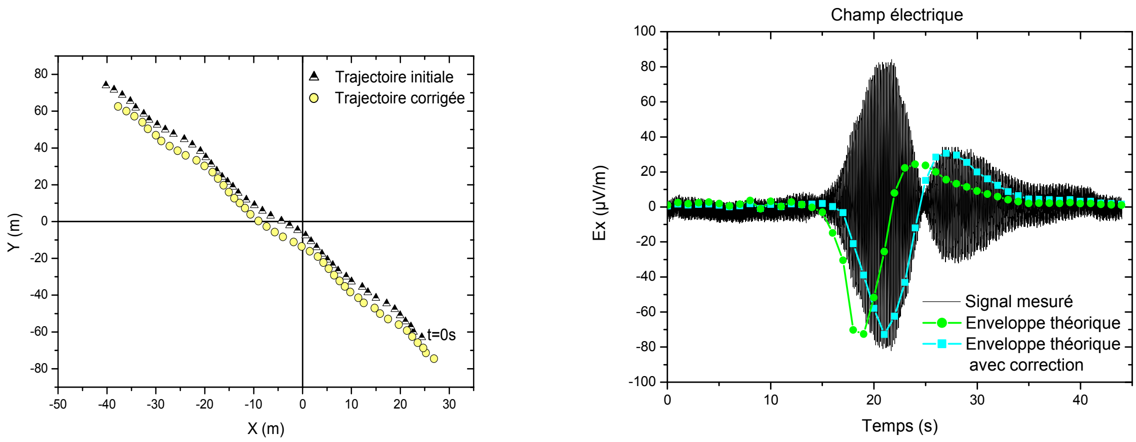


Figure 1- *A gauche :* Trajectoire du bateau par rapport au SIGMA situé en (0,0). *A droite :* Comparaison des résultats théoriques et expérimentaux obtenus pour la composante x du champ électrique avec et sans correction de trajectoire.

## 5. CONCLUSION

Dans ce papier, nous avons proposé, dans un premier temps, une méthode théorique permettant de calculer, sans approximation, les champs électromagnétiques de très basses fréquences dans un milieu conducteur à couches planaires multiples dans le cas où l'excitation est une source dipolaire oscillante située à l'intérieur des couches. Les résultats obtenus par cette méthode sont en accord avec d'autres modèles existants. Nous avons ensuite décrit la campagne de mesures qui nous a permis d'obtenir des résultats expérimentaux comparables à la théorie. Les mesures ont confirmé la validité du modèle théorique.

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***Commission D: Electronics and Photonics***

**INTEGRAL FUNCTION METHOD FOR THE ANALYSIS OF HARMONIC DISTORTION OF SOI MOSFETS**

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## 1. INTRODUCTION

It is well known that the DC drain current of MOSFETs exhibits a highly nonlinear characteristic as a function of the applied voltages. At high frequency of operation, the effect of (nonlinear) capacitances should no more be negligible and affect the distortions (HD). Nevertheless, previous work showed that HD is dominated by the *I-V* characteristic up to radio frequencies (RF) [1]. The most commonly used techniques to characterize the nonlinear behaviour of MOSFETs are based on the Fourier series expansion of this characteristic [2]. More recently, an integral function method (IFM) [3] was proposed with the advantage to be less sensitive to the measurements noise than Fourier series analysis. Unfortunately, both techniques neglect the non-linearities associated with memory and high-frequency effects. Other techniques based on RF measurements require more specific equipment as a sampling oscilloscope or a Large-Signal Network Analyser (LSNA) [4]. This last technique offers full characterization as we get both phase and amplitude of the signal, at the price of an expensive and non-widespread experimental set-up.

As Silicon-on-Insulator (SOI) is one of the most promising low cost technologies for integrated low-voltage, low-power circuits operating at microwaves [5], we focus on the distortion characterization of SOI nMOSFETs. The devices under test (DuT) that is considered through this analysis is a Fully-Depleted (FD) SOI nMOS composed of 12 fingers connected in parallel 6.6 µm width and 0.25 µm length each, built at CEA-LETI, France, following a 0.25 µm process. The extracted threshold voltage is 0.41 V. All measurements were performed on-wafer in common source configuration.

Up to now, IFM has been applied to evaluate the distortion associated with the gate transconductance $g_m$ *or* output conductance $g_d$ separately. In real circuit however, both of these contributions act together. The purpose of this work is to investigate the use of the IFM for the characterization of the nonlinearities associated with the drain current of a MOSFET that is controlled by both the drain and gate voltages, and that takes the influence of the load impedance $Z_L$ into account. Furthermore, we give an evaluation of the limit of the method in frequency computed by the Volterra approach. The validity of the methods is verified through LSNA measurements.

## 2. HD CHARACTERIZATION BASED ON DC MEASUREMENTS

The drain current of our DuT was measured in function of the gate and drain voltages with an HP4145 meter. The nonlinearities associated with the $I_D(V_G)$ relationship were first investigated. The THD of our DuT in saturation is plotted in function of the gate bias in Fig. 1. For gate voltages between 0.2 V and 0.8 V both IFM and LSNA methods give THD results in agreement. The THD minimum depends on the drain bias condition, however IFM shows minima at higher $V_G$ corresponding actually to the inflection point of the $I_D(V_G)$ curve, i.e. the maximum of the $g_m$, whereas the LSNA extractions take the whole behaviour of the MOSFET into account ($g_d$ and capacitances included). Therefore, the gate voltage at which the THD reaches its minimum extracted from RF measurements corresponds to the maximum gain of the transistor.

The differences between the results extracted by the two methods lies in the fact that only the nonlinear behaviour of the $g_m$ is taken into account for the IFM, while LSNA measures the device global distortion [6]. So we have to consider the $I_D(V_G, V_D)$ relationship instead of $I_D(V_G)$ and $I_D(V_G)$ separately. This two-dimensional problem may be reduced to a one-dimensional one by adopting appropriated boundary conditions. Indeed, for the common-source (CS) MOSFET in Fig. 2, the following relationship is always verified:

$$(V_{dd} - V_D)Y_L = I_D(V_D, V_G) \qquad (1)$$

Solving this equation for each $V_G$, we determine both the drain voltage in function of the gate voltage and the large-signal current characteristic that include the effect of the load $I(V_G,Z_L)$ (Fig. 3). The distortion may then be computed by IFM or Fourier series expansion on that characteristic. As we deal with measurements, IFM will be used but good agreements with Fourier expansion was obtained. Fig. 4 compares the $HD_2$ of a 50 Ω loaded FD transistor extracted by the previous mentioned methods. The supply voltage was set to $V_{dd}$ = 2 V and this value will be used throughout this paper. It shows good agreements between LSNA and IFM applied on $I(V_G,Z_L)$. For comparison purpose, we added on this plot the $HD_2$ obtained by Taylor series expansion, calculated by [7]:

$$HD_2 = \frac{V_o}{2}\left|\frac{g_{m2}}{g_{m1}} - A_{vDC}\frac{g_{d2}}{Y_L + g_{d1}}\right| \qquad (2)$$

where $g_{mi}$ ($g_{di}$) are the i$^{th}$ derivatives of *the* $I_D(V_G)$ ($I_D(V_D)$) and $A_{vDC}$ is the DC voltage gain. As expected, accurate characterization needs to take the effect of $g_d$ into account.

In order to analyse the impact of the load, $HD_3$ is plotted in Fig. 5 in function of $V_G$ for different values of $Z_L$. First, note that following (2), the transconductance is the dominant nonlinear source at low load impedance levels, while for high values of $Z_L$, the output conductance nonlinearity dominates. Second, the curves show several minima of $HD_3$. Fig. 6 shows the dependence of the minima on $V_G$ and $Z_L$. One is found for $V_G$ close to the threshold voltage (i.e., for $g_{m3}=0$), another one when the transistor is biased in triode. The third minima may be exploited for the design of high linear low-voltage circuits but tends to disappear at high input power level (Fig. 7).

## 3. IFM FREQUENCY LIMITATION

In order to set the limit frequency of the DC characterization of the distortion, we derived a simple analytical formulation of HD. This formulation is helpful to the circuit designer that wants to determine the evolution of HD in function of the load impedance and the frequency.

The method of nonlinear currents [7] is applied to the simple equivalent circuit of a MOSFET of Fig. 8. The drain current ID is assumed to be given by (3).

$$\begin{aligned} I_D &= g_{m1}V_G + g_{m2}V_G^2 + g_{m3}V_G^3 \\ &\quad + g_{d1}V_D + g_{d2}V_D^2 + g_{d3}V_D^3 \end{aligned} \qquad (3)$$

This means that the cross-derivatives are not taken into account. Furthermore, it is assumed that the only nonlinear elements are the gate transconductance and the output conductance. As discussed in [8], this model is well appropriated to analyse the distortion of a MOSFET in the saturation region. The fundamental limitation of the Volterra series lies on the fact that the nonlinear elements have to be described in term of the Taylor-series expansions of their current-voltage or charge-voltage characteristic. To ensure this representation to be valid, the nonlinearity should be weak enough and the excitation signals small enough.

The analysis shows that in practical cases, the dominant pole of $HD_2$ and $HD_3$ ($f_{pHD2}$ and $f_{pHD3}$, resp.) are limited by: $\left|f_{pHD_2}\right| \geq \frac{\left|f_{pAv}\right|}{3}$ and $\left|f_{pHD_3}\right| \geq \frac{\left|f_{pAv}\right|}{7}$ (4)

where $f_{pAv}$ is the pole of the voltage gain of the circuit. The $f_{pHD2}$ and $f_{pHD3}$ of our DuT loaded by $Z_L$ $=200\ \Omega$ are found to be about 10 GHz and 4.5 GHz, respectively. These relations give in function of $Z_L$ the limit of the validity range of DC methods to characterize HD.

## 4. CONCLUSIONS

This analysis shows that IFM is an easy way to characterize accurately HD of MOSFETs. The method has been extended in order to take the influence of the load impedance and both gate and drain controlling voltages into account. Results were in agreement with LSNA measurements. Also, the frequency limit of validity of the method was estimated through Volterra series analysis.

## 5. ACKNOWLEDGMENTS

The authors are grateful to Prof. D. Flandre from the Microelectronics Lab., UCL, and Prof. A. Cerderia, from Univ. Mexico, for helpful discussions. Thanks to D. Schreurs who performed the LSNA measurements. Financial support is provided by the FRIA.

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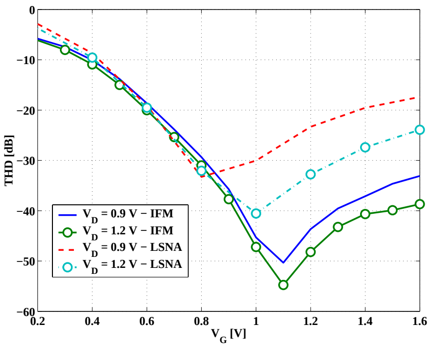


Fig. 1. THD measured with LSNA and calculated by IFM applied to $I_D(V_G)$ for various drain voltage; the AC input signal $V_o$ is 0.2 V.

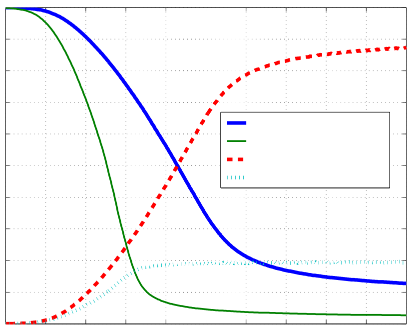

Fig. 3. DC characteristics of the CS FD device in function of the gate bias given for two load impedances.

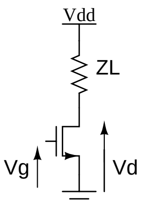


Fig. 2. MOSFET in common-source configuration

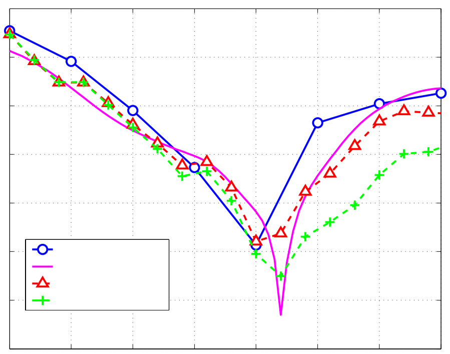

Fig. 4. Comparison of the techniques used to extract $HD_2$. A $V_o$=0.2 V sine-wave is applied to the gate, the drain voltage level follows a curve as in Fig. 3.

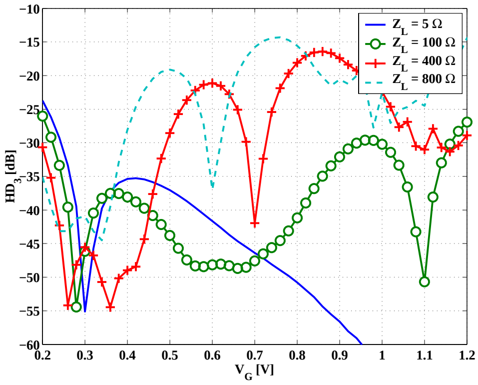


Fig. 5. $HD_3$ in function of the gate voltage for several load conditions extracted by IFM on $I(V_G,Z_L)$; $V_o$=0.2 V.

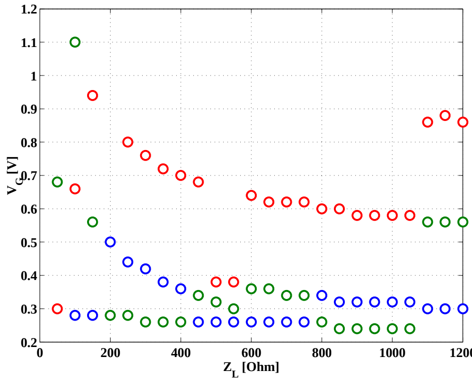


Fig. 6. Locus of the minima of $HD_3$ in the $(V_G,Z_L)$ plane extracted by IFM on $I(V_G,Z_L)$; $V_o$=0.2 V.

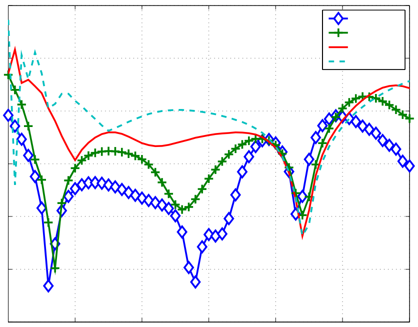

Fig. 7. $HD_3$ in function of the gate voltage for several input voltages $V_o$ extracted by IFM on $I(V_G,Z_L)$; $Z_L$ =200 Ω.

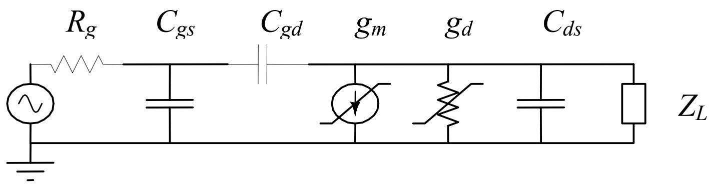


Fig. 8. MOS model used for the Volterra analysis.

# FERROMAGNETIC NANOWIRES FOR MICROWAVE FILTERS AND NONRECIPROCAL DEVICES

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## 1. INTRODUCTION

Ferromagnetic nanowires demonstrated a considerable potential for applications in microwave devices [1]. Numerous advantages make them strongly attractive as compared to classical materials. Devices using ferromagnetic nanowires are very compact due to their ferrite behaviour without applying any biasing field. Most ferrites are not suitable for applications at several Giga-Hertz because of their low resonance frequencies. Magnetic nanowires offer a good alternative to overcome this limitation: ferromagnetic resonance frequencies up to 40 GHz are obtained with Cobalt nanowires electrodeposited in special chemical conditions.
In this paper, we will first present an electromagnetic characterisation of a composite substrate based on magnetic nanowires, then a Magnetic PBG (Photonic Band Gap) material for filtering applications. Finally, a topology for the design of a microwave circulator based on the composite will be presented.

## 2. DIELECTRIC AND MAGNETIC PROPERTIES

We present in this section a theoretical investigation and an experimental validation of the dielectric constant of the magnetic nanowired composite. In addition, a volumetric model giving the magnetic permeability of the composite from its value for individual wires will be illustrated.
The magnetic nanowired composite is presented in Fig. 1a. It consists of a nanoporous polymer membrane. Pores are filled by a metal (ferromagnetic or not) via electrodeposition. In order to predict without ambiguity the dielectric constant of this composite, we study the electrical properties of non magnetic conductive nanowires. In this case, the magnetic permeability is equal to unity, and the product $\varepsilon_r \times \mu_r$ extracted from measurements is reduced to $\varepsilon_r$. Figures 1b and 1c give the simulations results of a microstrip line lying on Cu-nanowired substrate, obtained using IE3D (Method Of Moment-based electromagnetic simulator commercialized by Zeland Softwares). Figure 1b shows a considerable increase of the effective permittivity in the whole frequency band, when the nanowires height is increased. This effect is simply explained by a high capacitance created by the air gaps between the nanowires and the metallic strip (Fig. 1a). The evolution of the effective permittivity with the porosity is given in Fig. 1c, which shows a significant increase. This behavior is understood by an amplification of the air gap capacitive effect. Indeed, when we increase the porosity the individual capacitances being parallel to each other, are added resulting in a higher total capacitance. From the results of Fig. 1b and 1c we conclude that the conducting property of the magnetic nanowires induces an increase in the composite equivalent permittivity. The experimental validation of this conclusion is given by Fig. 1d, which gives the real part of the effective permittivity extracted from measurements for a polymer membrane without nanowires (--) and a membrane with Cu-nanowires (-). The asymptotic value of the effective permittivity is increased by 2.1, which confirms the theoretical results of Fig. 1b and 1c. This effect should be necessarily taken into account in the design of microwave devices based on ferromagnetic nanowires.
The estimation of the equivalent magnetic permeability of the composite is based on a combination of the classical magnetic susceptibility tensor components for one wire, with the volumetric proportion of nanowires. Figures 2a and 2b give for porosity = 6 %, the real and imaginary parts of respectively the composite equivalent permeability and the wire permeability. This model supposes a uniform distribution of the electromagnetic fields in the polymer and the wires. For low porosities (< 8 %) this approximation is rigorously satisfied.

The electromagnetic properties of the magnetic nanowired composite that have been presented above are essential for the design techniques, which will be described in the following sections.

## 3. MAGNETIC PBG FILTER

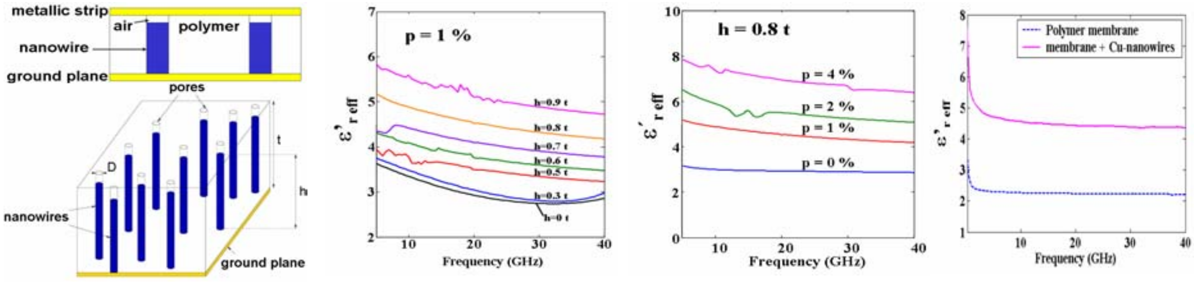
Classical PBG materials consist of a periodic medium, which exhibits stopband properties. This behaviour is only due to the periodicity of the physical characteristics such as the dielectric permittivity or the conductivity. We propose here to investigate the PBG properties when the periodic parameter is the magnetic permeability [2]. The strong dependence of the magnetic properties of the ferromagnetic nanowired composite on the applied static magnetic field gives additional tuning features to the classical PBG structure. In order to study the microwave behaviour of such kind of structure, an efficient variational formulation conbined with chain-matrix conversion (Var-ChM) has been established [3]. The model gives the S-matrix of a microstrip line lying on a 1-D MPBG (Magnetic PBG) as presented in Fig. 3a. For a given ferromagnetic material and taking into account the properties described in section II, this model allows to optimise the periodicity for maximal attenuation in the stopband. The MPBG sample, shown in Fig. 3b, is fabricated from a polymer membrane having a periodic arrangement of nanoporous rectangular areas filled with Permalloy nanowires. The porous areas are obtained using a patterned track-etching technique. Figure 4 shows the simulations and measurements results corresponding to the structure of Fig. 3b. The ferromagnetic resonance (FMR) frequency corresponding to Permalloy material is located around 15 GHz. The position of the MPBG peak located at 29 GHz is fixed by the geometrical parameters of the MPBG structure. The origin of each peak is well established if we look at the reflection coefficient around the corresponding frequencies. In the measurement and simulation curves, the FMR origin of the first peak is confirmed by -8 dB reflection around the FMR frequency: the electromagnetic power is absorbed by the magnetic material. The strong reflection around the second peak (-2 dB) proves its PBG origin: the electromagnetic power is not absorbed, it is reflected. The spurious losses on the transmission factors are mainly due to the high dielectric losses of the polycarbonate membrane and ohmic losses of the thin metallic strips. For the transmission and reflection curves the simulations are in very good agreement with the measurements.
Figure 5 shows the effect of a static magnetic field ($H_{dc}$) applied parallel to the nanowires of the MPBG. The position of the FMR absorption varies linearly with $H_{dc}$, while the MPBG peak remains almost at the same frequency. This effect can be exploited to tune the bandwidth of the filter. The Var-ChM model predicts very well the behavior of the MPBG under the DC magnetic field. Therefore, it can be used as a design tool for MPBG stopband filters since it is successfully validated by measurements [3].

## 4. CIRCULATOR

Nonreciprocal devices, such as circulators, working at the remanent state ($H_{dc}$ = 0 kOe) can be designed using this substrate. A theoretical model based on the Y-circulation theory and the properties described in section II has been established and validated. This model gives the S-matrix of a 3-port disc circulator lying on the magnetic nanowired substrate (Fig. 6). The circulation mechanism is strongly influenced by the remanent magnetisation and the matching network of the circulator disc to 50 Ω. In order to understand more deeply and to optimise the performances of the circulator, several designs are in progress. The preliminary results are very promising and interest some industrials.

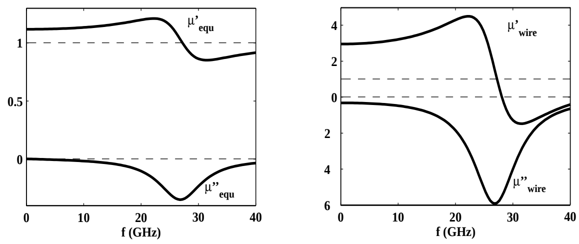
## 5. CONCLUSION

In conclusion, we have presented two possible applications for microwave devices based on the magnetic nanowired substrate. The MPBG effect offers new solutions to increase and tune the bandwidth or to create multiple zeros stopband filters. Very compact nonreciprocal devices can be designed using the magnetic nanowired substrate at the remanent state.




(a) (b) (c) (d)

Figure 1: (a) Topology of the magnetic nanowired substrate (b) simulated real part of the effective permittivity of Cu-nanowires versus the wire height for porosity = 1 %, and (c) versus the porosity for h = 0.8 t (d) real part of the effective permittivity extracted from measurements of polymer membrane without nanowires (dashed line) and a membrane (porosity = 4.4 %) filled with Cu-nanowires (solid line).




(a) (b)

Figure 2: (a) Equivalent magnetic permeability obtained with the volumetric model for p = 6 % (b) magnetic permeability of one wire.

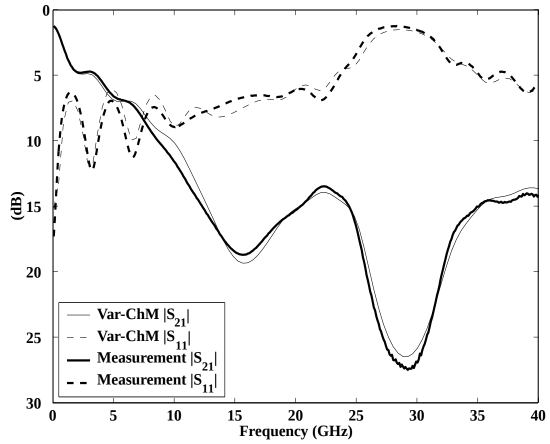
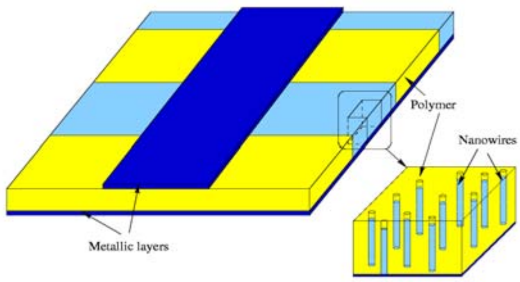



Figure 4: Measured (thick lines) and simulated (thin lines) transmission (-) and reflection (--) coefficients of MPBG using variational-chain matrix model.




(a

(b

Figure 3: (a) Schematic 3D view of microstrip line using the magnetic PBG as the substrate (b) Top-view picture of the MPBG sample.

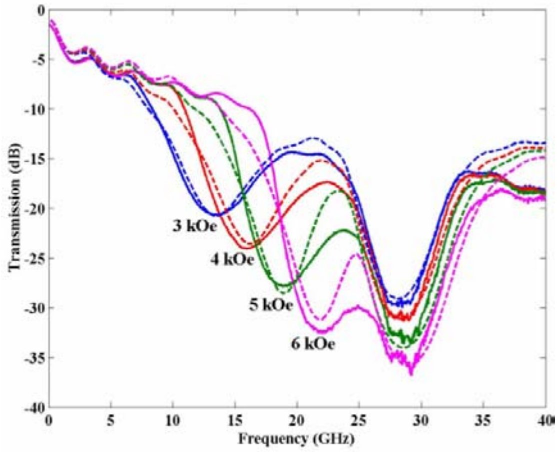



Figure 5: Measured (-) and simulated (--) microwave transmission at different values of the static magnetic field applied ($H_{dc}$) parallel to the nanowires of the MPBG.

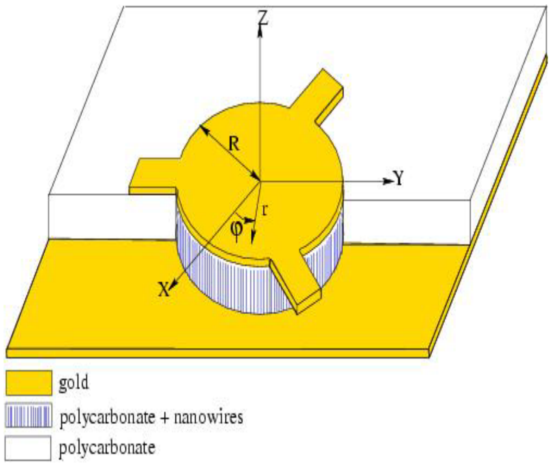



Figure 6: Schematic View of a microstrip 3-ports circulator on the magnetic nanowired substrate.

## 6. ACKNOWLEDGEMENTS

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**EMBEDDED THIN-FILM HEMTS IN MULTI-CHIP MODULES**

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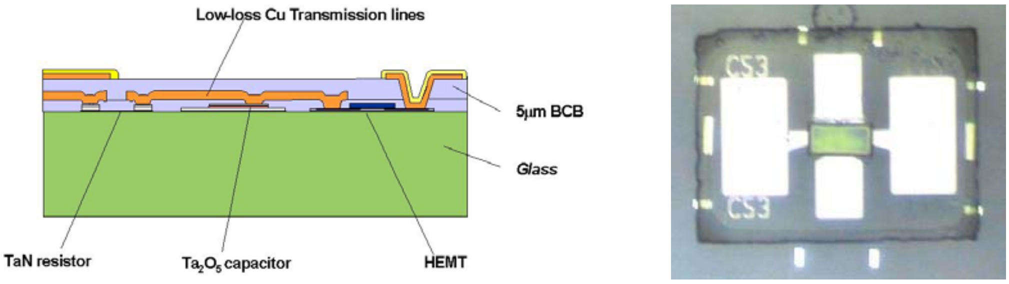
## 1. INTRODUCTION

Earlier developments have shown that MCM-D technology can be a viable candidate for the integration and interconnection of high quality microwave and RF applications [1]-[2]. This MCM-D integration of RF (sub-)systems leads to a new implementation paradigm called "System-in-a-Package". With this implementation approach, several RF components – each implemented in the most suitable IC technology – can be assembled in a relatively simple and economical way.
There are various possibilities to integrate active devices with MCM-D (passive) technology. First, IMEC started with wirebonding GaAs HEMTs on top of a MCM-D carrier substrate. As long wirebonds introduce large parasitic inductances, this method leads to rather poor RF performance. Shifting to flip-chip technology with gold bumps meant shorter interconnect lengths and thus less parasitics [3]. In this paper, a technique for the semi-monolithic integration of an embedded thin-film Ge-based MHEMT in MCM-D, is discussed. The very thin (< 3 μm) active device is embedded in the MCM-D substrate, together with the passives. In this technique, we use the selective substrate removal method as described in [4].

## 2. MCM-D TECHNOLOGY

Thin-film MCM-Ds are fabricated by a sequential deposition of conductors, typically Cu or Al, and dielectric layers, typically polyimide or BCB, on a substrate base made of ceramic, silicon or metal.




(a) (b)

Figure 1: (a) Schematic cross-section of MCM-D structure with integrated passive and active components; (b) A top view optical micrograph of a thin-film MHEMT after substrate removal.

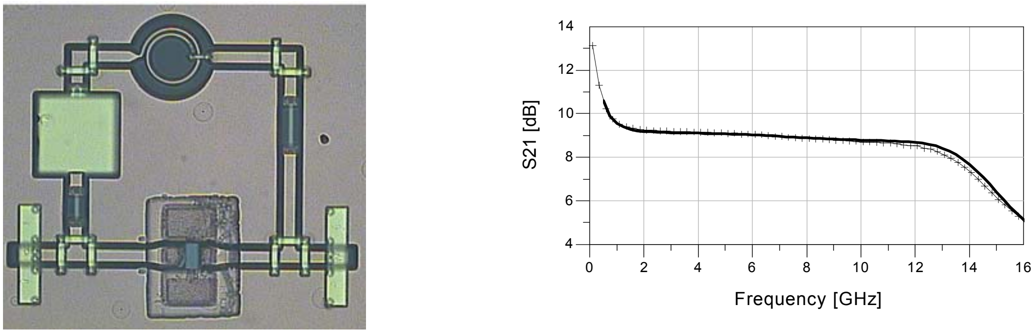
The IMEC MCM-D technology (figure 1 (a)) consists of alternating thin layers of photosensitive BCB and low loss Cu metallizations deposited on a borosilicate glass carrier substrate. The different metal layers are connected through via holes in the BCB dielectric. Due to the photosensitive property of BCB, the vias are immediately formed after the developing step.
The resistors indicated in figure 1 (a) are realized on the lower level, immediately on the carrier substrate [5]. The resistor material is TaN, with typical resistance values of 10-100 Ω per square and temperature coefficients of less than 20-150 ppm per centigrade.
Integrated capacitors may be realized in different ways [4]. The most common type is the classical MIM (Metal-Insulator-Metal) capacitor. The insulating dielectric may be BCB (capacitance/area ratio ≈ 5.5 pF/mm$^2$) or anodized tantalum (capacitance/area ratio ≈ 720 pF/mm$^2$) for large values. For the smallest values (< 100 fF), interdigital realizations may be used.
Spiral inductors in MCM-D on a low loss alumina or glass carrier substrate can exhibit very high quality factors at a low overall cost [5]. Quality factors above 100 @ 10 GHz have been realized.

## 3. INTEGRATED ACTIVE DEVICES

IMEC's MHEMT layers are grown by MBE [6]. The gate length of the devices is 0.2 μm and the gate width is 2 × 50 μm. In the case of MHEMTs on cheap Ge substrates, the conductive nature of the substrate makes such devices unsuitable for microwave applications. The substrate, having served its purpose as a platform for growth, is removed and the MHEMT device is transferred to the glass carrier substrate [4]. The thin-film MHEMTs exhibit excellent DC (gm = 700 mS/mm) as well as RF (fT = 80 GHz; fmax = 120 GHz) performance. Prior to the removal of the Ge substrate and after deposition of resistor and capacitor layers on the glass carrier substrate, the Ge wafer is diced. A chip containing a single MHEMT is embedded in a spin coated BCB glue layer using a device bonder in thermocompression mode. Chips as small as 400 × 500 μm2 can be positioned with 3 μm post-bonding accuracy.
Substrate removal and opening of the contact pads of the thin-film MHEMT are described in [7]. A BCB layer of about 5 μm thick is then spin coated in order to planarize the structure. In order to circumvent problems in the following processing steps, the total thickness of the devices has to be kept below 5 μm. This requirement is easily maintained due to the removal of the substrate and buffer layers. Via holes through the BCB dielectric allow to contact the thin-film transistor. A top view optical micrograph of a thin-film MHEMT on glass is shown in figure 1 (b).

## 4. CIRCUITS

To demonstrate the abilities of this combination of passive MCM-D technology and thin-film MHEMTs based on Ge, two integrated circuits have been designed and realized: a feedback amplifier and a C-band oscillator. The circuits have been designed using a measurement-based small-signal equivalent model of the thin-film MHEMT and IMEC's integrated passives design library containing parameterized scalable models [8]. The component library is integrated in ADS CAE environment. This allows easy and accurate co-design of the active and passive devices.




(a) (b)

Figure 3: (a) Optical micrograph of the feedback amplifier on glass (CPW lay-out); (b) Gain ($S_{21}$) of the amplifier, measurement (solid line), simulation (crosses).

A broadband feedback amplifier has been chosen since these amplifiers can be designed robust enough to allow small changes in the technology. Figure 3 (a) shows a photograph of the circuit. Figure 3 (b) shows the gain of the amplifier, 9 dB over a frequency band from 1 to 13 GHz. A very good agreement between simulation and measurement has been achieved.
Figure 4 shows a picture of the C-band oscillator. The oscillation frequency is centered around 7.6 GHz. When the embedded thin-film MHEMT is illuminated (1550 nm laser; 10 mW), the oscillation frequency can be tuned (optically) over more than 50 MHz. Light is absorbed in the $In_{.53}Ga_{.47}As$ channel layer of the transistor [9]-[11].

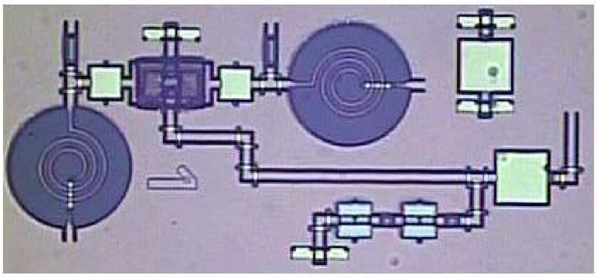


Figure 4: Optical micrograph of the C-band oscillator (CPW lay-out).

## 5. CONCLUSION

Thin-film multilayer MCM-D technology using the SiP concept is presented as a viable approach for the integration of high performance passives and embedded active devices. The active devices, i.e., thin-film MHEMTs based on Ge substrates, are embedded in the low-cost MCM-D substrate together with the passives. A C-band oscillator and a feedback amplifier integrated circuit have been fabricated to demonstrate the abilities of this combination of MCM-D and III-V-on-Ge technology.

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## SHORT ABSTRACTS (POSTERS)

***Commission B: Fields and Waves***

# CYLINDRICAL AND SPHERICAL OBSTACLES IN EPICS-GO

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## 1. ABSTRACT

Up till now, the UTD based software prediction tool EPICS-GO (Enhanced Propagation for Indoor Communications Systems), developed at the ESAT-TELEMIC division of the K.U. Leuven [1], like many other three-dimensional software [2], could only determine the signal in an environment that can be decomposed into (ir)regular hexahedral obstacles (obstacles with 6 side planes). In this paper, we want to give an overview of the geometrical problems involved in the determination of intermediate (penetration, reflection and diffraction) points of cylindrical [3] and spherical obstacles.

At first those intermediate points are computed in the geometrical part of the program. In the case of cylindrical obstacles all data is projected in a plane perpendicular to the rotation axis of the cylinder. The intermediate 2-dimensional point (reflection or diffraction points on a circle) is then computed and then transformed back by adding the relative height along the axis. The case of the spherical obstacles is even simpler because the exact points lye in the plane of transmitter, receiver and centre of the sphere. For the 2-dimensional calculations, efficient computational routines have been written and tested for this purpose.

The last step in the determination of the signal in presence of cylindrical and/or spherical obstacles, contains the computation of the electromagnetic fields [4]. For both cases, this is done by adding the loss due to penetration, reflection and/or diffraction as well as loss due to creeping waves travelling along the surface of the cylinder/sphere [5] to the free space loss.

The incorporation of curved surfaces will increase the accuracy of the predictions and remove the artefacts due to the polyhedral modelling of the environment.

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# CASE STUDY MODELLING OF GENERAL 3D DIELECTRIC VOLUMES IN HOMOGENEOUS MEDIA

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## 1. INTRODUCTION

The goal of my Ph.D. is to realize an accurate quasi 3D-model of *cylindrical dielectrics in multilayered media*. Additionally, it is the intention to couple these dielectric volumes with horizontal and vertical metallic surfaces. In that way all kind of interesting applications like planar and 3D antennas, embedded antennas, dielectric resonators, Microwave and Millimeter wave Integrated Circuits (MMIC's), Micro-Electromechanical Systems (MEMS) and electronics packaging can be simulated.

The modeling method is based on a spectral Volume Integral Equation (V-IE) formulation of the Maxwell laws for the electric and magnetic field. It has the enormous advantage that it calculates solutions relatively fast, due to the fact that only a small amount of unknowns need to be solved. An additional advantage is that the concept of Green's functions is used to obtain the solution. This implies that the solution is valid for the entire space and not just in a confined part of space, as is the case in FDTD and FE methods. As a consequence, the radiation aspects of antennas are more naturally present in IE solutions than they are in solutions of other methods. [1], [2]

## 2. CASE STUDY

To get acquainted with specific modelling issues and implementation problems, a simplified but characteristic problem is defined and solved first. The V-IE-formulation for dielectric volumes has been worked out in full detail for the case of homogeneous surroundings. Subsequently, a matrix equation is obtained using the corresponding free space Green's functions in easy Mixed Potential form (MPIE) and the method of moments for discretisation.

This has resulted in the implementation of a full wave solver for generally shaped dielectric volumes, with variable dielectric properties, exited by imposed volumetric currents or incident plane waves. Comparison of the simulated far field scattering of an incident plane wave on a dielectric sphere with the corresponding solution in closed form [3] should validate the new software.

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# AN EFFICIENT FMM-PML-MPIE FORMALISM FOR 2D MICROSTRIPS

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Many full-wave electromagnetic field simulators for planar circuits rely on Method of Moment (MoM) based integral equation solvers. These schemes use so-called Green functions to store all the characteristics of the layered medium. Only the system's metallic conductors are discretized into N segments in order to determine their currents. The N x N linear system that arises from the MoM can be solved by classical LU-decomposition or by an iterative method. Often one uses a mixed-potential integral equation (MPIE) to model the field problem and this is also the approach we adopt. Unfortunately, the calculation of both MPIE kernels – i.e. the Green dyadic GA and the Green function GV describing the vector and the scalar potentials – unavoidably calls for the time-consuming evaluation of Sommerfeld-type integrals. Building the dense moment matrix is also computationally expensive. Recently, several methods have been proposed to reduce the computational complexity associated with the iterative solution of the linear systems, e.g. the adaptive integral method (AIM), the thin-stratified medium fast-multipole algorithm (TSM-FMA) and the fast inhomogeneous plane wave algorithm (FIPWA).

A very efficient and elegant way to calculate the Green function, based on the use of Perfectly Matched Layers (PML), has been proposed in [1]. For a layered structure that is closed by a perfect electric conductor (PEC) plane at the top and bottom of the structure, the Sommerfeld integrals that arise in the Green function calculations can be expressed as a series of surface waves. By using a PML that is covered by a PEC plane, one can also close open layered media while maintaining the open character of the structure. In [1] this approach was used to obtain an analytic and easy to determine series representation for the Green functions of open layered media.

In this contribution, we combine this PML-MoM formalism with a Fast Multipole Method (FMM) in order to model scattering from two-dimensional (2D) microstrips. The proposed new FMM technique allows us to store the linear system with low memory requirements and to solve it fast and efficiently. Classical iterative methods have a computational complexity of order $O(N^2)$ to perform one matrix-vector multiplication. Since the terms in the series representation of the Green functions can be factorized exactly into a separate source and observation contribution, we achieve an operation count of $O(N)$ by using a recursive technique. Although the applications for 2D scattering at microstrip lines are limited, the development of this technique can be seen as the first step towards a PML-MLFMA technique for general 3D objects embedded in layered media, which is currently being investigated.

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***Commission C: Radio-Communication systems & signal processing***

# LOCALISATION USING CELL-ID TRILATERATION

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## 1. INTRODUCTION

Localisation is an essential part for Location Based Services (LBS), besides a communication and a database component. Possible applications can be found in the field of information (e.g. navigation), trigger (e.g. position dependent billing), track and trace (e.g. fleet management) and assistance (e.g. 112 emergency calls) services. In most of the current LBS applications localization happens by means of GPS. In dense urban areas, where navigational aid is most requested, contact with enough satellites can be lost due to the canyon effect. This paper describes briefly the research for a method for localization using cellular networks, which are heavily deployed in the same urban areas.
This research fits in a broader context of the fusing of different technologies (cellular, satellite and motion sensors) for localization purposes in the field of navigation.

## 2. CELLULAR LOCALIZATION

There exist many different techniques in the USA, due to some FCC 911-regulation, to get a position estimate along with the caller's identity. In Europe, where the deployment of cellular localization is expected to be market driven, the precision is limited to the cell where the user is calling from, ranging from some 100m (urban) till 32 km wide (countryside). The Cell-ID trilateration technique will use at least 3 Base Transceiver Stations (BTS) and the reception power of their broadcasted signals, to get a position estimate. Comparing with previous research and measurements, it is possible to differentiate between 2 locations that are 200m away from each other. This is still insufficient for the upcoming generation of LBS applications.

## 3. ADVANCED ERROR MODELS

By understanding the different error mechanisms better, the resolution can be raised – at the cost of more expensive equipment. Possible techniques are: position tracking, directive measurements, measurement post-processing, better cell-characterisation (models specifically for localisation – not the communication models)… Short term fading and possible extension of the techniques to UMTS will be specially considered.

# INNOVATIEVE ASPECTEN VOOR DYNAMISCHE DIENSTVERLENING IN NAVIGATIE

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Dit onderzoeksproject situeert zich in de discipline van de nautische en continentale navigatie. Het beoogt een sterk verbeterde flexibiliteit voor de dienstverleningen naar de gebruikers toe. De informatie in databases die toegeleverd worden, door zowel publieke als privé organisaties, verstrekken hoofdzakelijk informatie over enerzijds cartografie en topologie, die aangewend worden bij de uiteindelijke navigatie en anderzijds extra informatiefaciliteiten van een bepaalde omgeving.
Er stelt zich echter een probleem bij het updaten van deze databases naar de gebruiker toe. De updates die door de verschillende organisaties verschaft worden, gebeuren op statische en discrete wijze. Alvorens een update tot voltooiing wordt gebracht, is er vaak nog een interactie van de gebruiker vereist. Een gevolg hiervan is dat de consistentie van de database door de verstrekkende organisaties niet kan worden gegarandeerd, daar er een grote verantwoordelijkheid op de gebruiker van het systeem wordt afgewend.
Daarenboven is er een groot team van hooggeschoolde datacollectors nodig voor het inwinnen van alle informatie. Een continue controle is vereist voor de consistentie en nauwkeurigheid te waarborgen en is er een grote afhankelijkheid van heel wat officiële instanties. Deze informatie moet voortdurend worden bijgewerkt, desondanks er op dit ogenblik geen uniforme standaard voor deze gegevens beschikbaar is.
In een eerste fase dient er een uniforme vorm voor de datavoorstelling gevonden te worden, zodat deze de huidige diensten kan omvatten en er gemakkelijk nieuwe diensten aan kunnen worden toegevoegd. Kortom: een uniformisering van de informatievoorstelling voor navigatie en mobiele dienstverlening.
In een tweede fase zal gezocht worden naar een efficiënte en betrouwbare manier om deze voorstelling up-to-date te houden. Hiertoe dient een lokale updating service ontwikkeld te worden, zodat alle informatie in het gebied waar de gebruiker zich bevindt, betrouwbaar is.
Als derde fase van het onderzoeksproject zullen nieuwe faciliteiten voor de gebruikers ontwikkeld worden. Dit omvat het concretiseren van Location Based Services (LBS) van de conceptie en ontwikkeling naar de praktijk toe.
Er moeten dus krachtige en flexibele datarepresentaties worden ontwikkeld, die samen met de huidig beschikbare technieken, een sterke bouwsteen vormen om verder op te bouwen. Men moet tevens rekening houden met de gebruiksvriendelijkheid van de aangeboden diensten en het gemak van de updating. Hiernaast dient een robuuste en efficiënte techniek gevonden te worden voor de eigenlijke updating van de data op bestaande communicatie netwerken. Tot slot dient een representatie op het gebouwde systeem ontwikkeld te worden waarin men LBS kan gebruiken.

# GENERAL FRAMEWORK FOR DESCRIBING AND EVALUATING SUBOPTIMUM TECHNIQUES FOR RADAR STAP

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Radar Space-Time Adaptive Processing (STAP) is a technique designed for detecting slow-moving targets [1]. However, it is a well-known fact that optimum STAP detection implies a large computational cost, thereby preventing its use in operational settings.

As a result, many authors, e.g. [1-6], have proposed suboptimum methods that reduce the computational needs. First, we present a new framework that unifies all suboptimum methods we are aware of. Second, we present an original decomposition of the CSNR (Conditioned SNR) performance measure [7]. This decomposition can be used to understand the performance degradation that is specifically due to the use of a suboptimum method. It can thus be used to compare the performance of various suboptimum methods.

Those new ideas are applied to some well-known techniques, which are then compared to each other.

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# MOBILITY MANAGEMENT ON ALL-IP NETWORKS USING MULTIPROTOCOL LABEL SWITCHING

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## 1. INTRODUCTION

Today, two major trends in networking can be distinguished. First, the users' mobility is a key issue. Second, more and more real-time multimedia applications are being deployed on the Internet. Therefore, mobility management schemes, such as Mobile IPv6 [1], need to be adapted and integrated into architectures offering Quality of Service (QoS), e.g. DiffServ [2].
This paper describes PhD research of the K.U.Leuven on a QoS-enabled mobility management scheme, using MultiProtocol Label Switching (MPLS) [3] in the access networks for this purpose.

## 2. MOBILE MPLS CONNECTIONS IN THE ACCESS NETWORKS

The authors strongly believe that MPLS can provide QoS-enabled mobility management. Mobile Nodes (MNs) have to register themselves at one or more Mobility Anchor Points (MAPs), serving as local Home Agents. An IP address of the MAP, the so called Regional Care-of Address, is registered at the MN's Home Agent. When a Corresponding Node has data packets for the MN, these packets are sent over the IPv6 Internet using the standardized IPv6 type 2 Routing Header [1] until they reach the MAP. There, the Routing Header is mapped to an MPLS label and the packet is switched on the MPLS domain between MAP and MN. Actually, packets follow a kind of mobile MPLS connection from MAP to MN.
When the MN changes its point of attachment, this mobile connection can be rerouted easily. Only local changes and message flows are needed, which reduces hand-over latency and packet loss compared to [1]. A post hand-off optimization phase eases the next hand-off.
Integrating DiffServ in MPLS is very straightforward. This enables offering QoS on the mobile connections. Integrating multicasting in the hand-offs further reduces packet loss and hand-over latency.

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# ETUDE DES PERFORMANCES DES RESEAUX WLAN EN MILIEU INDOOR

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## 1. PROPAGATION INDOOR

Le déploiement de réseaux sans fils en indoor nécessite une connaissance approfondie du canal de transmission. L'effet le plus marquant de la transmission en milieu indoor est le phénomène de multi-trajets ou «multipath» bien connu. Au cours de ce travail, nous avons étudié la transmission en milieu indoor dans le cas de deux antennes fixes avec un mouvement de personnes entre les antennes. Ce type d'application est particulièrement adapté aux situations tels que des halls de gare ou d'aéroport. Afin d'établir la distribution statistique adéquate, des mesures de propagation ont été réalisées dans les couloirs de l'Université Libre de Bruxelles à 2 et 5 GHz en présence de nombreux étudiants.

## 2. TRAITEMENT STATISTIQUE DU SIGNAL ET APPLICATION

Nous avons trouvé que la meilleure caractérisation statistique du signal était obtenue pour une densité de probabilité de Rayleigh. Afin de mieux caractériser le signal, nous avons entrepris de mesurer la statistique du deuxième ordre. Pour ce faire, nous avons utilisé le variogramme. Ce dernier est défini par

$$\gamma(h) = \frac{1}{2N}\sum_{t=0}^{N-1}[(Z(t) - Z(t+h))^2]$$

où $Z(t)$ est la variable aléatoire et N est le nombre de doublets de points considérés espacés par un écart temporel h. A l'aide du variogramme et d'une approximation au sens des moindres carrés, nous avons pu mettre en évidence un temps de décorrélation du signal presque constant pour les différentes expériences aux alentours de 500ms. Afin de quantifier l'intérêt de ce nouveau modèle, nous avons utilisé le simulateur de réseaux NS2. Nous avons pu montrer que le débit obtenu pour différentes expériences est de manière significative inférieur au débit obtenu par un modèle classique de type gaussien( signal moyen calculé par la formule de path loss + une variable aléatoire gaussienne).

## 3. CONCLUSIONS

Grâce à ces différents résultats, nous avons pu montrer l'intérêt de développer un modèle plus précis pour l'étude du canal de transmission indoor en milieu très peuplé. L'influence des personnes nous a conduit à trouver une distribution de Rayleigh avec un temps de décorrélation presque constant. Une amélioration du simulateur de réseaux NS2 a été effectuée et a montré des débits inférieurs à ceux prédits par le canal gaussien prouvant ainsi l'intérêt d'un modèle plus précis.

**DESIGN OF A FULLY INTEGRATED DISTRIBUTED AMPLIFIER IN SOI TECHNOLOGY**

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## 1. INTRODUCTION

Broadband amplifiers find many applications in instrumentation, electronic warfare and broadband optical communications. Distributed amplifiers (DAs) were widely used for realizing broadband amplifiers in GaAs hybrid and MMIC technologies [1]. The last decade has witnessed the relentless increase in digital and analog integration of Si VLSI technology, with SOI CMOS as the candidate of choice for high performance digital applications. Opportunities for the nanometer SOI CMOS technology are the high gain and cut-off frequency, as well as the low RF noise and the low voltage and power consumption.

## 2. DISTRIBUTED AMPLIFICATION

Distributed amplification (Fig. 1) overcomes the gain bandwidth limitation. This method absorbs the FET's input/ output capacitance as part of the lumped elements of an "artificial transmission line", formed with the series inductance that connects adjacent drains and gates. The gain is a function of the drain/gate transmission line parameters, and is given by the following equation:

$$G_p = \frac{g_m^2 Z_0^d Z_0^g}{4(1-X_c^2)(1+X_g^2)} e^{-N(\alpha_g+\alpha_d)} \times \frac{\sinh^2\left[N(\alpha_d-\alpha_g)/2\right]}{\sinh^2\left[(\alpha_d-\alpha_g)/2\right]}$$

Where $g_m$ is the transistor transconductance, $Z_0^g$ and $Z_0^d$ are the characteristic impedances of the gate and the drain line, respectively. N is the number of the transistors. $\alpha_g$ et $\alpha_d$ are the attenuation factor of the gate and the drain line, respectively.
In the following, we have designed a 4 section DA in SOI CMOS process using a 0.12 floating body transistor and TFMS (Thing Film Micro Strip) lines as passive elements. The DA achieves a simulated gain around 5 dB over a bandwidth covering 0.4 – 25 GHz. The chip occupies an area of approximately 1500 x 500 □m$^2$.

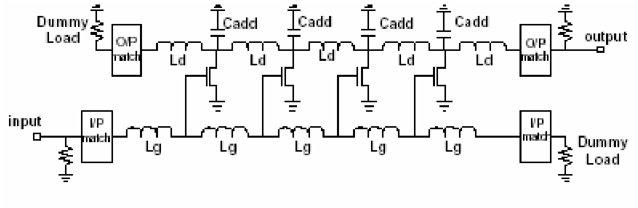


Figure 1. Distributed Amplifier Schematic

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## TRACKING OF OBJECTS IN VIDEO STREAMS USING POINTS OF INTEREST

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We are interested in the tracking of multiple objects in a video sequence in complex situations, where the objects can have different sizes, be of similar appearances, occlude each other or move with different velocities. Most current tracking systems do not deal well with these complex situations, in fact because the tracking is based on the difference between each image and a model of the background. A clue that this is not the right approach is that the human eye can detect objects moving across a “random” background.

The key idea of our work is to use the points of interest to robustly track multiple objects in complex situations. What are the points of interest? Sometimes called “corners” in the literature, these points are defined as any point in the image for which the signal changes two-dimensionally. Examples are “L-corners”, “T-junctions”, “Y-junctions”, black dots on white backgrounds and any location where the texture changes significantly. They are local features that is they are computed in a local neighborhood and are thus quite robust to partial occultation of the scene. Several detectors of points of interest have been proposed. The most popular is perhaps Harris’ detector. Points of interest detectors can be classified in three main categories [2]: contour-based, model-based and signal-based. Their classical applications are image matching, 3D modeling and object learning and recognition.

In our system, an object is defined via a set of points of interest and each point is characterized by the appearance of its neighborhood. The use of a set of interest points will allow us to track an object through partial occlusion as one or more points hopefully remain visible. In addition, we can also exploit potential relationships existing between the different points. For example, we can exploit the known spatial relationship between points located between the various parts of a football player: head, shirt, short, etc …

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# OPTIMAL ELEMENT SPACING IN LINEAR ARRAYS FOR SATELLITE COMMUNICATION

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## 1. DESIGN GOAL

As power is not abundantly available on satellites, the power consumed by onboard antennas should be as low as possible. This can be facilitated by using high-gain directive antennas on the satellite as well as at the ground station on earth, so that as few power as possible is wasted when sending and the power is picked up as efficiently as possible when receiving.
The free space pathloss and the atmospheric absorption are at worst when the satellite is low above the horizon, the link budget should thus be calculated for this case. As a rule of thumb satellite communication is started as soon as the elevation of the satellite is higher than 10°. For groundstation antennas, one thus needs to select the array that gives the highest gain at a scan angle of $\varphi$=10°.

## 2. THEORETICAL BACKGROUND

Suppose that identical currents are applied to the N antenna elements of an array. The maximum electric field strength of this array will be N times the electric field strength of one antenna element. But the power generated by these currents, depends on the inter element spacing as varying the element spacing will vary the electric fields at the element locations and hence vary the power that is consumed by the array. [R. C. Hansen, "Microwave Scanning Antennas vol. 2", Academic Press, 1966]
Research revealed that to obtain this maximal gain, the inter element spacing should be so that the first grating lobe lies just behind the edge of the visible interval. As a consequence, the optimal inter element spacing depends on the tapering of the array, the radiation pattern of its radiating elements and the number of elements in the array.
Choosing the number of elements in the array is a trade-off between cost and performance.

## 3. AN EXAMPLE

In the framework of the KULSAT project, a linear eigth elements receiving array of dipoles with a uniform tapering was designed. The concept that was developed for this optimization is only applicable to antenna elements with a rotation symmetric radiation pattern. For more general types of radiating elements an iterative optimization search is needed.
The concept was based on a far-field calculation of the power. In future work, this approach will be compared with a near-field approach and a measurement. The latter will reveal whether or not the theoretical supergaining is apparent in practice as well, despite the idealisations with respect to losses and mutual coupling.

**CONGESTIECONTROLE VOOR PAKKETGESCHAKELDE NETWERKEN**

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## 1. HET CONGESTIEPROBLEEM

De opmars van pakketgeschakelde netwerken lijkt niet te stoppen. Zowel het overweldigende succes van het Internet, als de overgang van het circuitgeschakelde GSM naar het pakketgeschakelde UMTS wijzen in die richting. Deze netwerken danken hun populariteit aan een mogelijk efficiënter gebruik van de beschikbare transmissiecapaciteit. Helaas is de kans op netwerkcongestie hier onlosmakelijk mee verbonden. Congestie definiëren we in deze context als de toestand waarin het nut van het netwerk afneemt als de belasting toeneemt.
Het congestiecontrolemechanisme van TCP (Transmission Control Protocol) [1] heeft het Internet tot nu toe grotendeels gevrijwaard van de meer ernstige vormen van congestie. De opkomst van toepassingen die TCP niet gebruiken, bedreigt echter de goede werking van het gedistribueerde TCP. Een ander probleem is dat deze toepassingen, die niet gevoelig zijn voor congestie, een onredelijk deel van de bandbreedte kunnen innemen.

## 2. ONZE OPLOSSING

Een congestieaanval door niet afgeleverde pakketten wordt algemeen beschouwd als de meest ernstige bedreiging voor de stabiliteit van het netwerk [2]. Ons algoritme zal dan ook vooral tegen deze vorm bescherming bieden. De basis van het algoritme ziet er als volgt uit: van elke trafiekstroom wordt in de laatste router van het netwerk het gegevensdebiet bepaald; dit debiet wordt dan vergeleken met het debiet bij de eerste router van het netwerk; op basis van die vergelijking regelt de eerste router het debiet van de respectieve stroom. Deze combinatie van terugkoppeling en ingangsdebietregeling zorgt ervoor dat een stroom die op een knelpunt in het netwerk beperkt wordt, al bij het binnenkomen in het netwerk wordt begrensd. De op die manier vrijgekomen bandbreedte kan nuttiger gebruikt worden door andere stromen.
Uit voorlopige simulaties blijkt dat dit algoritme in staat is stromen die niet gevoelig zijn voor congestie, terug te dringen, zodat het totale nuttige gegevensdebiet van het netwerk in congestiesituaties sterk toeneemt. Bovendien wordt door samenwerking met andere, reeds bestaande maatregelen een eerlijke verdeling van de beschikbare bandbreedte mogelijk.

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# NETWORK EMULATION: EXTENDING THE CLICK MODULAR ROUTER

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## 1. REAL-TIME APPLICATIONS: SIMULATION OR NETWORK EMULATION

The quality of real-time applications such as Voice over IP (VoIP) and video conferencing typically suffers from packet loss and high delays because of congested networks, high propagation delays on satellite links ... Developers of new multimedia applications for GPRS or UMTS need tools that enable them to evaluate the performance of their applications in a controllable, realistic and reproducible way.
Evaluation can be done with simulation programs, but this does not give much information about the end user experience. Evaluation is only based on a number of quantitative parameters that characterize the quality. Real-time emulation of network conditions is more useful for the evaluation of applications than simulation because the quality can be seen and heard directly. An emulation of the network conditions that mimics the behaviour of the real network as accurately as possible is therefore needed.

## 2. INTRODUCING A NEW AND BETTER NETWORK EMULATOR

We were unable to find any emulator that is capable of emulating realistic network conditions. Most commercial emulators are too expensive, and are also unable to dynamically emulate packet drop probabilities and delay because of their use of simple statistical models. Therefore we implemented a network emulator [1] in The Modular Click Router [2] language. This emulator is dynamical, flexible and extendible. The behaviour of the emulator is dynamic: time is divided in intervals. For each host that is attached to an interface, drop probability and delay from and to each interface can be set per interval.
The number of packets per second that can be handled, appears to be more than enough for normal multimedia applications. The accuracy of delay and drop rate that is imposed on the packets, is almost perfect. Delay can vary from one millisecond till a few seconds and drop rate can be obtained with an accuracy of one percent.
We are convinced that this implementation can be very useful to investigate whether certain technologies and conditions satisfy the specific requirements for real-time applications.

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**TRAFFIC MODELLING IN PACKET BASED NETWORKS**

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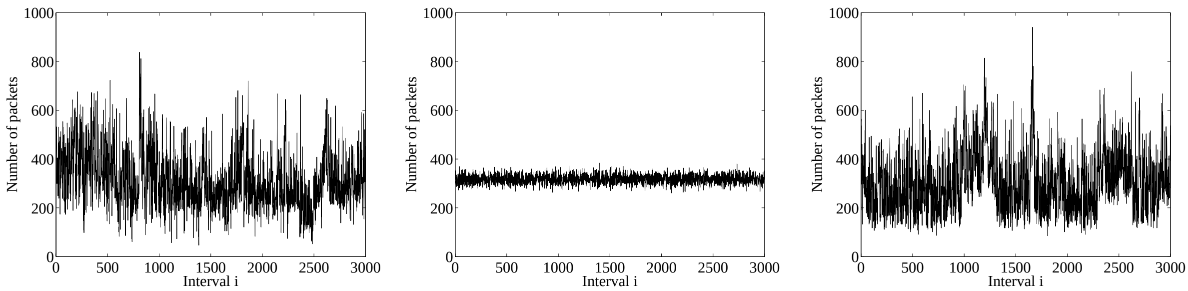
## 1. GENERAL OVERVIEW

Modern telecommunication networks are packet based, which means information is split into packets that travel independently from source to destination through the network. The aim of our research is to develop realistic models for aggregated packet streams, which consist of packets from different users and applications. These models are needed to predict the quality a network can offer. Especially real-time multimedia applications are very sensitive to quality measures like packet delay, delay variation and loss.
Today, the Poisson model is often used to model packet streams. However, this model is not able to capture the bursty behaviour of real traffic and will seriously underestimate the queuing delays and the losses. We developed a new discrete traffic model which is based on a hierarchical scheme of Bernoulli sources and which is able to fit variances and third order moments at different time scales. As indicated in Figure 1, our model closely resembles measured traffic. It also predicts the queuing behaviour of the measured traffic accurately.




(a) (b) (c)

Figure 1: Number of packets per interval of 1 millisecond:
(a) measured, (b) Poisson model, (c) our model.

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# ELECTRONICALLY CONFIGURED MCM-D INTEGRATED ANTENNA FOR 38 GHZ RADIO LINKS

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## 1. CONCEPT

The proposed antenna consists of 2 concentric subarrays of aperture coupled microstrip patches. By feeding the two subarrays with different amplitude and phase the beamwidth and sidelobe level of the antenna can be controlled. The application targeted is point-to-point radio links at 38 GHz.

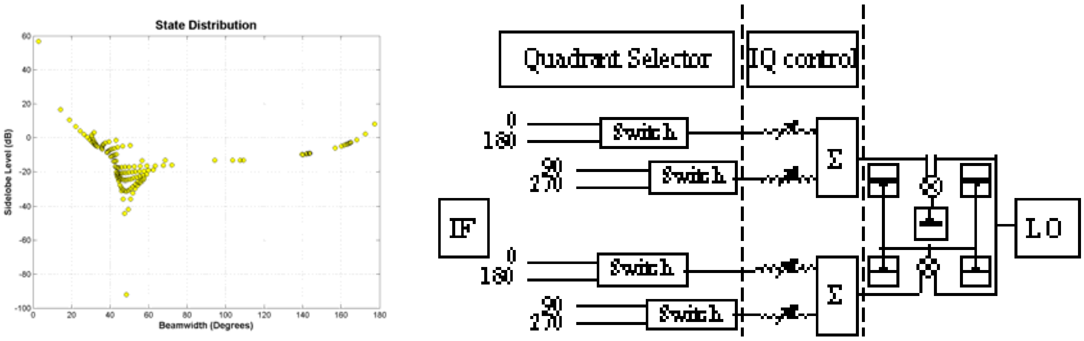


Figure 1: State distribution for 1 dB amplitude step and 22.5° phase step (left) and Proposed antenna architecture (right)

## 2. IMPLEMENTATION

The array will be implemented in the MCM-D technology available at Imec, where the patches are realised on the backside of a 700 µm glass substrate. The phase and amplitude control will be done at 2 GHz followed by an upconversion to 38 GHz. All subcircuits (attenuators, mixers, switches) are in chip form or in chip scale package so that the whole antenna will fit into an 8x16 $mm^2$ layout cell.
In the quadrant selector part of the circuit the coarse phase control is done. The IQ control part consisting of attenuators and a combiner then performs the fine control of the amplitude and phase. In this way we do a kind of QAM modulation where I(t) and Q(t) are the same signals 90 degrees out of phase.

**FAST MOVING HIGH BITRATE USERS AND TERMINALS**

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## 1. PROBLEEMSTELLING

Tal van multimedia-applicaties zijn tegenwoordig gemeengoed geworden in de vaste netwerken. Deze toepassingen vergen een hoge graad van Quality of Service (QoS) en worden doorgaans gekenmerkt door hoge bandbreedtevereisten. Dé uitdaging voor de toekomst is het onderzoeken hoe deze kwaliteit en bandbreedte ook kan worden geleverd in draadloze cellulaire netwerken. Organisaties zoals het IEEE en 3GPP leggen momenteel specificaties vast voor nieuwe draadloze technologieën die de honger naar bandbreedte van morgen moeten stillen (bv. UMTS en IEEE802.11g). Helaas blijven de snellere mobiele gebruikers, bv. in de wagen of trein, voorlopig nog steeds in de kou staan: hun beschikbare bandbreedte daalt naarmate ze zich sneller voortbewegen. In dichtbevolkte gebieden valt er bovendien een evolutie naar kleinere cellen te bemerken om aan de vraag naar hoge bandbreedte te kunnen blijven voldoen. Dit netwerkaspect heeft ingrijpende gevolgen voor snel bewegende gebruikers.

## 2. TOEKOMSTIGE DRAADLOZE BREEDBANDNETWERKEN

Opdat een service provider zijn breedband diensten met een voldoende kwaliteitsgraad zou kunnen aanbieden aan snel bewegende mobiele gebruikers, dient het netwerk tussen hen uitgebreid te worden met concepten zoals snelle handover aangevuld met dynamische traffic engineering voor de ondersteuning van gebruikersmobiliteit. In dit onderzoek wordt vertrokken van een aantal scenario's die elk hun specifieke problemen met zich mee brengen. Zo wordt er enerzijds gekeken naar gebruikers op een trein, die zich langs een vast traject voortbeweegt en die dus een erg voorspelbaar gedrag heeft. Bovendien is de gevraagde bandbreedte, wegens het grote aantal treinreizigers, heel groot. In dit scenario zal het netwerk heel lokaal een grote hoeveelheid aan dataverkeer moeten kunnen verwerken en bijgevolg zal het zijn capaciteit dynamisch moeten kunnen aanpassen aan de positie (en dus ook de snelheid) van de treinen. De handoverproblematiek zullen we hier trachten op te lossen door het frequentiepatroon te laten meebewegen met de trein. Naast het treinscenario beschouwen we ook de situatie waar de gebruikers zich op auto(snel)wegen voortbewegen. Hier is het verkeerspatroon veel minder voorspelbaar, maar is de vereiste bandbreedte per voertuig wel heel wat lager. Het dataverkeer zal nu doorgaans vrij uniform verspreid zijn over het volledige netwerk. Er zal een volledig IP gebaseerd handover protocol ontwikkeld worden dat rekening houdt met de voorspelbaarheid van de gebruikers, bv. m.b.v. GPS, en dat hoogwaardige breedband multimedia diensten en ITS mogelijk maakt in de wagen.

***Commission D: Electronics and Photonics***

## OVERCOMING DIFFICULTIES IN DIRECT EXTRACTION OF SI/SIGE HBT DEVICES

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### ABSTRACT

A small-signal equivalent circuit model is a physically based model. This means that it reflects the important physical characteristics of the device. The complete small-signal equivalent circuit as seen from the probe tips consists of two parts: the intrinsic core of the device and the external parasitics. Concerning the parasitics, we can deduce three types: pad parasitics, impedance and admittance of the access transmission lines. In case of direct extraction approach, the first step is to subtract the effect of these parasitics, then applying any one from the methods for direct extraction as described in Refs. [1,2]. This approach however has two disadvantages: firstly, the subtraction process increases measurement error propagation [3], and secondly, some residual access transmission line pad imminence may appear in the intrinsic core equivalent circuit. As a result we obtain frequency-dependent intrinsic elements, which complicates the subsequent construction of the corresponding large-signal model. Therefore, a possible alternative solution is to assume an equivalent circuit for the access transmission lines, and to find the element values using random-based search algorithms. Next, we fix these values of the parasitic parameters, and using the same algorithms, we try to find the best set of intrinsic parameters that minimizes the difference between the simulated and measured *S*-parameters. Applying the mentioned methodology to a 0.8 μm emitter long, and 9.6 μm wide AMIS Si/SiGe HBT device at the bias condition $V_{BE}$ =0.94 V and $V_{CE}$ =1.5 V, yields frequency-independent elements, meaning that we overcame the problems encountered in direct extraction techniques. A comparison between simulated and measured *S*–parameters shows us that the maximal deviation is 9%, and this for $S_{12}$.

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**FABRICATION OF SELF ALIGNED DOUBLE GATE MOS TRANSISTOR**

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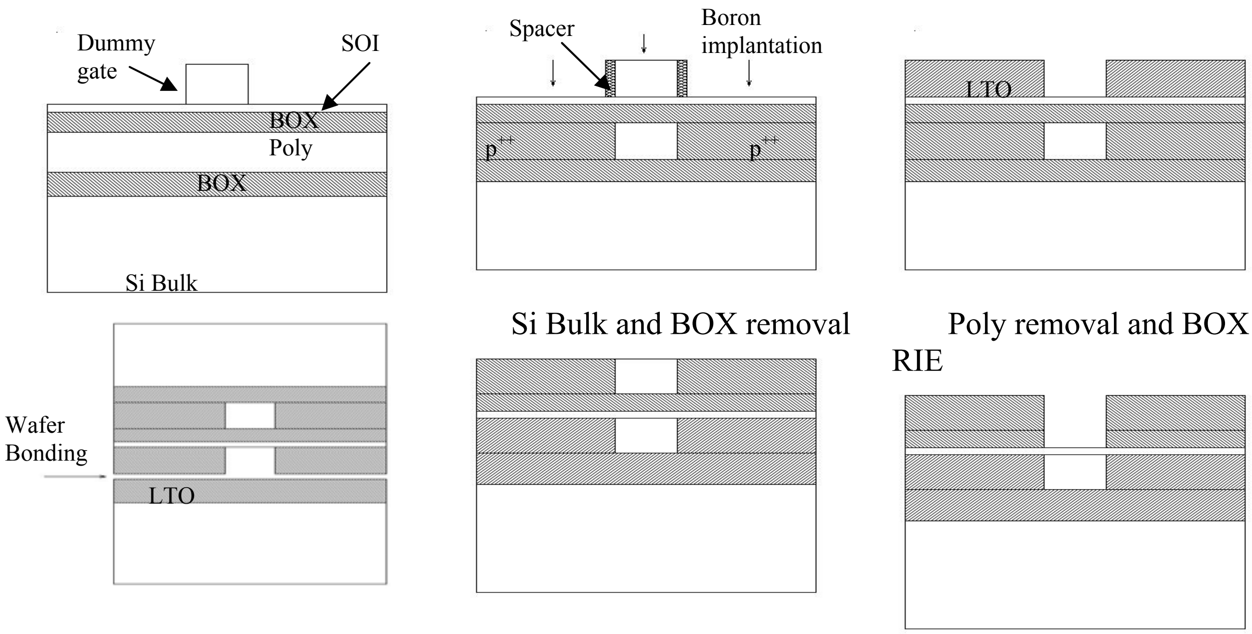
The fully depleted silicon-on-insulator (SOI) double gate MOS transistors are ideal structure to minimise short channel effect, parasitic capacitance and floating body effect. For very thin silicon films the inversion layer extends over the whole films giving rise to volume conduction. These kinds of structures have also the advantage to be able to increase the drain current and the transconductance. As the advantage of these structures becomes predominant over single gate technology for very small dimensions (in the order of a few deca nanometer), it is necessary to find a fabrication method to self align the top and the bottom gate.
The proposed method is based on the implantation of a $p^{++}$ polysilicon buried mask. Starting from a non conventional SOI wafer composed of a stack of BOX / polysilicon / BOX / SOI layers on a Si wafer, one deposits a polysilicon film which is patterned by electronic lithography to the dimension of the gate. Spacers are then added in order to compensate the lateral scattering of boron during the implantation and diffusion during the annealing steps. After high dose implantation of boron the spacers are removed, PECVD oxide is deposited, and then polished by CMP until one reaches the dummy gate which is removed by TMAH etching. The wafer is then bonded to a handle wafer and the backside is removed by TMAH etching until the first BOX layer. This BOX layer is removed by BHF etching and then the buried mask is revealed by TMAH etching which will preserve the doped zone and etch the undoped zone with a selectivity of 40. The second BOX layer is etched bye RIE in a $CF_4/H_2$ plasma. From this point the last fabrication steps are similar to classical CMOS process: growth of gate oxide, deposition of polysilicon gate, source and drain implantation, passivation, contact holes and metallization.



The poster will present all the steps of this process which have been characterized yet. The buried mask revelation has been qualified. TMAH etching selectivity of doped polysilicon is now well understood. CMP and wafer bonding are almost working. A lot of work has been done for stress reduction of the PECVD oxide layer which induces curvature of the wafer and prevents bonding and CMP. RIE of oxide with a mask of polysilicon has been characterized. It shows selectivity around 40. The electronic lithography step still to be performed.
An other approach to create self aligned structures is to perform an electronic lithography through a membrane. The membrane is composed of a five stack layers: poly 1, gate oxide 1, thin active silicon layer, gate oxide 2, poly 2. PMMA is vaporised on the both side of the membrane. The pattern of the two aligned gates is transferred to PMMA resist layer by electron beam lithography in controlling the beam energy; lower energy for the front gate and higher one for the bottom gate. This second method is analysed in collaboration with the University of Sherbrooke, CA.

# CHARACTERIZATION AND MODELLING OF QUASI DOUBLE-GATE SOI MOSFETS

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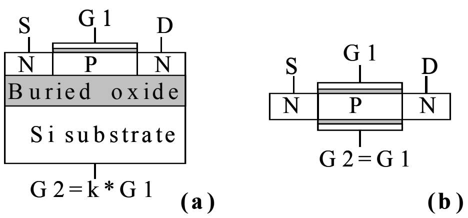
"Quasi" means to simulate/create something to resemble the object/device which we are interested in to have the properties/attributes of that certain object/device as close as possible. The main purpose of the so called quasi double gate (QDG) mode is to simulate as close as possible the characteristics of a real double gate SOI MOSFET device from the measurements of a single gate (SG) device. In order to emulate DG operation regime with a physical SG device, a back gate voltage, $V_{bg}$ is applied to the substrate of the SG device.
The determination of the $V_{bg}$ to be applied to the substrate can be determined via two methods. In the first method, $V_{bg}$ is determined by considering the thickness ratio between the front gate oxide and the buried oxide (BOX) of the device. *k* is multiplied with the corresponding applied front gate voltage, $V_{gs}$. Thus,

$$V_{bg} = k \times V_{gs}$$

The second method was proposed by *J.Pretet et al.*[1]. In this second method, the determination of $V_{bg}$ to be applied is not as straightforward as the first method. In this second method, the threshold voltage of the front gate, ($V_{thf}$) and back gate, ($V_{thb}$) are considered. Thus, the final expression is as follows,

$$V_{bg} - V_{thb} = k \times \left(V_{gs} - V_{thf}\right)$$



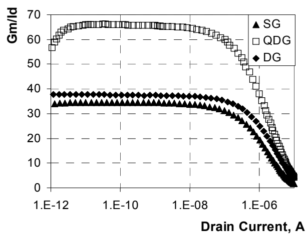



Fig. 2

Figure 1: (a) SG biased QDG mode (b) Double gate Figure 2: ATLAS simulation of $G_m/I_d$ curves for FD SOI MOSFET (L=0.25 μm) at $V_d$=50 mV

From the experiments and simulation that have been carried out, there is no major changes in both methods. However, it is clear that both methods cannot eliminate the dimensional effect whereby in weak operation, the QDG overestimates the transconducatance to current ratio (gm/Id)[2] whereas in strong inversion there is a good agreement between DG measurements and measurements of SG in QDG mode. Currently, through numerical simulations done with finite element method (FEM) software commercially available, such as ATLAS (SILVACO©) and analytical analysis, we are looking for an expression for the right coefficient, k valid from weak to strong inversion.

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# NONLINEAR MICROWAVE NANOBALLISTIC DEVICES FOR FREQUENCY DOUBLING

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The growing interest for systems operating at 80 GHz and 160 GHz for direct modulation/demodulation of optical telecommunication carrier essentially requires electronic devices operating up to THz frequencies. In the present work, we exploit the ballistic effects in nano-scaled InAlAs/InGaAs/InP based devices for RF applications up to the THz range. Owing to the large values of electron scattering length ($l_e$) (higher than 100 nm at room temperature) in InGaAs-based material system, it is possible to fabricate and operate ballistic devices of dimensions comparable to $l_e$ at room temperature [1], with sufficient repeatability and robustness. Those devices are fully compatible with the existing HEMT technology that has already demonstrated cut-off frequencies as high as 562 GHz for 25nm gate pseudomorphic $In_{0.52}Al_{0.48}As/In_{0.7}Ga_{0.3}As$ HEMT.
A Y-branch junction (YBJ) has been designed for RF operation. Nonlinear ballistic behavior of three-terminal ballistic junctions (TBJs) has been reported extensively in literature [2]. When finite dc voltages ($V_l$ and $V_r$) are applied to the left and right-hand side branches of a symmetric TBJ in a push-pull manner ($V_l = -V_r$), the voltage measured at the central branch ($V_b$) is always negative, thus the TBJ acts as a dc rectifier (Fig. 1). This ballistic effect is further enhanced in YBJs, depending on the angle between the two symmetric branches and is expected to offer excellent performance at extremely high frequencies [3,4]. In order to operate the Y-junction in push-pull manner in RF, a 180° phase-shift has to be introduced in one of the input branches. Here it is created around 10 GHz by means of a wideband hybrid coupler (Fig. 3 and Fig. 4)[5]. Such a feeder will enable to demonstrate the operation of the Y-junction as a frequency-doubler or power detector in the microwave range, in view of extrapolating its design for operation in the THz range. The intrinsic cut-off frequency, sensitivity and linearity of the YBJ frequency-doubler can indeed be improved by reducing the size and optimizing the device geometry.

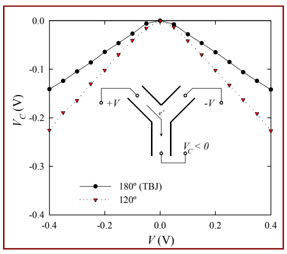


Fig. 1 Parabolic DC output of YBJ in push-pull mode.

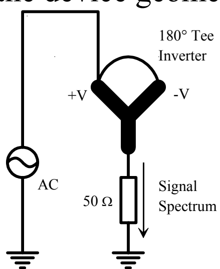


Fig. 2 Set-up of HF measurement.

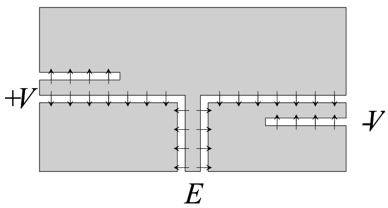


Fig. 3 180° Power divider.

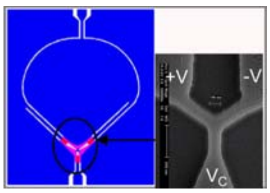


Fig. 4 Layout of RF feeder for push-pull mode.

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# USING MULTISINES FOR THE VALIDATION AND CONSTRUCTION OF RF BEHAVIOURAL MODELS

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## 1. ABSTRACT

The growing demand for higher performance from the devices leads to greater design complexity and the need for better methods to model RF components. One of the methods which has recently received much interest is behavioural modelling, mainly due to its relatively simple and unique approach regardless the device to be modelled, and also because it can be applied to a wide range of different components, including simple on-wafer transistors, entire circuits and even packaged components. Primarily, behavioural models were based on single-tone large-signal measurements, but following the trend towards more complex modulation schemes, the use of multisine excitations has been recently investigated. It was showed that this type of excitations not only accounts for the improvement of the model's accuracy as a result of good ability to approximate digitally modulated signals used in modern telecommunication systems [1], but also provides higher efficiency of experiment design by replacing many single-tone large-signal measurements [2]. This study elaborates the influence of the multisine excitation type on the accuracy and validity range of the RF behavioural model. For that purpose, state-space behavioural models for a packaged amplifier were derived according to four different multisine type large-signal measurements, performed on the Non-linear Network Measurement System set-up, and subsequently those models were evaluated for their accuracy. The comparison of the obtained results revealed a well-defined correlation between excitation type and model's accuracy. Moreover, this study showed that the multisine type that is the best for approximating a realistic digitally modulated signal is not automatically the best suited for accurate behavioural model generation.

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# ON THE ROBUSTNESS OF SPACE-TIME SIGNALING

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Multiple-Input Multiple-Output (MIMO) systems have drawn considerable attention in the communication society, by offering very high spectral efficiency and link reliability. Several communication techniques have been proposed to exploit the potential of MIMO systems. A common assumption in the design of space-time processing for MIMO systems is to consider the fading coefficients between the pairs of transmit-receive antennas as independent and identically Rayleigh distributed (i.i.d.). This is however an idealistic situation. In practice, the fading coefficients are correlated and this correlation depends on the antenna spacings and orientation, on the mutual coupling, the richness of scattering and the presence of dominant components. These effects highly influence the capacity as well as the performance of the space-time processing. High-rate codes, such as Spatial Multiplexing (SM), are significantly more affected by the propagation conditions than low-rate codes such as space-time orthogonal codes. In this communication [1], design criterion to guarantee the robustness of space-time codes in real-world channels is derived. No channel knowledge is assumed at the transmitter. We show that codes satisfying this criterion perform much better on real-world channels than codes only designed for i.i.d. channels. The criterion is totally general and can be applied to any kind of space-time code.

We consider a MIMO system with $n_t$ transmit and $n_r$ receive antennas, communicating through a frequency flat-fading channel. Over $T$ symbol durations, a codeword $\mathbf{S} = [\mathbf{s}_1 \ldots \mathbf{s}_T]$ of size $n_t$ x $T$ is transmitted through $n_t$ transmit antennas. At the $k^{\text{th}}$ time instant ($k$=1…$T$), transmitted and received signals are related by the following relationship

$$\mathbf{r}_k = \sqrt{E_s}\ \mathbf{H}\,\mathbf{s}_k + \mathbf{n}_k \tag{1}$$

where $\mathbf{r}_k$ is the $n_r$ x 1 received signal vector, $\mathbf{H}$ is the $n_r$ x $n_t$ channel matrix, $\mathbf{n}_k$ is a $n_r$ x 1 zero mean complex additive white Gaussian noise (AWGN) vector with $E\{\mathbf{n}_k\,\mathbf{n}_l^{\,H}\} = \sigma_n^2\,\mathbf{I}_{n_r}\,\delta[k-l]$ (the superscript $^H$ denotes conjugate transpose), $\mathbf{I}_{n_r}$ is the identity matrix of dimension $n_r$ x $n_r$ and $E$ denotes the expectation operator. $E_s$ is the average energy available at each transmit antenna. The channel is assumed constant over $T$ symbol durations. We assume that the instantaneous channel realizations are unknown at the transmitter and perfectly known at the receive side and that Maximum-likelihood (ML) decoding is performed.

Let $\mathbf{C} = [\mathbf{c}_1 \ldots \mathbf{c}_T]$ and $\mathbf{E} = [\mathbf{e}_1 \ldots \mathbf{e}_T]$ be two different codeword vectors of size $n_t$ x $T$. To make space-time codes robust in the presence of poor scattering conditions, it is shown in [1] that the code should be designed such that it performs well in terms of maximizing

$$\min_{\theta}\ \min_{\mathbf{C},\mathbf{E}} \left\| (\mathbf{C}-\mathbf{E})^T\,\mathbf{A}_T \right\|^2 \tag{2}$$

where $\mathbf{A}_T$ is the transmit array response in the direction of departure $\theta$.

On Figure 1, the cumulative distribution functions (cdf) of the Symbol Error Rate (SER) for the classical SM and a new SM code in the presence of measured channels have been displayed. It is clear that in real-world channels, the new code presents a better robustness since the cdf evolves at smaller SER compared to the classical SM. Moreover, the cdf of the new code is much steeper, showing that the performance are less dependent on the propagation conditions.

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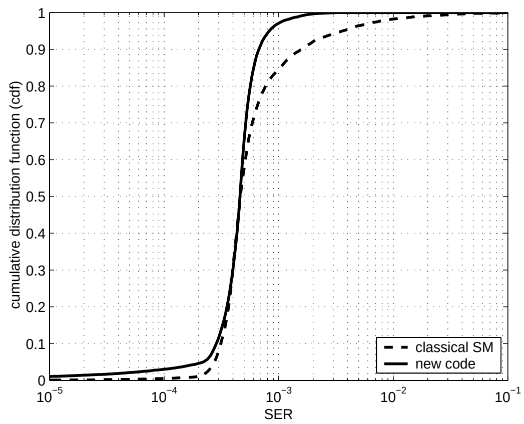


Figure 1: CDF of the SER in the presence of measured channels.

**FINITE ELEMENT METHOD SIMULATIONS OF MEMS BILAYERS**

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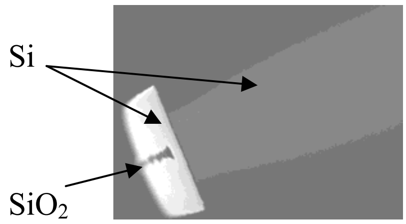


Figure 1 SEM picture of a stressed bilayer-cantilever

Cantilevers are structures classically used to extract mechanical performances of thin MEMS materials. Those structures allow the determination of stress gradients along the thickness of the characterized film. This is illustrated in Figure 1.

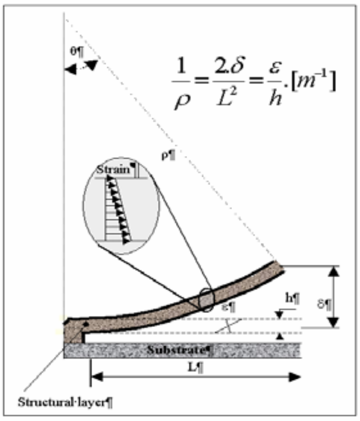


Figure 2 Stress gradient measurement from the shape of a cantilever

If a bilayer – a stack of two materials- is created, a stress gradient can appear. This is due to the mechanical properties mismatch for the two materials such as the thermal expansion coefficient.
This stress gradient will result in a circular shape of the released cantilever. This stress gradient can be due to:
- lattice misfit
- film growth effects
- thermal expansion coefficient misfits between the two layers

The starting point of our analysis was the observation of an oxide residue below silicon cantilevers released from a SOI –Silicon On Insulator- wafer. Those patterned and released cantilevers presented a circular shape (cfr. figure 2). Simulations taking into account only the thermal expansion coefficient misfit between the two materials (Si-$SiO_2$) reproduced nicely the mechanical phenomenon (Figure 3).

Values of the thermal expansion coefficient were found in the literature for Si ($\alpha_{Si}$= 2.33E-6 / ° C) and $SiO_2$ ($\alpha_{SiO2}$= 0.35E-6 / ° C). The growth temperature of the oxide was also considered. We did a simulation of a 200µm length cantilever using the ABAQUS software. The loading consisted in decreasing the temperature from the growth temperature (1050°C) to the ambient temperature (20°C). The results of this simulation are proposed in Figure 3.

Those results were confirmed by an analytical model proposed by Ghering *et al* [2].
Further work consist in developing methods to measure accurately the thermal expansion coefficients of materials in order to better predict the behavior of structures realized using our own deposition recipes and methods.

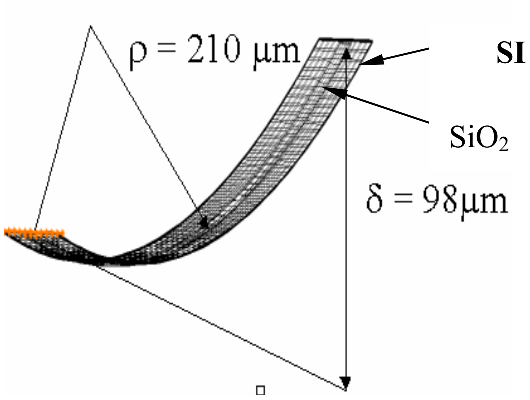


Figure 3 ABAQUS simulation of bilayer-cantilever

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# AC ANALYSIS OF FLOATING BODY EFFECTS IN PD SOI MOSFETS

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## 1. INTRODUCTION

As reported many times impact ionization ($I_i$) causes charging of the body region in floating Body (FB) Partially Depleted (PD) SOI MOSFETs, inducing a kink in the $I_d$-$V_d$ curve. As a consequence, a peak is observed in the DC output conductance ($G_d$) - $V_d$ curve when the device is measured in saturation regime. The AC analysis of this effect is fully explained in [1]. In more advanced technologies, gate tunneling through ultra-thin (~ 2 nm) oxide also provides parasitic body charging: electrons tunneling through the gate leave behind free holes that positively bias the body. Gate tunneling was demonstrated to induce a second peak in the device transconductance ($G_m$) - gate voltage ($V_g$) curve recorded in linear regime ($V_d$ =100 mV) [2]. We report here an AC analysis of this second peak and show that its AC behavior can be described with a small signal equivalent circuit that also includes impact ionization effects, and thus the AC observations reported in [1].

## 2. EXPERIMENTS AND MODEL

The small signal circuit proposed to combine and analyze the AC behavior of the two body effects mentioned above is depicted in Fig. 1, where it is seen that impact ionization is included in the model through the conductance $g_{bds} = \partial I_i/\partial V_d$ and gate tunneling through the complex admittance (i. e., a parallel combination of a conductance and a capacitance) between gate and body ($y_{gb}$), respectively. $y_{bs}$ and $y_{bd}$ are also complex and are respectively associated with source and drain junctions. According to that model, the extrinsic (measured) $G_m$ (resp. $G_d$) is a function of the device intrinsic tranconductance ($g_{mi}$) (resp., output conductance ($g_{di}$)) as well as an additional term related to the AC variations of the body potential ($v_b$):

$$G_m = g_{mi} + g_{mb} Re(v_b/v_g) \qquad (1)$$
$$G_d = g_{di} + g_{mb} Re(v_b/v_d) \qquad (2)$$

where $g_{mb}$ is the device body transconductance. The frequency dependence in both expressions is exclusively contained in the $v_b/v_x$ ratio ($x$ = $g$ or $d$). In linear regime, $I_i$ can be neglected and $g_{bds}$ removed from the model. In that case (1) becomes:

$$G_m = g_{mi} + g_{mb} Re\left[\frac{y_{bg}}{y_{bs} + y_{bd} + y_{bg}}\right] \qquad (3)$$

In saturation, $I_i$ must be considered and (2) becomes:

$$G_d = g_{di} + g_{mb} Re\left[\frac{y_{bd} + g_{bds}}{y_{bs} + y_{bd} + y_{bg}}\right] \qquad (4)$$

It is worth noticing that *both* expressions are characterized by a zero-pole doublet, which is consistent with the experimental data (Fig. 1a and b). Besides, the expression for the poles is the same:

$$f_p = \frac{1}{2\pi} \cdot \frac{Re(y_{bs} + y_{bd} + y_{bg})}{Im(y_{bs} + y_{bd} + y_{bg})} = \frac{g_{body}}{2\pi C_{body}} \quad (5)$$

where $g_{body}$ and $C_{body}$ are respectively the total conductance and the total capacitance seen by the body towards all external nodes. During $G_m$ (resp. during $G_d$) measurements in linear (resp. saturation) regime, gate tunneling (resp. impact ionization) biases the body, thereby increasing $g_{body}$. As this parasitic current is an increasing function of the DC $V_g$ (resp. $V_d$ ) bias, (5) demonstrates that the pole frequency increases at higher bias, which is what is observed for both $G_m$ and $G_d$ measurements. However, as impact ionization current biases the body to higher levels than gate tunneling, the observed pole frequencies are higher for $G_d$ than $G_m$ measurements. The validity and accuracy of this model were numerically confirmed with Eldo simulator and are presented in [3].

## 4. CONCLUSION:

Gate tunneling (resp. impact ionization) currents increases the transconductance (resp. output conductance) of FB PD MOSFETs in linear (resp. saturation) regime. In both cases, the frequency dependence of this increase exhibits very low frequency poles due to the high impedance seen by the body towards the external nodes of the device. The AC behavior of both effects can be predicted using a single small signal equivalent circuit that is proposed in this paper.

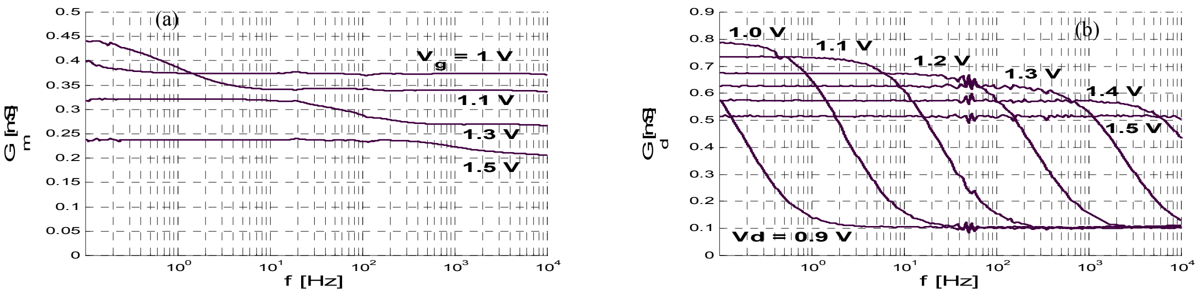


Figure 1: Experimental data (FB PD SOI MOSFET, L = 2 µm, W = 30 µm) (a): $G_m$ vs $f$ for $V_d$ = 50 mV and varyious $V_g$ (b): $G_d$ vs $f$ for $V_g$ = 0.6 V and varyious $V_d$.

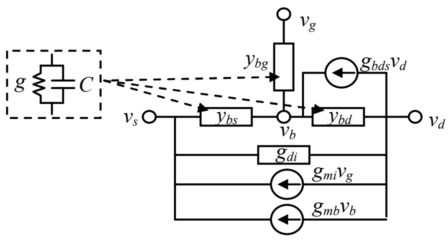


Figure 2: Small signal circuit accounting for body effects induced by gate tunneling and impact ionization in FB PD SOI MOSFETs.


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**DEVELOPMENT OF AN UNIFIED ARCHITECTURE MODEL FOR SOC'S IN DEEP SUB-MICRONIC TECHNOLOGIES**

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Today, we slowly enter in what could be called the "Post-PC Era" : the microelectronics market evolves now towards embedded systems (mobile phones, PDA, smart sensors, Body Area Network, …). Following this trend, portable devices require more and more hardware and software to support their new functionalities (MP3 playing, MPEG decoding, gaming, speech recognition …) leading to a serious increase of the design time and cost.
A new generation of electronic components called System on Chip (SoC) is being developed to face this problem. SoC's are expected to make intensive reuse of existing IP and also to be low powered and sufficiently heterogeneous in terms of resources (FPGA, microprocessor, DSP, memory, ASIC, …) to provide good performances to a large set of applications.
In the meantime, the progress in microelectronics enables faster chips to be produced and is mainly conducted in two separate areas :

- at the *Technological* level : we are crossing the "100 nm" barrier –hence entering deep sub-micronic technologies- and will be able to integrate more than one billion transistors on a single die in a near future. The complexity introduced by these changes in the silicon processes prevents us now from working at the gate level and will involve a revolution of the design tools forcing us to use a "top-down" approach –called System Level Design- rather than a "bottom-up" one.
- at the *Architectural* level : as the maximum clock frequency and silicon area are increasing, the principle of a single synchronous silicon die becomes obsolete. In a near future, we will be forced to use new structures called Coarse Grain Architectures where a few hundreds of synchronous grains will communicate through an integrated interconnection network on a single die. A lot of architectures based on the Coarse Grain principle are proposed in the literature but we actually have no easy and reliable way of comparing them in terms of performances.

The research we propose precisely tries to solve this design problem by building a unified model of SoC's for the different Coarse Grain Architectures to compare them and predict their performances a priori.
Such a work will provide designers with high level tools for helping them in their architectural choices and will attempt to bridge the gap between processes and design to understand their interaction and impact on performances.

**OPTIMIZING THE INTERCONNECT OF FUTURE SYSTEMS ON CHIP IN DEEP SUB MICRON TECHNOLOGY**

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The tremendous evolution of microelectronics is allowing the number of transistors on a silicon die to double each generation. Today the capacity of a chip is such that it is possible to integrate a whole complex system on a single chip, a concept known as System-on-Chip (SoC). Compared to multi-chip systems, SoCs offer better performance with reduced power consumption and size.
The integration of more and more transistors is made possible by constantly reducing their feature size (e.g. gate length). As feature sizes get smaller than 100 nm, we are about to enter a new era of design challenges called Deep Sub Micron (DSM). This era is mainly characterized by the surprising fact that communication will become more critical than computation: interconnect will be the dominating factor determining speed, noise and power.
By the end of the decade, the International Technology Roadmap for Semiconductors predicts chips containing billions of transistors, making possible the integration of a multitude of components: processor cores, large embedded memories and reconfigurable logic. One of the fundamental problems that remains is to determine which communication architecture should interconnect those components.
This research targets heterogeneous SoCs that will be used in future hand-held devices (PDAs, mobile phones). These devices will be used for various demanding applications such as advanced multimedia, cryptography and voice recognition. These applications will impose stringent requirements on the communication architecture (latency, bandwidth, jitter).
As we target battery-powered devices, energy consumption will be by far the most constraining resource. In particular the interconnect will be a huge power consumer in DSM technology. Unfortunately, with the exception of buses, power consumption has been neglected in interconnection networks for a long time. Only very recently have a few studies emerged in that domain.
Many on-chip communication architectures have been proposed from advanced buses to complex packet-switched networks known as Networks-on-Chip. However no in-depth studies and fair comparisons have been realized so far to determine which network performs the best under realistic traffic patterns.
A first step of the research consists of designing a high-level Networks-on-Chip simulator that will allow the comparison of the different communication architectures in terms of power consumption and performance, and eventually to identify interconnection networks that will satisfy the stringent requirements imposed by the application domain.

***Commission E: Electromagnetic noise and interference***

# DISTURBANCE LINES IDENTIFICATION IN PRINTED CIRCUIT BOARDS BY MEANS OF WAVELETS

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**SUMMARY**

In the last decade, a significant effort has been developed to increase the speed of digital circuits. However, At high frequencies, signal edge rates become faster giving rise to the electromagnetic interference phenomenon known as crosstalk. As a result, the integrity of the signal propagating down the interconnects of the printed circuit board (PCB) may be affected if the phenomenon is not taken into account in the early stage of design of electronic systems. It is thus of great importance to detect crosstalk and identify their source using simulations at design stage in order to eliminate or at least to reduce them to acceptable levels before constructing the prototype. This paper investigates a wavelet-based approach for disturbance lines identification by estimating the frequency of the source of disturbance and the separation between the coupled lines.
The procedure of frequency estimation of interference signals exploits the ability of the wavelets to detect singularities. The wavelet decomposition of a trapezoidal signal yields a set of zero coefficients except those corresponding to the transition, i.e., from high logical level to low logical level and vice versa. By inspecting the wavelet coefficients, one can detect the presence of the interference and separate it from the disturbed signal. The period of the crosstalk may be easily estimated by performing a wavelet decomposition of the reconstructed interference at an appropriate level and therefore detecting maxima of the autocorrelation of the wavelet coefficients.
The algorithm used to estimate the separation between two coupled utilizes two empirical laws based on physical ground and verified by simulations. It is assumed that the mean powers of two interferences caused by two signals with different frequencies and propagating down a line at some reference distance $s_{ref}$ from the disturbed line have been obtained by simulation during a preprocessing stage. The proposed approach is tested on two simulated coupled tracks in a PCB where the XFDTD (finite difference time domain method) simulator has been used to generate the crosstalk. It is shown that the frequency and the separation can be accurately estimated and used to identify the source of disturbance without making any assumption about the location of disturbance line.

# ON THE QUALITY OF SPLICES IN TRANSMISSION LINES

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## 1. INTRODUCTION

Since the introduction of digital services on the classical telephone lines, crosstalk has always been a major problem to cope with. To improve the quality of transmission, one of the fundamental conditions is a good understanding of the quantity of disturbance that is to be expected for a particular transmission system. As spectral management of access networks relies on reliable crosstalk models, and since crosstalk cancellation is the foundation for current improvements on the ADSL capacity, it is important to take into account all possible causes for the lather. Splices are one of the factors that seem to have been overlooked for quite some time.

Various attempts have been made to estimate the effects of the connecting of two or more lines with different crosstalk coupling constants, but what was not considered is the physical existence of splices (i.e. the untwisting of pairs in order to be able to connect them), and the existence of 'bad' splices (i.e. old splices that are untwisted over too much length, water leakage or sloppy connections).

## 2. THE MODEL AND MEASUREMENTS

In a previous paper by the first author, a model for the crosstalk, caused by splices, between two twisted pairs in a transmission line has been proposed. This model was verified on a lab-produced splice. Splices can be described as a local pointcoupling in a line, causing an extra amount of crosstalk, which is added to the total crosstalk on the line.

A series of measurements is proposed in which various factors influencing the quality of the spices are determined. First, the influence of the tightness of the splice is examined, followed by measurements determining the influence of soil humidity on the quality of the splice.

## 3. RESULTS

The results of the measurements are evaluated in an ADSL system simulation, to validate their influence on the achievable bitrate of the system. This allows quality boundaries to be set and provides with a method to evaluate 'bad' splices in a network.

# MEASUREMENTS IN RADIATED EMISSIONS IN CASE OF LARGE SYSTEMS

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This work is devoted to the research of alternating methods of measurements in radiated emission (between 30MHz and 1GHz) in case of large electrical systems.
According to the standards [1, 2], the measurements should be performed in an OATS (*Open Area Test Site*), which is clear of objects apart from the EUT (*Equipment Under Test*) and the measurement antenna. Besides, the measurements have to be carried out in the *far-field region* and for this reason, the antenna should be placed 10 meters (or 3 meters) away from the EUT.
Furthermore, the EUT should be put on a metallic surface to standardize the reflectives waves from the ground. Finally, the antenna has to be moved between 1 meter and 4 meters in height in order to obtain the maximum signal.

For large systems, it is quite impossible to follow the standards, because there are numerous problems with these recommandations.

Indeed, lots of large systems are very difficult to move and the measurements have to be performed *in situ*. In this case, the ambient radiated by other equipments placed nearby disturbs the emission signals coming from the EUT. Besides, the measurement antenna cannot be placed as far as 3 meters away from the EUT and the measurements must then be performed in the *near-field zone*, where the fields are more difficult to characterize.
Moreover, the ground surface is neither perfectly reflective nor completely transparent to the incoming electromagnetic waves. Consequently, it is very difficult to take into account the influence of the ground surface on the results obtained.
The main solutions proposed are: firstly, the noise coming from the equipments placed nearby will be removed by a *differential* method with a phase retrieval algorithm.
Secondly, to get far-field results from near-field measurements, a near-field $\rightarrow$ far-field transformation will be applied.
Thirdly, from measurements and a *genetic algorithm* [3], a set of electric and magnetic dipoles, which radiates the same near-field as the ones measured, will be substituted to the EUT. Then, the far-field results will be deduced from this set of elemental dipoles.
Consequently, by using these techniques, radiated emission signals coming from large EUT, which cannot be placed in an OATS or moved, will be known with a great accuracy.

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# MEASUREMENTS IN CONDUCTED EMISSIONS IN CASE OF LARGE SYSTEMS

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The research of new methods of measurements, in conducted emission (between 150kHz and 30MHz) in case of large systems, is the subject of the present work.
The standards describe two ways of measuring conducted emissions. The first one is to use a Line Impedance Stabilized Network (LISN) to perform these measurements. A LISN prevents the Equipment Under Test (EUT) from the noise coming from the mains and provides a defined impedance (e.g. 50 Ω) at the point of measurement. The second way is to use a passive voltage probe.
The LISN is preferred but the voltage probe is used when no other choice is available. Indeed, a LISN can only absorb limited currents and can be prohibitively costly.
The voltage probe method is a direct solution when the LISN cannot be inserted in the circuit but the two methods give very different results (for situations in which they can be both used) [1]. Indeed, the distributed impedances of the lines between the EUT and the power mains network are not negligible. Furthermore, the attenuation factor of the probe (-30dB) and the ambient coming from fields radiated by devices placed nearby are strong problems [1, 2]. Therefore, the results obtained with a voltage probe are not accurate.
A new method of measurement, taking into account the distributed impedance of the lines by using a six-pole model (generalization of the four-pole theory) allowed to enhance the results [3].
But the second problem still persists. Even if good results could be obtained with the "six-pole method", they cannot be compared with results obtained with a LISN. Indeed, the LISN acts as a filter regarding the noise coming from the power mains and the results obtained without LISN will be a *bad-estimation* of the actual signals.
Consequently, the next step of the study will be to take into account the impact of the LISN on the obtained results. In other words, a "virtual" LISN will be inserted in the circuit to overcome this over-estimation. Besides, the signal to noise ratio will be enhanced by preamplification techniques.

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***Commission F: Wave propagation and Remote Sensing***

# RAW-DATA SIMULATION FOR SAR SYSTEMS ON UAV-PLATFORMS

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## 1. UAV REFERENCE FRAME

High-resolution SAR imaging is most suitable from a large stable surveillance platform, SAR imaging from a UAV however is an altogether more challenging technical problem. This is due partly to the extra manoeuvrability of unmanned platforms. A simulation model of a complete radar system should allow the inclusion of models of platform dynamics and navigation systems to assess the effects on image formation. The existing end-to-end simulator from DLR, SETES, developed for spaceborne SAR systems, will be used to incorporate the specific aspects to be studied for UAV-based SAR systems. A first step in this simulation is the creation of a representative raw data set.
Therefore a UAV reference frame is established. It will limit the domain where the performance analysis and optimization of SAR systems will be performed. The reference frame will be defined by choosing a couple of existing UAV-systems (preferably complementary in their roles and missions) equipped with a SAR sensor.

## 2. CREATION OF FLIGHT DATA FOR RAW DATA SIMULATION

An appropriate flight data set (FDS) incorporates motion errors, controlled and uncontrolled (wind gust e.g.), as well as a vibrational environment (due to atmospheric turbulences e.g.), and will thus allow analysis of the effects of these errors. To create a wide variety of FDS, a program is developed with a number of variable input parameters. The two possible main inputs are 'user defined nominal flight path' (module 1) and 'flight path extracted from navigation data' (module 2).

Through the first module a three-dimensional standard flight path can be introduced. Additionally a vibrational environment can be chosen (by means of a user simulated spectral density or a spectral density chosen from the SETES database) as well as a set of motion errors (by means of user simulated motion errors or motion errors defined in the SETES database). If no errors need to be added the defined flight path will be converted to an earth based coordinate system and sent to the raw data generator (RDG). In the other case the flight path will be merged with the eigenfrequency information and the motion errors, converted to the spatial domain and then, as mentioned before, converted to the appropriate coordinate system and sent to the RDG. Through the second module the flight path will be created based on simulated or measured GPS/INS data. Since they will already include (un)controlled motions from the platform, these do not need to be added. The vibrational environment can be added and processed as described above.
In all cases the data sent to the RDG will indicate the exact antenna phase centre position for each transmitted pulse.

# ACCURATE MODELLING OF GPR SIGNAL TO MAP THE SOIL DIELECTRIC PROPERTIES

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For a large variety of environmental and civil engineering applications, the dielectric characterization of the shallow subsurface using remote sensing is a matter of concern. In that respect, ground penetrating radar (GPR) constitutes a promising high resolution characterization tool. However, notwithstanding considerable research has been devoted to GPR since the 1960s, its use for assessing quantitatively the subsurface properties is still constrained by the lack of appropriate GPR systems and signal analysis methods.

Research has focused on the development of a new integrated approach including GPR design, GPR signal forward modeling, and GPR signal inversion to estimate both the dielectric permittivity and electric conductivity of the shallow subsurface [1, 2]. We propose to use an ultrawide band stepped frequency continuous wave radar combined with an off-ground monostatic TEM horn antenna. This radar configuration is appropriate for real time mapping, and allows for an accurate forward modeling of the radar-antenna-subsurface system. Forward modeling is based on linear system response functions for describing the antenna, and on the exact solution of the three-dimensional Maxwell equations for wave propagation in a horizontally multilayered medium representing the subsurface. Model inversion, formulated by the classical least square problem, is carried out iteratively using advanced global optimization techniques.

The proposed approach is validated under laboratory conditions on a tank filled with a two-layered sand subject to different water content levels. The inverse estimation of the soil dielectric permittivity is remarkably well in accordance with each water content level. The identification of the electric conductivity led to less satisfactory results. However, a sensitivity analysis demonstrated the good stability properties of the inverse solution, and put forward the necessity to reduce the remaining measurement and modeling errors by a factor 10. This may partly be achieved through a better characterization of the antenna transfer functions, and by performing measurements in an environment without close extraneous scatterers.

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# COMBINING RADAR AND RADIOMETRY FOR MINE DETECTION

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*Resorting thoroughly to the physical and mathematical understanding of the radar signal is of significant importance to landmine detection efforts, both in supporting identification techniques, as well as in predicting their performance. Within this context, research has focused on the development of an integrated approach including GPR design, antenna modeling, soil characterization, and mine signature identification. Concepts and methods developed for the radar will be extended to the radiometer.*

The adopted GPR system consists in an ultrawide band (UWB) stepped frequency continuous wave (SFCW) radar combined with a TEM horn antenna to be used off-ground in monostatic mode. This radar configuration is of practical interest since it responds to subsurface mapping requirements and allows for an efficient and more realistic modeling of the radar-antenna-subsurface system. Modeling of the system is based on linear system response functions and on the exact solution of the three-dimensional Maxwell equations for wave propagation in a horizontally multilayered medium representing the shallow subsurface.

Promising advances have further been done on the identification of the soil dielectric properties from the radar signal. The approach is based on the inversion of the radar signal, which is formulated by the classical least square problem. The objective function is minimized using combined global and local optimization algorithms. The method was validated in laboratory conditions on a tank filled with a two-layered sand subject to different water content levels. In this case where EM wave propagation phenomena are fairly complex, the model agreed very well with the measurements and the soil dielectric properties were accurately determined from the GPR signal inversion.

Research is currently aiming at extending the above approach for extracting the signature of an inhomogeneity (perfectly conducting or dielectric) embedded in the multilayered medium. The inhomogeneity is included in the radar-antenna-subsurface model using a specific transfer function. This last is computed using the Method of Moments. Measured and computed target's radar signature agree favorably. The computer code can handle arbitrary PEC bodies embedded in stratified medias and will be extended to dielectrics.

Henceforth the antenna model constitutes a field usable tool to improve subsurface sensing using monostatic GPR, yet research on the soil dielectric characterization and mine identification still pertain to concept demonstrators that have to be progressively improved and validated in conditions closer to the reality.

Once the radar-soil-model will have been validated, the theoretical concepts and numerical methods developed for the radar will be extended to model accurately the noise temperature of a scene constituted by a body buried in a soil.

# A PROBALISTIC SIMULTANEOUS ITERATIVE RECONSTRUCTION TECHNIQUE APPROACH APPLIED TO SUBSURFACE SEISMIC TOMOGRAPHY

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Problems related to landscape management, natural hazards and civil engineering involve subsurface structures that can be delineated by geophysical imaging. Seismic tomography can accurately characterize a medium according to its velocity variations.

Traditional seismic traveltime tomography based on ray-tracing methods assumes that the waves' frequency is infinite. As a consequence, only the medium located along the ray path has an impact on the wave propagation. In subsurface tomography, the infinite frequency assumption does not hold, as targets have about the same size as the wavelength. The seismic waves' propagation is affected not only by the medium along the shortest traveltime path but also by the medium located in its vicinity.

JaTS, our Java seismic tomography software, implements original algorithms achieving optimal accuracy with reasonable computing costs. A second order Fast Marching Method is used for solving the eikonal equation, therefore enabling a fast and robust computation of seismic traveltimes between sources and receivers. The wavepaths are represented by Fresnel volumes rather than by conventional rays. This approach accounts for complex velocity models and has the advantage of considering the effects of the wave frequency on the velocity model resolution. The model is computed by a Simultaneous Iterative Reconstruction Technique which has been reformulated to integrate Fresnel wavepaths by using a probabilistic approach. In addition, various utilities are implemented in order to decrease artifacts occurring in the vicinity of the sources and receivers. JaTS also offers the possibility of reconstructing the velocity field on a grid larger than the one used for the wave propagation computation. This contributes to stabilize the estimated values. All of the seismic processing tools have been integrated using a user-friendly graphical interface. JaTS represents a tightly integrated tool suite that supports the entire process of importing SG2 field records, first-break picking, forward modeling and velocity-field computing across multiple platforms.

Future development is under way for automatic first-break picking and extension of the algorithms to 3D processing.

*Commission K: Electromagnetics in Biology and Medicine*

**NEW STRATEGIES OF ACQUISITION AND PROCESSING OF ENCEPHALOGRAPHIC POTENTIALS**

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The measurement of the EEG (electroencephalogram) is a medical technique of diagnose that started to be explored in the thirties and has been used on large scale since the sixties. It consists to measure the biopotentials emitted by the upper layers of the brain at different standardized places on the scalp. It contributes to diagnose diseases (like epilepsy), brain tumors, sleep disorders, etc.
Three main problems are connected to the recording of the EEG:
First, the potentials being of the order of magnitude of the microvolt, those measurements are very sensible to noise and hum from the surrounding.
Moreover, because of the trend to measure those signals on larger periods (many hours or even days) it becomes uncomfortable for the patient and especially for young children to stay immobile during the whole recording. Therefore a wireless system would be an improvement.
Also because of those large periods of recording, the use of signal processing techniques to detect some patterns of clinical interest becomes nearly mandatory to reduce the human work.
Although some researches have been done on each of those three topics, few have been done on the incidence and the interaction of one topic on the other. Those topics are however tightly bounded. For example, the wireless part of the EEG can degrade the signal by polluting it. It is also obvious that the signal acquisition and processing are bounded because the SNR required for the acquisition depends on the ability of the signal processing to reject artifacts and hum, and the quality of the pattern detection depends on the SNR of the signal acquisition.
The aim of the thesis is to quantify those interactions and determine what is the best combination in terms of signal quality, patient comfort and pattern detection. Our approach is to study, develop and validate a portable EEG that would have an improved signal quality, and to see how its enhancement can improve automatic pattern detection. The later will be applied to the case of the “continuous spike-and-wave discharges during slow sleep” - a kind of infantile epilepsy - to limit the field of investigation. Some tools like a brain wave generator will also be developed.
The thesis is done at the Laboratory of Electronic and Microelectronic (ULB) and is financed by the FRIA. We benefit from the support of the firm MEDATEC.

# A STRESS REDUCING DESIGN TO SET-UP A LONG-TERM STUDY ON THE ATHERMAL BIOLOGICAL EFFECTS OF MICROWAVES

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## 1. INTRODUCTION

Today, very few people are exposed to thermally significant levels of microwaves. The great majority of exposures occur at levels at which weak-field interactions would be the only possible source of any adverse health response. Data to assess human health hazard in terms of pulse peak power, repetition frequency, pulse width and carrier frequency, are needed in view of the widening application of systems using modulated signals like cellular phones and radar systems, involving both occupational and general population exposures. It is of a vast importance to assess the range of microwave effects in relationship to human health. An increased or decreased presence of a biological parameter does not necessarily indicate any pathological process, but may be a useful index for biological dosimetry of microwave exposure.

## 2. OBJECTIVES

The objective of the present study consists in verifying on an animal model the cellular alterations and possible biological modifications due to microwave exposure (1 GHz and 10 GHz), using both continuous waves and pulsed waves. The immune system of the body is mainly based on the good functioning of the different parts of it: this study focuses on physiological changes of selected blood and hormonal parameters.

## 3. MATERIAL AND METHODS

In our project, the *Wistar albino* rat is used as a model to study the possible biological effects of low power microwaves. The different groups of rats are exposed 2 hours a day, 7 days a week during 12 months to microwaves at 1 GHz and 10 GHz, continuous and pulsed mode at an average power density of respectively 200 $\mu W/cm^2$ and 500 $\mu W/cm^2$. We selected a series of tests based on the following parameters: blood cells, stress-induced hormones (adrenocorticotrophic hormone and cortisol) and cytokines (interleukine 1, 6 and tumor necrosis factor a). Enzyme-linked immunosorbent assays (ELISA) are used to quantify those hormones in the blood plasma. Body weight is monitored with minimal disturbance to the animal. It is a reasonably sensitive indicator of stress, especially chronic stress, and therefore it is strongly recommended that animals which may be stressed should weight regularly and their body weights be compared with those of controls. Daily observations of the rats behaviour are part of the protocol.
The *Wistar albino* rats are exposed in groups of 30 in an exposure unit (1.11 m x 0.60 m x 0.71 m). This method constitutes an innovating element. In most of the other studies, the rats are kept separately in plexiglass cylinders during exposure, where free space to move is limited. Such a narrow housing causes an enormous stress response in the rat [1]. This stressor

creates a supplementary variable, which could possibly mask secretion of certain stress-related hormones or corticosteroids [2].
We use 6 month-old rats with an average body weight of 600 grams. Animals are kept in normal vivarium conditions and are made accustomed to experimental procedures, environment and handling 3 months prior to the beginning of the experiments. It is important to run the experiment always on the same period of the day because of the circadian rhythm in the secretion of corticosteroids.
To identify the rats during the whole experiment in general and during blood sampling and observation in particular, an unmistakable permanent method to distinguish one rat from another has been thought out. The ear of the rat is pierced following a formerly established pattern of figures. This procedure is carried out under total anaesthesia on the base of sevoflurane.

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***Omissions: The next two papers belonging to commission B and K disappeared for unknown reasons from the files of the editors. They were submitted in due time. The editors apologise for this error and rectify this omission in the electronic version. The other pages have been left unchanged.***

# THE SHORTED ANNULAR RING REDUCED SURFACE WAVE MICROSTRIP ANTENNA

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This paper presents some basic properties and comments on the shorted annular ring (SAR) reduced surface wave microstrip antenna. This antenna was introduced by Jackson [1] in 1993 stating that no surface waves in the dominant TM0 mode are excited if certain design criteria are taken into account. This results into a high radiation efficiency antenna design. Another advantage is the fact that there is no radiation at low elevation [2] resulting in suppression of interference caused by reflections. This is a useful property for GPS antennas since this interference has an influence on position determination.
The design of the SAR involves rather complicated mathematics such as an equation containing multiple Bessel functions. To overcome this problem a simple design procedure was developed [3] based on an asymptotical approximation of the ring width normalized to the substrate wavelength. The accuracy of this first approximation was further improved by introducing correction terms. These correction terms were found using curve fitting. This simplified design procedure allows the determination of the dimensions of the SAR with pen and paper.

Although it was originally stated that the radiation efficiency is very high for a SAR since the dominant TM0 surface wave mode is not excited, a clear connection between radiation efficiency was found by simulation. The radiation efficiency degrades to about 30% for the thinnest possible ring. To determine the cause of this radiation efficiency degradation an array of 2 SAR antennas was simulated for multiple antenna spacing. The coupling between the 2 antennas is related to a square root of the antenna spacing, indicating a surface wave coupling between the 2 antennas. The surface waves are probably due to the vertical wall closing in on the radiating outer edge of the SAR antenna. Further investigation of this phenomenon is being conducted.

Since surface waves are excited despite the special design a small perturbation is allowable. This enables the placement of a hybrid coupler in the centre of the antenna. A single layer circular polarized design of a SAR with hybrid feed was introduced in [4].

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## DEVELOPMENT OF OPTIMAL RADIATING STRUCTURES IN THE SURROUNDING OF BIOLOGICAL TISSUES

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Traditionally an antenna is not in the immediate surroundings of an interfering structure. Therefore in most cases this structure is not taken into account in traditional antenna design. However, if the antenna is placed inside or surrounded by a biological tissue, the interaction between antenna and this interfering structure will become very important. In these cases, the antenna is usually very small. The question automatically popping up is: what is the best way to establish a design involving a radiating structure in combination with a biological tissue? This doctoral research tries to give an answer to this question. The aim is to come to an integrated design strategy that allows the designer to decide in which way a radiated structure topology can be chosen and optimized to obtain a minimal exposure risk in the biological tissue.

A very important aspect is the numerical simulation of the problem. It must solve the radiating structure as well as the exposure problem. At present, different numerical techniques exists: Method of Moments (MoM), Finite Element Method (FEM) and Finite-Difference Time-Domain (FDTD), ... However, every method has his advantages and disadvantages. The MoM is the best method to solve accurate wire and curved antenna structures but is not suited to model large complex inhomogeneous biological structures. The FDTD method is ideal for solving large complex inhomogeneous structures, but it is less accurate in solving wire and curved antennas. A hybridization of the FDTD with the frequency domain MoM can be the solution 0. In this doctoral research an FDTD-MOM "tool" will be constructed and used in combination with measurements and some other commercial tools to solve two problems:

I. In view of new technology, such as UMTS (Universal Mobile Telecommunication System) it is assumed that mobile transmitters will be places more frequently in the lower stomach area. At present, no applicable data is available for risk assessment of Specific Absorption Rates (SAR) that will be expected in related tissues. In the research envisaged, exposure of organs near transmitters will be investigated and a new very small antenna will be constructed which gives less exposure risk, low SAR and temperature, to these organs. In the design, mainly microstrip antennas with circular polarization will be investigated.

II. In recent years an increasing trend towards sensor systems in medical health surveillance has occurred, which is further enhanced by the application of wireless communication. In the research presented, the exposure risk of a communication link between a transmitter inside and a receiver outside the body will be investigated. A small radiating structure will be constructed and a frequency dependent model for tissues will be incorporated to investigate the risk and possibility of direct digital pulsed communication.

### REFERENCE

[1] Mohab A. Mangoud, Raed A. Abd-Alhameed, Peter E. Excell, "Simulation of Human Interaction with Mobile Telephones Using Hybrid Techniques Over Coupling Domains", IEEE Trans. Microwave Theory Tech., vol. 48, pp. 2014-2021, Nov. 2003.

# INHOUDSOPGAVE

In het kader van het Vlaams Kennis- en Cultuurforum coördineert de Koninklijke Vlaamse Academie van België voor Wetenschappen en Kunsten jaarlijks tot 25 wetenschappelijke bijeenkomsten, ook contactfora genoemd, in de domeinen van de natuurwetenschappen (inclusief de biomedische wetenschappen), menswetenschappen en kunsten. De contactfora hebben tot doel Vlaamse wetenschappers of kunstenaars te verenigen rond specifieke thema's.

De handelingen van deze contactfora vormen een aparte publicatiereeks van de Academie.

## URSI FORUM 2003 RADIO SCIENCE ON THE MOVE (18 december 2003)

Het doel van het " URSI FORUM 2003 Radio Science on the move", dat in 2003 voor de elfde maal (telkens onder een verschillende vorm) georganiseerd wordt, bestaat erin een kruisbevruchting van ideeën over alle universitaire en taalgrenzen binnen Europa tot stand te brengen in alle domeinen die met o.a. met mobiele communicatie te maken hebben. De gelegenheid wordt geboden aan Europese onderzoekers, die actief zijn op de onderzoeksdomeinen van URSI, hun werk voor te stellen aan een jongere gemeenschap, die de kans krijgt om niet alleen na de formele voorstellen, maar ook tijdens de pauzes en de lunch de voorgestelde onderwerpen meer in detail te bespreken en de nodige contacten kunnen leggen om vruchtbare samenwerkingen tot stand te brengen, wat. Het onverhoopte succes van de vorige uitgaven, vooral sinds het openstellen naar Europa in 2000 bewijst dat zulk een informatie- en discussieforum een noodzaak is. Bovendien vormen de proceedings van dit forum een extensief overzicht van het onderzoek op draadloze communicatie aan de universitaire instellingen uit België en de ons omringende landen.
Dit jaar staat werd het opzettelijk breed van opzet gehouden en laten we alle aspecten, die behoren tot de onderzoeksgebieden of "commissies" van URSI aan bod komen, met name (originele Engelse benamingen):

A : Electromagnetic metrology
B : Fields and waves
C : Radio-Communication systems & signal processing
D : Electronics and photonics
E : Electromagnetic noise and interference
F : Wave propagation and remote sensing
G : Ionospheric radio and propagation
H : Waves in plasma
J : Radio astronomy
K : Electromagnetics in biology and medicine