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Techniques champ proche : du système antenne à l'application sur le terrain

Mohammed Serhir

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Université Paris-Saclay

HABILITATION À DIRIGER DES RECHERCHES

préparée à CentraleSupélec

présentée et soutenue publiquement par

Mohammed SERHIR

le 10 octobre 2019

TECHNIQUES CHAMP PROCHE: DU SYSTÈME ANTENNAIRE À L'APPLICATION SUR LE TERRAIN

Jury

M. Jean-Yves Dauvignac,	Professeur à l'Université de Nice-Sophia Antipolis	Rapporteur
M. Raphael Gillard,	Professeur à l'INSA de Rennes	Rapporteur
M. Sébastien Lambot,	Professeur à l'Université Catholique de Louvain	Rapporteur
M. Dominique Lesselier,	Directeur de Recherche CNRS, L2S	Examineur
M. Lionel Pichon,	Directeur de Recherche CNRS, GeePs	Examineur
M. Jean-Michel Geffrin ,	Ingénieur de Recherche CNRS, Institut Fresnel	Examineur
M. Régis Guinvarc'h,	Professeur à CentraleSupélec	Invité

AVANT PROPOS

Ce document synthétique, à mi-chemin entre la synthèse et le recueil de publications, donne un aperçu de mes activités de recherche entreprises depuis 2008. Cela me permet de donner un résumé détaillé de mes travaux en m'appuyant sur les publications les plus représentatives dans mes différents axes de recherche. Ce tapuscrit est composé de 5 chapitres.

- Le premier chapitre est un curriculum vitæ complet décrivant les étapes principales de mon parcours, mes activités d'enseignement, d'encadrement et de recherche.
- Le deuxième chapitre donne un aperçu de mes activités de recherche sur la modélisation comportementale d'antennes développées après ma thèse (soutenue en 2007). J'ai dressé un portrait de cet axe de recherche et précisé mes contributions les plus représentatives. Les articles majeurs sont produits à la fin du chapitre pour éclairer le lecteur et apporter les détails complémentaires.
- Le troisième chapitre aborde mes apports dans le thème de la caractérisation d'antennes en champ proche. Avant de présenter les spécificités de mes travaux autour de la méthode matricielle pour le calcul du champ lointain et la caractérisation d'antennes dans le domaine temporel, j'établis un état des lieux où j'exprime ma perception de cet axe de recherche avant de souligner mes apports dans chaque point soulevé. Les développements théoriques ne sont pas intégrés dans le texte pour préserver une certaine sobriété dans la présentation du sujet. Néanmoins, je mets à la disposition du lecteur les articles majeurs, généreux en détails, pour apporter les compléments nécessaires et les preuves suffisantes des conclusions avancées. À la fin de ce chapitre, je présente une description succincte des activités collaboratives tissées le long de mon parcours de recherche.
- Le quatrième chapitre développe mes travaux sur l'imagerie micro-onde appliquée au radar de sol. J'y présente les conclusions de la modélisation du problème direct par la méthode DGTD et la résolution du problème inverse par Linear Sampling Method. Les applications étudiées sont la détection des systèmes racinaires des arbres et la localisation de câbles souterrains en situation réelle. Je propose aussi un bref aperçu sur mes activités autour du développement antennaire dédié au radar de sol. Les travaux de recherche sur l'imagerie du proche sous-sol sont pluridisciplinaires. Ils représentent le point de convergence de mes activités précédentes. Les résultats et les conclusions de ce chapitre sont appuyés par le choix des publications représentatives de chaque activité entreprise.
- La conclusion de ce travail de synthèse est abordée dans le chapitre 5 où je développe mes perspectives et mon projet de recherche pour les cinq années à venir.

*à
Claire et Sara,
lumières de ma vie*

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Chapitre 1

Parcours Professionnel

*« [Si l'Homme] se vante, je l'abaisse.
S'il s'abaisse, je le vante Et le contredit
toujours Jusqu'à ce qu'il comprenne
Qu'il est un monstre
incompréhensible. »*

Blaise Pascal

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1.1 Fiche synthétique

MOHAMMED SERHIR - 38 ans

Professeur adjoint à CentraleSupélec

Chercheur au laboratoire *Génie Électrique et Électronique de Paris* (GeePs)

Formations et diplômes

2004-2007	Thèse de doctorat à l'IETR soutenue le 12 décembre 2007
2003-2004	D.E.A. Électronique, Conception & Architecture
2000-2003	Élève à l'École Mohammadia d'Ingénieurs, Rabat, diplômé en juillet 2003

Activités d'enseignement

2018-2019	310 heures à CentraleSupélec Cursus Supélec, CentraleSupélec, Système Communicants et Objets Connectés Département Électromagnétisme, majeure Électromagnétisme et Communication
2012-2018	220 heures par an, cursus ingénieur Supélec Département Électromagnétisme, majeure Électromagnétisme et Communication
2004-2007	192 heures sur 3 ans à l'INSA de Rennes, Dép. Système Réseau de Communication

Activités de recherche

Mes travaux de recherche au sein du GeePs s'insèrent dans les thématiques « ondes et milieux complexes » du pôle **PIEM** (Physique et Ingénierie de l'ÉlectroMagnétisme) et « sondage et imagerie des sols » du pôle **CEMES** (Contrôle Électromagnétique des Matériaux Et des Structures). Dans le pôle **PIEM** mes travaux s'articulent autour de la conception, l'optimisation et la caractérisation d'antennes, qu'il s'agisse d'antennes spirales en collaboration avec la chaire SONDRRA ou d'antennes pulsées ultra large bande dédiées aux radars de sol. Mes travaux sur la caractérisation d'antennes visent à développer des méthodes de mesure en champ proche pour proposer des extensions aux moyens de test du laboratoire par l'exploitation de mesures sur des surfaces arbitraires où l'échantillonnage du champ proche est irrégulier. Dans le pôle **CEMES**, mes travaux sont axés sur l'analyse électromagnétique des potentialités de l'approche multi-statique pour un radar de sol. Ces travaux s'insèrent dans la *COST Action TU1208* (2013-2017) et plus particulièrement dans des problématiques d'instrumentation et de modélisation électromagnétiques associées au *Ground Penetrating Radar*. Mes activités dans ce domaine ont pour objectif d'étudier les problèmes directs et inverses dans le contexte GPR pour des configurations bi/multi-statiques. Actuellement, nous développons les outils pour la caractérisation électromagnétique non destructive des systèmes racinaires d'arbres en situation réelle, à des fins d'ingénierie civile et d'estimation de biomasse. Un partenariat s'inscrivant dans l'Institut Risegrid est en cours de consolidation et de développement sur la modélisation du réseau de distribution basse tension pour la communication par courant porteur en ligne à destination des compteurs intelligents (Linky).

Publications

Articles dans des revues internationales à comité de lecture :	22
Chapitres d'ouvrages scientifiques :	2
Brevet :	1
Actes de colloques internationaux avec comité de lecture :	37
Colloques nationaux avec actes :	14
Rapports de recherches :	8

Activités d'animation scientifique et/ou d'intérêt collectif

Responsable de la mise en oeuvre opérationnelle des systèmes de mesure en environnement anéchoïque

Management Committee Substitute à l'action COST 1208 Civil Engineering Applications of GPR

Responsable scientifique de contrats industriels

Rapporteur pour *IEEE TAP*, *IEEE AWPL*, *IET Microwave Antenna & Propagation*, *PIER*, *Radio Science*, *Sensors*, *Construction and Building Materials*

1.2 Curriculum Vitae

MOHAMMED SERHIR né le 8 janvier 1981 (38 ans)

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Professeur adjoint à CentraleSupélec
Chercheur au laboratoire *Génie Électrique et Électronique de Paris (GeePs)*

FORMATION

2004-2007	Thèse à PIETR-INSA (INSA de Rennes, CNRS UMR 6164, Univ. Rennes 1) Groupe Antenne & Hyperfréquence, sous la direction de P. Besnier et M. Drissi
Décembre 2007	Doctorat en Électronique, mention <i>Très honorable avec les félicitations du jury</i>
2003-2004	D.E.A. en Électronique, INSA de Rennes Option Conception & Architecture
Juillet 2004	Diplôme de DEA
2000-2003	Élève à l' École Mohammadia d'Ingénieurs , Rabat
Juillet 2003	Diplôme d' ingénieur EMI
1998-2000	Lycée Mohamed V, Casablanca - Classes préparatoires MPSI
1997-1998	Baccalauréat, série sciences mathématiques

EXPÉRIENCES PROFESSIONNELLES

depuis 2014	Professeur adjoint à CentraleSupélec (département Électromagnétisme) Chercheur au GeePs (CNRS UMR 8507, CentraleSupélec, Université Paris-Sud, Univ Paris-Saclay, Sorbonne-Université) Pôle PIEM (Physique et Ingénierie de l'ÉlectroMagnétisme) Pôle CEMES (Contrôle Électromagnétique des Matériaux Et des Structures) Porteur de plusieurs projets (200k€) avec : chaire SONDRRA, EDF R&D, Rincet BTP, Spirtech, MBDA, Detect Réseaux
2010 - 2014	Professeur assistant à Supélec (département Électromagnétisme) Chercheur au L2S (CNRS UMR 8506, CentraleSupélec, Univ Paris-Saclay) Département DRE , Département de Recherche en Électromagnétisme
2009 - 2010	Post-Doctorat au L2S , Laboratoire des Signaux et Systèmes DRE , Financement DIGITEO, sous la direction de N. Ribière-Tharaud Sujet : « Investigation sur la caractérisation d'antennes par des techniques de champ proche dans le domaine temporel »
2008 - 2009	Post-Doctorat à l' Institut Fresnel , Marseille (CNRS UMR 6133) Équipe : HIPE , Financement ANR CESAR, sous la direction de A. Litman Sujet : « Contrôle radiofréquence de l'écoulement de l'eau dans le sol et de l'absorption racinaire »
2004 - 2007	Allocataire de recherche (MENRT) à l' INSA de Rennes Enseignant vacataire à INSA de Rennes, dép. Système Radio et Communication
févr.-juil. 2004	Stage de D.E.A. à l' IETR-INSA de Rennes , CNRS UMR 6164 Sujet : « Évaluation d'un procédé de mesures d'antennes longues dans une base champ proche sphérique », sous la direction de P. Besnier
mars-juin 2003	Projet de fin d'études au Laboratoire IETR - INSA de Rennes Sujet : « La caractérisation champ proche pour une meilleur optimisation des performances des antennes intégrées », sous la direction de M. Drissi

LANGUES

Anglais	Courant
Arabe	Bilingue

1.3 Titres universitaires - Diplômes

Doctorat de l'Institut National des Sciences Appliquées de Rennes

Année	2007
Spécialité	Électronique, Électromagnétisme, Hyperfréquence
Laboratoire	Institut d'Électronique et Télécommunication de Rennes (IETR)
Titre	Développement de modèles de rayonnement électromagnétique à partir d'une mesure en champ proche sphérique
Mention	<i>Très honorable avec félicitations du jury</i>

Composition du jury

Rapporteurs :	Laurent Chusseau, Directeur de recherche CNRS, IES, Montpellier Luis Jofre, Professeur à l'Université Politècnica de Catalunya
Examineurs :	Abdelhak Ziyat, Professeur à l'Université Mohammed Premier, Oujda François De Daran, Ingénieur de recherche CEM, Valeo
Président :	Jean Charles Bolomey, Professeur à l'Université Paris-Sud, L2S
Encadrants :	Philippe Besnier, Chargé de recherche CNRS, HDR, IETR Mhamed Drissi, Professeur à INSA Rennes, IETR

Résumé :

Mes travaux de thèse portent sur le développement d'une méthodologie de modélisation destinée à reproduire le rayonnement électromagnétique d'antennes à partir d'une mesure en champ proche sphérique. Cette étude a été abordée sur le plan du développement des méthodes, de l'élaboration d'un modèle comportemental, et de la mise en œuvre des techniques numériques. Il s'agit d'exploiter l'ensemble des données issues d'un banc de caractérisation sphérique pour produire un modèle d'antennes composé de sources élémentaires (dipôles électriques et magnétiques) qui rayonnent de façon équivalente en champ lointain, mais aussi, dans une certaine mesure, en champ proche. L'une des applications visées consiste à établir l'image d'une antenne par le biais de la théorie des images ce qui permet d'exploiter la mesure en champ proche dans un contexte normatif en compatibilité électromagnétique (CEM).

Mots-clés : Modélisation d'antennes, développement en harmoniques sphériques, champ proche, champ lointain, identification modale, traitement de signal, CEM, théorème de translation et de rotation des harmoniques sphériques, mesure d'antenne.

Diplôme d'Études Approfondies

Année	2004
Spécialité	Électronique
Option	Conception et Architecture
Établissements	INSA de Rennes - Université Rennes I - Supelec

Diplôme d'Ingénieur

Année	2003
Spécialité	Génie électrique
Option	Électronique et Communication
Établissements	École Mohammadia d'Ingénieurs (EMI), Rabat

Baccalauréat

Année	1998
Série	sciences mathématiques, Casablanca

1.4 Bilan d'activités : Enseignement

Cursus ingénieur Supélec - Département Électromagnétisme

Période 2012 - 2019 (220 heures équivalent TD par an)

- TD, **Champs et Propagations** - 1^{ère}A cursus Supélec - 6 heures
- Cours, TD, TP, **Électronique Haute Fréquence** - 2^{ème}A cursus Supélec - 60 heures
- Cours, TD, **Propagation Guidée** - 3^{ème}A cursus Supélec - 20 heures
- Cours, TD, **Circuits Hyperfréquences** - 3^{ème}A cursus Supélec (Master **CAT**) - 36 heures
- Encadrement de projets 2^{ème}A et 3^{ème}A (CEI et PEI) - 60 heures
- Membre de l'équipe pédagogique du cours d'Électronique Haute Fréquence du cursus de Supélec
- Membre du Jury de passage de la deuxième année du cursus ingénieur Supélec
- Membre du Jury d'admission sur titre du cursus ingénieur Supélec
- Intervenant pour l'association **Espérance en Béton** pour des visites de vulgarisation pour lycéens
- Participation et animation des **cafés pédagogie** dont l'objectif est de partager les retours d'expériences sur les bonnes pratiques d'un enseignement efficace et pédagogique.
- Formation continue (15 heures/an) : «Introduction aux antennes», «Techniques champ proche»

Cursus ingénieur CentraleSupélec - Systèmes Communicants et Objets Connectés

Période 2018 - 2019 (167 heures équivalent TD)

- Responsable du pôle projet « **Ingénierie Électromagnétique** » - 1^{ère}A - 40 heures
- Animateur des ateliers PPI (Pratiques Pour l'Ingénieur) - 1^{ère}A - 36 heures
- Co-porteur de la ST2 de la dominante SCOC - 1^{ère}A - 15 heures
- Porteur d'un enseignement d'intégration 1^{ère}A - 27 heures
- TD Électromagnétisme 1^{ère}A - 33 heures
- TD Antennes 1^{ère}A - 6 heures
- Encadrement de projets 1^{ère}A - 10 heures

Descriptif sommaire des enseignements

Je participe aux travaux dirigés du cours d'**Électromagnétisme** pour les étudiants de 1^{ère} année du cursus de CentraleSupélec. Ce cours traite du rayonnement, de la propagation des ondes électromagnétiques en espace libre et en milieu diélectrique, des réflexions et des réfractions, de l'optique de Fourier et de l'imagerie. Pendant leur formation, les élèves de CentraleSupélec doivent développer leur projet (1^{ère} et 2^{ème}) au sein des 25 pôles projets proposés dans l'École. Je suis porteur du pôle **ingénierie électromagnétique** (40 étudiants/semestre). Pour sa première occurrence (2018-2019), nous avons proposé des sujets sur le contrôle non-destructif, la recharge de véhicule sans contact et la caractérisation des débris spatiaux. Ces projets sont jumelés avec des **ateliers PPI** (Pratique Pour l'Ingénieur) que j'anime avec un intervenant extérieur pour aider les élèves à développer des aptitudes de travail en équipe, de communication non-violente, de gestion de projet, communiquer la valeur ajoutée.

Je co-porte, avec Jacques Antoine, la séquence thématique 2 associée à la dominante SCOC (systèmes communicants et objets connectés). Il s'agit d'organiser/programmer des interventions/séminaires données par des acteurs majeurs du domaine des télécommunications à destination des élèves de 1^{ère} année (100 élèves) afin de les sensibiliser aux enjeux présents et futurs de la thématique. Ces interventions sont accompagnées de cours spécifiques dont la matérialisation est faite pendant l'enseignement d'intégration (projet d'une semaine) qui clôt la séquence. Je suis porteur d'un enseignement d'intégration en partenariat avec Nokia. En 2^{ème} année du cursus ingénieur de Supélec, je suis responsable du cours d'**électronique haute fréquence** à destination des étudiants qui suivent la voie par apprentissage. J'y traite les lignes de transmission, les paramètres S, les circuits d'amplificateurs et d'oscillateurs, le bruit dans les montages électroniques. Le volet pratique de ce cours est traité pendant les TPs sur machine en utilisant le logiciel ADS (Analog Design System). En 3^{ème} année du cursus ingénieur de Supélec j'interviens dans le cours de **propagation guidée**. J'y aborde l'analyse des lignes de transmission pour l'étude des coupleurs directifs, des filtres et des résonateurs en technologie imprimée. Aussi, je suis responsable du cours **circuits hyperfréquences** où je propose une vue d'ensemble des composants micro-ondes réciproques et non-réciproques. On analyse les limitations qui impactent le fonctionnement des blocs fonctionnels.

Dans le cursus ingénieur Supélec, les élèves doivent valider un projet de fin d'études qui occupe 80 demi-journées. J'ai encadré en moyenne deux groupes d'élèves par an depuis 2011. Ces projets peuvent être des conventions d'études industrielles (**CEI**) caractérisées par un aspect contractuel qui engage le porteur du projet auprès d'un client (industriel) pour encadrer les élèves afin de réaliser une mission technique soumise à un cahier des charges clairement identifié.

1.5 Bilan d'activités : Encadrement

Encadrement de doctorants

1.	<p>2018-2021 Encadrement Financement Sujet</p>	<p>Arbia Haded, soutenance prévue en octobre 2021 directeur : D. Picard (50%), co-encadrant : M. Serhir 50% CIFRE ANRT, EDF R&D, dép. MIRE, École doctorale EOBE de Paris-Saclay Modélisation du réseau de distribution basse tension dans la bande de fréquence utilisée pour la communication par courant porteur en ligne en bande étroite</p>
2.	<p>2017-2020 Encadrement Financement Sujet Publications</p>	<p>Abderrahmane Aboudourib, soutenance prévue en octobre 2020 directeur : M. Serhir 50% (dérogation 2017), co-encadrant : D. Lesselier (50%) Contrat doctoral de CentraleSupélec, École doctorale EOBE de Paris-Saclay Imagerie en champ proche de systèmes racinaires en contexte de génie civil et estimation de biomasse [CI.35] [CI.36] [CI.37]</p>
3.	<p>2017-2020 Encadrement Financement Sujet Publications</p>	<p>Peipei Ran, soutenance prévue en décembre 2020 directeur : D. Lesselier (55%), co-encadrant : M. Serhir 45% China Scholarship Council, École doctorale STIC de Paris-Saclay Resonant micro-structured 3-dimensional antenna systems – from mathematical analysis to laboratory-controlled experimentation [CI.30] [CI.32] [CI.33] [CI.34]</p>
4.	<p>2014-2017 Encadrement Financement Sujet Publications Situation actuelle</p>	<p>Xiang Liu, soutenue le 13 décembre 2017 directeurs : M. Lambert (20%), L. Pichon (20%), co-encadrants : M. Serhir 35%, A. K. Ntichi (25%) Fondation Supélec, École doctorale EOBE de Paris-Saclay Contrôle non destructif du sol et imagerie d'objets enfouis par des systèmes bi- et multi-statiques : de l'expérience à la modélisation [RI.19] et [CI.17] [CI.18] [CI.19] [CI.20] [CI.26] [CI.27] [CI.28] Ingénieur de recherche à Valeo, Créteil (94)</p>
5.	<p>2014-2017 Encadrement Financement Sujet Publications : Situation actuelle :</p>	<p>Mariam Ait Ou Kharraz, soutenue le 5 mai 2017 directeur : D. Picard (50%), co-encadrant : M. Serhir 50% CIFRE ANRT, EDF R&D, dép. MIRE, École doctorale EOBE de Paris-Saclay Caractérisation du réseau de distribution basse tension dans la bande de fréquence utilisée pour la communication par CPL en bande étroite [RI.21] et [CI.24] [CI.25] [CI.29] Ingénieur de recherche à EDF, Palaiseau (91)</p>
6.	<p>2012-2015 Encadrement Financement Sujet Publications Situation actuelle</p>	<p>Mohamed Farouq, soutenue le 28 septembre 2015 directeur : D. Picard (40%), co-encadrant : M. Serhir 60% Allocation de Paris-Sud, École doctorale STITS de Paris-Sud Transformation de fronts d'onde par des méthodes matricielles [RI.12], [RI.13], [RI.15] et [CI.14], [CI.15] [RI.16], [CI.21], [CI.23] Post-doctorant au L2S, Gif-sur-Yvette (91)</p>

Encadrement de stagiaires M1R et M2R

1.	<p>2012 Formation Encadrement Sujet</p>	<p>Mohamed Farouq, stage M2R, durée : 5 mois Master IST, « Composants et Antennes pour les Télécommunications », Univ. Paris-Sud M. Serhir (70%), D. Picard (30%) Développement d'une procédure de correction de sonde pour la mesure du champ électromagnétique au voisinage de la source du rayonnement</p>
2.	<p>2014 Formation Encadrement Sujet</p>	<p>Salwa Himya, projet de fin d'études, durée : 4 mois École Mohammadia d'Ingénieurs, Rabat M. Serhir (100%) Conception et pilotage d'un radar MIMO à pénétration de sol avec reconstruction 3D des objets enfouis</p>

2016	Ding Tao, stage M1R , durée : 3 mois
Formation	Polytech Paris-Sud, cycle ingénieur (2 ^{ème} A), Électronique Énergie Système
3. Encadrement	M. Serhir (100%)
Sujet	Création et validation d'un code de pilotage d'un robot 5 axes pour la mesure des antennes
2016	Gonçalves Lima Paulo Janderson, stage M1R , durée : 2 mois
Formation	Supélec, cycle ingénieur (2 ^{ème} A), Gif-sur-Yvette
4. Encadrement	M. Serhir (100%)
Sujet	Développement et comparaison des techniques d'imagerie dédiées à la détection d'objets enfouis par les radars de sol
2019	Leone Almendro Mendes, stage M1R , durée : 6 semaines
Formation	Supélec, cycle ingénieur (2 ^{ème} A), Gif-sur-Yvette
5. Encadrement	M. Serhir (100%)
Sujet	Développement et mise en œuvre de routines de pilotage d'un système robotisé 3 axes dédié à la mesure d'antennes en coordonnées sphériques

Responsabilité scientifique dans le cadre de Convention d'Étude Industrielle (CEI)

2011	Étude pour Rincet BTP , durée : 2 mois
Formation	3 ^{ème} année du cursus Supélec (M2)
1. Porteurs	N. Ribière-Tharaud (60%), M. Serhir (40%)
Sujet	Étude et optimisation hyperfréquence du radar GPR Euradar
2012	Étude pour Rincet BTP , durée : 2 mois
Formation	3 ^{ème} année du cursus Supélec (M2)
2. Porteurs	F. Jouvie (50%), M. Serhir (50%)
Sujet	Expérimentation et comparaison de deux antennes pour des radars de sol
2017	Étude pour Spirtech , durée : 2 mois
Formation	3 ^{ème} année du cursus Supélec (M2)
3. Porteur	M. Serhir (100%)
Sujet	Étude de faisabilité, prototypage et expérimentation d'un lecteur de cartes sans contact ISO 14443 adapté au péage d'autoroute urbaine
2018	Étude pour MBDA , durée : 2 mois
Formation	3 ^{ème} année du cursus Supélec (M2)
4. Porteurs	M. Serhir (50%), D. Picard (50%)
Sujet1	Réalisation de carte d'émission avec un amplificateur 10W en CW
Sujet2	Réalisation d'un modem UHF

1.6 Activités d'animation scientifique et/ou d'intérêt collectif

- Responsabilité de la mise en œuvre opérationnelle des systèmes de mesure dans les chambres anéchoïques du GeePs, participation au groupe interdisciplinaire du GDR ondes sur la mesure micro-onde pour l'inter-comparaison des mesures d'antennes.
- European cooperation in science and technology : « Management Comity Substitute » à l'**action COST 1208** : Civil Engineering Applications of Ground Penetrating Radar (**2013 – 2017**) (www.gpradar.eu/). Nos travaux s'insèrent dans : WG1 (Novel GPR Instrumentation) et WG3 (electromagnetic methods for near-field scattering problems by buried structures ; data processing techniques).
- Membre du comité d'organisation de **NUMELEC 2017** : 9^{ème} conférence européenne sur les méthodes numériques en électromagnétisme 15-17 novembre 2017 Paris.
- Membre du comité d'organisation de **MMS 2017** : 17^{ème} Mediterranean Microwave Symposium, 28-30 novembre 2017, Marseille.
- Rapporteur pour des éditeurs de journaux :
 - IEEE Transaction on Antenna and Propagation
 - IEEE Antenna and Wireless Propagation Letter
 - IET Microwave Antenna & Propagation
 - Progress In Electromagnetic Research

- Radio Science, Sensors
- Elseviers Journal, Construction and Building material
- International Journal of RF and Microwave Computer-Aided Engineering.
- Évaluation doctorale et participation à des jurys de thèses en tant qu'examinateur :
 - Thèse de **M. A. Moutaouekkil**, Université Mohamed Ier, Oujda le 24 juin 2013
 - Thèse de **M. Farouq**, Université Paris-Sud, Gif-sur-Yvette, le 28 septembre 2015
 - Thèse de **M. Ait Ou Kharraz**, Université Paris-Saclay, Gif-sur-Yvette, le 5 mai 2017
 - Thèse de **X. Liu**, Université Paris-Saclay, Gif-sur-Yvette, le 13 décembre 2017
 - Thèse de **S. Ouedraogo**, Université Paris-Saclay, Gif-sur-Yvette, le 9 janvier 2018
- Responsabilité scientifique de contrats industriels : recherche partenariale
 - **Rincent BTP**, Caractérisation des radars pour le contrôle de la chaussée [Rap.1] [Rap.2] [Rap.3].
 - **Spirtech**, étude de faisabilité et prototypage d'un lecteur RFID pour une longue portée [Rap.5].
 - **MBDA**, développement de cartes de transceiver pour les plages UHF et ISM [Rap.6] et [Rap.7].
 - **Detect Réseaux**, développement d'antennes ULB pour un radar multi-statique [Rap.8].
 - **Sondra, 2010 - 2019**, expérimentations et validations en chambre anéchoïque.
- Collaborations internationales
 1. Collaboration avec l'**Université Mohamed I^{er}** Oujda (A. Ziyat). Dans ce cadre, le doctorant **M. A. Moutaouekkil (2013)**, a fait un séjour de 6 mois au laboratoire pour valider expérimentalement sa méthode de correction de sonde pour la caractérisation des éléments rayonnants en champ proche réactif ([RI.5], [CI.6]).
 2. Collaboration avec **National University of Singapore** (K. Mouthaan) et **SONDRA** (R. Guinvarc'h) dans le cadre des thèses de **H. R. Fang (2014)** et **M. R. Balakrishnan (2014)**. Cette collaboration concerne la conception et la caractérisation d'antennes spirales multi-polarisées ([RI.11], [CI.11], [CI.13]).
 3. Collaboration avec **University of Electronic Science and technology of China** (S. Xiao) et le **L2S** (D. Lesselier) dans le cadre de la thèse de **H. L. Tu (2017)** ([RI.14], [CI.22], [CI.30]). Cette collaboration s'intéresse au rôle de la résonance électromagnétique dans la super-résolution.
- Collaborations nationales
 1. **GDR ondes** pour la campagne de recensement des chambres anéchoïques menée par J. M. Geffrin (Institut Fresnel) pour constituer le Réseau des Utilisateurs de Chambres anéchoïques Hyperfréquence (<http://www.fresnel.fr/rca/>).
 2. Collaboration avec l'équipe **HIPE** de l'**Institut Fresnel** durant mon post-doctorat (2009) [RI.4], [CI.4] : cette collaboration inclut le thème de radar à pénétration de sol en partenariat avec C. Eyraud et Detect Réseau en 2019 [Rap.8]
 3. Collaboration avec D. Lesselier du (**L2S**), L. Pichon et L. Bernard du (**LGEP**) dans le cadre de la thèse de **M. Benhamouch (2009 - 2012)** [RI.6] [RI.8] : il s'agit d'exploiter le retournement temporel dans la localisation d'objets en espace libre en se basant sur une cartographie d'énergie.
 4. Collaboration avec D. Lesselier du **L2S** dans les thèses de **Peipei Ran (2017 - 2020)** et **Abderrahmane Aboudourib (2017 - 2020)** ([CI.30] à [CI.37])
 5. Collaboration avec **EDF R&D** (P. Jensen, C. Lavenue, A. Audebert, A. Jeandin) du département MIRE (Mesures et systèmes d'informations pour les réseaux électriques intelligents) dans le cadre de la thèse **M. Ait Oukharraz (2014 - 2017)**. Les travaux développés s'articulent autour de la caractérisation du réseau de distribution basse tension dans la bande fréquentielle utilisée pour les communications par courants porteurs en ligne en bande étroite [9 kHz – 500 kHz]. Cette thèse s'insère dans le cadre de l'**Institut Risegrid** (Research Institute for Smarter Electric Grids) et les travaux se poursuivent dans le cadre de la thèse CIFRE de **Arbia Haded (2018 - 2021)** [RI.21], [CI.24], [CI.29].
 6. Collaboration avec S. Pellevrault de (**Rincent BTP R&D**) pour l'optimisation d'antennes pour le radar de sol (2011 - 2014) [B.1], [Rap.1], [Rap.2], [Rap.3].
 7. Collaboration avec C. Mirry et E. Amador de **EDF R&D site Moret-sur-Loing** : localisation des jonctions des câbles souterrains (2014) [Rap.4].

1.7 Bilan d'activités : Recherche

Introduction

Les travaux de recherche qui figurent dans ce bilan ont été sélectionnés pour souligner les points essentiels et les lignes de force de mon parcours de recherche depuis ma thèse soutenue en 2007. Sur le plan institutionnel, mes travaux de recherche se sont déroulés au Département de Recherche en Électromagnétisme du Laboratoire des Signaux et Systèmes (L2S) et se sont poursuivis à partir de 2015 au laboratoire de Génie Électrique et Électronique de Paris (GeePs) ; anciennement appelé LGEP (Laboratoire de Génie Électrique de Paris), suite à la fusion de Centrale Paris et Supélec. La construction de cette synthèse reprendra mes contributions dans (i) la modélisation du rayonnement des antennes d'un point de vue méthodologique, des applications et des ouvertures, (ii) le développement instrumental pour la caractérisation en champ proche d'antennes et les traitements associés dans le régime temporel et fréquentiel, (iii) l'imagerie du proche sous-sol dans une approche multi-statique et multi-antenne couplée avec l'algorithme d'inversion approprié (LSM) et le développement antenne pour le GPR. Cette synthèse décrira aussi les travaux développés dans le cadre de trois collaborations. La première englobe les échanges avec le Laboratoire Sondra et concerne l'optimisation d'antennes spirales à polarisation contrôlée. La deuxième collaboration se déroule en partenariat avec le département MIRE d'EDF site de Saclay et concerne la caractérisation et la modélisation du réseau basse tension dans la bande de fréquence utilisée par les courants porteurs en ligne en bande étroite. En dernier lieu, je donnerai un aperçu des travaux développés avec le L2S sur l'analyse électromagnétique d'antennes en présence d'une structure périodique.

Modélisation comportementale d'antennes : passerelle entre simulation et mesure

Mes travaux de thèse (soutenue en 2007) ont été développés au sein du groupe Antennes et Hyperfréquences de l'IETR UMR 6164. À l'issue de la thèse, j'ai développé une méthodologie de modélisation destinée à reproduire le rayonnement électromagnétique d'antennes à partir d'une mesure en champ proche sphérique (sphère partielle autour de l'antenne). Cette étude a été abordée essentiellement sur le plan du développement des méthodes et de la mise en œuvre des techniques numériques. Cette représentation idoine de l'antenne est composée de sources élémentaires (dipôles électriques et magnétiques) qui rayonnent de façon équivalente en champ lointain, mais aussi, dans une certaine mesure, en champ proche. Sur le plan numérique, cette méthode permet d'établir l'image d'une antenne par le biais de la théorie des images. Ces travaux ont été validés en collaboration avec l'Institut Fresnel et l'équipe Hyperfréquence Instrumentation Processing et Expérimentation (HIPE) durant mon post-doctorat entre 2008 et 2009. Ce dernier s'inscrivait dans le cadre du projet ANR : CESAR « Contrôle radiofréquence de l'Écoulement de l'eau dans le Sol et de l'Absorption Racinaire ». Ce projet visait à mettre au point et à exploiter une instrumentation micro-onde dédiée à la caractérisation du champ diffracté d'objets placés dans un environnement complexe. Durant ce post-doctorat, j'étais responsable de la mise en œuvre opérationnelle du nouveau système de mesure. J'ai commencé à développer une appétence pour le développement instrumental et un intérêt naissant pour l'étude de la diffraction électromagnétique. Cela concerne la modélisation électromagnétique du problème direct et la résolution du problème inverse de diffraction électromagnétique. J'ai réalisé que ma méthode de modélisation comportementale pourrait apporter une meilleure connaissance de l'antenne émettrice (dipôles équivalents) et du champ incident (champ proche) dans l'analyse des problèmes directs et inverses. L'antenne est remplacée par une constellation de sources élémentaires qui procure une description rigoureuse de la fonction de Green du problème étudié. Ce travail concrétisa la passerelle entre mes travaux de thèse et ceux de mon post-doctorat aussi bien que le passage entre l'expérimentation et la modélisation électromagnétique du problème. Cette collaboration avec l'Institut Fresnel et l'équipe HIPE se poursuit avec C. Eyraud dans le cadre de mes activités actuelles autour du radar de sol (en collaboration avec Detect Réseau).

Techniques et instrumentations pour la caractérisation en champ proche

Mes activités de recherche associées aux techniques de mesure d'antennes en champ proche ont commencé durant ma thèse et se sont poursuivies après mon recrutement à Supélec. Mes activités actuelles concernent la mise en place de nouvelles méthodologies de mesure dans le domaine fréquentiel d'une part et le développement instrumental associé à la caractérisation d'antenne dans le domaine temporel d'autre part.

Dans le domaine fréquentiel, ma contribution majeure concerne la mise en œuvre d'une méthodologie de caractérisation d'antenne basée sur le développement modal du champ proche mesuré sur des surfaces non canoniques ou avec un échantillonnage irrégulier suivant les trois variables de description de l'espace.

Cette technique se base sur la transformation de fronts d'ondes par des méthodes matricielles dont l'objectif est de déterminer le champ lointain rayonné par une antenne à partir de la mesure du champ proche sur une surface à géométrie quelconque. Le point clé de cette technique est de désolidariser la géométrie de la surface d'acquisition du champ proche (sphérique, cylindrique ou plane) du système de coordonnées dans lequel est exprimé le développement modal du champ.

Il en résulte l'avantage de pouvoir exploiter une surface de mesure quelconque, pour des raisons de compacité, de simplicité de mise en œuvre, de disponibilité du système de mesure ou de suppression d'erreurs de troncature. Cette technique a été mise en œuvre en se basant sur les trois principaux développements modaux (cartésien, cylindrique et sphérique) et a été validée numériquement et expérimentalement pour des surfaces de mesures différentes. Plus particulièrement, deux applications ont été étudiées : la prise en compte d'erreurs de positionnement de la sonde sur des surfaces canoniques de mesure de champ proche et la mesure sur des surfaces arbitraires. Dans la seconde application, il s'agissait d'optimiser la surface de mesure pour en diminuer l'aire et le nombre de points de mesure. La flexibilité de la méthode matricielle assure une extension des modes d'usage des systèmes de mesure existants comme en témoignent les résultats développés durant la thèse de **Mohamed Farouq (2012 - 2015)**.

Dans le cadre de mes activités expérimentales en matière d'instrumentation hyperfréquence, j'ai développé une plateforme micro-onde pour la caractérisation des sources rayonnantes dans le domaine temporel. Il s'agissait de concevoir, dimensionner, automatiser le système de déplacement (robot 5 axes) et élaborer un logiciel d'acquisition pour permettre la caractérisation en champ proche en coordonnées cartésiennes des antennes pulsées. Les transformations champ proche/champ proche et champ proche/champ lointain ont été développées selon le formalisme temporel pour déterminer à partir d'une mesure sur un plan en face de l'antenne, le champ rayonné à d'autres distances. Pour cela, j'ai exploité le formalisme de Green pour l'élaboration de ces transformations et j'ai établi une validation expérimentale de l'ensemble du système.

Le développement de cette plateforme avait confirmé mon intérêt pour la caractérisation impulsionnelle des radars de sol. Ce savoir-faire expérimental pour la mesure d'antennes ultra-large bande est à l'origine de plusieurs collaborations avec des partenaires industriels. Nous en avons eu plusieurs avec Rincet BTP entre 2011 et 2014 en vue de développer des antennes ultra large-bande caractérisées par des réponses impulsionnelles courtes. La plateforme mise en place a été également exploitée pour réaliser les premières mesures de diffraction en espace libre dans le régime temporel où nous avons testé le retournement temporel comme approche de localisation basée sur une cartographie de l'énergie. Ces travaux ont été développés dans le cadre d'une collaboration entre le LGEP et le L2S durant la thèse de M. Benhamouche (2009 – 2012).

Imagerie micro-onde appliquée au radar de sol

Mon intérêt pour l'imagerie du proche sous-sol est né durant mon post-doctorat à l'Institut Fresnel et mes travaux dans cette thématique correspondent à une ouverture sur un nouvel axe de recherche vu à travers le prisme du développement antennaire, la modélisation/caractérisation en champ proche et l'expertise expérimentale.

J'ai identifié, comme première réflexion sur le sujet, des axes potentiellement intéressants pour la communauté du radar de sol (Ground Penetrating Radar). Il s'agit, d'une part, de placer l'analyse expérimentale sur la base de mesures récoltées dans le domaine fréquentiel et d'utiliser les méthodes d'inversions compatibles avec les modes d'acquisition des données (configuration bi ou multi-statique). D'autre part, adopter une approche multi-statique lors des acquisitions accroîtrait substantiellement l'information extraite lors du contrôle non destructif des milieux auscultés.

J'ai organisé mes activités de recherche (depuis 2014) sur l'imagerie micro-onde appliquée au radar de sol autour de deux thèmes : (i) la modélisation électromagnétique du problème direct et (ii) le développement de méthodes pour l'imagerie qualitative. Mes contributions dans ces deux thèmes résident dans (i) l'analyse du problème direct en utilisant la méthode du Galerkin discontinu dans le domaine temporel et (ii) la résolution du problème inverse de diffraction en développant la méthode d'échantillonnage linéaire (Linear Sampling Method). Cette méthode d'inversion qualitative et non itérative est adaptée pour des acquisitions multi-statiques. Elle permet d'inverser les données en temps réel. Pour l'étude du problème direct nous avons étudié l'efficacité de la méthode du Galerkin discontinu dans le domaine temporel pour la modélisation « fullwave » des scènes prenant en compte les interactions entre les antennes émettrices et réceptrices avec le milieu ausculté. Cette phase numérique du travail s'appuie sur des validations expérimentales des scénarios modélisés. Notre contribution réside aussi dans l'amélioration du formalisme classique de la LSM pour prendre en compte le diagramme de rayonnement des antennes émettrices et en exploitant les paramètres S mesurés au niveau des ports des antennes lors des inversions. Ces travaux

ont été développés durant la thèse de **Xiang Liu (2014 - 2017)** intitulée : « Contrôle non destructif du sol et imagerie d'objets enfouis par des systèmes bi- et multi-statiques : de l'expérience à la modélisation ».

Dans le cadre d'un volet applicatif particulier nous étudions les potentialités du Ground Probing Radar pour fournir une cartographie approximative des racines principales des arbres et cela à des fins d'estimation de leur impact potentiel lors de la préparation ou la conduite de travaux de génie civil à proximité. Ces résultats peuvent fournir des informations utiles sur la biomasse et la santé de l'arbre. Ces travaux sont développés dans le cadre de la thèse de **Aberrahmane Aboudourib (2017 - 2020)** : « Imagerie en champ proche de systèmes racinaires : de la modélisation électromagnétique à l'acquisition et au traitement des données en contexte de génie civil et estimation de biomasse ». Ils sont menés dans le cadre d'une coopération entre les laboratoires GeePs et L2S.

Ce projet doit conduire à proposer une preuve de concept solide dans le cadre de la caractérisation non destructive de systèmes racinaires d'arbres en situation réelle. A l'issue de cette thèse, notre objectif est de développer un logiciel d'imagerie en temps réel basé sur une acquisition multi-statique que l'on validera dans des situations contrôlées de laboratoire aussi bien qu'avec des données acquises sur le terrain. Nous avons établi une connaissance consolidée des propriétés diélectriques des structures racinaires et modélisé le problème direct avec la méthode FDTD proposée dans le logiciel open source « GprMax » en considérant un modèle hétérogène du sol. Le traitement des données des radargrammes est établi en se basant sur les techniques de migration de Kirchhoff et du filtrage adapté pour produire une reconstruction 3D des objets enfouis (modèles simplifiés de racines) à partir d'un C-scan. Nous travaillons actuellement sur le développement de la technique du SAR multi-statique comme alternative à la méthode du LSM.

Mes travaux sur le radar de sol s'appuient sur un savoir-faire d'optimisation d'antennes ultra-large bande et une maîtrise de la chaîne expérimentale qui va de l'instrumentation au post-processing des données. J'aborde ce thème de recherche pluridisciplinaire de manière globale en m'appuyant sur la collaboration de numériciens, inverseurs, expérimentateurs afin d'apporter des contributions dans : (i) la modélisation électromagnétique de l'interaction des ondes/antennes avec des milieux inhomogènes (sol), (ii) le développement de techniques d'inversion/imagerie et (iii) la mise au point de l'instrumentation adéquate pour les validations expérimentales.

Nous avons développé des antennes ultra-large bande destinées à alimenter les phases expérimentales de nos travaux sur le contrôle non destructif du sol. Une collaboration avec EDF R&D (site de Moret-sur-Loing) nous a offert la possibilité de faire des mesures sur leur site de test pour une validation en situation réelle des techniques d'imagerie du sous-sol pour détecter la jonction entre deux câbles électriques souterrains. Nos travaux de recherche s'inséraient naturellement dans la COST Action TU1208 (2013 - 2017) (Transport and Urban Development, Civil Engineering Applications of Ground Penetrating Radar) dans les groupes de travail sur l'instrumentation et la modélisation électromagnétiques associées au radar de sol.

Ecosystème institutionnel et collaborations

L'écosystème de CenraleSupélec est propice au développement de collaborations académiques. Nous allons en lister quelques-unes pour illustrer cette dimension. En effet, les collaborations menées ont un point commun, celui de s'appuyer sur mon expertise expérimentale et l'infrastructure de mesure du laboratoire. La plus remarquable des collaborations concerne la caractérisation d'antennes ultra-large bande en partenariat avec des chercheurs du laboratoire franco-singapourien SONDRRA (Supélec, ONERA, Defense Science and Technology Agency (DSTA, Singapour), National University of Singapore). Cette collaboration (2010 - 2016) avec R. Guinvarc'h est articulée autour de la conception, l'optimisation et la caractérisation d'antennes spirales à double polarisation circulaire.

Avec l'University of Electronic Science and Technology of China et le L2S nous avons développé des travaux qui ont pour finalité l'analyse électromagnétique à la fois numérique et expérimentale de la super-résolution dans des milieux structurés basée sur le retournement temporel. Dans ce travail, la visée première était de montrer l'apport d'une structure périodique, de périodicité finie, dans la qualité de localisation de l'élément rayonnant. Les phénomènes de résonance (dans le sens de cavité complexe) de la structure périodique considérée apporteront une meilleure résolution spatiale dans l'identification de l'élément rayonnant parmi d'autres. Cette collaboration se poursuit avec D. Lesselier du L2S dans le cadre de la thèse de **Peipei Ran** « Resonant micro-structured 3-dimensional antenna systems - from mathematical analysis to laboratory-controlled experimentation » financée par le China Scholarship Council pour la période 2017 - 2020.

Une troisième collaboration englobe les travaux avec le département MIRE de EDF R&D autour de la caractérisation du réseau de distribution basse tension dans la bande de fréquentielle utilisée pour la communi-

cation par courants porteurs en ligne en bande étroite. Une première thèse portant sur la caractérisation de la propagation du signal CPL (courants porteurs en ligne) dans les lignes basse-tension en rapport avec le déploiement des compteurs intelligents Linky a été soutenue dans le cadre de cet échange, celle de **M. Ait Ou Kharraz (2014 - 2017)**. Les travaux de cette thèse sont étroitement liés à l'Institut RISEGRID (Research Institute for Smarter Electric Grids). Cette collaboration se poursuit dans le cadre de la thèse de **Arbia Haded (2018 - 2021)** et concerne désormais la mise en œuvre d'un logiciel de modélisation de la propagation des signaux CPL dans un réseau électrique. Ce logiciel incorpore les briques développées dans la thèse de M. Ait Ou Kharraz (modèles de câbles, modèles de charges et de bruit).

1.8 Publications et communications

Articles dans des revues internationales avec comité de lecture

- [RI.22] | M. Teniou, H. Roussel, M. Serhir, N. Capet, G. Piau and M. Casaletti, « Experimental Validation of Tensorial Metasurfaces for the Implementation of Radiating Aperture Field Distributions », *IEEE Transactions on Antennas and Propagation*. doi : 10.1109/TAP.2019.2911328
- [RI.21] | M. Ait Ou Kharraz, D. Picard, M. Serhir, P. Jensen and C. Lavenu, « Measurement Methods of Outdoor Low Voltage Cable Characteristics for Narrowband Power Line Communication », *IEEE Transactions on Power Delivery*. doi : 10.1109/TPWRD.2018.2876248
- [RI.20] | M. Serhir, D. Lesselier, « Wideband reflector-backed folded bowtie antenna for ground penetrating radar », *IEEE Transactions on Antennas and Propagation*, vol. 66, no. 3, pp. 1056-1063, 2018
- [RI.19] | X. Liu, M. Serhir, M. Lambert, « Detectability of junctions of underground electrical cables with a ground penetrating radar : Electromagnetic simulation and experimental measurements », *Construction and Building Materials*, vol. 158, pp. 1099-1110, 2018
- [RI.18] | M. Teniou, H. Roussel, M. Serhir, N. Capet, G. Piau, M. Casaletti, « Tensorial metasurface antennas radiating polarized beams based on aperture field implementation », *International Journal of Microwave and Wireless Technologies*, 10(2), pp. 161-168, 2018
- [RI.17] | A. Cozza, F. Masciovecchio, C. Dorgan, M. Serhir, F. Monsef, D. Lecoine, « A dielectric low-perturbation field scanner for multi-path environments », *IEEE Transactions on Antennas and Propagation*, vol. 65, no. 4, pp. 1978-1987, 2017
- [RI.16] | R. Guinvarc'h, M. Serhir and F. Boust, « A compact dual-polarized 3 :1 bandwidth omni-directional array of spiral antennas », *IEEE Antennas and Wireless Propagation Letters*, vol. 15, pp. 1909-1912, 2016
- [RI.15] | M. Farouq, M. Serhir, D. Picard, « Antenna far-field assessment from near-field measured over arbitrary surfaces », *IEEE Transactions on Antennas and Propagation*, vol. 64, no. 12, pp. 5122-5130, 2016
- [RI.14] | H. Tu, S. Xiao, D. Lesselier, M. Serhir, « Super-resolution characteristics based on time-reversed single-frequency electromagnetic wave », *Journal of Electromagnetic Waves and Applications*, 30 (13), pp. 1670-1680, 2016
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Chapitre 2

Modélisation comportementale, passerelle entre simulation et mesure

« Un vieil étudiant, c'est quelque chose de sinistre; un raté, un médiocre, c'est plus sinistre encore. »

Georges Perec

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2.1 Introduction et positionnement

La modélisation comportementale d'une antenne a pour objectif de proposer une représentation équivalente de celle-ci à partir du champ mesuré sur un domaine d'observation. Cette représentation est exploitable dans un contexte applicatif ou, potentiellement réutilisable dans la prédiction du comportement de l'antenne dans des scénarios de simulations électromagnétiques plus complexes. Avant de présenter mes contributions dans cette thématique je propose de définir le paysage dans lequel se développe ma réflexion.

La modélisation comportementale des structures rayonnantes est présente dans la littérature électromagnétique selon deux schémas de pensées. Dans le premier on aborde les finalités de la modélisation (pourquoi modéliser), tandis que le second s'attarde davantage sur la méthodologie développée pour mettre en oeuvre cette modélisation (comment modéliser). Je vous propose une photographie synthétique mais non exhaustive de l'état de l'art au moment de rédiger cette partie du présent rapport. Dans ma classification, je distingue les finalités de la modélisation, en lien avec un contexte applicatif, des méthodes utilisées pour la définition des modèles. Les exemples choisis sont illustrateurs et résultent de mon interaction avec des domaines plus ou moins proches de mes intérêts scientifiques.

Une représentation équivalente d'une antenne (structure rayonnante) à base de sources élémentaires peut présenter une diversité de finalités. **Le diagnostic des structures rayonnantes** est un premier exemple [ALVAREZ et collab. \[2007\]](#) [ALVAREZ LOPEZ et collab. \[2009\]](#) [NISHINA et CHEN \[2016\]](#) [MIKKI et ANTAR \[2012\]](#). Il s'agit de traduire le champ mesuré sur un domaine d'observation en un ensemble de sources/courants placés sur la structure originelle (l'antenne). La nature (électrique, magnétique, polarisation) et la distribution spatiale des sources équivalentes ainsi déterminées nous renseignent sur l'état de fonctionnement de l'objet à diagnostiquer. La visualisation des courants équivalents reconstruits sur une antenne réseau, par exemple, nous renseigne sur l'état de fonctionnement de l'élément unitaire constitutif du réseau à diagnostiquer. La modélisation comportementale est exploitable dans l'**analyse des problèmes directs et inverses** [LOPERA et collab. \[2007\]](#) [SOLDOVIERI et collab. \[2011\]](#) [LAMBOT et collab. \[2004\]](#) [LAMBOT et ANDRÉ \[2014\]](#) [NOUNOUH et collab. \[2015\]](#). En effet, une meilleure connaissance de la source du champ incident (surtout proche) améliore la qualité de la résolution des problèmes électromagnétiques directs et inverses. L'extraction de l'effet de l'antenne et la détermination de la fonction dyadique de Green associée au problème étudié, améliorent grandement la résolution du problème de diffraction électromagnétique. Dans l'analyse du problème direct, le modèle équivalent de l'antenne est pris en compte dans le calcul du champ diffracté et aucune approximation n'est faite (onde plane ou cylindrique comme approximation du champ incident). La connaissance de l'élément émetteur/récepteur, par le biais de son modèle équivalent, offre des possibilités de filtrage (isoler la fonction de transfert de l'antenne) pour une meilleure analyse des champs diffractés dans le domaine du radar de sol par exemple.

La représentation simplifiée de l'antenne offre un maillage plus efficace [OKONIEWSKI et STUCHLY \[1998\]](#). En cherchant à modéliser électromagnétiquement des scénarios complexes, comportant plusieurs objets en interaction et dont les caractéristiques géométriques présentent des disparités importantes (détails fins pour une partie de la scène), il est nécessaire de définir une stratégie de maillage efficace et peu coûteuse en temps de calcul. Si l'élément rayonnant, en l'occurrence l'antenne, est remplacé par un ensemble de sources élémentaires simples (type dipôles), cela nous exempte de mailler finement l'antenne dans la mesure où elle a été remplacée par un modèle équivalent discret (sources dipolaires). Cette stratégie accélère grandement l'analyse numérique. Cette approche est particulièrement prisée dans l'analyse des interactions des ondes électromagnétiques avec le corps humain pour des études dosimétriques de l'exposition des personnes ou dans l'étude des interactions des antennes dans leur environnement de fonctionnement (voiture, train, avion,...).

La notion de **sparsité** ou de **parcimonie** [ALQADAH et collab. \[2014\]](#) dans les problèmes électromagnétiques exprime le souhait d'établir des écritures qui prennent en compte tous les degrés de libertés du problème en un nombre réduit de paramètres. Cette projection, dans le sens algébrique du terme, est destinée à réduire la taille des systèmes d'équations à résoudre, à améliorer l'indépendance des vecteurs considérés et à contraindre plus efficacement les résolutions des problèmes linéaires. Modéliser une antenne avec un nombre réduit de dipôles équivalents rentre parfaitement dans cette définition.

Les finalités présentées ci-dessus sont tributaires des méthodologies de modélisation. J'en ai identifié certaines que je souhaite décrire brièvement. La méthode des **équations intégrales du champ électrique** [ALVAREZ et collab. \[2007\]](#) [ALVAREZ LOPEZ et collab. \[2009\]](#) [NISHINA et CHEN \[2016\]](#) [MIKKI et ANTAR \[2012\]](#) [KIAEE et collab. \[2018\]](#) est de loin la plus usitée. Elle exprime, via une écriture intégrale, le champ rayonné qui résulte d'une distribution de courants électriques et/ou magnétiques. La transcription discrète de cette relation intégrale est combinée avec la méthode des moments pour définir un système linéaire. Ce dernier

lie le champ rayonné aux fonctions des courants équivalents à déterminer. Le système linéaire ainsi reconstruit est résolu par la méthode du gradient conjugué.

Par ailleurs, l'**algorithme génétique ou plus généralement les algorithmes évolutionnaires (évolutionnistes)** [SIJHER et KISHK \[2005\]](#) [IZQUIERDO et RUBIO \[2011\]](#) sont exploités pour déterminer les positions et les pondérations des sources/dipôles équivalents à partir du champ proche mesuré sur un domaine d'observation. On trouve dans cette même catégorie de méthodes, les techniques basées sur l'**apprentissage machine** [HUANG et collab. \[2017\]](#). Cette approche se base sur une puissante capacité de calcul pour transcrire le champ proche mesuré en une distribution de sources équivalentes (positions et amplitudes complexes des sources). Néanmoins, cette technique nécessite une phase d'apprentissage, essentielle à ce mode de résolution, définie en se basant sur des configurations théoriques. Enfin, je termine avec la **technique d'optimisation récursive convexe** [YANG et collab. \[2018\]](#). Cette méthode de modélisation à base de dipôles infinitésimaux fait appel à une optimisation récursive convexe. Son objectif est de réduire le nombre de dipôles nécessaires sans dégrader la précision du modèle. Les positions et les polarisations sont fixées *a priori* et le résultat de l'optimisation récursive est comparé à celui issu d'une optimisation contrainte.

2.2 Identification modale et modélisation d'antennes

Afin de mieux situer mes contributions dans cet axe de recherche, j'ai souhaité donner une vue d'ensemble des développements existants en termes de finalités et de méthodes de modélisations comportementales d'antennes. Dans ce qui suit, je donnerai les contextes des travaux réalisés et soulignerai mes contributions.

Mes travaux sur la modélisation comportementale d'antennes présentés dans ce rapport ont été développés durant mon doctorat (2004 - 2007). En effet, à partir des composantes tangentielles (amplitude et phase) du champ proche mesuré sur une sphère autour de l'élément rayonnant, nous construisons un modèle équivalent composé de dipôles électriques et magnétiques. La méthode proposée concilie l'interprétation implicite et explicite du principe de Huygens-Fresnel. L'interprétation implicite est traduite par le développement en ondes sphériques du champ rayonné tandis que l'interprétation explicite de ce même principe suppose la substitution de l'élément rayonnant par une constellation de sources secondaires qui reproduit le même champ que l'objet initial.

J'ai pensé la méthodologie de modélisation en imaginant le parallèle entre la décomposition en modes sphériques du champ rayonné et la décomposition en dipôles élémentaires. Il se trouve que le rayonnement d'un dipôle centré dans son repère est porté par un unique mode sphérique. En conséquence, j'ai basé mon analyse sur la construction d'un système linéaire qui assure la passerelle entre les courants des dipôles élémentaires équivalents et les coefficients de la décomposition en modes sphériques du champ proche mesuré. Les outils majeurs qui ont servi dans la construction du système linéaire sont les théorèmes de translation et de rotation des harmoniques sphériques que j'ai explicités dans le cas le plus général dans [SERHIR et collab. \[2008\]](#).

Ce travail de recherche s'est poursuivi pendant mon post-doctorat (2008 - 2009) où j'ai établi une validation expérimentale de la méthode à partir des mesures effectuées dans la base sphérique de l'Institut Fresnel. Cette apport expérimental au travail a mis en lumière la robustesse de la méthodologie de modélisation vis-à-vis du bruit et de la troncature de la surface de mesure du champ proche. Nous avons construit le modèle équivalent à base de dipôles placés sur l'ouverture d'une antenne cornet à partir des données du champ proche rayonné sur une partie de la sphère de mesure (troncature). Les résultats sont remarquablement précis [SERHIR et collab. \[2010\]](#) et le modèle équivalent reproduit le rayonnement de l'antenne dans toutes les régions de l'espace à l'exception de la sphère minimale.

Pour accéder au champ à l'intérieur de la sphère minimale, nous avons placé les dipôles qui composent le modèle équivalent directement sur la surface de l'antenne réseau considérée [SERHIR et collab. \[2010\]](#). Dans cette application, nous mettons en avant l'utilisation de l'information *a priori* concernant l'antenne, pour définir un modèle compact décrit par un nombre réduit de dipôles (parcimonie). Dans la suite de ces travaux, j'ai procédé à une simplification du modèle pour réduire le nombre de dipôles nécessaires. Il s'agit de comparer la contribution en puissance de chaque dipôle pour décider de sa conservation. Selon le seuil considéré, leur nombre est réduit, ce qui, naturellement, impacte la précision du modèle.

Par rapport au paysage scientifique présenté précédemment, ma principale contribution dans cet axe de recherche est méthodologique : je définis la pondération des dipôles équivalents sur la base d'une identification modale [SERHIR et collab. \[2008\]](#). Cette représentation parcimonieuse du champ mesuré contribue à la définition d'un système d'équations linéaires à résoudre de taille réduite. La résolution est faite au sens des moindres carrés. Cette méthode de modélisation propose une écriture idoine qui se veut simple d'im-

plémentation et d'intégration dans un scénario de simulation numérique. En effet, j'ai exploité le modèle équivalent d'une antenne cornet construit à partir des mesures en champ proche dans un code de simulation électromagnétique basé sur la méthode des moments (FEKO). Le scénario étudié visait à déterminer le champ diffracté suite à une illumination en champ proche d'une cible étendue. Les résultats des simulations qui exploitent les dipôles équivalents présentent un très bon accord avec les résultats des mesures. En termes applicatifs, je me suis intéressé au diagnostic d'une antenne réseau dans [SERHIR et collab. \[2010\]](#) pour identifier l'élément défaillant du réseau. Cependant une confirmation expérimentale des résultats avancés est à développer dans le très court terme car les techniques de diagnostic des structures réseau sont tributaires du rapport signal à bruit des données utilisées.

2.3 Bilan des contributions et perspectives

Cet axe de recherche propose des ouvertures dans les thématiques associées à la modélisation électromagnétique et au traitement du problème inverse de diffraction. Dans les perspectives de ce sujet, je propose de poursuivre cette réflexion autour des 4 points suivants :

1. **Généralisation de la méthode :** Je souhaite généraliser cette méthodologie de modélisation pour une caractérisation dans le domaine temporel. Le modèle équivalent doit reproduire la réponse impulsionnelle de l'antenne dans toutes les directions de l'espace. Dans la définition actuelle, je calcule la pondération des dipôles équivalents à partir du champ mesuré à une fréquence donnée. Peut-on alors exploiter un nombre fini de fréquences pour définir un modèle équivalent qui reproduit la réponse impulsionnelle de l'antenne?
Pour répondre à cette question, je propose d'élargir la discussion au cas où l'on mesure le champ proche dans le régime temporel et que l'on définisse un autre formalisme pour dimensionner les dipôles équivalents qui rayonnent le même champ. Ce modèle équivalent serait utilisable dans un contexte de radar de sol.
2. **Interaction du modèle équivalent :** L'interaction du modèle équivalent avec l'environnement proche. Il serait très intéressant d'analyser les rectifications à apporter aux pondérations des dipôles équivalents déterminés pour une antenne seule quand celle-ci présente un fort couplage avec son environnement.
3. **Simplification de la procédure :** Actuellement, cette technique de modélisation comportementale exploite les coefficients du développement en ondes sphériques pour définir un modèle équivalent de l'antenne. Cela présente des avantages de stabilité et de précision. Néanmoins, l'inconvénient majeure réside dans la nécessité de disposer du champ proche sur une sphère autour de la structure rayonnante. Cette contrainte ajoute une complexité dans la mise en œuvre dans les cas pratiques. Il est important de simplifier la procédure en utilisant un nombre réduit de points de mesure en se basant sur l'information *a priori* sur l'antenne.
4. **Prise en charge des dipôles équivalents :** Cette thématique de recherche nécessite qu'on s'y intéresse du point de vue de la simulation numérique. J'entends par cela la prise en charge des dipôles équivalents dans les codes de modélisation numérique et plus particulièrement les logiciels du commerce (qui touchent un large public). Dans le formalisme intégral des équations de Maxwell, le modèle équivalent est parfaitement intégrable dans le logiciel FEKO par exemple. Cependant, les logiciels comme HFSS ou CST MWS n'acceptent pas les excitations à base de dipôles infinitésimaux co-localisés et non couplés. Il serait intéressant de proposer des solutions pour améliorer l'intégrabilité du modèle équivalent.

Publications représentatives de la thématique

Les documents joints à ce rapport présentent les éléments théoriques de la méthode et les résultats obtenus. Les références ci-dessous sont représentatives du travail accompli dans cette thématique de recherche.

- [RI.4] M. Serhir, J. M. Geffrin, A. Litman, P. Besnier, « Aperture antenna modeling by a finite number of elemental dipoles from spherical field measurements », *IEEE Transactions on Antennas and Propagation*, vol. 58, no. 4, pp. 1260-1268, 2010
- [RI.3] M. Serhir, P. Besnier, M. Drissi, « Antenna modeling based on a multiple spherical wave expansion method : application to an antenna array », *IEEE Transactions on Antennas and Propagation*, vol. 58, no. 1, pp. 51-58, 2010

Aperture Antenna Modeling by a Finite Number of Elemental Dipoles From Spherical Field Measurements

Mohammed Serhir, Jean-Michel Geffrin, Amélie Litman, and Philippe Besnier

Abstract—A method to determine a distribution of a finite number of elementary dipoles that reproduce the radiation behavior of the antenna under test (AUT) from truncated spherical field measurements is proposed. It is based on the substitution of the actual antenna by a finite number of equivalent infinitesimal dipoles (electric and magnetic), distributed over the antenna aperture. This equivalent set of elementary dipoles is optimized using the transmission coefficient involving the spherical wave expansion of the measured field and using an appropriate matching method. Once the current excitation of each dipole is known, the radiated field of the antenna at different distances can be rapidly determined. Moreover, using an iterative simplification procedure, the number of equivalent dipoles is reduced, which eases the implementation of the antenna equivalent model in any existing electromagnetic code. The feasibility, the reliability and the accuracy of the method are shown using experimental data issued from the measurement of an X-band horn antenna, in two different measurement setups.

Index Terms—Antenna measurement, antenna modeling, elemental dipoles, spherical wave expansion.

I. INTRODUCTION

FOR most of the commercial antennas, the geometric parameters and the material characteristics may not be accessible or may even be unknown. Measurements, such as near-field (NF) techniques, are therefore the only way to determine accurately the radiation characteristics of the antenna. In particular, the recent development of multi-probe near-field system contributes to reduce the measurements duration, enabling real-time 3D complex characterization [1]. The issue is to derive from these measurements the radiation pattern of the antenna in its real environment.

When the geometrical parameters and the material properties of the antenna are well-known, the antenna modeling in its operating environment is still a challenge using conventional

full-wave techniques (finite elements method, finite difference time domain, or method of moments). This is due to the fact that the meshing criterion depends on the finest and the smallest element of the antenna. Consequently, the computation cost increases rapidly with increasing antenna size relative to wavelength.

In this paper, we propose a hybrid three-step approach. Firstly, the radiation pattern of the antenna under test is measured in a completely controlled environment (anechoic chamber). Secondly, from these measurements, a simple equivalent model is achieved in terms of elementary equivalent sources that reproduce the antenna radiation pattern. Finally, this equivalent model is incorporated into a commercial electromagnetic (EM) algorithm to assess the behavior of the modeled antenna in different environments (operating conditions).

In the literature, many authors have proposed different strategies where they determine currents/charges distributions that radiate an electromagnetic field matching the AUT radiation. In [2]–[6] an electric field integral equation method (EFIE) relating the measured electric field (tangential components) to equivalent electric/magnetic currents on the aperture of the antenna has been proposed. The EFIE is solved using the conjugate gradient method. This method has mainly been used to perform near-field to far-field transformation and for antenna diagnosis applications [7], [8].

In [9]–[11], the authors used the genetic algorithm (GA) technique to substitute the AUT by a set of infinitesimal dipole sources. These dipoles are distributed inside the volume enclosing the AUT. The infinitesimal dipoles have been selected due to their implementation simplicity in any EM code, and the GA demonstrates its ability to figure out the type (electric or magnetic), the position, the orientation and the excitation of each dipole. The limitation of this method appears when dealing with complex antennas (electrically large structures), which require a large number of equivalent dipoles, as well as convergence issues and prohibitive CPU computing time.

In [12] the authors have developed a modeling technique, based on the equivalence principle. The AUT is substituted by a set of equivalent dipoles distributed over the minimum sphere circumscribing the antenna (closed surface). The excitations of the antenna equivalent dipoles are derived from the AUT transmission coefficients and the equivalent dipoles spatial distribution is well-adapted to build up a well-conditioned matrix system. Computations with EM simulation data illustrate the accuracy of the method. Nevertheless, these equivalent

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sources distribution is not optimum for large radiating structures, particularly, for aperture antennas (horn antennas). The number of equivalent sources grows considerably when the sources are distributed over the antenna minimum sphere. Thus, the equivalent model loses its simplicity when implemented in an EM code.

In the present paper, an improvement of the analysis of [12] is presented. The antenna equivalent current sources are placed over a planar surface with finite extent in front of the antenna aperture (S_{equi}) and only tangential dipoles are considered. For this purpose, spherical field measurement data is used to determine the spherical wave expansion (SWE) of the horn antenna radiated field. Then, taking into account the antenna geometrical *a priori* information, and by adopting a convenient equivalent sources spatial distribution over the planar surface, the SWE of the antenna radiated field is written in terms of tangential electric and magnetic dipoles placed over the antenna aperture. Hence, one can represent a large structure using a finite number of equivalent dipoles, wherein the excitations of the dipoles are correlated to the antenna transmission coefficients.

The most significant aspect studied in the present paper is the reliability and the accuracy of the method when using experimental data issued from the measurement of a real antenna (horn antenna) in the far-field region. Firstly, using measurement data over a sphere and secondly using truncated measurement data. The viability of the proposed modeling method is determined by comparison with experimental measurement data, where we have to deal with the measurement errors, the position inaccuracies and the signal to noise ratio of the measurement system. Once the equivalent model is achieved, a simplification procedure is presented. This procedure aims to find a finite number of elementary dipoles whose radiation pattern fits accurately the actual AUT radiation pattern by eliminating the insignificant current sources and dipoles. Then, the EM field can be rapidly computed at any distances. In this paper a comparison between the SWE and the proposed modeling technique in terms of computation time is presented.

To assess the overall procedure, the radiation pattern of an X-band horn antenna has been measured over a truncated sphere using a single probe spherical range measurement in the large anechoic chamber of the Institut Fresnel [13]. The antenna radiation pattern in that measurement setup is measured in the far field region. Once the equivalent model is defined, we compute the radiation pattern at a different distance. In particular, the estimated pattern is compared to another set of measurements performed this time with a single probe planar measurement setup, which corresponds to the new anechoic chamber of the Institut Fresnel.

The paper is structured as follows: In Section II, following a review of the spherical wave expansion formulation, the basis of the proposed technique is detailed. Section III presents the modeling process applied to the X-band horn antenna radiating at the frequency of 12 GHz. In order to illustrate the robustness of the proposed modeling technique we use truncated experimental measurement data using a part of the measurement sphere to validate the efficiency of the method. The setups, as well as the measurement procedure, are described in Section III. Finally, some concluding remarks are outlined in Section IV. All

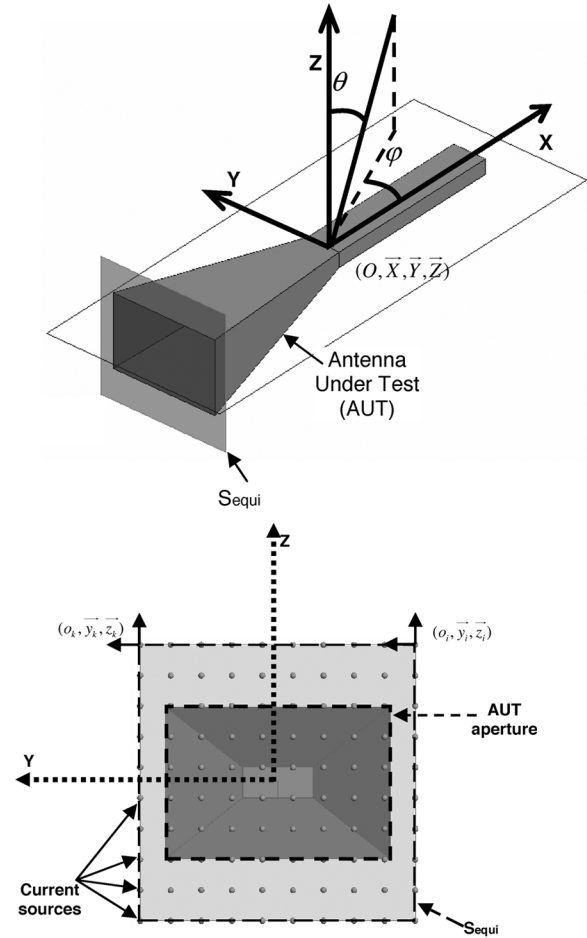


Fig. 1. Original horn antenna (top) and the equivalent current sources distribution over the equivalent surface S_{equi} (bottom).

theoretical quantities are expressed in the S.I. rationalized unit system with $e^{-j\omega t}$ time dependence.

II. CONSTRUCTION OF THE EQUIVALENT MODEL

The goal of the proposed modeling technique is to find the excitation of a given number (L_{source}) of equivalent current sources that reproduce the antenna radiation pattern. In particular for a horn antenna, these sources are placed over the planar surface representing the horn aperture S_{equi} (Fig. 1). We make use of the theoretical development, detailed in [12], regarding the new spatial distribution of the equivalent sources adopted in this work.

Contrarily to [12], in this paper, we define a current source as the combination of four tangential co-localized short dipoles: 2 electric ones and 2 magnetic ones. The i th current source (for $1 \leq i \leq L_{\text{source}}$), is associated to a coordinate system $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$, where the origin coincides with the position of this source in the AUT global coordinate system $(O, \vec{X}, \vec{Y}, \vec{Z})$. The choice of this current source composition is justified by the fact that using two orthogonal electric (magnetic) dipoles, we make sure that whatever the dipole orientation over the (AUT) aperture is, the dipole can be represented by two orthogonal components. The aim of this work is to use the AUT transmission coefficients to determine the

excitation of the equivalent model sources. These coefficients are determined from truncated measurement data in contrast to [12], where they were computed from a complete sphere of synthesized data (using a simulation software). For this reason, we recall the formulation of the spherical wave expansion (SWE) of the radiated field.

A. Spherical Wave Expansion

In a source free region, outside the minimum sphere circumscribing the AUT, the SWE of the radiated field \vec{E}_{AUT} in the spherical coordinate system associated to the global coordinate system $(O, \vec{X}, \vec{Y}, \vec{Z})$, where the origin $O(0,0,0)$ coincides with the centre of the measurement sphere (Fig. 1), is expressed in terms of truncated series of spherical vector wave functions [14] as

$$\begin{aligned} \vec{E}_{\text{AUT}}(\vec{r}) &= \frac{k}{\sqrt{\eta}} \sum_{s=1}^2 \sum_{n=1}^{N_{\text{tr}}} \sum_{m=-n}^{m=n} Q_{s,m,n} \vec{F}_{s,m,n}^{(3)}(\vec{r}) \\ &= \sum_{j=1}^{J_{\text{max}}} Q_j \vec{F}_j^{(3)}(\vec{r}) \end{aligned} \quad (1)$$

where $N_{\text{tr}} = \lceil kr_{\text{min}} \rceil + 10$ and k corresponds to the wave number, $\eta = \sqrt{\varepsilon/\mu}$ the intrinsic admittance of the medium, $Q_{s,m,n}$ are the spherical wave coefficients (transmission coefficients), and $\vec{F}_{s,m,n}^{(3)}$ represent the power-normalized spherical vector wave functions. The truncation number N_{tr} depends on the antenna dimensions and the operating frequency [14]. Using the compact index j such as $j = 2(n(n+1)+m-1)+s$, we simplify the SWE expression where $J_{\text{max}} = 2(N_{\text{tr}}(N_{\text{tr}}+2))$. These coefficients are uniquely determined from the knowledge of the tangential components of either \vec{E} or \vec{H} on the measurement sphere (due to the uniqueness theorem). In [15], [16] truncated near-field measurement data has been shown to be sufficient to calculate the Q_j coefficients. Once the Q_j are determined, the field outside the measurement sphere is completely characterized by (1).

Our aim is to determine an equivalent model of a given antenna from its transmission coefficients Q_j using experimental data. We take advantage of the compactness of the SWE expression for the field radiated by a short dipole ($N_{\text{tr}} = 1$ for electric or magnetic dipole) and we relate the field radiated by an antenna described by J_{max} spherical vector wave functions to the set of infinitesimal dipoles using translational and rotational addition theorems.

B. Short Dipole Excitation

In what follows, the superscripts “ e ” and “ m ” stand for the electric and magnetic dipoles, respectively. The subscripts “ y ” and “ z ” denote \vec{y} , \vec{z} polarized dipoles, respectively. The subscript “ i ” designates the i th current source located at the position $o_i(x_i, y_i, z_i)$ in $(O, \vec{X}, \vec{Y}, \vec{Z})$.

The field $\vec{E}_i^{e,z}$ radiated by an infinitesimal \vec{z} -directed electric dipole located at the origin of $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ is expressed

in the spherical coordinate system $\vec{r}_i^{\rightarrow}(r_i, \theta_i, \varphi_i)$ associated with $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ as

$$\begin{aligned} \vec{E}_i^{e,z}(\vec{r}_i^{\rightarrow}) &= \frac{k}{\sqrt{\eta}} \alpha_i^{e,z} \vec{F}_4^{(3)}(\vec{r}_i^{\rightarrow}), \text{ with} \\ \alpha_i^{e,z} &= -\frac{k}{\sqrt{6\eta\pi}} I_i^{e,z} l. \end{aligned} \quad (2)$$

In the global spherical coordinate system $\vec{r}(r, \theta, \varphi)$ associated with $(O, \vec{X}, \vec{Y}, \vec{Z})$ we have

$$\vec{F}_4^{(3)}(\vec{r}) = \frac{k}{\sqrt{\eta}} \sum_{j=1}^{J_{\text{max}}} A_{j,i}^{e,z} \vec{F}_j^{(3)}(\vec{r}). \quad (3)$$

From (2) and (3) it follows that the field in the global spherical coordinate system $\vec{r}(r, \theta, \varphi)$ is given by

$$\vec{E}_i^{e,z}(\vec{r}) = \frac{k}{\sqrt{\eta}} \alpha_i^{e,z} \sum_{j=1}^{J_{\text{max}}} A_{j,i}^{e,z} \vec{F}_j^{(3)}(\vec{r}). \quad (4)$$

Detailed formulations and expressions for \vec{x} and \vec{y} -dipoles are provided in [12].

Accordingly, the i th current source is composed of 4 co-localized tangential dipoles, this will be characterized by 4 transmission coefficients $\alpha_i^{e,y}, \alpha_i^{m,y}, \alpha_i^{e,z}, \alpha_i^{m,z}$. These coefficients are proportional to the dipoles length l and to the current excitations $I_i^{e,y}, I_i^{m,y}, I_i^{e,z}, I_i^{m,z}$.

We associate to the i th current source the row vector $[S_i]$ where

$$\begin{aligned} [S_i] &= [\alpha_i^{e,y}, \alpha_i^{m,y}, \alpha_i^{e,z}, \alpha_i^{m,z}] \\ &\text{in } (o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i) \\ &= [\alpha_i^{e,y} A_{j,i}^{e,y}, \alpha_i^{m,y} A_{j,i}^{m,y}, \alpha_i^{e,z} A_{j,i}^{e,z}, \alpha_i^{m,z} A_{j,i}^{m,z}]^{1 \leq j \leq J_{\text{max}}} \\ &\times \text{in } (O, \vec{X}, \vec{Y}, \vec{Z}). \end{aligned} \quad (5)$$

The coefficients $A_{j,i}^{e,y}, A_{j,i}^{m,y}, A_{j,i}^{e,z}, A_{j,i}^{m,z}$ are determined using translations and rotations of the vector spherical wave functions related to each dipole from $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ to $(O, \vec{X}, \vec{Y}, \vec{Z})$ [see Appendix in [12]].

C. Antenna Modeling Method

In order to determine the excitation of each elemental dipole composing the equivalent model, we have to define the adequate matching method. This can only be done in the global coordinate system. This is why we have used translational and rotational addition theorems [14], [17], [18] to express the field radiated by the i th current source located at $o_i(x_i, y_i, z_i)$ in the antenna global coordinate system $(O, \vec{X}, \vec{Y}, \vec{Z})$.

Then, the superposition of all the L_{source} current sources field $\vec{E}_{\text{mod}}(\vec{r})$ expressed in $(O, \vec{X}, \vec{Y}, \vec{Z})$ is identified with \vec{E}_{AUT} . In the same way as [12] we have to solve the linear equations

$$A \cdot X = Q \quad (6)$$

where, $\mathbf{A} = [A_{j,i}^{e,y}, A_{j,i}^{m,y}, A_{j,i}^{e,z}, A_{j,i}^{m,z}]_{1 \leq j \leq J_{\max}, 1 \leq i \leq L_{\text{source}}}$, $\mathbf{X} = [S_1 \dots S_{L_{\text{source}}}]^T$ and $\mathbf{Q} = [Q_1 \dots Q_{J_{\max}}]^T$. The superscript “ T ” denotes the matrix transpose.

The lsqr routine from MatLab (least square data fit) [19] has been used to solve the over-determined matrix equation ($4 \times L_{\text{source}} \leq J_{\max}$). Once Equation (6) is solved, the equivalent model characterization is finalized and each current excitation $(I_i^{e,y}, I_i^{m,y}, I_i^{e,z}, I_i^{m,z})_{1 \leq i \leq L_{\text{source}}}$ is fully computed.

D. Equivalent Model Simplification

The equivalent model of the antenna can be furthermore simplified by reducing the number of infinitesimal dipoles. In [12] authors present an iterative process which aims at reducing the number of dipoles based on the evaluation of the dipoles’ radiation power. In this paper we have improved the elimination process of the dipoles having a non significant radiation power by eliminating as well the current sources having an insignificant radiation power. Let us first recall the elimination procedure of the dipoles having a too small radiation power.

By solving (6), the vector rows $[S_i]_{1 \leq i \leq L_{\text{source}}}$ are fully determined. Then, for each current source, the power contribution of each dipole in the total radiated power P_i is evaluated.

Let define the normalized row vector power contribution

$$[S_i^{\text{nor}}] = \frac{\frac{1}{2} \left[|\alpha_i^{e,y}|^2, |\alpha_i^{m,y}|^2, |\alpha_i^{e,z}|^2, |\alpha_i^{m,z}|^2 \right]}{P_i}$$

where

$$P_i = \frac{1}{2} \left(|\alpha_i^{e,y}|^2 + |\alpha_i^{m,y}|^2 + |\alpha_i^{e,z}|^2 + |\alpha_i^{m,z}|^2 \right).$$

Using different thresholds P_{dT} , the number of dipoles can be reduced. First, the number of current sources and their spatial position are chosen and a solution of (6) is built with the contribution of all the dipoles as depicted in Section II.C. By calculating the normalized row vectors $\{S_i^{\text{nor}} \forall i \in [1 \dots L_{\text{source}}]\}$, we identify the dipoles whose corresponding terms in S_i^{nor} are less than P_{dT} . A new solution of (6) is then computed without these negligible dipoles. This procedure is pursued as long as no more negligible dipoles are detected. When the power contribution of each dipole is greater than the threshold P_{dT} , the power contribution of each current source in the antenna radiation power is evaluated.

The next paragraph describes the elimination of the current sources with negligible radiation power.

The current sources have to reproduce the antenna radiation pattern. In particular, the superposition of the current sources radiated power has to be equal to the radiated power of the antenna P_{AUT} . Let us define the radiated power by the antenna and the current sources, respectively by

$$P_{\text{AUT}} = \frac{1}{2} \sum_{j=1}^{J_{\max}} |Q_j|^2 \quad (7)$$

$$P = \frac{1}{2} \sum_{i=1}^{L_{\text{sources}}} \sum_{j=1}^{J_{\max}} \left(|\alpha_i^{e,y} A_{j,i}^{e,y}|^2 + |\alpha_i^{m,y} A_{j,i}^{m,y}|^2 + |\alpha_i^{e,z} A_{j,i}^{e,z}|^2 + |\alpha_i^{m,z} A_{j,i}^{m,z}|^2 \right). \quad (8)$$

Since

$$\sum_{j=1}^{J_{\max}} |A_{j,i}^{e,y}|^2 = 1, \quad \sum_{j=1}^{J_{\max}} |A_{j,i}^{m,y}|^2 = 1$$

$$\sum_{j=1}^{J_{\max}} |A_{j,i}^{e,z}|^2 = 1, \quad \sum_{j=1}^{J_{\max}} |A_{j,i}^{m,z}|^2 = 1 \quad (9)$$

and identifying (7) and (8), we obtain

$$P_{\text{AUT}} = P = \frac{1}{2} \sum_{i=1}^{L_{\text{sources}}} \left(|\alpha_i^{e,y}|^2 + |\alpha_i^{m,y}|^2 + |\alpha_i^{e,z}|^2 + |\alpha_i^{m,z}|^2 \right). \quad (10)$$

Equation (9) is justified by the fact that wherever an infinitesimal dipole is placed in the global coordinate system, its radiated power is invariable. Let now define P_i^{nor} for each current sources such that

$$P_i^{\text{nor}} = P/P_{\text{average}}, \quad \forall i \in [1 \dots L_{\text{source}}]$$

where

$$P_{\text{average}} = P_{\text{AUT}}/L_{\text{sources}}.$$

While supposing that the antenna radiated power is equidistributed over the equivalent current sources (P_{average}), we can evaluate, by calculating P_i^{nor} , the contribution (weighting) of the i th current source in the radiation power. Afterward, we can isolate and eliminate current sources for which P_i^{nor} are less than the given threshold P_{csT} . Finally, a solution of (6) is re-computed without taking into account these insignificant current sources and dipoles.

III. RESULTS

A. Spherical Measurement Setup

To illustrate the effectiveness of the proposed methodology, the modeling process is applied to a commercial X-band horn antenna, (Fig. 2) operating at the frequency of 12 GHz. The waveguide dimensions are 124 mm \times 23 mm and the aperture size is 73.5 mm \times 50 mm ($3\lambda \times 2\lambda$). The associated minimum sphere circumscribing the antenna has a radius of $r_{\text{min}} \simeq 175$ mm (7λ). This antenna is measured using the spherical field system in the anechoic chamber of the Centre Commun des Ressources Micro-ondes de Marseille (CCRM) managed for this topic by Institut Fresnel researchers, which is presented in Fig. 3 [13]. The tangential complex components of the electric field (E_θ, E_φ) of this antenna have been measured at the distance $R_{\text{meas}} = 1.79\text{m} = 71.6\lambda$. To carry out the equivalent current sources model, the measurement data are collected over a spherical surface ($0 \leq \theta \leq 165^\circ, 0^\circ \leq \varphi \leq 357^\circ$). The field sampling is of 3° both in θ and φ coordinates, with 56×120 measurement points.

B. Full Spherical Measurement Data Set

The measurement data matrix is completed (with null field values) in order to fulfil the measurement over a sphere $0 \leq \theta \leq 180^\circ, 0^\circ \leq \varphi \leq 357^\circ$. Based on these data and using the

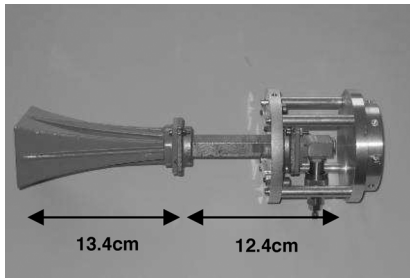


Fig. 2. The AUT description: an X-band horn antenna with the exciting waveguide.

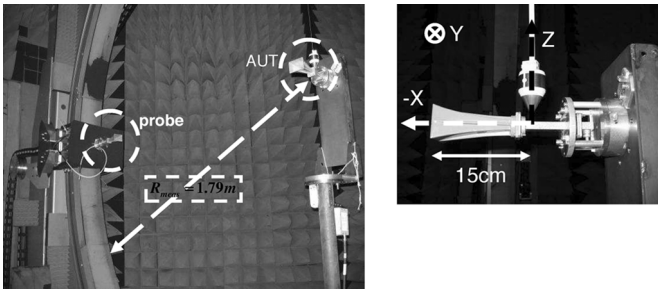


Fig. 3. Measurement setup using the single probe spherical near field range at CCRM (left). Description of the measured horn antenna (AUT) position in the measurement coordinate system (right).

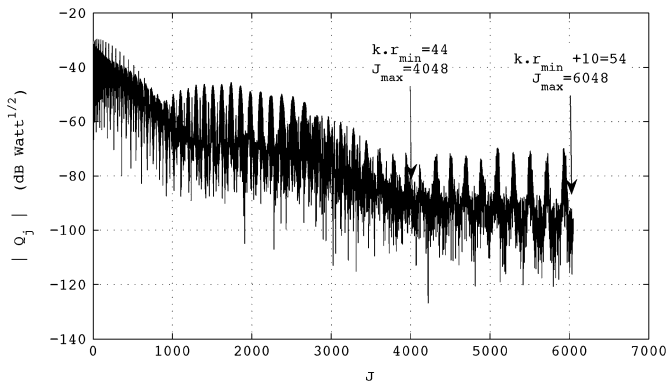


Fig. 4. Magnitude of the AUT transmission coefficients calculated using the measurement data collected over the spherical surface $R_{\text{meas}} = 1.79$ m, for $0 \leq \theta \leq 165^\circ$ and $0^\circ \leq \varphi \leq 357^\circ$.

matrix method [15] we compute the antenna transmission coefficients. These are presented in Fig. 4. The SWE is truncated at $N_{\text{tr}} = [kr_{\text{min}}] + 10 = 54$, such that $J_{\text{max}} = 2 \times 54 \times 56 = 6048$ transmission coefficients are considered for the current configuration.

The AUT equivalent model is composed of 10×10 current sources separated by 0.4λ . These sources are chosen to be distributed uniformly over a planar domain S_{equi} of size $90 \text{ mm} \times 90 \text{ mm}$ positioned at the antenna aperture plane $x \simeq -150 \text{ mm}$ (Fig. 3). From a theoretical point of view, the aperture plane should be infinite. However, it is possible to truncate it as the radiated power of the antenna, and consequently, the equivalent dipoles, is concentrated in a finite region in front of the antenna aperture.

Once the transmission coefficients of the antenna are known and the equivalent current sources positions defined, we solve

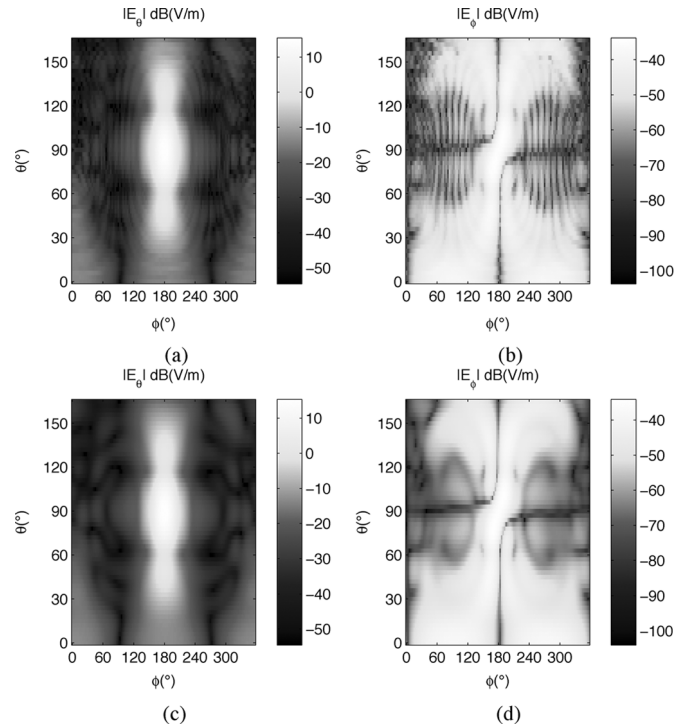


Fig. 5. Magnitude of the E-field components E_θ and E_φ in dB (V/m) at the measurement distance $R_{\text{meas}} = 1.79$ m. In (a) and (b) the measured field radiated by the actual horn antenna and in (c) and (d) the field radiated by the antenna equivalent model.

(6) and we compute the current excitation of each dipole. Thereafter, we rapidly calculate the E-field at different distances in straightforward manner using the equivalent dipoles radiation.

The E-field radiated by the AUT equivalent model is compared to the actual antenna radiation pattern in the measurement area $0 \leq \theta \leq 165^\circ, 0^\circ \leq \varphi \leq 357^\circ$ in Fig. 5 at the measurement distance. The measured field and the field obtained from the equivalent model are in a good agreement. In addition, the radiation pattern obtained from the equivalent model fits very well with the one resulted from the SWE method as presented in Fig. 6. Also, as shown in Fig. 5, the field radiated by the equivalent model tends to cancel the ripples observed in the measured field. These ripples are principally caused by measurement errors as for example: multiple reflections between the AUT and the measurement probe, the interaction between the antenna and its mounting system . . .

C. Truncated Spherical Measurement Data Set

In order to further verify the effectiveness of the proposed method, a comparative study is presented by determining the horn antenna equivalent model using a truncated measurement data set. For the comparison, the previously used measurement data (X-band horn antenna) have been truncated to the region of significant radiated field $15^\circ \leq \theta \leq 165^\circ, 105^\circ \leq \varphi \leq 255^\circ$ with angular steps of $\Delta\theta = 3^\circ, \Delta\varphi = 3^\circ$ ($\Delta\theta = \Delta\varphi = \pi/60 \approx \pi/N_{\text{tr}}$). This resulted in 51×51 field points. Then, the measurement data matrix is completed (with null field values) in order to fulfil the measurement over a complete sphere. Using these measurement data, the antenna transmission coefficients

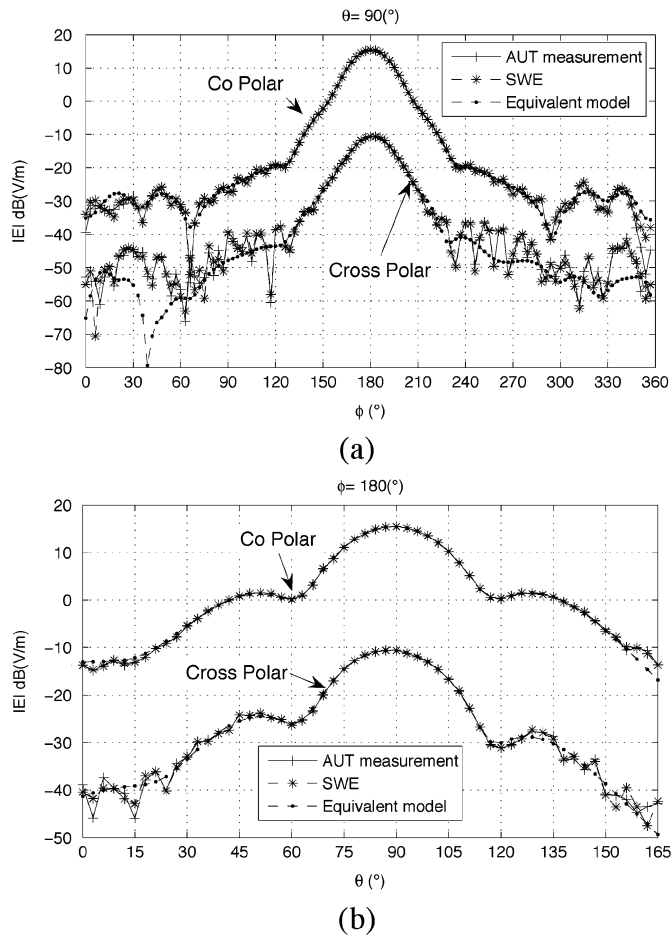


Fig. 6. Comparison of co-polarized and cross-polarized components of the fields at the measurement distance $R_{\text{meas}} = 1.79$ m. The comparison includes: the AUT measurement, the field calculated using the SWE and the field radiated by the equivalent model. (a) For $\theta = 90^\circ$ and (b) for $\varphi = 180^\circ$.

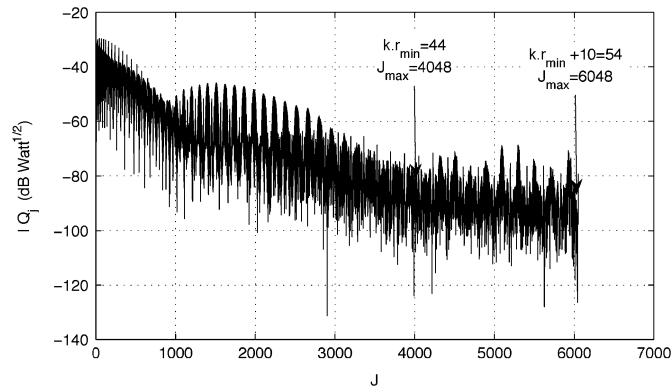


Fig. 7. Magnitude of the AUT transmission coefficients calculated using the measurement data collected over the spherical surface $R_{\text{meas}} = 1.79$ m, for $15^\circ \leq \theta \leq 165^\circ$ and $105^\circ \leq \varphi \leq 255^\circ$.

are recomputed. The SWE is truncated at $N_{\text{tr}} = 54$ and the resulting transmission coefficients are presented in Fig. 7.

While considering the same equivalent current sources spatial distribution, we solve (6) and we determine the excitation of each equivalent dipole regarding the new calculated transmission coefficients (truncated measurement data). At the measurement distance (R_{meas}), in the area $15^\circ \leq \theta \leq 165^\circ$, $105^\circ \leq$

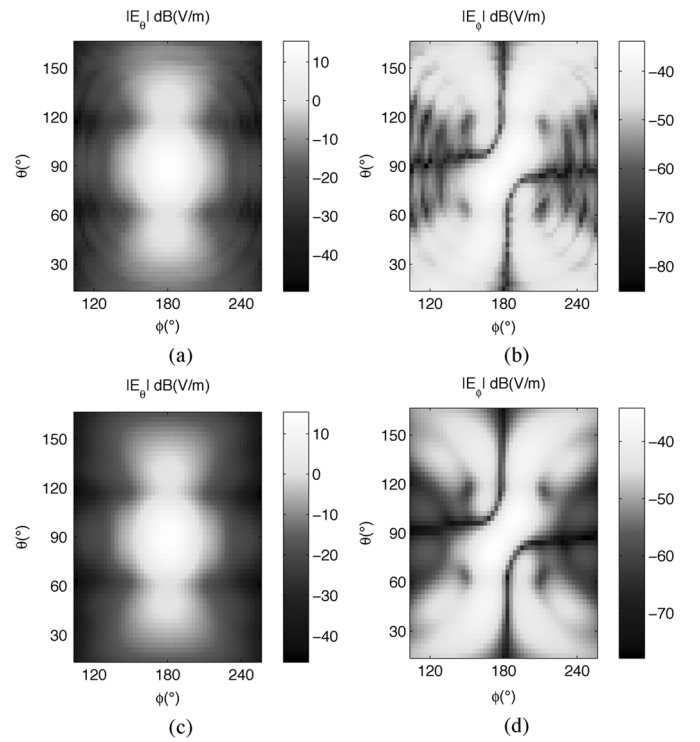


Fig. 8. Magnitude of the E-field components E_θ and E_φ in dB (V/m) at the measurement distance $R_{\text{meas}} = 1.79$ m. In (a) and (b) the measured field radiated by the actual horn antenna and in (c) and (d) the field radiated by the antenna equivalent model determined using truncated measurements.

$\varphi \leq 255^\circ$, the comparison between the radiation patterns of the actual AUT and the equivalent model are presented in Fig. 8. As it can be seen, the radiation pattern of the equivalent model fits very well with the actual field. Moreover, the E-field of the equivalent model is in good agreement with the E-field resulted from the SWE as presented in Fig. 9.

In Fig. 10 we present the equivalent dipoles distribution over S_{equi} resulting from different thresholds ($P_{\text{dT}} = P_{\text{csT}} = 0\%, 2\%, 4\%$). As it can be seen the region where the magnitude of the equivalent dipoles is significant coincides with the horn aperture (dashed rectangle). The horn antenna being off-centred during the measurement, this can justify the fact that we have considered an equivalent aperture (S_{equi}), whose dimensions are greater than the horn actual aperture. Nevertheless, the equivalent dipoles magnitude is consistent with the actual E-field in the aperture of the antenna, since the significant equivalent dipoles are concentrated in the aperture area (dashed rectangle). In addition, using different thresholds ($P_{\text{dT}} = P_{\text{csT}} = 2\%, 4\%$), the polarization of the equivalent dipoles (electric and magnetic) is improved. In other words, the actual antenna E-field polarization is not purely vertical. This corresponds to the equivalent model comprising all dipoles. However, eliminating non-significant dipoles tends to enhance the E-field polarization purity. Consequently, substituting the antenna by an equivalent model with $P_{\text{dT}} = P_{\text{csT}} = 2\%$, and implementing this model in an EM code, we make sure that the polarization of the radiated field is a closer to a purely vertical polarization (the cross-polarization is then completely cancelled for $\theta = 90^\circ$ and $\varphi = 180^\circ$).

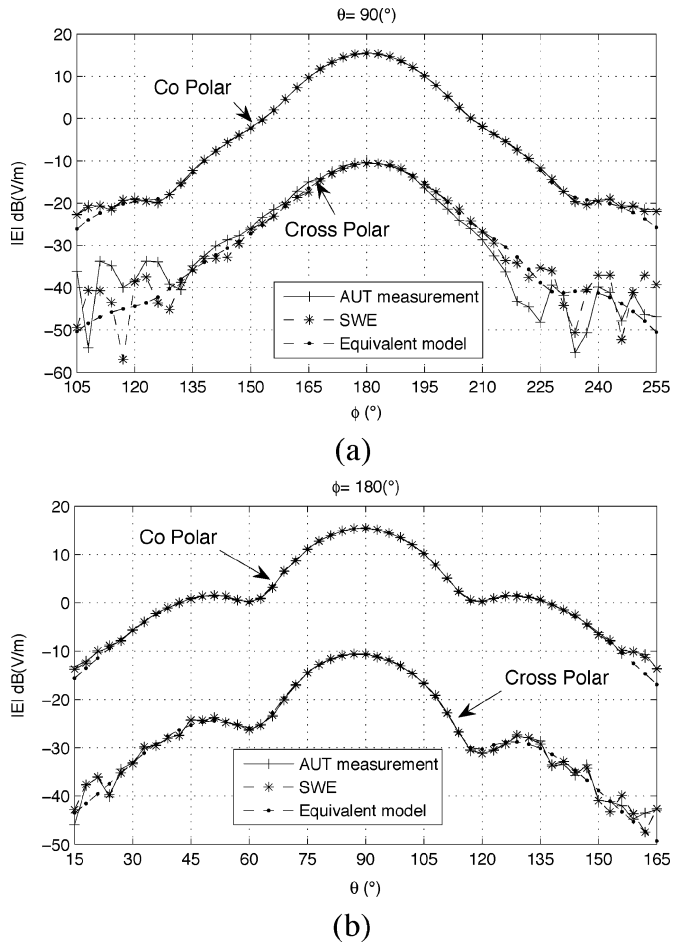


Fig. 9. Comparison of co-polarized and cross-polarized components of the fields at the measurement distance $R_{\text{meas}} = 1.79$ m. The comparison includes: the AUT measurement, the field calculated using the SWE and the field radiated by the equivalent model based on truncated spherical measurement. (a) For $\theta = 90^\circ$ and (b) for $\varphi = 180^\circ$.

In Table I, we show the composition of each simplified equivalent model corresponding to different thresholds, the initial number of equivalent current sources is reduced twice for $P_{dT} = P_{cst} = 4\%$ compared with the initial current sources distribution.

D. Discussion: The Number of the Antenna Equivalent Sources

The number of the equivalent dipoles and their spatial distribution depend on the antenna aperture size and the antenna equivalent model targeted accuracy. In fact, the spacing between sources is directly correlated with the accuracy of the antenna equivalent model and has a direct effect on the condition number of the matrix \mathbf{A} in (6). We have tried out different spacing ($0.25\lambda, 0.4\lambda, 0.5\lambda,$) between equivalent sources and two general remarks arise.

Using the Nyquist criterion $\Delta x = \Delta y = 0.5\lambda$ to distribute the equivalent dipoles over the antenna aperture, the matrix \mathbf{A} in (6) is very well-conditioned. However, the antenna equivalent model, based on this criterion, is not accurate. Otherwise, using a smaller spacing, for example $\Delta x = \Delta y = 0.25\lambda$, the accuracy of the antenna equivalent model is improved but the number of equivalent dipoles increases. Furthermore, this

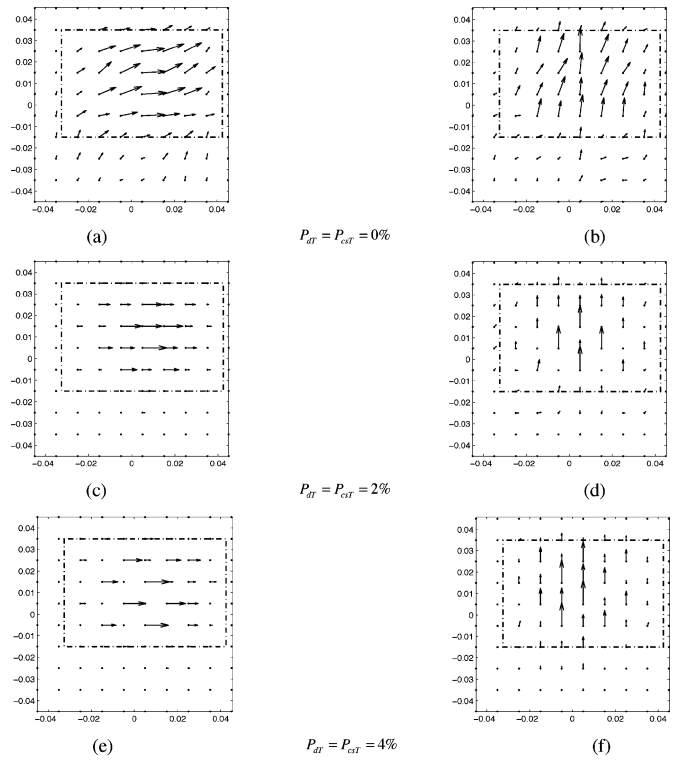


Fig. 10. The magnitude of the equivalent dipoles distributed over the equivalent surface (S_{equi}) for the filtering thresholds $P_{dT} = P_{cst} = 0\%$, 2% and 4% (left) magnetic dipoles. (right) Electric dipoles. The dashed rectangle represents the AUT actual aperture.

equivalent dipoles distribution tends to deteriorate the condition number of \mathbf{A} . Since the equivalent model is intended to be implemented in an EM code, its simplicity (number of dipoles) is very important. As a matter of fact, we have to compromise between the accuracy of the equivalent model and the conditioning of \mathbf{A} . The spacing criterion adopted for the example presented in the paper ($\Delta x = \Delta y = 0.4\lambda$), is a good compromise. The results obtained here show that the antenna equivalent model is accurate for the criterion $\Delta x = \Delta y = 0.4\lambda$. Further work will be required to determine whether this criterion is applicable for various arrangements of dipoles and to broader range of antennas (different sizes and geometries).

IV. VALIDATION

A. The Planar Measurement Setup and Experimental Validation Process

The performances of the previously described simplification process are evaluated by comparing the simplified equivalent models E-fields with the actual radiation pattern over a line ($x = -45$ cm, -40 cm $\leq y \leq 40$ cm, $z = 0$) at a distance of 30 cm from the horn antenna aperture which is closer than the measurement made for the equivalent model determination (1.79 m). To achieve this measurement, we have exploited the new measurement setup of the Institut Fresnel which consists in a single probe planar measurement system in an anechoic environment (Fig. 11).

The planar scanner has a total dual scan length of 2.1 m in X and Y direction placed in an anechoic environment. The same

TABLE I
EQUIVALENT MODELS AFTER THE SIMPLIFICATION PROCEDURE FOR $P_{dT} = P_{csT} = 0, 1, 2, 3$ AND 4%

$P_{dT} = P_{csT} = 0\%$ 100 current sources 400 dipoles		$P_{dT} = P_{csT} = 1\%$ 92 current sources 215 dipoles		$P_{dT} = P_{csT} = 2\%$ 74 current sources 178 dipoles		$P_{dT} = P_{csT} = 3\%$ 71 current sources 124 dipoles		$P_{dT} = P_{csT} = 4\%$ 53 current sources 95 dipoles	
$dipole^{e,y}$	100	$dipole^{e,y}$	70	$dipole^{e,y}$	38	$dipole^{e,y}$	9	$dipole^{e,y}$	2
$dipole^{m,y}$	100	$dipole^{m,y}$	63	$dipole^{m,y}$	61	$dipole^{m,y}$	55	$dipole^{m,y}$	48
$dipole^{e,z}$	100	$dipole^{e,z}$	73	$dipole^{e,z}$	60	$dipole^{e,z}$	52	$dipole^{e,z}$	41
$dipole^{m,z}$	100	$dipole^{m,z}$	9	$dipole^{m,z}$	19	$dipole^{m,z}$	8	$dipole^{m,z}$	4

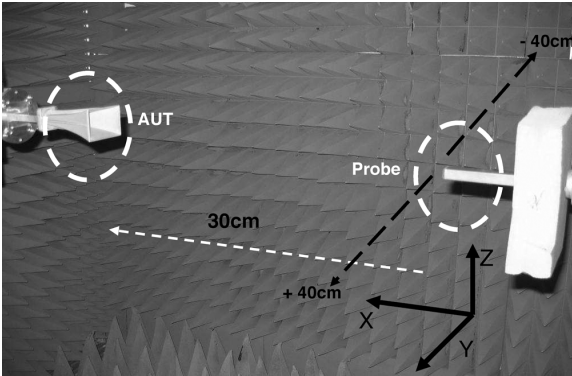


Fig. 11. Measurement setup using the single probe planar near field range at Institut Fresnel.

horn antenna is used in the transmit mode. The probe, which is mounted on one of the two carriages of the linear scanner, consists of an open ended wave guide. The probe is connected to a vector network analyzer (VNA) and the magnitude and the phase of the antenna radiated field are measured. A computer monitors the probe displacement over the line in front of the antenna and acquires the data provided by the VNA.

In Fig. 12 we present the E_z field radiated by different simplified AUT equivalent models ($P_{dT} = P_{csT} = 0\%, 1\%, 2\%, 3\%$) compared to the measured E_z field and the field estimated using the SWE method over a line at a distance of 30 cm from the aperture. As it can be seen the agreement is excellent. In addition, we present in Fig. 12 the computation time needed to calculate E_z at the measurement distance (30 cm from the AUT aperture). We show that the proposed simplification procedure eliminates insignificant current sources and dipoles while preserving a good accuracy. Once the equivalent model is fully defined the E field is rapidly estimated at different distances. Generally, the computation time and the accuracy needed to derive the E field radiated by a simplified equivalent model depend on the chosen threshold. This flexibility allows one to make a compromise between the accuracy (computation time) and the complexity (number of considered equivalent current sources).

B. The SWE Versus the AUT Equivalent Model

The SWE is a very accurate way to describe the radiated field by means of truncated summation of spherical wave functions and the accuracy of the SWE cannot be questioned. In addition, a convenient choice of the truncation number can help to cancel the interaction between the antenna and the mounting system during the measurement, to filter out the rapid variation of the

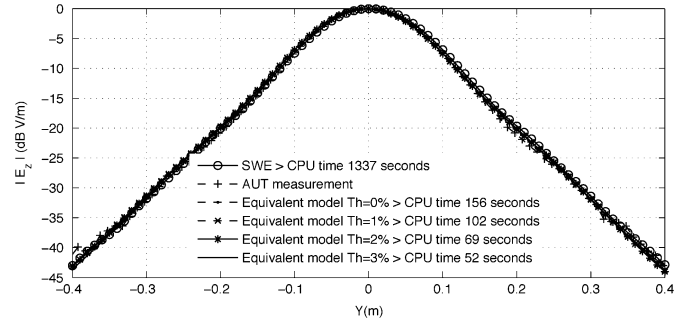


Fig. 12. The magnitude of the E_z field measured at the distance of 30 cm from the AUT aperture compared with the E_z fields calculated from the SWE and the simplified equivalent models.

electric field caused by the mounting system or by the interaction between the antenna under test and the probe.

Moreover, the use of the SWE to assess the antenna equivalent model allows to reduce of the number of the linear equations to consider in (6). In fact, the horn antenna radiated field modelled in the paper is completely characterized by $\mathbf{J}_{\max} = 6048$ spherical wave coefficients. Consequently we have to solve a linear system $\mathbf{AX} = \mathbf{Q}$ with 6048 equations. If the SWE were not used, the number of equations would increase to 13440, the number of data points on the measurement sphere ($0 \leq \theta \leq 165^\circ, 0 \leq \varphi \leq 357^\circ$).

Thus, rewriting the transmission coefficients in terms of equivalent dipoles using the proposed method, enhanced with the simplification procedure, can be a good way to express the whole information contained in the transmission coefficients with a limited number of dipole excitations. The rearrangement of the SWE into a multiple SWE expressed in different local coordinate systems (using *a priori* information), which is the principle of our method, is a time consuming procedure (translational and rotational addition theorems). Furthermore, if we want to achieve the accuracy of the SWE with a truncation number $N_{tr} = [kr_{\min}] + 10 = 54$ we need to use a large number of equivalent current sources (large number of unknowns and ill-posed problem). Consequently, the modeling procedure is time consuming for large structures.

Once the AUT equivalent model is achieved, the computation of the radiation pattern at different distances from the AUT is faster than using the SWE. This is due to the fact that, the number of equivalent dipoles is smaller than the number of the AUT transmission coefficients.

In addition, implementing the equivalent model in an EM code, one is able to calculate the E-field propagation properties

in complex and inhomogeneous environments. As an example, one can study the antenna radiation pattern through an interface between the free space and a homogeneous media.

V. CONCLUSION

The presented method aims to substitute the measured AUT by a set of infinitesimal dipoles regularly distributed over the antenna aperture surface, simply from the knowledge of the measured field over a truncated spherical surface. Based on the orthogonal properties of the spherical vector wave functions, the presented matching method has been shown to be a convenient tool to define a set of elementary structures (with $N_{tr} = 1$) from the antenna transmission coefficients. The accuracy of the equivalent model depends on the current sources distribution over an equivalent surface of finite extent which makes the method flexible. This is one of the advantages of the proposed modeling procedure. We have shown from the presented example (X-band horn antenna) that from the collected tangential electric field, the equivalent principle can be applied to an equivalent planar surface of finite size. The equivalent model can reproduce rapidly the actual radiation pattern at different distances from the considered antenna. The overall procedure has been assessed by exploiting two different measurement systems, a spherical one and a planar one, both being in anechoic environments.

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Antenna Modeling Based on a Multiple Spherical Wave Expansion Method: Application to an Antenna Array

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Abstract—A method to derive an equivalent radiation source for planar antennas is presented. This method is based on spherical near-field (NF) data (measured or computed) to ascertain an equivalent set of infinitesimal dipoles placed over the main antenna aperture. These produce the same antenna radiation field, both inside and outside the minimum sphere enclosing the antenna. A spherical wave expansion (SWE) of the NF data is written in terms of infinitesimal dipoles using a transition matrix. This matrix expresses the linear relations between the transmission coefficients of the antenna and the transmission coefficients of each dipole. The antenna *a priori* information are used to set the spatial distribution of the equivalent dipoles. The translational and rotational addition theorems are exploited to derive the transmission coefficients of the dipoles. Once the excitation of each dipole is known, the field at any aspect angle and distance from the antenna is rapidly calculated. Computations with EM simulation data of an antenna array illustrate the reliability of the method.

Index Terms—Antenna modeling, spherical wave expansion, translational and rotational addition theorems.

I. INTRODUCTION

THE antenna modeling in its operating environment is still a challenge using full-wave solutions. Indeed, the cost of such solutions grows exponentially with the size and the complexity of the problem. To cope with these difficulties, a two-steps hybrid approach is proposed. The antenna under test (AUT) is separately characterized and a simple equivalent model is defined. This comprises point sources that emulate the antenna radiation everywhere.

At the same time, when large antennas (antenna array) are hardly characterized by full-wave solutions, the measurement alternative can be of great interest, especially, the spherical multi-probe near-field technique [1]. This technique enables real-time 3D complex characterization all over the sphere surrounding the antenna provided the Nyquist sampling criterion is satisfied [2], [3]. The antenna measured near field data are then used to define an equivalent model that allows the knowledge of the radiated EM field both inside and outside the minimum sphere enclosing the antenna. Thereafter, the antenna

equivalent model is incorporated into an existing EM algorithm to simulate multiple interaction scenes.

However, typical constraints arise, while modeling an antenna by an equivalent current/charge distribution. First, the equivalent model must reproduce the same radiated field as the antenna in near and far field regions. Secondly, the equivalent current/charge constituting the equivalent model must be consistent with the actual antenna current /charge distribution. Thirdly, we have to define a minimization technique that allows an efficient and well-conditioned identification process.

In the literature, different approaches were adopted. In [4]–[9] an electric field integral equation method (EFIE) is developed to relate the measured electric field to equivalent magnetic /electric current on the antenna aperture. The EFIE is solved using the conjugate gradient method. Equivalent currents are written as linear combinations of two-dimensional pulse basis functions, which can be approximated by Hertzian dipoles distributed over any enclosing arbitrary surface. This method has been investigated for near field to far field transformation, inverse electromagnetic radiation applications, antenna diagnosis. In [10] different strategy for antenna diagnosis is proposed. It expresses the relation between the plane wave expansion and the spherical wave expansion, which permits the calculation of the aperture field accurately.

In [11]–[13] evolutionary algorithms were introduced. Using genetic algorithm (GA) technique, the AUT is substituted by a set of infinitesimal dipoles. These are distributed inside the volume enclosing the antenna. The choice of infinitesimal dipole sources is justified by the simplicity of its implementation in any EM code, and the GA shows its ability to determine the type (electric or magnetic), the position, the orientation and the excitation of each dipole. The GA shows some difficulties to determine the optimal solution, when dealing with a large number of dipoles, including convergence issues and prohibitive CPU computing time, while increasing the antenna size. In [14] quantum particle swarm optimization algorithm has been used to define a set of equivalent dipoles for antenna modeling. This method is based on an expansive iterative process, which depends on $7 \times N$ degrees of freedom, where N is the number of dipoles.

In a previous paper [15] the authors have introduced a modeling technique depending on $6 \times N$ unknowns, where N is the number of current sources. It consists in the substitution of the antenna with a set of infinitesimal dipoles distributed over the minimum sphere circumscribing the antenna. These equivalent dipoles distribution is well adapted to build up a

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well-conditioned matrix system but is not convenient to provide the near-field within the minimum sphere for planar antennas for example.

In the present paper, an improvement of the analysis presented in [15] is proposed. Starting with the *a priori* knowledge of the antenna geometry, the equivalent dipoles are placed over the main radiating sources. Then, the spherical wave expansion (SWE) of the antenna radiated field is rewritten as the total contribution of small number of equivalent electric and magnetic dipoles, providing adequate changes of coordinate systems. Once assessed, this set of dipoles allows the knowledge of the near-field information inside the minimum sphere circumscribing the antenna and the EM field can be rapidly calculated at any distances in a straightforward manner (faster than the SWE), using the free space Green's function.

The paper is organized as follows. Section II describes the theoretical procedure of the method. Section III presents the application process of the method for an antenna array. Particular attention is paid to the determination of equivalent dipoles number and locations. Finally, concluding remarks summarizing the potential of this method are provided in Section IV.

All theoretical results are expressed in the S.I. rationalized system with $e^{-j\omega t}$ time dependence.

II. FORMULATIONS AND THEORETICAL DEVELOPMENTS

The complete formulation is given in [15]. We therefore, recall only the essential parts of this development, with a particular attention to the modifications introduced by the new arrangement of elementary source locations.

Here, a current source is defined as a combination of 4 tangential uncoupled and co-localized infinitesimal dipoles: 2 electric and 2 magnetic. We attach a local coordinate system $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)_{1 \leq i \leq L_{\text{source}}}$ to each current source, where the origin $o_i(r_i, \theta_i, \varphi_i)$ coincides with the position of the i th source. Explicitly, we intend to rewrite the spherical wave expansion (SWE) expressed in $(o, \vec{x}, \vec{y}, \vec{z})$ by means of multiple and local SWE that are expressed in $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)_{1 \leq i \leq L_{\text{source}}}$.

Our purpose is to emulate the electromagnetic field radiated from an antenna by a set of current sources (L_{source}) placed over the antenna main radiating surface (Fig. 1). Consequently, the field inside the minimum sphere is reached and the number of equivalent sources required to emulate accurately the antenna E-field, is reduced, while preserving the whole SWE information.

A. Spherical Wave Expansion

In source free region outside the minimum sphere of radius r_{min} circumscribing the antenna, the spherical wave expansion (SWE) of the radiated electric field \vec{E}_{AUT} in spherical coordinates system is expressed in terms of truncated series of spherical vector wave functions [16] as

$$\vec{E}_{AUT}(\vec{r}) = \frac{k}{\sqrt{\eta}} \sum_{j=1}^{J_{\text{max}}} Q_j \vec{F}_j^{(3)}(\vec{r}) \quad (1)$$

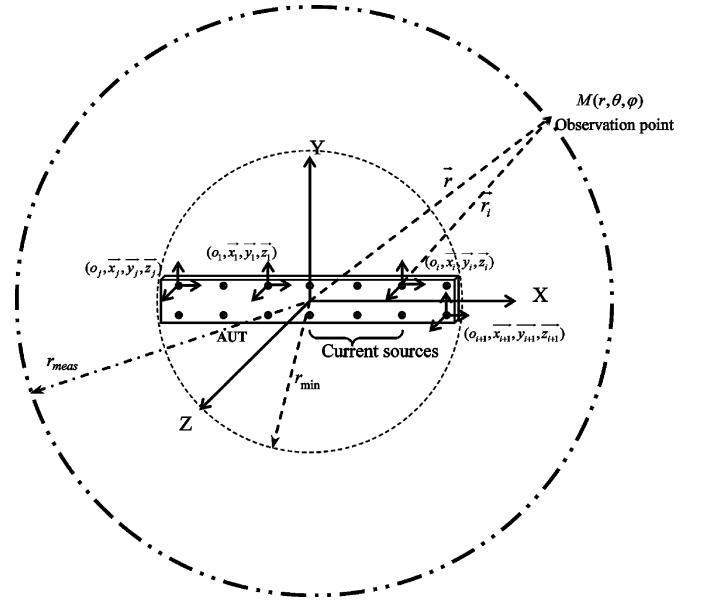


Fig. 1. Measurement sphere of radius r_{meas} , AUT minimum sphere and current sources configuration.

where, Q_j are the transmission coefficients, and $\vec{F}_j^{(3)}(\vec{r})$ are the power-normalized spherical vector wave functions. The truncation number $J_{\text{max}} = 2(N_{tr}(N_{tr} + 2))$ with, $N_{tr} = [kr_{\text{min}}] + 10$ depends mainly on the antenna dimensions and the operating frequency [16]. The Q_j are determined from the knowledge of the tangential components of either \vec{E} or \vec{H} on the measurement sphere. In [16], the Q_j coefficients are determined using the orthogonal properties of spherical wave functions $\vec{F}_j^{(3)}(\vec{r})$. Once these coefficients are known, the field outside the minimum sphere is completely characterized by (1).

The SWE of the field radiated by an infinitesimal x -directed electric dipole $\vec{E}_i^{e,x}$ placed at the origin of $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ is expressed in the spherical coordinate system $\vec{r}_i(r_i, \theta_i, \varphi_i)$ associated with $(o_i, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ as

$$\vec{E}_i^{e,x}(\vec{r}_i) = \frac{k}{\sqrt{\eta}} \frac{\sqrt{2}}{2} \alpha_i^{e,x} \left(\vec{F}_2^{(3)}(\vec{r}_i) - \vec{F}_6^{(3)}(\vec{r}_i) \right) \quad (2)$$

with $\alpha_i^{e,x} = -(k/\sqrt{6\eta\pi})I_i^{e,x}l$.

Expressions for z and y -directed dipoles either electric or magnetic ones are provided in [15].

In the case of a planar antenna, the antenna equivalent dipoles are distributed over the antenna main surface and only the tangential dipoles are considered. Therefore, each current source (i th) is characterized by 4 transmission coefficients $\alpha_i^{e,x}$, $\alpha_i^{m,x}$, $\alpha_i^{e,y}$, $\alpha_i^{m,y}$. We refer to the i th current source by the row vector $S_i = [\alpha_i^{e,x}, \alpha_i^{m,x}, \alpha_i^{e,y}, \alpha_i^{m,y}]$.

B. The Problem Synthesis

Using translational and rotational addition theorems [16]–[18], we express the field radiated by the i th current

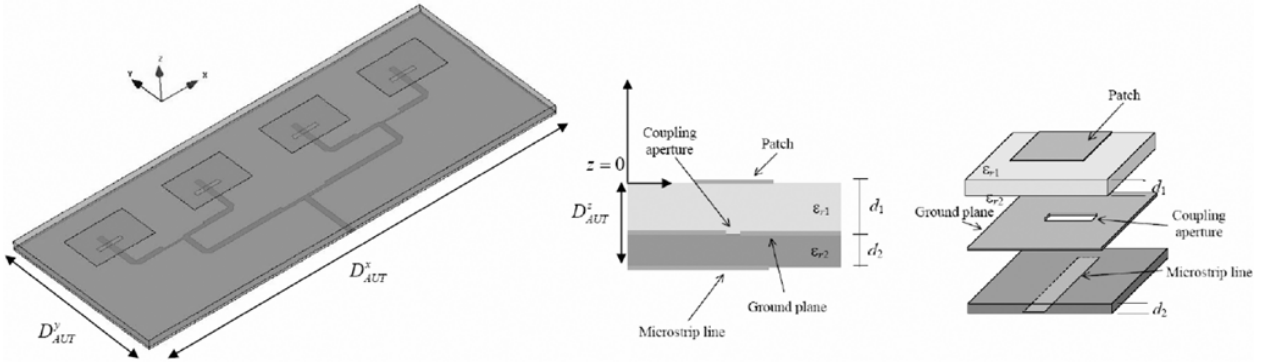


Fig. 2. 3D view of the microstrip array structure. Dielectric stratification (middle) and exploded view of the structure (right).

source in the antenna coordinate system $(o, \vec{x}, \vec{y}, \vec{z})$. Hence, the row vector S_i is expressed in $(o, \vec{x}, \vec{y}, \vec{z})$ by

$$[M_{i,j}]_{j=1 \dots J_{\max}} = [\alpha_i^{e,x} A_{j,i}^{e,x}, \alpha_i^{m,x} A_{j,i}^{m,x}, \alpha_i^{e,y} A_{j,i}^{e,y}, \alpha_i^{m,y} A_{j,i}^{m,y}]_{j=1 \dots J_{\max}} \quad (3)$$

where the coefficients $A_{j,i}^{e,x}, A_{j,i}^{m,x}, A_{j,i}^{e,y}, A_{j,i}^{m,y}$ are developed in [15, Appendix]. Let define $\mathcal{A}_{j,i} = [A_{j,i}^{e,x}, A_{j,i}^{m,x}, A_{j,i}^{e,y}, A_{j,i}^{m,y}]$.

The superposition of the fields radiated by all equivalent dipoles \vec{E}_{mod} is written in $(o, \vec{x}, \vec{y}, \vec{z})$ as

$$\begin{aligned} \vec{E}_{\text{mod}}(\vec{r}) &= \frac{k}{\sqrt{\eta}} \sum_{i=1}^{L_{\text{source}}} \sum_{j=1}^{J_{\max}} \left(\alpha_i^{e,x} A_{j,i}^{e,x} + \alpha_i^{m,x} A_{j,i}^{m,x} \right) \cdot \vec{F}_j^{(3)}(\vec{r}) \\ &= \frac{k}{\sqrt{\eta}} \sum_{j=1}^{J_{\max}} \sum_{i=1}^{L_{\text{source}}} [\mathcal{A}_{j,i}] \cdot S_i^T \cdot \vec{F}_j^{(3)}(\vec{r}). \end{aligned} \quad (4)$$

Identifying \vec{E}_{AUT} with \vec{E}_{mod} , and from (1) and (4) we get

$$\sum_{j=1}^{J_{\max}} Q_j \vec{F}_j^{(3)}(\vec{r}) = \sum_{j=1}^{J_{\max}} \sum_{i=1}^{L_{\text{source}}} [\mathcal{A}_{j,i}] \cdot S_i^T \cdot \vec{F}_j^{(3)}(\vec{r}). \quad (5)$$

Due to the orthogonal properties of $\vec{F}_j^{(3)}$, the transmission coefficients of both sides of (5) have to be equal.

Let us denote $\mathcal{A} = (\mathcal{A}_{j,i})_{\substack{1 \leq i \leq L_{\text{source}} \\ 1 \leq j \leq J_{\max}}}$, $\mathcal{X} = [S_1 \dots S_{L_{\text{source}}}]^T$ and $\mathcal{Q} = [Q_1 \dots Q_{J_{\max}}]^T$ where, where the superscript “ T ” expresses the matrix transpose. Equation (5) becomes

$$\mathcal{A} \cdot \mathcal{X} = \mathcal{Q}. \quad (6)$$

Made up of $4 \times L_{\text{source}}$ columns and J_{\max} lines, \mathcal{A} is the transition matrix, which expresses linear relations between the transmission coefficients of current sources (\mathcal{X}) and the transmission coefficients \mathcal{Q} of the actual antenna. To solve (6), we use the least square (LSQR) code of MatLab to determine the $4 \times L_{\text{source}}$ unknowns composing the row vector \mathcal{X} .

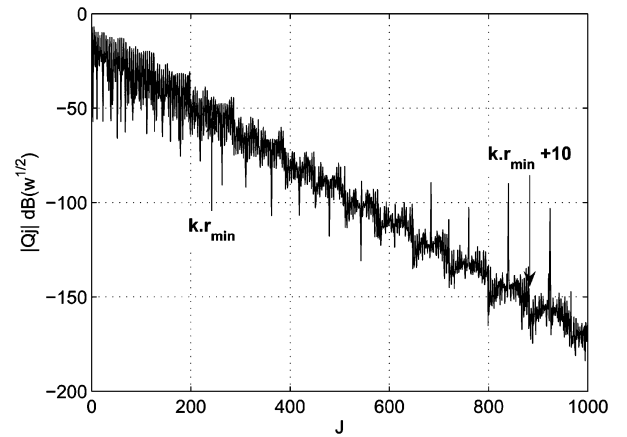


Fig. 3. Magnitude of the antenna array transmission coefficients.

III. RESULTS

A. Example of Equivalent Models of an Antenna Array

In this section, we report a numerical example by simulation of a linear microstrip array of aperture-coupled patches operating at the frequency 5.82 GHz. The antenna dimensions are $D_{\text{AUT}}^x = 0.1464$ m (2.84λ) along the x -axis, $D_{\text{AUT}}^y = 0.06$ m (1.16λ) along the y -axis, and $D_{\text{AUT}}^z = 0.0031$ m along the z -axis with $r_{\min} = 0.085$ m (1.65λ) (Fig. 2). The antenna is composed of a linear array of 4 rectangular patches printed on a dielectric support. Each patch is aperture-coupled to the microstrip line. A common ground plane separates the radiating part from the microstrip feeding network (Fig. 2). The antenna is issued from Antenna Center of Excellence benchmarking data base, where the detailed description is provided in [19].

The antenna equivalent model construction is based on near field data obtained through the electromagnetic simulation of the antenna in a 3D simulator that uses the finite elements method (FEM). We recover at the distance of $r_{\text{meas}} = 15\lambda$ (≈ 0.75 m), the amplitude and phase of both E-field components E_θ and E_φ over a spherical surface ($0 \leq \theta \leq \pi$, $0 \leq \varphi \leq 2\pi$), with the angular resolution of $\Delta\theta = \Delta\varphi = 5$ deg. Based on these data, we assess the antenna transmission coefficients that are presented in Fig. 3. The SWE is truncated at $N_{tr} = [kr_{\min}] + 10 = 20$, so $J_{\max} = 2N_{tr}(N_{tr} + 2) = 880$ transmission coefficients Q_j is considered.

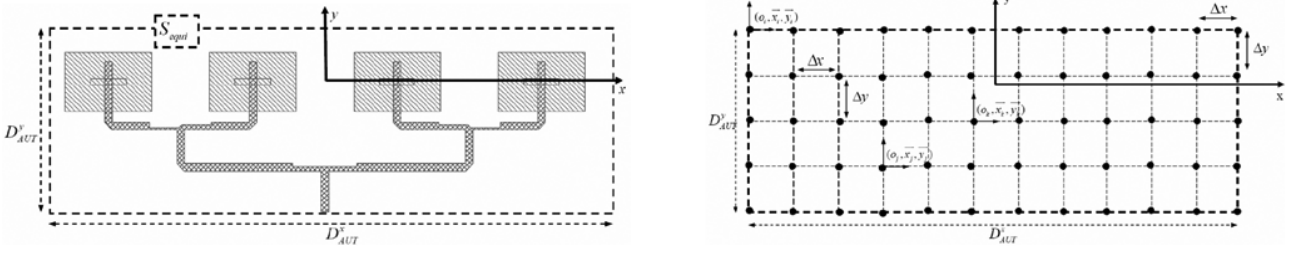


Fig. 4. The description of the equivalent current sources distribution over the antenna surface. The dashed rectangle represents the antenna equivalent surface. (right) The black points represent the equivalent current sources distributed regularly over the antenna equivalent surface respecting the criterion $\Delta x = \Delta y = 0.25\lambda$.

The antenna being planar, we have chosen the natural way to place the current sources (rectangular grid). The antenna is seen as the rectangular surface

$$S_{\text{equi}} = (x, y, z) \\ \in \{-0.073 \text{ m} \leq x \leq 0.073 \text{ m}, \\ -0.042 \text{ m} \leq y \leq 0.017 \text{ m}, z = 0\}$$

over which, the equivalent dipoles are uniformly distributed (Fig. 4). The antenna equivalent model is associated with a spacing criterion Δx and Δy between the current sources. This results in a number of current sources given by $L_{\text{source}} = [(D_{\text{AUT}}^x/\Delta x) + 1] \cdot [(D_{\text{AUT}}^y/\Delta y) + 1]$ where $[X]$ means the integer part of X . Thereafter, the positions of equivalent current sources in $(o, \vec{x}_i, \vec{y}_i, \vec{z}_i)$ are geometrically determined.

We attach a local coordinate system $(o_i, \vec{x}_i, \vec{y}_i)$ to each current source, we determine the corresponding transition matrix \mathcal{A} and we solve (6) to calculate the excitation of each dipole. Then, the E-field radiated from the equivalent dipoles E^{mod} is calculated at different spherical observation domains, near the minimum sphere and in the far field regions and compared with the actual antenna radiation pattern E^{act} using the following error function:

$$RMS_E = \sqrt{\frac{\sum_{\theta, \varphi} |E^{\text{act}}(\theta, \varphi) - E^{\text{mod}}(\theta, \varphi)|^2}{\sum_{\theta, \varphi} |E^{\text{act}}(\theta, \varphi)|^2}}$$

where

$$E = E_\theta, E_\varphi.$$

Comparisons between the antenna radiation pattern (finite elements method) and the radiation pattern of the equivalent dipoles in near and far field regions are shown in Fig. 5 for the $\Delta x = \Delta y = 0.25\lambda$ and $\Delta x = \Delta y = 0.4\lambda$. As it can be seen, the figures show good agreements and the equivalent models radiation pattern fit very well with the one resulted from the spherical wave expansion method. In Fig. 6, the error function is presented as a function of the distance from the antenna. It is seen that the equivalent model corresponding to $\Delta x = \Delta y = 0.25\lambda$ results in $0.23\% \leq RMS_{E_\theta} \leq 0.29\%$ and $0.16\% \leq RMS_{E_\varphi} \leq 0.31\%$. Using the criterion $\Delta x = \Delta y = 0.4\lambda$, the

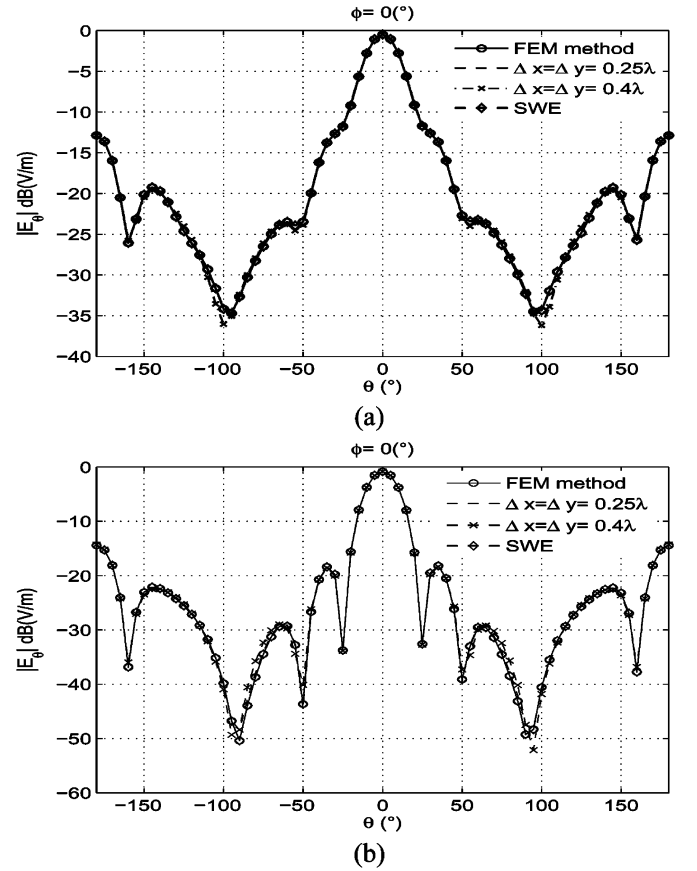


Fig. 5. At the plane $\phi = 0^\circ$, (a) Near-field observed at the distance of 4λ . (b) Far-field radiation pattern. The actual field (FEM) (o-marker line), the field resulted from the SWE (\diamond -marker line), and the field radiated from the equivalent models associated with $\Delta x = \Delta y = 0.25\lambda$ and 0.4λ .

equivalent model results in $0.77\% \leq RMS_{E_\theta} \leq 0.88\%$ and $0.75\% \leq RMS_{E_\varphi} \leq 1\%$. For a smaller spacing $\Delta x = \Delta y = 0.2\lambda$, the accuracy of the antenna equivalent model is improved ($0.04\% \leq RMS_{E_\theta} \leq 0.1\%$, $0.08\% \leq RMS_{E_\varphi} \leq 0.17\%$) but the number of equivalent sources increases ($\Delta x = \Delta y = 0.2\lambda$ corresponds to 90 current sources). These equivalent models reproduce accurately the antenna radiation everywhere outside the minimum sphere ($2\lambda \leq \text{distance} \leq 40\lambda$).

In conclusion, the spacing between the current sources is related to the accuracy of the equivalent model. Consequently, the choice of the spacing between the equivalent current sources $\Delta x, \Delta y$ may be based on a targeted error budget for a given application.

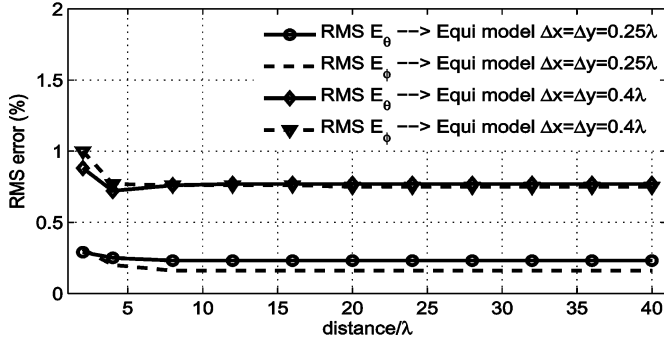


Fig. 6. The error function RMSE versus the distance for $\Delta x = \Delta y = 0.25\lambda$ and 0.4λ associated equivalent models.

From a numerical point of view, the spacing between sources has a direct effect over the condition of the matrix \mathcal{A} in (6). Using a large spacing $\Delta x = \Delta y \geq 0.5\lambda$ the matrix \mathcal{A} is very well-conditioned ($\Delta x = \Delta y = 0.6\lambda$ corresponds to $\text{cond}(\mathcal{A}) = 3.17$), but the antenna equivalent model is not accurate ($11\% \leq \text{RMS}_{E_\theta} \leq 14\%$, $4.2\% \leq \text{RMS}_{E_\varphi} \leq 7.4\%$). Inversely, the conditioning of the matrix \mathcal{A} deteriorates as the value of the spacing Δx and Δy is decreased ($\text{cond}(\mathcal{A})=122$ for $\Delta x = \Delta y = 0.4\lambda$ and $\text{cond}(\mathcal{A}) = 6.574 \cdot 10^5$ for $\Delta x = \Delta y = 0.25\lambda$), while the accuracy of the equivalent model is improved. As a matter of fact, we have to compromise between the accuracy and the conditioning of \mathcal{A} .

Inside the minimum sphere, over the plane $\{-0.075 \text{ m} \leq x \leq 0.075 \text{ m}, -0.045 \text{ m} \leq y \leq 0.02 \text{ m}, z = 0.02 \text{ m}\}$ located at the distance $z = 0.02 \text{ m}$ (0.4λ) from the antenna, the E_y -field component estimated by the 3D simulation software (Finite elements method) and the E_y -field radiated by the equivalent dipoles are presented in Fig. 7. A good agreement is observed, between the actual antenna radiation pattern and the equivalent model radiation pattern. For the near-field presented in Fig. 7, the error function (E_y component) reaches 1.42%.

As presented in Fig. 8, the equivalent models ($\Delta x = \Delta y = 0.25\lambda$ or 0.4λ) provide a good approximation of the field inside the minimum sphere for distances greater than 0.3λ from the antenna. Otherwise, the field in the vicinity of the antenna (distance $< 0.2\lambda$) is mainly reactive and the information of the evanescent modes are not available at the measurement distance ($r_{\text{meas}} = 15\lambda$). Consequently, the equivalent models presents some discrepancies ($\text{RMS}_{E_y} \geq 13\%$ $\Delta x = \Delta y = 0.25\lambda$), while trying to reproduce the E-field in the vicinity of the antenna (distance $= 0.2\lambda$). At the distance of 0.3λ , RMS_{E_y} takes the values 4.32% and 7.28% for the equivalent models associated with $\Delta x = \Delta y = 0.25\lambda$ and $\Delta x = \Delta y = 0.4\lambda$ respectively.

The aim of the examples presented here, is to show the feasibility of the modeling method for the case of a finite open surface (equivalence principle for an open surface) and to figure out how to compromise between the accuracy and the number of equivalent sources (simplicity of the equivalent model). The flexibility of the presented modeling methodology is outlined.

Exploiting the *a priori* information concerning the antenna geometry, we have defined an equivalent model, which reproduces accurately the antenna radiation pattern. We have established that this modeling technique is able to define different

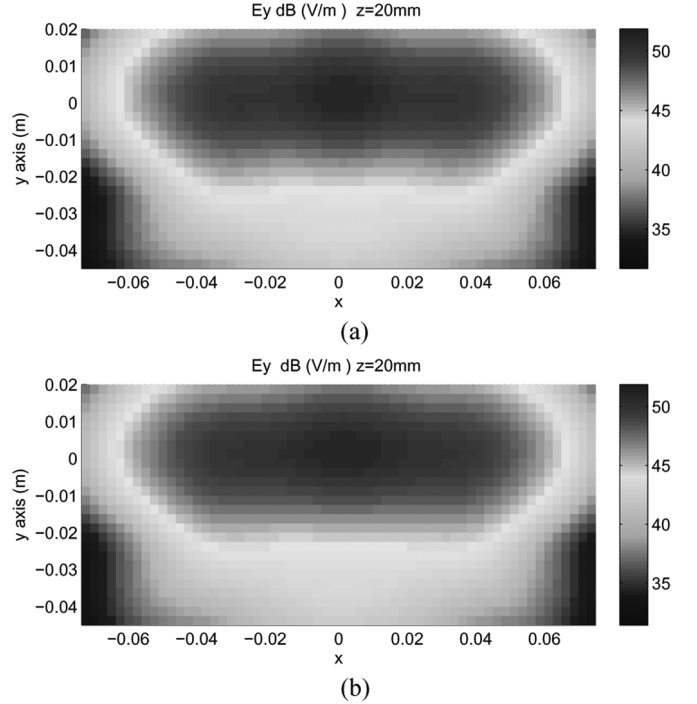


Fig. 7. The aperture field E_y (dB V/m) at the plane $z = 0.02 \text{ m}$ (0.4λ) from the AUT: (a) Resulted from the FEM simulation software. (b) Produced by the equivalent model associated to $\Delta x = \Delta y = 0.25\lambda$.

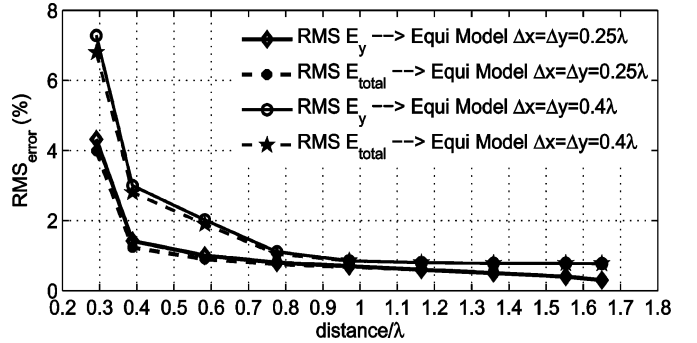


Fig. 8. The error function RMSE evaluated inside the minimum sphere versus the distance for $\Delta x = \Delta y = 0.25\lambda$ and 0.4λ associated equivalent models.

kind of equivalent models where the choice of the spacing between equivalent current sources (complexity) depends on the desired accuracy, so, on the application for which the equivalent model is intended.

Definitely, the criterion $\Delta x = \Delta y = 0.4\lambda$ provides a good compromise between the accuracy, the number of equivalent current sources and for which the matrix \mathcal{A} is well-conditioned.

B. The Reduction of the Number of Equivalent Model Dipoles

Solving (6), we determine the row vectors $S_i = [\alpha_i^{e,x}, \alpha_i^{m,x}, \alpha_i^{e,y}, \alpha_i^{m,y}]$ for all current sources. By means of the normalized row vector $S_{i\text{-nor}}$ expressed as

$$S_{i\text{-nor}} = \frac{1}{2P_i} [|\alpha_i^{e,x}|^2, |\alpha_i^{m,x}|^2, |\alpha_i^{e,y}|^2, |\alpha_i^{m,y}|^2]$$

where

$$P_i = \frac{1}{2} (|\alpha_i^{e,x}|^2 + |\alpha_i^{m,x}|^2 + |\alpha_i^{e,y}|^2 + |\alpha_i^{m,y}|^2) \quad (7)$$

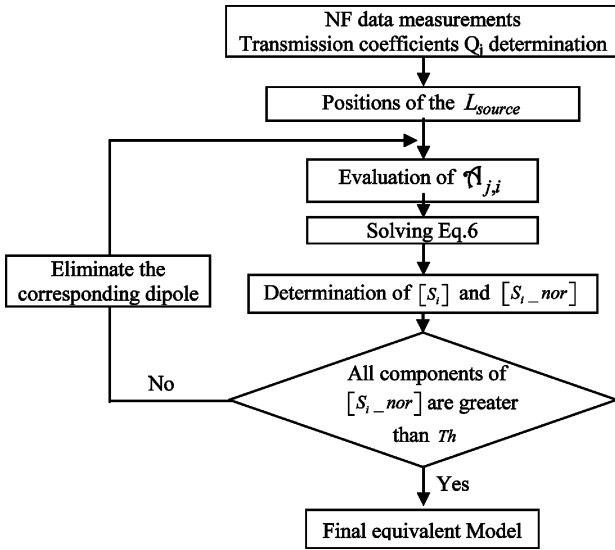


Fig. 9. The description of the reduction of the number of equivalent model dipoles procedure.

we evaluate the power contribution of each dipole constituting the current source. The elements of S_{i_nor} correspond to the power contribution of each dipole in the current source radiation power P_i .

By a proper choice of the threshold Th , we remove the insignificant dipoles, so that the number of the dipoles constituting the equivalent model is reduced. An iterative process is carried out.

First, we choose a spacing criterion between the equivalent current sources. Second, respecting a uniform distribution, we determine the spatial position of the current sources over the planar antenna main radiating surface. Third, we solve (6) taking into account all the dipoles and we evaluate the row vectors $\{S_{i_nor} \text{ for } i \in [1 \dots L_{source}]\}$ so that one can eliminate the dipoles whose corresponding terms in S_{i_nor} are less than Th . Finally, a new solution of (6) is performed providing the new configuration (without the insignificant dipoles). This procedure is pursued as long as non-significant dipoles are detected. The flowchart describing the complete proposed modeling procedure is outlined in Fig. 9.

Results of the antenna equivalent model based on $\Delta x = \Delta y = 0.25\lambda$ are investigated. The reduced equivalent models configurations corresponding to different thresholds Th are presented in Table I. Then, the E-field radiated from these reduced equivalent models is compared with the actual E-field estimated by FEM at different observation distances $2\lambda \leq \text{distance} \leq 40\lambda$.

From Fig. 10, where we present RMS_{E_e} as function of the distance, we observe that the same level of accuracy is maintained after the reduction procedure for $Th = 2\%, 4\%$ ($0.52\% \leq RMS_{E_e}, RMS_{E_\phi} \leq 0.73\%$), while reducing considerably the number of the equivalent dipoles (240 dipoles for $Th = 0\%$ to 126 dipoles for $Th = 4\%$).

Since the actual radiating currents are due to the patches, which are aperture-coupled to the microstrip line, the actual current sources depends on the geometry of the apertures and

TABLE I
EQUIVALENT MODELS AFTER THE REDUCTION PROCEDURE CASE:
 $\Delta x = \Delta y = 0.25\lambda$ FOR $Th = 2, 4\%$

Number of current sources		WITHOUT FILTERING	$Th = 2\%$	$Th = 4\%$
$L_{source} = 60$ current sources	Total	240 dipoles	184 dipoles	126 dipoles
	$dipole^{e,y}$	60	60	50
	$dipole^{e,x}$	60	36	33
	$dipole^{m,y}$	60	44	5
	$dipole^{m,x}$	60	44	38

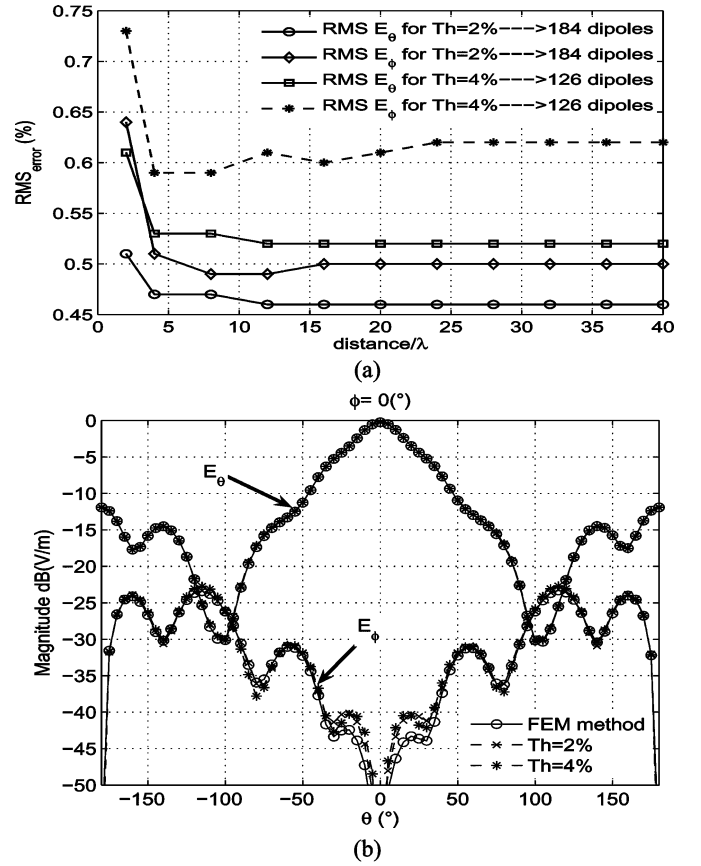


Fig. 10. (a) The error function RMSE corresponding to the thresholds $th = 2$ and 4% as a function of the distance resulting from the reduced equivalent models associated with $\Delta x = \Delta y = 0.25\lambda$. (b) The near field observed at the distance 2λ radiated from the reduced equivalent models compared to the one issued from FEM (actual antenna radiation pattern).

the patches. The apertures are aligned with the x -axis, so the antenna equivalent dipoles must include x -directed magnetic dipoles. On the other hand, the rectangular patches include x -directed and y -directed electric dipoles.

As presented in Table I, after the reduction procedure, the dipoles types constituting the reduced equivalent models are: y -directed, x -directed electric dipoles, and x -directed magnetic dipoles. These dipoles are consistent, from a qualitative point of view, with the antenna actual current sources.

For the equivalent model based on $\Delta x = \Delta y = 0.25\lambda$, we converge toward a reduced configuration of the equivalent

model (using $Th = 4\%$) while fitting accurately (99%) the actual antenna radiation pattern. This threshold allows us to reproduce accurately the radiated E-field as the actual antenna using only 126 dipoles (126 coefficients). In other words, the use of *a priori* information concerning the antenna geometry to define the multiple-SWE can be of great interest since it permits to rewrite the transmission coefficients of the antenna in limited number of equivalent dipole excitations. The reduction procedure tends to cancel dipoles that do not contribute effectively in the radiation pattern.

Contrarily to the spherical wave expansion, the minimum number of transmission coefficients is related to the antenna dimensions. At least we should consider the truncation number $N_{tr} = \lceil kr_{\min} \rceil = 10$, which leads to the use of $J_{\max} = 2N_{tr}(N_{tr} + 2) = 240$ transmission coefficients. The SWE is a very accurate way to describe the radiated field and the accuracy of the SWE cannot be questioned. Nonetheless, the SWE is valid outside the minimum sphere and the implementation of the SWE representation of the radiated field in an electromagnetic code is not trivial. Rewriting the transmission coefficients (SWE) in terms of equivalent dipoles, enhanced with the reduction procedure, can be a good way to express the whole information contained in the spherical wave coefficients in a limited number of dipole excitations.

The rearrangement of the SWE into a multiple SWE expressed in different local coordinate systems (using *a priori* information) is a time consuming procedure (translational and rotational addition theorems). Nevertheless, once the equivalent model is determined, the calculation of the radiation pattern in different distances from the antenna is faster than using the SWE. This is due to the fact that the number of equivalent dipoles is smaller than the number of the transmission coefficients.

IV. CONCLUSION

A method for antenna modeling has been presented. This method is based on the substitution of the original antenna by a set of equivalent infinitesimal dipoles that reproduce the same radiating field. Since the equivalent model includes a set of dipoles placed over the main antenna radiating surface, the information given by these dipoles gives qualitative information concerning the origin of the radiation. Using the antenna *a priori* information, it has been shown that the number of infinitesimal dipoles can be significantly reduced.

The use of the transmission coefficients to elaborate the antenna equivalent model is interesting. This allows a straightforward matching procedure, where the whole antenna intrinsic information are contained in a finite number of complex values. Once the antenna equivalent model is defined, the electromagnetic radiation of the antenna can be calculated at any point both inside and outside (FF region) the minimum sphere except in the reactive region of the antenna. Also, the antenna *a priori* information concerning the AUT geometry have been used to express the multiple spherical wave expansion, which is of great interest since it has allowed the rewrite the SWE with a limited and reduced number of dipole excitations.

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Chapitre 3

Techniques et instrumentation pour la caractérisation en champ proche

« Reviens en toi-même et regarde : si tu ne vois pas encore la beauté en toi, fais comme le sculpteur d'une statue qui doit devenir belle; il enlève une partie, il gratte, il polit, il essuie jusqu'à ce qu'il dégage de belles lignes dans le marbre; comme lui, enlève le superflu, redresse ce qui est oblique, nettoie ce qui est sombre pour le rendre brillant, et ne cesse pas de sculpter ta propre statue. »

Plotin

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3.1 Introduction et contexte

Le laboratoire GeePs dispose d'une importante configuration expérimentale micro-onde. Cette infrastructure a été constituée par J. C. Bolomey et son équipe pour la caractérisation d'antennes en champ proche dès les années 80. La technique de mesure multi-sondes basée sur la diffusion modulée reste un bon exemple des développements accomplis [BOLOMEY et collab. \[1988\]](#), [BOLOMEY et GARDIOL \[2001\]](#). C'est dans ce contexte que j'inscris mes travaux de recherche sur la caractérisation d'antennes en champ proche. Ma contribution est portée sur le développement de nouvelles techniques de mesures flexibles nécessitant moins de contraintes sur les distributions spatiales des points de mesure. Pour cela, nous avons mis en place dans le cadre de la thèse de M. Farouq (2012 - 2015) [FAROUQ \[2015\]](#) dirigée par D. Picard et encadrée par moi-même, une méthode matricielle de calcul du champ lointain à partir de la mesure du champ proche sur une surface quelconque. Nous avons conduit l'approche matricielle exclusivement dans le domaine harmonique.

Parallèlement à ces travaux, je me suis penché sur la caractérisation en champ proche d'antennes ultra-large bande dans le domaine temporel. J'ai constitué l'instrumentation dédiée aux mesures pulsées et les transformations champ proche - champ proche ou lointain pour le traitement des données mesurées. J'ai focalisé mon intérêt sur l'impact du critère d'échantillonnage dans le domaine temporel sur le calcul du champ lointain. Ce travail a été appréhendé dans ses deux facettes numériques et expérimentales.

Dans les parties qui suivent je présenterai une photographie (subjective) de l'état de l'art de cet axe de recherche. Cela concerne une description du principe de la mesure en champ proche, l'échantillonnage régulier et irrégulier sur la surface de mesure et les techniques de transformation champ proche - champ lointain. Il s'agit de resituer mes travaux de recherche dans le spectre de la littérature existante. Je terminerai par une description focalisée sur mes contributions majeures dans cet axe de recherche à la fois à travers la méthode matricielle et la caractérisation des antennes pulsées.

3.2 Méthodes de caractérisation d'antennes en champ proche

La caractérisation d'antennes est un sujet qui intéresse grandement la communauté des antennistes qui exigent des processus expérimentaux peu coûteux, flexibles et fiables pour déterminer le diagramme de rayonnement, le gain et la directivité des antennes. Pour cela, j'ai identifié deux familles de méthodes. La mesure directe, dite de champ lointain, où le rayonnement de l'antenne est capté en zone de Fraunhofer [BALANIS \[1989\]](#). Dans ce cas, on cherche à déterminer la réponse de l'antenne suite à une illumination en onde plane. Cela exige de se placer à grande distance de l'antenne à tester. Cette idée est reprise dans le cadre de la caractérisation dans une base compacte. Celle-ci est constituée d'une source primaire et d'un ou plusieurs réflecteurs qui permettent la création d'une onde plane d'extension limitée à petite distance du réflecteur. En plaçant l'antenne testée dans la zone de l'onde plane, on caractérise son diagramme de rayonnement.

Les méthodes indirectes de caractérisation utilisent les données complexes du champ électromagnétique mesuré dans la zone de Fresnel (en champ proche) pour ensuite calculer le diagramme de rayonnement. La zone de Fresnel étant le domaine situé entre la région du champ proche réactif (voisinage immédiat de l'antenne) et celle du champ lointain de la structure rayonnante (Fraunhofer). Dans mon analyse, je m'intéresse exclusivement au champ proche rayonné où la partie évanescence n'est pas prise en compte.

Dans les méthodes de caractérisation d'antennes en champ proche, le diagramme de rayonnement est calculé via une transformation champ proche - champ lointain. Cette transformation exploite les composantes tangentielles du champ proche mesuré (amplitude et phase) sur une surface échantillonnée régulièrement autour de l'antenne pour déterminer le comportement asymptotique du champ à grande distance. Les méthodes indirectes présentent des avantages en terme de coûts (chambres anéchoïques petites) et de précision (rapport signal à bruit).

Depuis un demi-siècle, des travaux sont développés autour de ces techniques et l'instrumentation dédiée existe selon différents systèmes de coordonnées. Les techniques de transformation champ proche - champ lointain sont largement développées et commentées dans la littérature. Plus récemment, des travaux sur le champ proche exploitent des axes robotisés ou des drones pour déplacer la sonde de mesure pour relever le champ selon une surface quelconque autour de l'objet rayonnant [GEISE et collab. \[2019\]](#).

Pour appréhender la thématique de caractérisation d'antennes en champ proche, il est, selon ma compréhension, indispensable d'aborder deux aspects. Le premier concerne la stratégie d'échantillonnage du champ mesuré. Le deuxième s'intéresse aux méthodes d'exploitation des données mesurées.

Si les stratégies d'échantillonnage du champ proche sont très diverses, les algorithmes de transformation champ proche - champ lointain se basent essentiellement sur deux familles de solutions : la reconstruction

des sources équivalentes et le développement modal du champ mesuré (développement en ondes planes, cylindriques, sphériques, sphéroïdales...). Cette distinction est en réalité fictive car on peut écrire une relation explicite entre les sources et les paramètres du développement modal du champ [HANSEN \[1988\]](#).

3.2.1 Transformation champ proche - champ lointain par reconstruction de sources

La méthode des sources équivalentes traduit explicitement l'équation intégrale du champ électrique (Electric Field Integral Equation) [BALANIS \[1989\]](#). Cette équation relie le champ proche mesuré à une distribution de courants électriques et/ou magnétiques placés sur un domaine occupé par la source. L'écriture intégrale du champ exploite le formalisme de Green pour faire rayonner une distribution donnée de courants électriques/magnétiques. Cette méthode est une transcription du principe d'équivalence de Huygens-Fresnel où des sources réelles ou équivalentes rayonnent la même répartition de champ sur la surface de mesure.

En traduisant l'équation intégrale du champ en une écriture matricielle, on formule un problème linéaire dont la résolution permet de calculer les pondérations complexes des sources équivalentes. Une fois identifiées, ses sources permettent de calculer le champ à proximité de l'antenne (pour des fins de diagnostic) ou à grande distance pour caractériser le diagramme de rayonnement. Les références les plus commentées de cette méthode sont associées de près ou de loin à Tapan K. Sarkar de l'Université de Syracuse, New York. On peut citer [SARKAR et TAAGHOL \[1999\]](#) [TAAGHOL et SARKAR \[1996\]](#) [PETRE et SARKAR \[1992\]](#) [LAS-HERAS et collab. \[2006\]](#) [ALVAREZ LOPEZ et collab. \[2009\]](#). Ces références partagent la même analyse du problème (écriture intégrale du champ). Ces travaux utilisent des courants équivalents magnétiques et/ou électriques avec des distributions bi/tri-dimensionnelles selon les fonctions de base exploitées dans la méthode des moments (fonctions de base de Rao-Wilton-Glisson pour le cas 3D). Le système matriciel qui résulte de cette méthode est résolu via une décomposition en valeurs singulières (SVD) ou la méthode du gradient conjugué.

J'ai identifié dans [PEREZ et BASTERRECHEA \[2007\]](#) un inventaire des méthodes heuristiques exploitées pour la résolution des problèmes de reconstruction de sources : l'algorithme génétique, l'optimisation par essais particuliers et la méthode du recuit simulé. Les comparaisons des champs lointains obtenus par les différentes méthodes donnent un léger avantage à l'optimisation par essais particuliers pour son compromis précision/complexité.

Les méthodes de transformation champ proche - champ lointain basées sur la reconstruction des sources équivalentes n'imposent aucune contrainte sur la distribution spatiale du champ proche. La résolution de la formulation matricielle du problème exploite essentiellement des situations sur-déterminées où le nombre de points de mesure est supérieur au nombre de sources équivalentes à identifier. Ce problème inverse de sources est mal-posé (la régularisation est indispensable) et les solutions obtenues sont au sens des moindres carrés. Se basant sur une identification point à point du champ mesuré, cette méthode présente l'avantage d'être compatible avec l'échantillonnage irrégulier du champ proche. L'information *a priori* sur l'élément rayonnant apporte une grande aide pour le choix des surfaces vectrices des sources équivalentes.

Pour les besoins de nos recherches sur le sujet, nous avons implémenté la méthode des sources équivalentes basée sur l'écriture intégrale du champ. Elle est certes très précise, néanmoins elle est particulièrement sensible au choix du nombre et de la nature des courants équivalents considérés. Cette particularité réduit son caractère systématique en comparaison avec les méthodes basées sur le développement modal du champ mesuré. Les résultats de cette méthode sont présentés et comparés aux techniques modales dans la thèse de M. Farouq (2012 - 2015) [FAROUQ \[2015\]](#).

3.2.2 Transformation champ proche-champ lointain et développement modal

Les algorithmes de transformation champ proche - champ lointain basés sur le développement modal sont étroitement liés à la stratégie d'échantillonnage du champ sur la surface de mesure. Les données mesurées sont exprimées sous la forme d'une somme pondérée de modes. Ces derniers vérifient l'équation de Helmholtz vectorielle ainsi que les conditions aux limites sur la surface de mesure et à l'infini. Les pondérations des modes résultent du calcul d'une intégrale de Fourier des valeurs complexes du champ mesuré. Les systèmes de coordonnées adoptés dans les développements modaux (plan cartésien, cylindrique, sphérique...) sont dictés par les géométries des surfaces de mesure du champ. Les coefficients de pondération de ces développements sont calculés en se basant sur les propriétés d'orthogonalités des modes.

Yaghjian [YAGHJIAN \[1986\]](#) propose une vue d'ensemble des développements canoniques : cartésien, cylindrique et sphérique, ainsi que les critères d'échantillonnage du champ proche associés à chaque système de coordonnées. Le développement en spectre d'ondes planes est abordé dans [JOY et PARIS \[1972\]](#)

PARIS et collab. [1978] KERNS [1981] [GREGSON et collab. [2007]]. Le développement en ondes cylindriques est abordé dans LEACH et PARIS [1973] et celui en ondes sphériques est traité dans LUDWIG [1971] HANSEN [1988].

Plaçons nous dans le cas d'un champ proche mesuré sur une surface plane selon une grille régulière qui respecte le critère d'échantillonnage de Nyquist-Shannon. En effet, pour discrétiser un signal tout en conservant la totalité de l'information, il faut montrer que son spectre est borné. Or, le champ présent au voisinage immédiat d'une antenne n'est pas à spectre borné. Ceci est dû à la présence d'ondes évanescentes JOY et PARIS [1972]. La décomposition du champ mesuré en spectre d'ondes planes met en évidence une part visible du spectre et une autre invisible. La notation « invisible » est justifiée par son caractère fugitif jusqu'à sa disparition complète au bout de quelques longueurs d'ondes, et donc invisible en champ lointain. Ceci peut expliquer les limites de la résolution spatiale obtenue par des techniques de rétro-propagation du champ (techniques holographiques). Les modes évanescents porteurs des fréquences spectrales supérieures ne sont pas présents dans les données mesurées initialement. Il serait invraisemblable de pouvoir les reconstituer par rétro-propagation.

En coordonnées cylindriques, le champ est sondé sur un cylindre infini ou suffisamment grand (mesure en zone de Fresnel). Le champ mesuré est décrit par une série pondérée d'ondes cylindriques (harmoniques cylindriques), où les coefficients de pondération décroissent très vite quand l'ordre des harmoniques augmente. Il existe un ordre N au-delà duquel la pondération des harmoniques cylindriques devient négligeable. Dans LEACH et PARIS [1973] le nombre N de modes nécessaires à la description du champ rayonné est lié aux dimensions de l'antenne (rayon du cylindre minimal qui englobe la source) et à la fréquence de travail. Le nombre N considéré dicte l'échantillonnage angulaire du champ à adopter.

L'échantillonnage en coordonnées sphériques est calqué sur le modèle cylindrique. Le champ rayonné par une antenne contenue dans une sphère minimale de rayon r_{min} est exprimé par une somme pondérée de modes sphériques. Le nombre de ces derniers dépend du rayon r_{min} et de la fréquence de travail. Dans HANSEN [1988], on détermine le nombre de modes $N = \lceil \frac{2\pi}{\lambda} r_{min} \rceil$, où λ est la longueur d'onde. L'analyse du comportement des pondérations des modes sphériques en fonction de leur ordre a été présentée dans un premier temps dans LUDWIG [1971].

Dans la réalité des faits, la surface de mesure du champ proche est limitée et cela impacte irréversiblement la qualité du champ lointain calculé. Les coefficients de pondération des développements modaux résultent d'une transformée de Fourier tronquée. Cette troncature de la surface de mesure délimite l'étendue spatiale où le champ lointain est correctement calculé. En réalité, la troncature impacte toutes les directions spatiales du champ lointain. Les zones où l'information du champ proche venait à manquer sont davantage impactées par rapport aux autres directions spatiales.

Pour le besoin de nos travaux sur le sujet nous avons implémenté les trois développements modaux : plan cartésien, cylindrique et sphérique dans FAROUQ et collab. [2015a] FAROUQ et collab. [2015b] FAROUQ et collab. [2016a]. Les détails mathématiques y sont exposés et commentés. Nous avons calculé les coefficients de pondération de chaque développement en utilisant la transformée de Fourier et en exploitant la méthode matricielle. Cette dernière a été mise au point car nous nous intéressons à une mesure du champ proche sur une surface quelconque ou sur une grille irrégulière où l'utilisation de la transformée de Fourier classique n'est plus possible.

3.2.3 Échantillonnage irrégulier du champ proche et calcul du champ lointain

Quand je me suis intéressé à l'échantillonnage irrégulier du champ proche, je me suis posé la question de l'intentionnalité de ses irrégularités. Le choix de travailler avec un échantillonnage irrégulier (sur un plan) du champ proche est motivé par le désir d'acquérir rapidement et en un minimum de points les données nécessaires et suffisantes pour déterminer le champ lointain. C'est le cas lors des mesures polaires, bi-polaires ou spirales du champ proche.

Les travaux sur l'échantillonnage en champ proche sont reliés directement ou indirectement à Ovidio Bucci de l'Université de Naples Federico II en Italie BUCCI et FRANCESCHETTI [1987] BUCCI et GENNARELLI [1988] BUCCI et collab. [1988] BUCCI et FRANCESCHETTI [1989] BUCCI et collab. [1991a] BUCCI et collab. [1991b] BUCCI et collab. [1998b]. Cette approche simple et intuitive est présentée de façon synthétique dans BUCCI et MIGLIORE [2006]. On y trouve les fondements des assertions sur l'échantillonnage des fonctions à bande limitée. Dans BUCCI et FRANCESCHETTI [1987] et BUCCI et FRANCESCHETTI [1989], Ovidio Bucci introduit la notion de nombre de degrés de liberté d'un champ rayonné. Cela consiste à déterminer le nombre minimum de points de mesure, sur un domaine qui englobe l'élément rayonnant, nécessaires et suffisants pour la description de la fonction du champ avec une précision donnée. Le nombre de degrés de liberté du

champ rayonné est corrélé aux dimensions de l'antenne.

Dans le cas où le nombre de points mesurés était inférieur aux degrés de liberté de l'antenne, la reconstruction du champ, dans le sens du théorème d'échantillonnage, est moins précise. On parle de recouvrement de spectre. Ces résultats reprennent les conclusions de Shannon-Whittaker et ne sont guère en contradiction avec le critère d'échantillonnage de Nyquist-Shannon. O. Bucci exploite dans son formalisme un champ réduit (utilisation d'un terme de phase pour exclure la dépendance à la distance de mesure) qu'il exprime par le biais de fonctions à bande limitée. Cette fois encore Bucci ne fait que réactualiser les travaux de YEN [1956] et KNAB [1979] sur l'échantillonnage non uniforme des fonctions à bande limitée. Le lien avec le développement modal du champ en ondes cylindriques et sphériques est immédiat. Le nombre de modes nécessaires et suffisants pour écrire le champ avec une précision de 99.99% est égale au nombre de degrés de liberté du champ rayonné.

Pour parvenir à utiliser des données de champ proche irrégulièrement distribuées sur un plan, nous avons identifié trois solutions. Une première solution consiste à interpoler les données afin de les reconstruire sur une grille régulière. Le champ lointain résulte de la transformée de Fourier rapide du champ interpolé. À cette occasion, je cite les méthodes d'interpolation de Lagrange (plan polaire) GATTI et RAHMAT-SAMI [1988], la méthode des splines cubiques et l'interpolation à échantillonnage optimal que l'on rencontre dans BUCCI et collab. [1991b] D'AGOSTINO et collab. [2016]. La méthode d'interpolation à échantillonnage optimal exprime le champ comme une série pondérée de fonctions à bande limitée. Parmi ces fonctions on trouve : sinus cardinal, la fonction d'approximation sphéroïdale de Dirichlet ou la fonction d'échantillonnage de Tschebyscheff.

Une deuxième solution a été proposée dans YACCARINO et collab. [1994] pour le calcul du champ lointain à partir des mesures polaires et bi-polaires (sans phase d'interpolation). Cette méthode exploite le développement du champ mesuré en séries de Jacobi-Bessel et la transformation de Fourier Bessel.

La troisième solution est en réalité située entre le développement modal du champ proche et la méthode des sources équivalentes. Il s'agit de la méthode multipolaire rapide (Fast Multipole Method). La FMM a été développée pour résoudre les équations intégrales pour les problèmes de Helmholtz en 2D ROKHLIN [1990] puis étendue au cas 3D dans COIFMAN et collab. [1993]. Song SONG et CHEW exploite cette méthode pour résoudre l'équation intégrale du champ électrique et en propose la généralisation en développant une méthode multipolaire rapide multi-niveaux assurant une complexité de l'ordre de $O(N \log N)$ dans SONG et collab. [1997]. N étant le nombre de points de mesure. Thomas F. Eibert de l'Université de Stuttgart et son équipe ont remis au goût du jour la FMM SCHMIDT et collab. [2008] SCHMIDT et EIBERT [2009] QURESHI et collab. [2013]. Cette méthode, labellisée Fast Irregular Antenna Field Transformation Algorithm (FIAFTA) a été utilisée pour calculer le champ lointain à partir des données mesurées sur des surfaces arbitraires.

Je m'inscris dans la première approche proposée par O. Bucci relative aux techniques d'interpolations du champ proche irrégulier. Nous avons comparé dans FAROUQ et collab. [2014b] les résultats du champ lointain après avoir interpolé les données du champ proche irrégulier. Ce dernier est arbitrairement irrégulier sur un plan où, localement, le critère d'échantillonnage de Nyquist-Shannon n'est pas respecté. Nous avons utilisé les fonctions à bande limitée telle que *Approximate Prolate Series* KNAB [1979] et la fonction sinus cardinal. Nous avons comparé les résultats obtenus avec ceux de la méthode d'interpolation des splines cubiques. Nous en avons conclu que l'interpolation exploitant la fonction sinus cardinal est plus précise notamment pour différents rapports signal à bruit du champ proche irrégulier utilisé.

Dans le cas où les irrégularités sont non-intentionnelles et interviennent suite à des erreurs de positionnement de la sonde de mesure (dysfonctionnement mécanique), le problème est abordé selon deux aspects. Dans le premier, on suppose que l'amplitude des erreurs de positionnement de la sonde est acceptable (critère à définir) et les conséquences en champ lointain de ses incertitudes sont établies à l'image des travaux NEWELL [1988] et MUTH [1988]. Dans le second, les positions exactes des points de mesure sont prises en compte et le champ lointain est calculé en conséquence.

Je positionne mes travaux dans cette deuxième perspective où les erreurs de positionnement de la sonde de mesure affectent les différents systèmes de coordonnées selon les trois variables de l'espace pour le cas planaire (x, y, z) , sphérique (r, θ, ϕ) et cylindrique (ρ, ϕ, z) . Les déplacements de la sonde de mesure ne sont pas réguliers (dans le sens de l'échantillonnage classique) sur la surface de mesure mais la position de chaque point de mesure est connue.

Dans le cas des irrégularités 3D, les techniques d'interpolations du champ proche ne sont plus appliquées et les méthodes avérées pour le calcul du champ lointain sont : la reconstruction des sources équivalentes et

la méthode multipolaire rapide (FMM). Cette dernière n'est qu'une variante de la première, dans la mesure où elle utilise l'équation intégrale du champ comme point de départ. Ma contribution dans cette thématique consiste à proposer une alternative pour exprimer le développement en ondes planes, cylindriques et sphériques à partir des données irrégulières.

3.3 La méthode matricielle pour le calcul du champ lointain

Dans le cadre de la thèse de Mohamed Farouq (2012 - 2015) FAROUQ [2015], nous avons développé la méthode matricielle pour le calcul du champ lointain. Cette méthode nous permet d'exprimer le développement modal du champ proche mesuré sur un domaine quelconque y compris en présence d'irrégularités selon les trois variables de l'espace. Les coefficients modaux du champ mesuré sont obtenus par la résolution d'un système d'équations linéaires reliant les fonctions d'ondes planes / cylindriques / sphériques aux données mesurées. Cette résolution matricielle est faite au sens des moindres carrés. Cette approche du problème a été initiée dans les travaux de Ronald C. Wittmann du NIST (National Institute of Standards and Technology) pour la géométrie plane WITTMANN et collab. [1998] et sphérique WITTMANN et collab. [2004]. Ma contribution a été de généraliser l'analyse matricielle du développement modal du champ pour des géométries quelconques.

Nous présentons la méthode matricielle comme une alternative à la méthode intégrale (sources équivalentes) pour calculer le champ lointain à partir d'une distribution irrégulière de points de mesure où, localement, le critère d'échantillonnage de Nyquist-Shannon n'est pas respecté. D'ailleurs, pour le besoin de notre étude, nous avons développé la méthode des sources équivalentes dans FAROUQ et collab. [2015a] pour pouvoir comparer des diagrammes de rayonnement (champ lointain) déterminés à partir d'un champ proche irrégulier selon (x, y, z) . Nous avons considéré des irrégularités de différentes amplitudes et pour des données ayant différents rapports signal à bruit. Les sources équivalentes sont composées de courants électriques et magnétiques que nous avons placés sur l'ouverture de l'antenne cornet utilisée lors des mesures. Dans FAROUQ et collab. [2015a] (joint avec ce rapport), nous avons utilisé le développement en ondes planes pour déterminer le champ lointain à partir des échantillons irréguliers selon (x, y, z) . L'analyse comparative des méthodes est appréhendée numériquement et expérimentalement. La méthode matricielle donne des résultats très similaires à ceux de la méthode des sources équivalentes. Néanmoins, l'approche modale (méthode matricielle) est plus rapide et repose sur la résolution d'un système d'équations linéaires présentant un meilleur conditionnement. La méthode matricielle est systématique et aucune information *a priori* sur la structure de l'antenne n'est requise. Les deux méthodes sont robustes vis-à-vis du bruit de mesure. Cependant la méthode des sources équivalentes est plus précise pour des rapports signal à bruit inférieurs à 30dB. Je précise que la méthode des courants équivalents est tributaire du nombre de sources considérées. Ce nombre est choisi de façon empirique et non systématique. Les validations expérimentales des deux méthodes se sont déroulées dans la base de mesure en champ proche plan cartésien du laboratoire.

Dans FAROUQ et collab. [2015b], nous avons abordé les irrégularités selon les coordonnées cylindriques (ρ, ϕ, z) et sphériques (r, θ, ϕ) . Nous avons calculé le champ lointain à partir d'une distribution irrégulière de points de mesure autour d'un cylindre ou d'une sphère où, localement, le critère d'échantillonnage de Nyquist-Shannon n'est pas respecté. Dans les développements en ondes cylindriques et sphériques, la méthode matricielle est utilisée pour calculer les coefficients modaux en résolvant un système d'équations linéaires. Nous avons montré que les diagrammes de rayonnement obtenus sont en conformité avec le champ lointain de référence. Cette conformité est tributaire du nombre de modes considérés. Notre étude est validée numériquement et expérimentalement.

Nous avons utilisé dans nos validations expérimentales une antenne réseau (Kathrein) mesurée dans la base cylindrique et une antenne cornet caractérisée dans la base sphérique du laboratoire. Dans les cas des irrégularités 3-D, l'utilisation des techniques d'interpolation du champ proche ne sont plus exploitables et par conséquent la transformée de Fourier rapide n'est pas utilisable pour le calcul du champ lointain. La méthode matricielle est une alternative sérieuse et les développements mathématiques pour les cas cylindriques et sphériques sont présentés et commentés dans FAROUQ et collab. [2015b].

Dans la deuxième phase de l'étude, nous avons démontré la possibilité d'exploiter un développement en ondes planes pour calculer le champ lointain à partir d'un champ proche mesuré sur un hémisphère FAROUQ et collab. [2014a] sur lequel les deux composantes tangentielles (E_θ et E_ϕ) du champ proche ont été mesurées. La précision du champ lointain obtenu dépend du rayon de l'hémisphère, de l'échantillonnage angulaire (régulier) choisi et des pas spectraux adoptés. L'utilisation de la demi-sphère améliore le résultat

du diagramme de rayonnement par rapport à l'erreur de troncature. Il s'agit de démontrer que le développement modal peut être indépendant de la surface de mesure et nous avons choisi le développement en ondes planes pour sa simplicité.

Dans FAROUQ et collab. [2016a] (joint avec ce rapport), nous avons démontré la possibilité d'établir le développement en ondes cylindriques à partir des composantes tangentielles du champ proche mesuré sur un cylindre à section carrée. Pour cela, nous avons exploité notre banc de caractérisation en champ proche plan cartésien pour mesurer les composantes tangentielles du champ (E_x et E_y) sur quatre plans autour de l'antenne réseau étudiée (Kathrein). En exploitant la méthode matricielle, nous calculons le champ lointain à partir des données mesurées. Notre reconstruction du champ lointain est complète en azimut. En revanche, pour la correction de l'erreur de troncature (due à la taille des plans de mesure), nous avons considéré un champ proche mesuré sur un cylindre fermé associé à un développement en ondes sphériques [FAROUQ et collab. [2016b]].

Publications représentatives des travaux sur la méthode matricielle

Les trois articles apportent les compléments théoriques et les détails des validations numériques et expérimentales.

- [RI.15] M. Farouq, M. Serhir, D. Picard, « Antenna far-field assessment from near-field measured over arbitrary surfaces », *IEEE Transactions on Antennas and Propagation*, vol. 64, no. 12, pp. 5122-5130, 2016
- [RI.13] M. Farouq, M. Serhir, D. Picard, « Matrix method for antenna plane wave spectrum calculation using irregularly distributed near-field data : application to far-field assessment », *Progress in Electromagnetics Research M*, 42, pp.71-83, 2015
- [RI.12] M. Farouq, M. Serhir, D. Picard, « Matrix method for far-field calculation using irregular near-field samples for cylindrical and spherical scanning surfaces », *Progress In Electromagnetics Research B*, 63, pp.35-48, 2015

Matrix Method for Antenna Plane Wave Spectrum Calculation Using Irregularly Distributed Near-Field Data: Application to Far-Field Assessment

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Abstract—The matrix method for the calculation of antenna far-field using irregularly distributed near-field measurement data is presented. The matrix method is based on the determination of the plane wave expansion (PWE) coefficients from the irregular near-field samples using a matrix form that connects the radiated field with the corresponding plane wave spectrum. The plane wave spectrum is used to determine the far-field of the antenna under test (AUT). The matrix method has been implemented, and its potentialities are presented. The validations using analytical radiating model (dipoles array) and experimental measurement (X band standard gain horn antenna) results have demonstrated the efficiency and stability of the proposed method.

1. INTRODUCTION

The characterization of antenna far-field pattern is an important task of antenna measurement community. Among the existing methods for far-field assessment, we can find direct and indirect measurement techniques. The direct technique measures the antenna response when illuminated by a plane wave. This is reached directly at the far-field distance or using a compact range. The indirect technique is based on the near-field measurement over canonical surfaces in an anechoic chamber from which the radiation pattern is calculated through near-field to far-field transformation (NFFFT) algorithms. An overview of these algorithms is presented in [1]. The NFFFT are based on an expansion of the electromagnetic field into plane wave (PWE), cylindrical wave (CWE) or spherical wave (SWE). The geometry of the used scanning near-field surface imposes which wave expansion to use [2–7]. The planar near-field measurement technique is used to characterize directive antennas and have enjoyed a great popularity due to the relative simplicity involved in the data acquisition and processing. For the plane wave expansion, Wang [7] has presented a detailed examination of the theory and practice of this technique in 1988.

As an alternative to the wave expansion methods, it is possible to model the AUT through a set of equivalent currents distribution ([8–13]). These currents are set from the near-field measurement data, by solving the inverse radiation problem using the method of moments (MoM) [14]. This is the solution of the integral equations relating the equivalent currents and their radiated field that matches the measured one. From these equivalent currents, it is possible to calculate the AUT radiated field at any point of the space outside the equivalent sources domain. This method needs, however, the AUT a priori information: AUT location, dimensions, type... The quality of the field reproduced (also the far-field) by the equivalent currents depends on the chosen distribution and spatial positioning of these equivalent electric or magnetic sources.

Newell [15] identified 18 error sources in the antenna planar near field measurements procedure and the impact of each near-field error on the far-field pattern was estimated. In [16], Muth evaluated

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the effect of probe displacement errors in the planar near-field measurement procedure on the far-field pattern. In [16] not only the maximum far-field errors caused by systematic position errors or the dependence of this maximum on the wave-number were studied, but also it deduced the exact error that contaminated the far-field when the probe displacements were known. Moreover, the theoretical study was carried out with enough generality so that the analysis can be extended to study position errors in cylindrical and spherical scanning geometries. In [17] authors analyzed the same kind of errors but in bi-polar measurements. The NF measured with position errors may be considered either as an erroneous NF or as a NF measured on an irregular meshing. Considering this second situation the position errors may be random but have to be known (laser interferometry is often used for this purpose).

Dealing with irregular measurement data, the near-field plane wave expansion (PWE) based on 2-D Discrete Fourier Transform (DFT) allowing the far-field calculation is not possible. Using the fact that the measured field is a band limited function [18], the irregular near-field is interpolated to reconstruct a regular near-field grid, then, the data are processed using classical 2-D DFT algorithm to determine the far-field radiation pattern. This solution based on interpolating the near-field has been presented by many authors and exists in a variety of coordinate systems [19–22]. Wittmann in [23] have proposed an approach to deal with these irregularities by combining unequally spaced DFT [24], interpolation, and the iterative conjugate gradient algorithm. More recently an interesting approach has been presented in [25–27] in which a plane wave expansion is used to calculate near-field to far-field transformation with similar paradigm of equivalent-currents method. This method permits the far-field calculation of the antenna from its near-field sampled over irregular grid. To overcome the computational complexity of this method the Multilevel Fast Multipole Method has been used [28] to perform matrix-vector products and accelerate the resolution of the linear system. The implementation of the fast multipole method proved to be a rather difficult task in part because of its complexity and because of the need to optimize all the steps of this method.

In this paper we present the matrix method to calculate the antenna PWE and consequently the far-field using non-uniformly spaced samples distributed randomly in 3-D rectangular surface (irregularities in x , y and z directions), where the Shannon criterion may be broken in limited area of the measurement surface [18]. The principle of the matrix method proposed here is based on the matrix formulation of the PWE that expresses the linear relationship between the irregularly distributed near-field data and the plane wave spectrum. Solving this linear system, we calculate the plane wave spectrum and the far-field in the half space in front of the AUT.

We aim at introducing the matrix technique for PWE calculation where no interpolation is used and no a priori information are needed to calculate the AUT far-field. The equivalent currents technique, where the a priori information is essential, is popular and we are interested in comparing the two techniques using synthetic and experimental data.

The paper is structured in the following sections. In Section 2 the mathematical developments of the matrix method are presented. Results issued from this method are presented in Section 3 for different antennas. The first studied AUT is an array of 10×10 infinitesimal dipoles to generate the irregular near-field data analytically. Then, the far-field results of the proposed method are presented for an X band standard gain horn antenna measured in a planar near-field range at 12 GHz. Finally, a conclusion is outlined in Section 4.

2. THE MATRIX METHOD FOR PLANE WAVE EXPANSION

In this section we present the mathematical basis of the matrix method to calculate the PWE from a rectangular irregular near-field samples distributed randomly in a 3-D domain.

We start by recalling the theory of classical plane wave spectrum for regularly distributed near field data. Then we introduce the matrix method for such case. Finally we extend the matrix method to the case where the samples are irregularly distributed. Throughout this paper, $\exp(j\omega t)$ time dependence is suppressed in all the E -field expressions.

The plane wave expansion (PWE) expresses the electric field vector \vec{E} at the point (x_{meas}, y, z) of the measurement surface as a function of the plane wave spectrum vector \vec{F} . In our analysis we suppose that only the tangential components E_y and E_z of the electric field are measured at the plane x_{meas} where the antenna boresight is in the x -direction. The field \vec{E} is expressed as a function of the plane

wave spectrum \vec{F} following

$$\vec{E}(x_{\text{meas}}, y, z) = \frac{1}{4\pi^2} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \vec{F}(k_y, k_z) e^{-j(k_x x_{\text{meas}} + k_y y + k_z z)} dk_y dk_z \quad (1)$$

$$\vec{F}(k_y, k_z) = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \vec{E}(x_{\text{meas}}, y, z) e^{j(k_x x_{\text{meas}} + k_y y + k_z z)} dy dz \quad (2)$$

In practice, the measurement surface can not be infinite and $y_{\min} \leq y \leq y_{\max}$ and $z_{\min} \leq z \leq z_{\max}$. Accordingly we present the truncated form of (2) as

$$\vec{F}(k_y, k_z) = \int_{y_{\min}}^{y_{\max}} \int_{z_{\min}}^{z_{\max}} \vec{E}(x_{\text{meas}}, y, z) e^{j(k_x x_{\text{meas}} + k_y y + k_z z)} dy dz \quad (3)$$

The near-field is discretized using the sampling steps Δy and Δz over y and z directions.

$$\vec{F}(k_y^n, k_z^m) = \sum_{p=1}^P \sum_{q=1}^Q \vec{E}(x_{\text{meas}}, y_p, z_q) e^{j(k_x^{n,m} x_{\text{meas}} + k_y^n y_p + k_z^m z_q)} \Delta y \Delta z \quad (4)$$

$$\vec{E}(x_{\text{meas}}, y_p, z_q) = \frac{1}{4\pi^2} \sum_{n=1}^N \sum_{m=1}^M \vec{F}(k_y^n, k_z^m) e^{-j(k_x^{n,m} x_{\text{meas}} + k_y^n y_p + k_z^m z_q)} \Delta k_y \Delta k_z \quad (5)$$

where, $k_x^{n,m} = \sqrt{k^2 - (k_y^n)^2 - (k_z^m)^2}$, k is the wave-number in free space, $\Delta k_y = 2\pi/(y_{\max} - y_{\min})$, $\Delta k_z = 2\pi/(z_{\max} - z_{\min})$, $P = (y_{\max} - y_{\min})/\Delta y + 1$, $Q = (z_{\max} - z_{\min})/\Delta z + 1$, $N = (k_{y_{\max}} - k_{y_{\min}})/\Delta k_y + 1$ and $M = (k_{z_{\max}} - k_{z_{\min}})/\Delta k_z + 1$ with $k_{y_{\min}} \leq k_y \leq k_{y_{\max}}$ and $k_{z_{\min}} \leq k_z \leq k_{z_{\max}}$.

The field and the plane wave spectrum can be expressed as follow: $\vec{E} = E_x \vec{e}_x + E_y \vec{e}_y + E_z \vec{e}_z$ and $\vec{F} = F_x \vec{e}_x + F_y \vec{e}_y + F_z \vec{e}_z$.

$$E_y(x_{\text{meas}}, y_p, z_q) = \frac{1}{4\pi^2} \sum_{n=1}^N \sum_{m=1}^M F_y(k_y^n, k_z^m) e^{-j(k_x^{n,m} x_{\text{meas}} + k_y^n y_p + k_z^m z_q)} \Delta k_y \Delta k_z \quad (6)$$

$$E_z(x_{\text{meas}}, y_p, z_q) = \frac{1}{4\pi^2} \sum_{n=1}^N \sum_{m=1}^M F_z(k_y^n, k_z^m) e^{-j(k_x^{n,m} x_{\text{meas}} + k_y^n y_p + k_z^m z_q)} \Delta k_y \Delta k_z \quad (7)$$

We express the matrix form of (6) and (7) as

$$\begin{cases} E_y(x_{\text{meas}}, y_p, z_q) = A_{\text{regular}} F_y(k_y^n, k_z^m) \\ E_z(x_{\text{meas}}, y_p, z_q) = A_{\text{regular}} F_z(k_y^n, k_z^m) \end{cases} \quad (8)$$

with

$$A_{\text{regular}} = \frac{\Delta k_y \Delta k_z}{4\pi^2} \begin{pmatrix} e^{-j(k_x^{1,1} x_{\text{meas}} + k_y^1 y_1 + k_z^1 z_1)} & \dots & e^{-j(k_x^{N,M} x_{\text{meas}} + k_y^N y_1 + k_z^M z_1)} \\ \vdots & \ddots & \vdots \\ e^{-j(k_x^{1,1} x_{\text{meas}} + k_y^1 y_p + k_z^1 z_q)} & \dots & e^{-j(k_x^{N,M} x_{\text{meas}} + k_y^N y_p + k_z^M z_q)} \end{pmatrix} \quad (9)$$

The matrix method consists in the inversion of the equation system (8) where F_y and F_z are the unknowns.

As an extension, we generalize the matrix method to deal with the irregular distributed data over a known grid, which are due to the errors caused by the probe displacement. In other terms, the irregular grid provides from a slightly modified regular grid. The Near-Field (NF) data are collected over a 3-D grid defined by $(x_{\min}^{\text{irreg}} \leq x_l^{\text{irreg}} \leq x_{\max}^{\text{irreg}}, y_{\min}^{\text{irreg}} \leq y_l^{\text{irreg}} \leq y_{\max}^{\text{irreg}}$ and $z_{\min}^{\text{irreg}} \leq z_l^{\text{irreg}} \leq z_{\max}^{\text{irreg}}$), for $1 \leq l \leq L$, with L is the number of measured points.

In this situation (irregular distributed data), (8) can be rewritten as follow:

$$\begin{cases} E_y^{\text{irreg}}(x_l^{\text{irreg}}, y_l^{\text{irreg}}, z_l^{\text{irreg}}) = A_{\text{irreg}} F_y(k_y^n, k_z^m) \\ E_z^{\text{irreg}}(x_l^{\text{irreg}}, y_l^{\text{irreg}}, z_l^{\text{irreg}}) = A_{\text{irreg}} F_z(k_y^n, k_z^m) \end{cases} \quad (10)$$

where,

$$A_{\text{irreg}} = \frac{\Delta k_y \Delta k_z}{4\pi^2} \begin{pmatrix} e^{-j(k_x^{1,1} x_1^{\text{irreg}} + k_y^1 y_1^{\text{irreg}} + k_z^1 z_1^{\text{irreg}})} & \dots & e^{-j(k_x^{N,M} x_1^{\text{irreg}} + k_y^N y_1^{\text{irreg}} + k_z^M z_1^{\text{irreg}})} \\ \vdots & \ddots & \vdots \\ e^{-j(k_x^{1,1} x_L^{\text{irreg}} + k_y^1 y_L^{\text{irreg}} + k_z^1 z_L^{\text{irreg}})} & \dots & e^{-j(k_x^{N,M} x_L^{\text{irreg}} + k_y^N y_L^{\text{irreg}} + k_z^M z_L^{\text{irreg}})} \end{pmatrix} \quad (11)$$

$$E_y^{\text{irreg}} = \begin{pmatrix} E_y^{\text{irreg}}(x_1^{\text{irreg}}, y_1^{\text{irreg}}, z_1^{\text{irreg}}) \\ \vdots \\ E_y^{\text{irreg}}(x_L^{\text{irreg}}, y_L^{\text{irreg}}, z_L^{\text{irreg}}) \end{pmatrix}, \quad E_z^{\text{irreg}} = \begin{pmatrix} E_z^{\text{irreg}}(x_1^{\text{irreg}}, y_1^{\text{irreg}}, z_1^{\text{irreg}}) \\ \vdots \\ E_z^{\text{irreg}}(x_L^{\text{irreg}}, y_L^{\text{irreg}}, z_L^{\text{irreg}}) \end{pmatrix} \quad (12)$$

$$F_y = \begin{pmatrix} F_y(k_y^1, k_z^1) \\ \vdots \\ F_y(k_y^N, k_z^M) \end{pmatrix}, \quad F_z = \begin{pmatrix} F_z(k_y^1, k_z^1) \\ \vdots \\ F_z(k_y^N, k_z^M) \end{pmatrix} \quad (13)$$

In our problem the near-field measurement is devoted to antenna far-field determination. Indeed, we take into account only the propagating plane waves ($k_x^{n,m} = \sqrt{k^2 - (k_y^n)^2 - (k_z^m)^2}$ real) for the calculation of the plane wave spectrum F . This stay true, since the distance between the probe and the antenna under test is beyond the evanescent near-field zone. As a consequence, the spatial sampling steps are chosen at a value slightly lower than half a wavelength.

Equation (10) is over determined since the number of equations exceeds the number of unknowns, we look for the least square solution (LSQR) of (10). The LSQR solution passes by the solution of (14).

$$\begin{aligned} A_{\text{irreg}}^H E_y^{\text{irreg}} &= A_{\text{irreg}}^H A_{\text{irreg}} F_y \\ A_{\text{irreg}}^H E_z^{\text{irreg}} &= A_{\text{irreg}}^H A_{\text{irreg}} F_z \end{aligned} \quad (14)$$

The operator A_{irreg}^H is the Hermitian (conjugate) transpose of A_{irreg} . The least square solution F_y and F_z of (14) minimize $\|A_{\text{irreg}}^H E_y^{\text{irreg}} - A_{\text{irreg}}^H A_{\text{irreg}} F_y\|$ and $\|A_{\text{irreg}}^H E_z^{\text{irreg}} - A_{\text{irreg}}^H A_{\text{irreg}} F_z\|$. In the given example, different direct and iterative methods, the pseudo inverse, conjugate gradient (CG) and LSQR method [29] have been applied. LSQR and CG achieve a small residual of the same order, whereas LSQR shows a slightly faster convergence. The matrix inversion using the pseudo inverse method reveals, as expected, very long computation time, even for rather small matrices. The LSQR method is based on the bidiagonalization procedure of Golub and Kahan [29]. It is analytically equivalent to the standard method of conjugate gradients, but possesses more favorable numerical properties.

Once F_y and F_z are determined, the normal component F_x of the vector spectrum \vec{F} is calculated using the property that the region contains no free charges ($\text{div}(\vec{E}) = 0$)

$$\begin{aligned} \vec{k} \cdot \vec{F} &= 0 \\ k_x F_x + k_y F_y + k_z F_z &= 0 \\ F_x &= -\frac{k_y F_y + k_z F_z}{k_x} \end{aligned} \quad (15)$$

Once the measured near-field plane wave spectrum is calculated the far-field in the spherical coordinates system (r, θ, ϕ) is directly determined based on the following expression

$$\vec{E}_{\text{far-field}}(r, \theta, \phi) = \frac{jk \sin \theta \cos \phi e^{-jkr}}{r} \vec{F}(k \sin \theta \sin \phi, k \cos \theta) \quad (16)$$

As it can be seen the mathematical development of the matrix method for the plane wave expansion from irregular near-field data is easy to implement.

3. RESULTS

The matrix method is tested in two situations. In the first situation a numerical radiating antenna is considered. It is composed of 10×10 infinitesimal electric dipoles. In this case, the NF irregular

positions are modeled considering the following procedure. First, we create a regular grid in both y and z directions using respectively the sampling steps Δy and Δz . Then, we add random functions (Random_y , Random_z defined in Matlab) varying between -1 and $+1$. These functions allow us to create an irregular grid in two dimensions. The irregularities are controlled using the weighting factors χ_y and χ_z . Finally, to generate a three dimensions irregularities we consider a given x_{meas} to which we add a random function Random_x varying between 0 and 1 and multiplied by a controlled irregularity factor χ_x . This procedure is formulated in (17)

$$\begin{aligned} y_{l(p,q)}^{\text{irreg}} &= p\Delta y + \text{Random}_y\chi_y \\ z_{l(p,q)}^{\text{irreg}} &= q\Delta z + \text{Random}_z\chi_z \\ x_{l(p,q)}^{\text{irreg}} &= x_{\text{meas}} + \text{Random}_x\chi_x \end{aligned} \tag{17}$$

In the second situation, we measure an X-band standard gain horn antenna in SUPELEC planar near-field range. This measurement setup allows only the near-field measurement over regular grid at certain distance from the AUT. In order to simplify the experimental validation of the matrix method, we just consider the irregularities created in x direction, because the majority of the existing methods are not able to take into account this irregularity. This is done by measuring the near field in different planes $x = x_{\text{meas}}$. Then, experimental irregular near field data are constructed by choosing randomly the samples from the different measurement planes for each (y_p, z_q) .

In both situations, the far-field pattern is calculated using the matrix method, equivalent currents method and by DFT while ignoring irregularities and the different obtained far-field patterns are compared with the reference far-field (Fig. 1). In the numerical study the reference far-field is calculated directly. The reference far-field for the experimental study is obtained by DFT of the measured regular near-field at $x_{\text{meas}} = 12 \text{ cm}$.

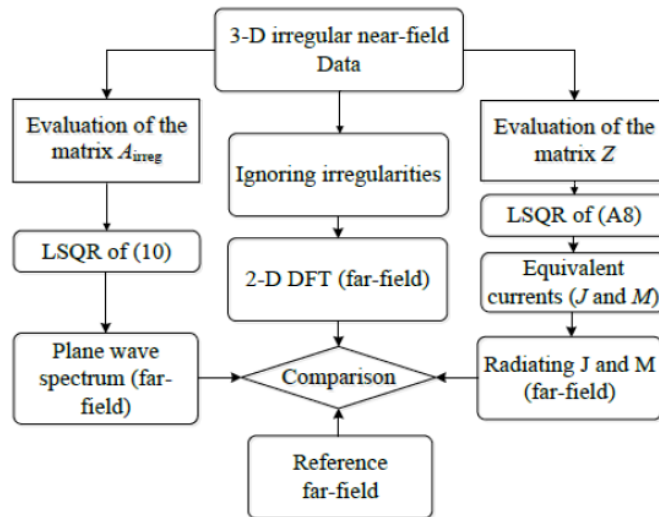


Figure 1. Validation procedure of the matrix method.

3.1. Numerical Study

The 10×10 z -dipoles array under consideration is composed of equally excited dipoles distributed regularly over the plane yOz . The dipoles are $\lambda/2$ spaced along the y and z directions (λ is the working wavelength). The near-field measurement surface is centered on the axis of the AUT and the field is collected over a square surface of side 20λ ($y_{\text{max}} = -y_{\text{min}} = z_{\text{max}} = -z_{\text{min}} = 10\lambda$). The near field is collected at a distance of x^{irreg} with $x_{\text{meas}} = \lambda$ from the antenna at the frequency 2 GHz . The finite size

of the measurement surface limits the reliable region of the calculated far-field. The reliable far-field angular region (θ_{valid} and ϕ_{valid}) is computed as

$$\begin{cases} \theta_{\text{valid}} = \tan^{-1} \left(\frac{(z_{\text{max}} - z_{\text{min}}) - d}{2x_{\text{meas}}} \right) \\ \phi_{\text{valid}} = \tan^{-1} \left(\frac{(y_{\text{max}} - y_{\text{min}}) - d}{2x_{\text{meas}}} \right) \end{cases} \quad (18)$$

where, d is the diameter of the AUT and x_{meas} is the distance between the AUT and the measurement plane. In this case of study $\theta_{\text{valid}} = 80^\circ$ and $\phi_{\text{valid}} = 80^\circ$.

In order to evaluate the accuracy of the proposed method, we quantify the deviation between the co-polar far-field E_{calc} issued from the different methods (matrix method and equivalent currents method) and the co-polar of the exact one E_{exact} calculated directly from the dipoles radiation pattern, by means of the root mean square difference of the two far-fields. This quantity is evaluated either on the 2D or 1D far-field pattern. For example, the 2D far-field pattern deviation, i.e., error, is defined as below:

$$\text{Error}(\%) = 100 \sqrt{\frac{\sum_{\theta, \phi} |E_{\text{calc}}(\theta, \phi) - E_{\text{exact}}(\theta, \phi)|^2}{\sum_{\theta, \phi} |E_{\text{exact}}(\theta, \phi)|^2}}, \quad 10^\circ \leq \theta \leq 170^\circ \quad \text{and} \quad -80^\circ \leq \phi \leq 80^\circ \quad (19)$$

The results are presented and compared with the reference field (E_{exact}) in Figs. 2(a)–(b)–(c)–(d). Fig. 2 shows the far-field pattern calculated from a near field measured in irregular positions using $\chi_{y,z} = \chi_x = \lambda/10$. This means that the irregularity magnitude is $\lambda/5$ peak to peak. It is seen that the use of the matrix method based on the PWE gives a comparable accuracy with the equivalent currents method. Both methods fit very well the actual radiated far-field. In the case where we don't take into account the irregularities, we use directly the discrete Fourier transform to calculate the plane wave spectrum and the far-field radiation pattern. This is presented in Fig. 2(d). The differences between this far-field and the actual one are important.

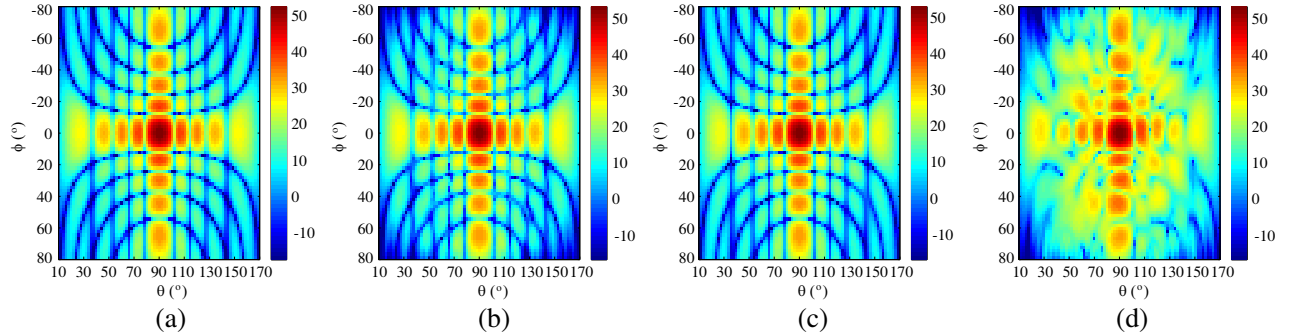


Figure 2. Amplitude of the radiation pattern (co-polar) as a function of θ and ϕ . (a) Reference pattern, (b) matrix method, (c) equivalent currents method, (d) ignoring irregularities.

The principal plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ of the co-polar component of the E -field pattern are presented in Fig. 3 for irregularities associated with $\chi_{y,z} = \chi_x = \lambda/10$. This shows the comparison between the reference and the far-field pattern obtained using the matrix method, equivalent currents method and when the irregularities are ignored. We note that both methods present a good agreement with the reference radiation pattern with an error of order 1.1% for $\theta = 90^\circ$ and 1.6% for $\phi = 0^\circ$ for the matrix method. The use of the equivalent currents method leads to errors of 1.6% for $\theta = 90^\circ$ and 1.4% for $\phi = 0^\circ$. In the situation where the irregularities are not taken into account, significant errors in the side lobes of the radiation pattern are induced.

We study the influence of the weighting factor $\chi_{y,z}, \chi_x$. Fig. 4 shows the far-field pattern determined using irregular near-field with $\chi_{y,z} = \chi_x = \lambda/5$. In this figure we present the results of the calculated

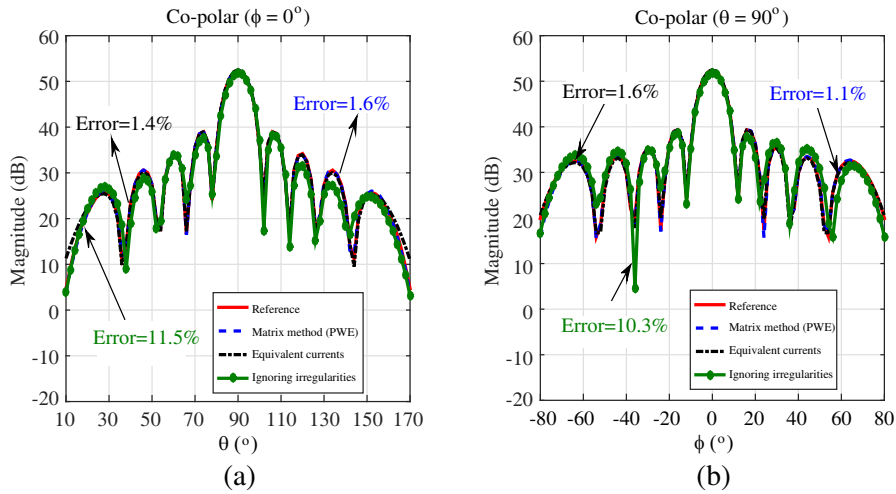


Figure 3. The far-field (co-polar) comparison at the principal plane cuts. The comparison includes: reference radiation pattern, matrix method, equivalent currents methods and when ignoring irregularities.

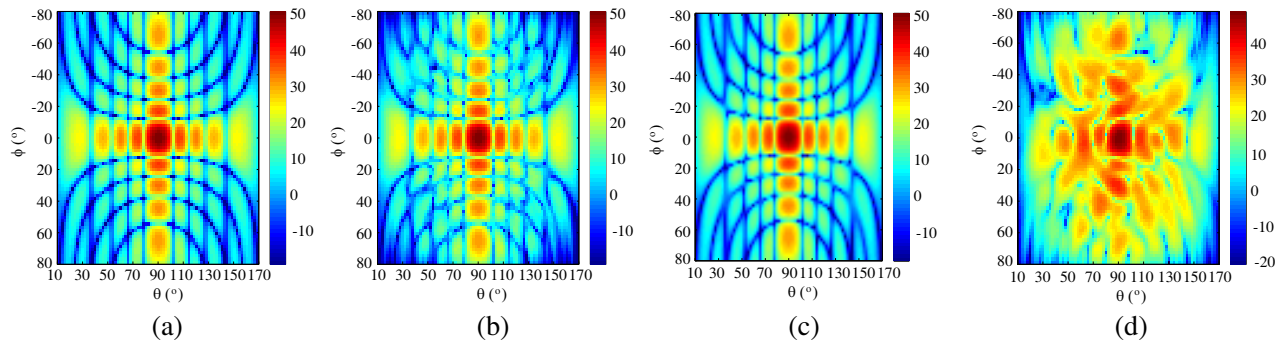


Figure 4. Amplitude of the radiation pattern (co-polar) as a function of θ and ϕ . (a) Reference pattern, (b) matrix method, (c) equivalent currents method, (d) ignoring irregularities.

far-field using the matrix method, the equivalent currents method and when ignoring the irregularities. As it can be seen, increasing the weighting factor influences directly the far-field accuracy especially in the low level far-field areas. Moreover, in the case where we ignore the irregularities the radiation pattern is strongly modified.

In Fig. 5 we present the plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ for $\chi_{y,z} = \chi_x = \lambda/5$. In these plane cuts the error between the reference far-field and the far-field pattern obtained using the matrix method is of order 2.3% for $\theta = 90^\circ$ and 1.4% for $\phi = 0^\circ$. The equivalent currents method presents the errors 1.9% for $\theta = 90^\circ$ and 1.5% for $\phi = 0^\circ$ plane cuts. Considering $\chi_{y,z,x} = \lambda/5$ makes the determination of the far-field pattern in the case where irregularities are not taken into account very critical (green curve). This leads to high errors in the main and side lobes of order 31.4% for $\theta = 90^\circ$ and 31.1% for $\phi = 0^\circ$.

Finally, we are interested in studying the effect of the signal to noise ratio (SNR) over the stability of the matrix and the equivalent currents methods. In fact, this can influence the accuracy of the matrix inversion when calculating the plane wave spectrum or for the equivalent currents determination. To do so, the near-field calculated at irregular positions $(x_l^{irreg}, y_l^{irreg}, z_l^{irreg})$ with $\chi = \lambda/10$ and $\chi = \lambda/5$ is contaminated with a controlled white Gaussian noise to reduce the SNR.

Figure 6 presents the error values for $\chi = \lambda/10$ and $\lambda/5$ as a function of the near-field data SNR for the far-field calculated over a hemi-sphere (2π steradian). It is seen from Fig. 6 that the error takes

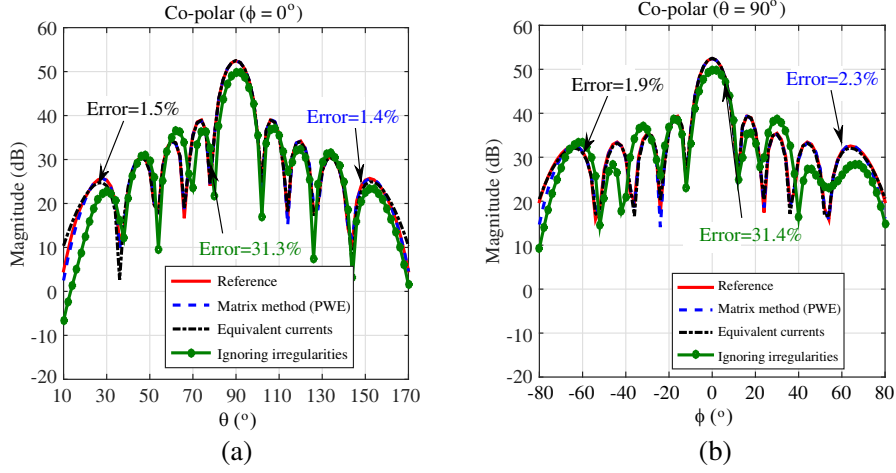


Figure 5. Principal far-field (co-polar) plane cuts. The comparison includes: reference radiation pattern, matrix method, equivalent currents methods and when ignoring irregularities.

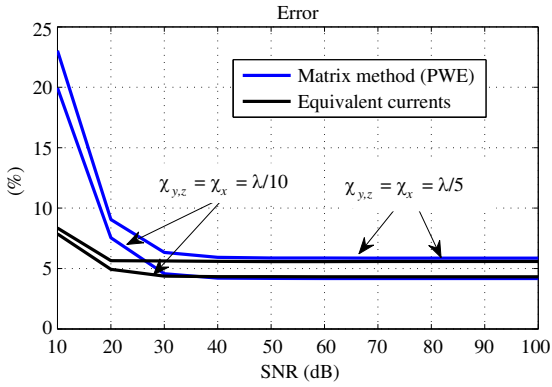


Figure 6. 2D error values for $\chi_{y,z} = \chi_x = \lambda/10$ and $\chi_{y,z} = \chi_x = \lambda/5$ versus the SNR for the matrix and the equivalent currents methods.

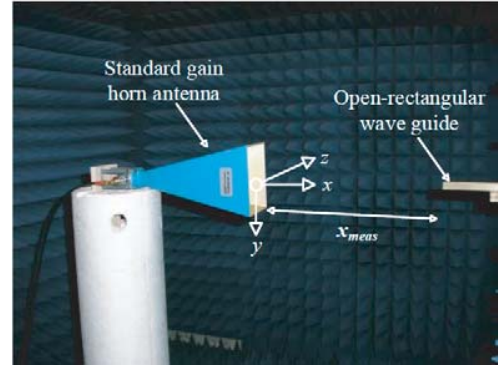


Figure 7. The horn antenna measured in the planar near-field facility.

high values for low SNR near-field data. However, beyond the threshold of 30 dB the error level stays stable. Moreover, we can see that increasing χ changes slightly the far-field results for both methods (+3%). These increasing values influence the low level far-field areas.

3.2. Experimental Study

Our purpose is to illustrate the capability of the matrix method for the determination of the far-field of a standard gain pyramidal horn antenna (Fig. 7) measured in SUPELEC planar near-field setup. We consider the operating frequency of 12 GHz. The horn aperture dimensions are 13 cm \times 9 cm and the measurement has been performed using an open-ended rectangular waveguide.

Using the planar translation axes set in the anechoic chamber we measure the tangential components of the near-field E_y and E_z radiated by the horn antenna over a regular grid ($-60 \text{ cm} \leq y_{\text{meas}} \leq 60 \text{ cm}$ and $-30 \text{ cm} \leq z_{\text{meas}} \leq 30 \text{ cm}$) with $\Delta y = \Delta z = \lambda/2.5$ sampling criterion in each dimension over 5 different planes ($x_{\text{meas}} = 10 \text{ cm}, 11 \text{ cm}, 12 \text{ cm}, 13 \text{ cm}$ and 14 cm). Then, the irregular near-field is constructed by choosing randomly the near-field samples from the 5 measurement grids for each (y, z) positions.

Figure 8 presents the near-field magnitude for the case of regularly distributed near-field data ((a)

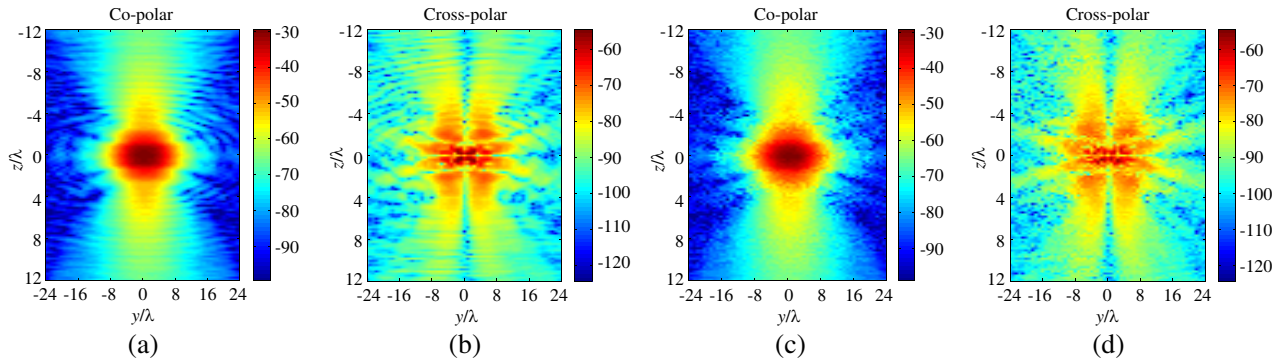


Figure 8. Near-field magnitude measured (a), (b) over a regular planar surface and (c), (d) over irregular surface.

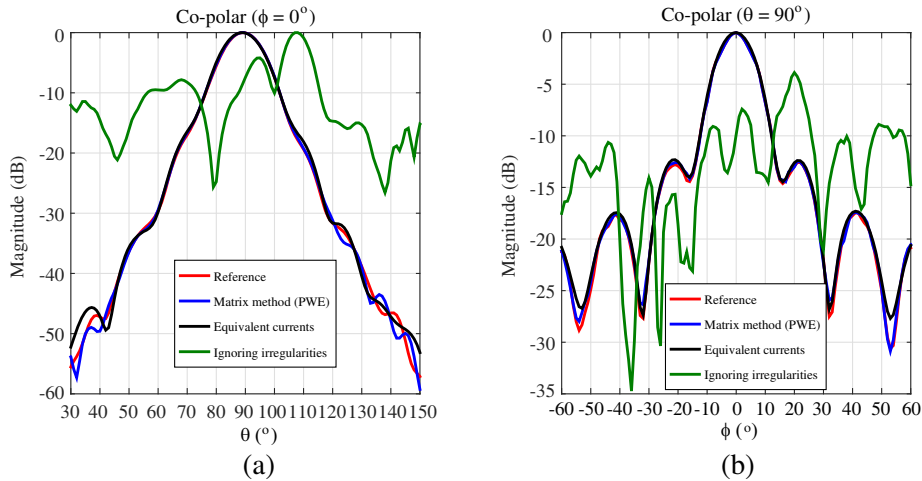


Figure 9. Far-field (co-polar) comparison of the principal plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ in the far-field validity zone ($\theta_{\text{valid}} \approx 60^\circ$ and $\phi_{\text{valid}} \approx 60^\circ$). The comparison includes: reference radiation pattern, the matrix method, the equivalent currents method and when ignoring irregularities.

Table 1. Computation time (s) and condition number for matrix method and equivalent currents method.

	Computation time		Condition number	
	Matrix method	Equivalent currents	Matrix method	Equivalent currents
Numerical validation	6.8	30	2.2	$3 \cdot 10^{11}$
Experimental validation	572	1628	2.7	$1.4 \cdot 10^{15}$

co and (b) cross polarization) and for the case of irregularly distributed data ((c) co and (d) cross polarization).

The far-field calculated using the matrix method is compared with the one calculated directly using DFT of regular near-field data measured at $x_{\text{meas}} = 12$ cm. This is considered as the reference far-field pattern. The green curve is associated with irregular near-field data to which a classical near-field to far field transformation is applied while neglecting the irregularities and by considering $x_{\text{meas}} = 12$ cm (medium value). For the equivalent currents assessment (electric and magnetic) we have defined a planar surface at the horn antenna aperture. The dimensions of this source surface are $-9 \text{ cm} \leq y_s \leq 9 \text{ cm}$ and $-6 \text{ cm} \leq z_s \leq 6 \text{ cm}$ and the number of equivalent currents to assess is $20 \times 30 = 600$.

The comparison concerns the principal plane cuts as shown in Fig. 9. The far-field radiation patterns when neglecting irregularities are strongly affected. Using the proposed matrix method and the equivalent currents method the far-field results fits very well the reference one for the co-polarization plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$.

To get more insight into the efficiency of the proposed matrix method, we have presented in Table 1 the computation times and the condition numbers of the proposed method and the equivalent currents method. As it can be seen the matrix method converges rapidly to the desired solution and presents a low condition number which means that the problem is well posed.

4. CONCLUSION

The matrix method allowing the plane wave expansion of near-field over 3-D irregular grid has been proposed. The technique has shown a good accuracy and high calculation stability. The numerical validation examples have shown that for the near-field data measured at known irregular positions, the far-field can be accurately calculated. Experimental results have been presented using a standard gain horn antenna measured in the SUPELEC planar near-field facility. It has been shown that the radiation pattern of the antenna under test can be reconstructed accurately using the proposed method. However, in the case where we do not take into account the irregularities, the radiation pattern is strongly modified. The matrix method has shown interesting features: a low computational complexity, implementation simplicity, fast convergence to the solution, low condition number and also no a priori information about the AUT is needed in comparison with the well-known equivalent currents method.

APPENDIX A. THE EQUIVALENT CURRENTS METHOD

According to the equivalence theorem, the field radiated by an antenna can be reproduced from an equivalent set of electric and magnetic equivalent currents over the virtual surface S' placed in the aperture of the radiating antenna (Fig. A1). The electric field integral equation [30] expresses the AUT radiated field, in the half space in front of the AUT ($x > 0$), as the superposition of the contribution of electric (J) and magnetic (M) currents.

$$\vec{E}_{\text{meas}}(\vec{r}) = \vec{E}_J(\vec{r}) + \vec{E}_M(\vec{r}) \quad (\text{A1})$$

$$\vec{E}_J(\vec{r}) = \frac{-j\eta}{4\pi k} \int \int_{S'} k^2 \vec{J}(\vec{r}') \frac{e^{-jkR}}{R} + \nabla \left(\nabla \left(\vec{J}(\vec{r}') \frac{e^{-jkR}}{R} \right) \right) dS' \quad (\text{A2})$$

$$\vec{E}_M(\vec{r}) = \frac{-1}{4\pi} \nabla \times \int \int_{S'} \vec{M}(\vec{r}') \frac{e^{-jkR}}{R} dS' \quad (\text{A3})$$

In (A2) and (A3) $\vec{r}(x, y, z)$ is the position of the observation point and $\vec{r}'(x', y', z')$ is the position of the source point (equivalent currents). $R = |\vec{r} - \vec{r}'|$ is the distance between these two points. The equivalent currents are expressed as $\vec{M} = M_y \vec{e}_y + M_z \vec{e}_z$ and $\vec{J} = J_y \vec{e}_y + J_z \vec{e}_z$.

The Cartesian components of the electric field can be expressed according to [30] as

$$E_{M_y}(x, y, z) = -\frac{1}{4\pi} \int_{y'} \int_{z'} (x - x') M_z \frac{1 + jkR}{R^3} e^{-jkR} dy' dz' \quad (\text{A4})$$

$$E_{J_y}(x, y, z) = -\frac{j\eta}{4\pi k} \int_{y'} \int_{z'} (G_1 J_y + (y - y') G_2 ((y - y') J_y + (z - z') J_z)) e^{-jkR} dy' dz' \quad (\text{A5})$$

$$E_{M_z}(x, y, z) = -\frac{1}{4\pi} \int_{y'} \int_{z'} (x' - x) M_y \frac{1 + jkR}{R^3} e^{-jkR} dy' dz' \quad (\text{A6})$$

$$E_{J_z}(x, y, z) = -\frac{j\eta}{4\pi k} \int_{y'} \int_{z'} (G_1 J_z + (z - z') G_2 ((y - y') J_y + (z - z') J_z)) e^{-jkR} dy' dz' \quad (\text{A7})$$

with $G_1 = \frac{-1 - jkR + k^2 R^2}{R^3}$ and $G_2 = \frac{3 + 3jkR - k^2 R^2}{R^5}$.

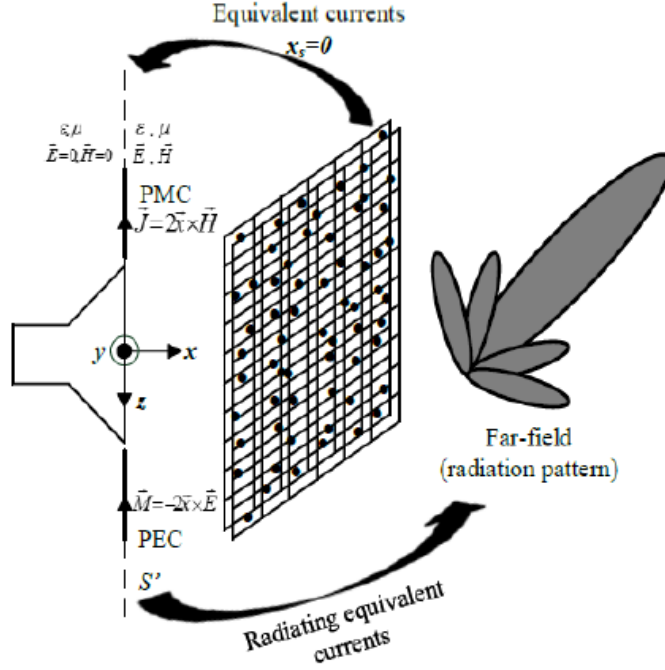


Figure A1. The equivalent currents method for the Far-Field calculation from irregular near-field measured samples.

The numerical representation of the integral equations is a system of linear equations, where the currents M_y , M_z , J_y and J_z are the unknowns. We can write the Equations (A4)–(A7) in a matrix form

$$\begin{pmatrix} E_{y_{meas}} \\ E_{z_{meas}} \end{pmatrix} = \begin{pmatrix} Z_{y,J_y} & Z_{y,J_z} & Z_{y,M_y} & Z_{y,M_z} \\ Z_{z,J_y} & Z_{z,J_z} & Z_{z,M_y} & Z_{z,M_z} \end{pmatrix} \begin{pmatrix} J_y \\ J_z \\ M_y \\ M_z \end{pmatrix} \quad (A8)$$

where for example:

$$Z_{y,J_y} = -\frac{j\eta\Delta y'\Delta z'}{4\pi k} e^{-jkR} \left(G_1 + (y - y')^2 G_2 \right) \quad (A9)$$

where, $\Delta y'$ and $\Delta z'$ are the dimensions of the source rectangular patches. The solution of (A8) is performed as (10) using LSQR method. As presented in this paper, the mathematical formulation of the equivalent current method seems to be more complicated to handel than the matrix method presented above. Indeed, the accuracy of the equivalent currents method depends on the definition of the currents surface (S') where these sources are uniformly distributed. This surface is defined based on the AUT a priori information and the number of considered equivalent sources is of crucial importance. The matrix method is easy and systematic in its application where no a priori information are needed.

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Antenna Far-Field Assessment From Near-Field Measured Over Arbitrary Surfaces

Mohamed Farouq, Mohammed Serhir, and Dominique Picard

Abstract—A matrix method is developed to calculate the antennas radiation pattern from near-field data measured over arbitrary surfaces. The principle of the matrix method is to determine the modal expansion coefficients by solving a system of linear equations. The number of required measured samples is deduced from the number of modes considering the cylindrical or the spherical wave expansion of antenna radiated fields. Once the modal coefficients are known, the far field of the antenna under test (AUT) is obtained. The matrix method has been implemented, and its potentialities are evaluated using two specific arbitrary surfaces (square section cylinder surrounding the AUT and a closed cylinder). Numerical (dipoles array) and experimental (base station antenna) results are provided to illustrate the efficiency and the stability of the proposed method.

Index Terms—Antenna measurement, arbitrary surfaces, iterative least square method (LSQR), matrix method, modal expansion, near-field to far-field transformation (NFFFT).

I. INTRODUCTION

TWO different methods are used in near-field (NF) techniques: the modal expansion and the equivalent sources [1]. The modal expansion is the most commonly used method because of its simplicity and its numerical efficiency. The fields radiated by the antenna under test (AUT) are expanded in terms of wave functions and the measured NFs are used to determine the coefficients of the expansion. The modal expansion exists in three main coordinate systems: planar [2], cylindrical [3], and spherical [4]. The measuring surface is scanned with a uniform distribution of points. This exploration allows to use the orthogonality properties of the modes to get the set of coefficients from the tangential components of the measured field. The modal expansion technique has the advantage of being stable on the one hand and to be very fast numerically due to the use of the fast Fourier transformation (FFT) on the other hand. However, the classical modal expansion technique cannot be used to compute the antenna radiation pattern from the measured NF over irregular distribution samples or arbitrary surfaces.

The second method based on the equivalent sources [5]–[10] can be employed to evaluate the radiation pattern of the AUT from the NF data collected over irregular samples or arbitrary surfaces. The basis of this technique is to shape the AUT through a set of equivalent-currents

distribution. These equivalent currents are defined from the NF measurement data, by solving the inverse radiation problem using the method of moments. This is the solution of the integral equations connecting the equivalent currents and their radiated field that matches the measured one. From these equivalent currents, it is possible to calculate the AUT radiated field at any point of the space outside the equivalent sources domain. Nevertheless, this method requires the AUT *a priori* information: location, dimensions, and type. The quality of the far field (FF) reproduced (also the NF) by the equivalent currents depends on the chosen distribution and spatial positioning of these equivalent electric or magnetic sources. Furthermore, the equivalent-currents method is a costly numerical operation and it limits the size of the problem that can be handled to a few wavelengths.

An approach has been presented in which a plane wave expansion is used to calculate NF to FF transformation (NFFFT) with a similar paradigm of equivalent-currents method. This technique is based on the fast multipole method (FMM). It permits the FF calculation of the AUT from its NF sampled over regular [11], irregular grid [12], or arbitrary surfaces [13]. To reduce the computational complexity of the FMM, a multilevel FMM (MLFMM) has been used to perform matrix-vector products and accelerate the resolution of the linear system. The implementation of the MLFMM proved to be a rather difficult task in part because of its complexity as well as the need to optimize all the steps of this method in other part [14].

An alternative technique to the equivalent-currents method has been proposed in [15] and [16]. This technique permits to handle the limit of the classical modal expansion techniques (irregular distributed samples and arbitrary surfaces). In these recent approaches, the authors have suggested a matrix method based on the plane wave expansion [15] to compensate the errors in the probe displacement in Cartesian coordinates. Moreover, to assess the radiation pattern from irregular distribution samples (probe displacement errors) in cylindrical and spherical coordinates, [16] introduces a matrix method to calculate the cylindrical and the spherical expansion coefficients from an irregular distribution of measured points.

In this paper, we analyze the application of the matrix method for calculating the modal expansion coefficient (MEC) from arbitrary measurement surfaces. The calculation of these MEC allows the assessment of the radiation pattern of the AUT in all spatial directions. The technique is derived from the modal expansion method, which allows the connection, in a matrix form, between the measured electric NF and the MEC. The number of samples required to reconstruct the antenna

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field is directly linked to the number of unknowns used in the linear system of equations derived using cylindrical or spherical expansions of AUT radiated fields.

The matrix method will be applied in the first case to determine the cylindrical modal coefficients calculated from the measured tangential components over square section cylinder surrounding the AUT. This measurement configuration can be obtained using an NF planar measurement system with an AUT mounted on an axis manually rotated. It allows to extend the usability of a planar NF measurement for nondirective antennas radiation pattern assessment. In the second case, we consider a cylindrical measurement surface closed by two disks. The matrix method is used for the calculation of spherical modal coefficients. This measurement configuration can be obtained using an NF measurement system in cylindrical coordinates, with a further axis of radial translation of the probe to explore the two end disks closing the cylinder. The radiation pattern calculated is valid over 4π steradian without any truncation error.

This paper is organized as follows. In Section II, the mathematical developments of the matrix method are presented. Results issued from these methods are presented in Section III for different antennas. The first studied AUT is an array of infinitesimal dipoles to generate the NF data. Then, the FF results of the proposed method are presented for a base station antenna measured in a planar NF range at 1.9 GHz. Finally, a conclusion is presented in Section IV.

II. MATHEMATICAL DEVELOPMENT

A. Cylindrical Wave Expansion and a Square Section Cylinder as Measurement Surface

In the classical case, we consider a measurement cylinder of radius ρ_{meas} , outside the minimum cylinder of radius ρ_{min} , surrounding an arbitrary AUT. The measured NF is described in a cylindrical coordinates system (ρ, φ, z) . Over the measurement cylinder, the measured tangential components of the electric NF can be presented [3] as a truncated summation of cylindrical vector waves. These components are expressed as follows:

$$E_\varphi(\rho_{\text{meas}}, \varphi_p, z_q) = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [a_n(k_z^m) e_{1,\varphi}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q) + b_n(k_z^m) e_{2,\varphi}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q)] \Delta k_z \quad (1)$$

$$E_z(\rho_{\text{meas}}, \varphi_p, z_q) = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [a_n(k_z^m) e_{1,z}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q) + b_n(k_z^m) e_{2,z}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q)] \Delta k_z. \quad (2)$$

The cylindrical normal component of the electric NF can be expressed as follows:

$$E_\rho(\rho_{\text{meas}}, \varphi_p, z_q) = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [a_n(k_z^m) e_{1,\rho}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q) + b_n(k_z^m) e_{2,\rho}^{n,k_z^m}(\rho_{\text{meas}}, \varphi_p, z_q)] \Delta k_z \quad (3)$$

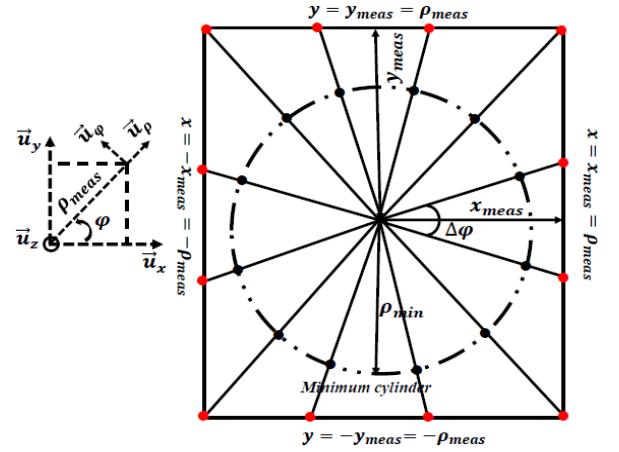


Fig. 1. Projection of the square section cylinder over the xy plane. Red points: measurement locations.

where φ_p and z_q are the measurement points of the NF over the cylinder of radius ρ_{meas} . $\Delta\varphi$ and Δz are, respectively, the sampling steps in the φ - and z -directions.

The number of cylindrical modes is calculated using $N_{\text{cyl}} = \text{int}(k\rho_{\text{min}}) + n_{\text{cyl}}$ with $n_{\text{cyl}} \in \mathbb{N}$ and int is the integer function. $-\pi/\Delta z \leq k_z^m \leq \pi/\Delta z$, with $M = (k_{z_{\text{max}}} - k_{z_{\text{min}}})/\Delta k_z + 1$.

The functions $e_1^{n,k_z^m} = e_{1,\rho}^{n,k_z^m} \vec{u}_\rho + e_{1,\varphi}^{n,k_z^m} \vec{u}_\varphi + e_{1,z}^{n,k_z^m} \vec{u}_z$ and $e_2^{n,k_z^m} = e_{2,\rho}^{n,k_z^m} \vec{u}_\rho + e_{2,\varphi}^{n,k_z^m} \vec{u}_\varphi + e_{2,z}^{n,k_z^m} \vec{u}_z$ represent the cylindrical wave functions solution of the Helmholtz equation presented in [3]; \vec{u}_ρ , \vec{u}_φ , and \vec{u}_z are the unit vectors of the cylindrical coordinates. $a_n(k_z)$ and $b_n(k_z)$ are the cylindrical modal coefficients (MEC).

Here, we aim at developing a formulation, which permits to calculate the cylindrical MEC from a measurement over square section cylinder enclosing the AUT. These measurement surfaces will be scanned using a planar NF range. The mathematical development of the chosen configuration can be extended to any arbitrary surfaces provided that the measurement surface is outside of the minimum cylinder ρ_{min} .

Over the square section cylinder surrounding the AUT (Fig. 1), we measure the tangential components of the NF (E_y, E_z) at $x_{\text{meas}} = \pm\rho_{\text{meas}}$ and (E_x, E_z) at $y_{\text{meas}} = \pm\rho_{\text{meas}}$ over planar surfaces, outside the minimum cylinder of radius ρ_{min} . We desire at linking the measured tangential components of the electric NF to the cylindrical MECs $a_n(k_z^m)$ and $b_n(k_z^m)$. To do this, we use the expressions relating the Cartesian coordinates and the cylindrical coordinates. These relations are expressed as follows:

$$\begin{cases} x = \rho \cos \varphi \\ y = \rho \sin \varphi \\ z = z \end{cases} \quad \text{and} \quad \begin{cases} E_x = E_\rho \cos \varphi - E_\varphi \sin \varphi \\ E_y = E_\rho \sin \varphi + E_\varphi \cos \varphi \\ E_z = E_z \end{cases} \quad (4)$$

By replacing (4) in (1)–(3) and grouping the terms of the cylindrical MECs $a_n(k_z^m)$ and those of $b_n(k_z^m)$, we can express the components of the electric field E_x, E_y , and E_z in planar

surface as follows:

$$E_x = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [(e_{1,\rho}^{n,k_z^m} \cos \varphi - e_{1,\varphi}^{n,k_z^m} \sin \varphi) a_n(k_z^m) + (e_{2,\rho}^{n,k_z^m} \cos \varphi - e_{2,\varphi}^{n,k_z^m} \sin \varphi) b_n(k_z^m)] \Delta k_z \quad (5)$$

$$E_y = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [(e_{1,\rho}^{n,k_z^m} \sin \varphi + e_{1,\varphi}^{n,k_z^m} \cos \varphi) a_n(k_z^m) + (e_{2,\rho}^{n,k_z^m} \sin \varphi + e_{2,\varphi}^{n,k_z^m} \cos \varphi) b_n(k_z^m)] \Delta k_z \quad (6)$$

$$E_z = \sum_{n=-N_{\text{cyl}}}^{N_{\text{cyl}}} \sum_{m=1}^M [(e_{1,z}^{n,k_z^m}) a_n(k_z^m) + (e_{2,z}^{n,k_z^m}) b_n(k_z^m)] \Delta k_z. \quad (7)$$

Note that (5)–(7) can be expressed in the following matrix formulations:

$$\begin{cases} E_x = C_{11} a_n(k_z^m) + C_{12} b_n(k_z^m) \\ E_y = C_{21} a_n(k_z^m) + C_{22} b_n(k_z^m) \\ E_z = C_{31} a_n(k_z^m) + C_{32} b_n(k_z^m) \end{cases} \quad (8)$$

with C_{11} , C_{12} , C_{21} , C_{22} , C_{13} , and C_{23} are the coupling matrices.

These matrices contain the cylindrical wave functions. Solving (8) allows us to calculate the cylindrical MECs $a_n(k_z^m)$ and $b_n(k_z^m)$. The calculated MEC are used to evaluate the FF (radiation pattern) of the AUT using

$$E_\theta(r, \theta, \phi) = -j2k \sin \theta \frac{e^{-jkr}}{r} \sum_{n=-N_{\text{cyl}}}^{n=N_{\text{cyl}}} j^n b_n(k \cos \theta) e^{jn\phi} \quad (9)$$

$$E_\phi(r, \theta, \phi) = -2k \sin \theta \frac{e^{-jkr}}{r} \sum_{n=-N_{\text{cyl}}}^{n=N_{\text{cyl}}} j^n a_n(k \cos \theta) e^{jn\phi}. \quad (10)$$

B. Spherical Wave Expansion and a Closed Cylinder as Measurement Surface

In this section, we apply the matrix method to calculate the FF from NF data measured over a cylinder closed by two disks. This measurement configuration will be exploited with the spherical modal expansion. The use of the spherical sampling on a cylindrical surface reduces the number of measurement points in the z -direction in comparison with the classical cylindrical NF measurement. Moreover, closing the cylinder by two disks, the extent of the measurement cylinder can be reduced and we can avoid the truncation error.

We aim at extending the possibilities of the cylindrical NF facility. For this, we consider spherical coordinates (r , θ , and ϕ) and we choose to measure the electric field outside the minimum sphere of radius r_{min} circumscribing the AUT. Throughout this paper, $\exp(j\omega t)$ time dependence is suppressed in all the E -field expressions. The spherical wave expansion (SWE) of the radiated electric field components is expressed in terms of truncated series of spherical vector wave functions [4] as follows:

$$E_\theta(r_{\text{meas}}, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sum_{l=1}^{L_{\text{max}}} Q_l F_{\theta,l}^{(4)}(r_{\text{meas}}, \theta, \phi) \quad (11)$$

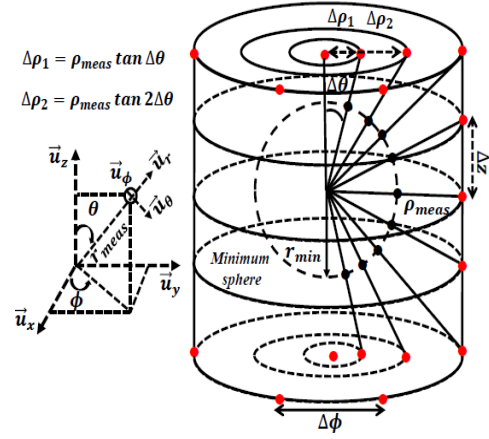


Fig. 2. Measurement surface is a cylinder closed by two disks. Red points: measurement locations.

$$E_\phi(r_{\text{meas}}, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sum_{l=1}^{L_{\text{max}}} Q_l F_{\phi,l}^{(4)}(r_{\text{meas}}, \theta, \phi). \quad (12)$$

The spherical normal component of the electric NF can be expressed as follows:

$$E_r(r_{\text{meas}}, \theta, \phi) = \frac{k}{\sqrt{\eta}} \sum_{l=1}^{L_{\text{max}}} Q_l F_{r,l}^{(4)}(r_{\text{meas}}, \theta, \phi) \quad (13)$$

where $\eta = (\epsilon_0/\mu_0)^{(1/2)}$ is the intrinsic admittance, r_{meas} is the radius of measurement sphere, k is the wavenumber in the free space, $\vec{F}_l = F_{r,l} \vec{u}_r + F_{\theta,l} \vec{u}_\theta + F_{\phi,l} \vec{u}_\phi$ represents the spherical wave functions of the outward propagating fields [4] with \vec{u}_r , and \vec{u}_θ and \vec{u}_ϕ are the unit vectors of the spherical coordinates. Q_l values are the spherical wave coefficients. $L_{\text{max}} = 2N_{\text{sph}}(N_{\text{sph}} + 2)$ where $N_{\text{sph}} = \text{int}(kr_{\text{min}}) + n_{\text{sph}}$ is the number of spherical modes depending on the AUT dimensions and the operating frequency with $n_{\text{sph}} \in \mathbb{N}$.

In the case of the closed cylindrical surface (Fig. 2), we measure the tangential components of the NF E_ϕ and E_z in the cylindrical surface of radius ρ_{meas} and in the two disks, which are placed at $z_{\text{meas}} = \pm \rho_{\text{meas}}$ with $\Delta \rho = \rho_{\text{meas}} \tan(\Delta \theta)$, with $\rho_{\text{meas}} \geq r_{\text{min}}$.

Our objectif is to calculate the spherical modal coefficients Q_l from the tangential E_ϕ and E_z components measured on a closed cylinder. The mathematical development of the method is based on the relationship between the following spherical and cylindrical coordinates:

$$\begin{cases} r_{\text{meas}} = \frac{\rho_{\text{meas}}}{\sin \theta} \\ \theta = \arctan\left(\frac{\rho_{\text{meas}}}{z}\right) \\ \phi = \phi \end{cases} \quad \text{and} \quad \begin{cases} E_\rho = E_r \sin \theta + E_\theta \cos \theta \\ E_\phi = E_\phi \\ E_z = E_r \cos \theta - E_\theta \sin \theta. \end{cases} \quad (14)$$

Substituting (11)–(13) in (14), we get a more compact form connecting the two tangential components of the NF $E_\phi(\rho, \varphi, z)$ and $E_z(\rho, \varphi, z)$ with spherical modal

coefficients Q_l

$$E_\varphi(\rho, \varphi, z) = \frac{k}{\sqrt{\eta}} \sum_{l=1}^{L_{\max}} Q_l F_{\phi,l}^{(4)}(r, \theta, \phi) \quad (15)$$

$$E_z(\rho, \varphi, z) = \frac{k}{\sqrt{\eta}} \sum_{l=1}^{L_{\max}} [F_{r,l}^{(4)}(r, \theta, \phi) \cos \theta - F_{\theta,l}^{(4)}(r, \theta, \phi) \sin \theta] Q_l \quad (16)$$

(15) and (16) can be rewritten in the following matrix form:

$$\begin{cases} E_\varphi(\rho, \varphi, z) = S_1 Q_l \\ E_z(\rho, \varphi, z) = S_2 Q_l \end{cases} \quad (17)$$

with S_1 and S_2 are the coupling matrices. These matrices contain the spherical wave functions, as defined in [4]. The inversion of these matrices permits to calculate the spherical MECs Q_l . The mathematical development presented earlier can be used for any arbitrary surface provided that the measurement surface is outside the minimum sphere encompassing the AUT. The MEC is used to evaluate the FF of the AUT using

$$\vec{E}(r, \theta, \phi) = \frac{k}{\sqrt{\eta}} \frac{1}{\sqrt{4\pi}} \frac{e^{-jkr}}{kr} \sum_{l=1}^{L_{\max}} Q_l \vec{K}_l^{(4)}(\theta, \phi) \quad (18)$$

where $\vec{K}_l^{(4)}$ is the FF radiation pattern developed in [4].

The inversion of both linear system equations (8) and (17) is done using MATLAB iterative least square method (LSQR) method [17]. In the following study, we aim at evaluating the efficiency of the matrix method for FF assessment in different situations.

III. MATRIX METHOD EVALUATION

The proposed method is tested using numerical simulations considering two configurations: the square section cylinder circumscribing the AUT and the closed cylinder. Furthermore, the experimental measurements will be carried out for the first one. In the numerical simulations, the NF data are computationally generated using analytical expression of an infinitesimal dipoles array. In the experimental validation, the NF data of a base station antenna are measured in the Supec planar NF measurement setup [18]. The efficiency of the proposed method to assess the radiation pattern of the AUT is studied.

The matrix method accuracy is evaluated by comparing the reference FF (E_{ref}) and the FF resulting from the matrix method (E_{calc}). The reference FF is determined directly using the dipoles array in the case of numerical simulations. Two different errors are calculated: the global error and the spatial distribution of the local error

$$\begin{cases} \text{Error} = 100 \sqrt{\frac{\sum_{\theta, \phi} |E_{\text{calc}}(\theta, \phi) - E_{\text{ref}}(\theta, \phi)|^2}{N_\theta N_\phi \max_{\theta, \phi} |E_{\text{ref}}(\theta, \phi)|^2}} \\ \text{2-D error} = 100 \frac{|E_{\text{calc}}(\theta, \phi) - E_{\text{ref}}(\theta, \phi)|}{\max_{\theta, \phi} |E_{\text{ref}}(\theta, \phi)|} \end{cases} \quad (19)$$

where N_θ and N_ϕ are, respectively, the number of samples in the θ - and ϕ -directions.

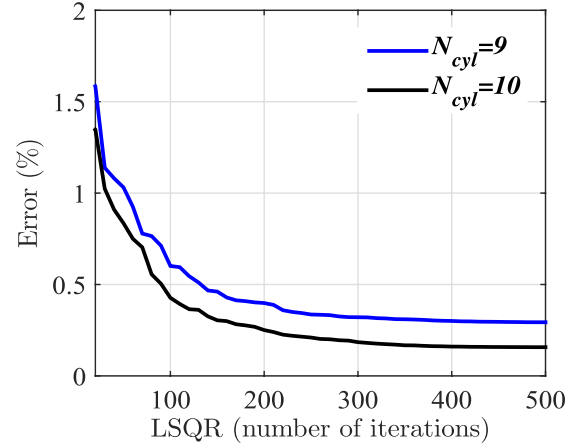


Fig. 3. Error function versus the number of LSQR iterations and the used number of cylindrical modes N_{cyl} .

A. Numerical Study: Square Section Cylinder and CWE

The AUT is composed of two layers of 4×8 infinitesimal dipoles array placed in the yz plane. The dipoles of the AUT are tilted with 6° with respect to the z -axis to generate a cross-polarization level with 20 dB below the maximum copolarization magnitude. The dipoles in both layers are $\lambda/2$ spaced along the y - and z -dimensions. The first layer is located at $x = 0$ and the second one at $x = -\lambda/4$. For each layer, the excitation of the dipoles is constant in magnitude and phase. A phase offset of $\pi/2$ is considered in the second layer to minimize the radiation in the backward direction.

The square section cylinder is scanned according to $z_{\max} = -z_{\min} = 15\lambda$ with a sampling step of $\Delta z = \lambda/2$. In the x - and y -directions, the faces are sampled with a constant angular azimuthal step $\Delta\phi = \pi/N_{\text{cyl}}$. Each face has a width of 6λ . The choice of a constant azimuth angular sampling criterion is resulting from the use of the cylindrical modal expansion.

The procedure consists in measuring the tangential components (E_y, E_z) at $x_{\text{meas}} = \pm 3\lambda$ and (E_x, E_z) at $y_{\text{meas}} = \pm 3\lambda$ of the NF radiated by the AUT on each face. This configuration allows a critical angle $\theta_{\text{valid}} = 78^\circ$. The NF data are used in (8) to determine the cylindrical modal coefficients $a_n(k_z^m)$ and $b_n(k_z^m)$. Knowing the modal coefficients, we can deduce the FF (radiation pattern) using (9) and (10).

To evaluate the influence of the considered number of samples in the x - and y -directions, we choose different integers n_{cyl} in $N_{\text{cyl}} = \text{int}(k\rho_{\min}) + n_{\text{cyl}}$. In Fig. 3, we quantify the error function defined in (19) by considering two different numbers of modes versus the LSQR iterations. It is seen from Fig. 3 that the matrix method converges in both cases $N_{\text{cyl}} = \text{int}(k\rho_{\min}) + 4 = 9$ and $N_{\text{cyl}} = \text{int}(k\rho_{\min}) + 5 = 10$ with a low FF error level.

The 2-D error distributions considering $N_{\text{cyl}} = 9$ and $N_{\text{cyl}} = 10$ are presented in Fig. 4. As it is seen, the errors are very low in both cases. These errors appear at low level field region.

We compare the main FF cut planes considering $N_{\text{cyl}} = 9$ and $N_{\text{cyl}} = 10$ with the actual one in Fig. 5. It can be seen that

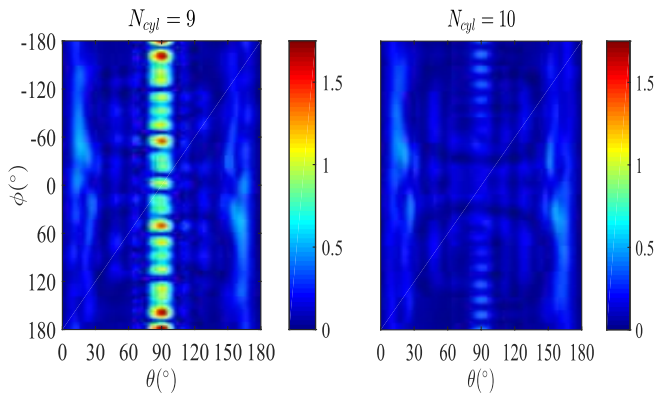


Fig. 4. 2-D error (%) distribution in copolar FF component.

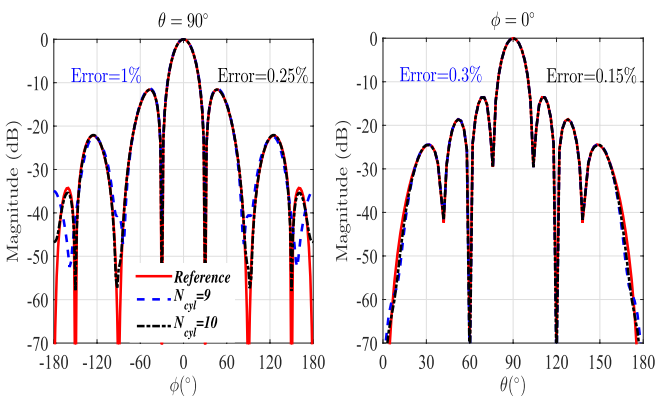
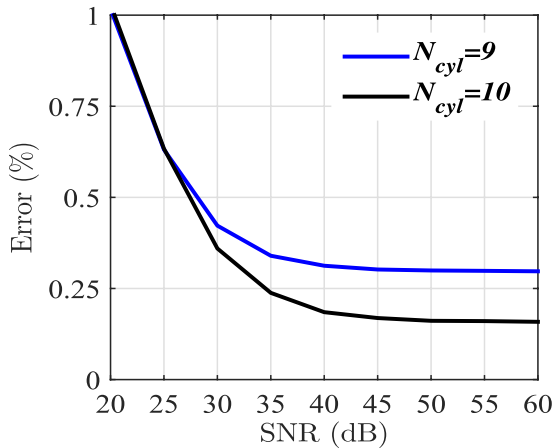


Fig. 5. Copolar FF component.

Fig. 6. Error function versus SNR and the considered number of cylindrical modes N_{cyl} .

the matrix method results present an excellent agreement with the reference radiation pattern in both directions for $N_{cyl} = 9$ and $N_{cyl} = 10$.

In addition, we study the effect of the number of cylindrical modes and the signal-to-noise ratio (SNR) over the matrix method accuracy. The NF data are contaminated with a controlled white Gaussian noise level to reduce the data SNR. Fig. 6 presents the error values as a function of the SNR of the NF data. It is seen from Fig. 6 that the error function takes high values for low SNR NF data. However, beyond

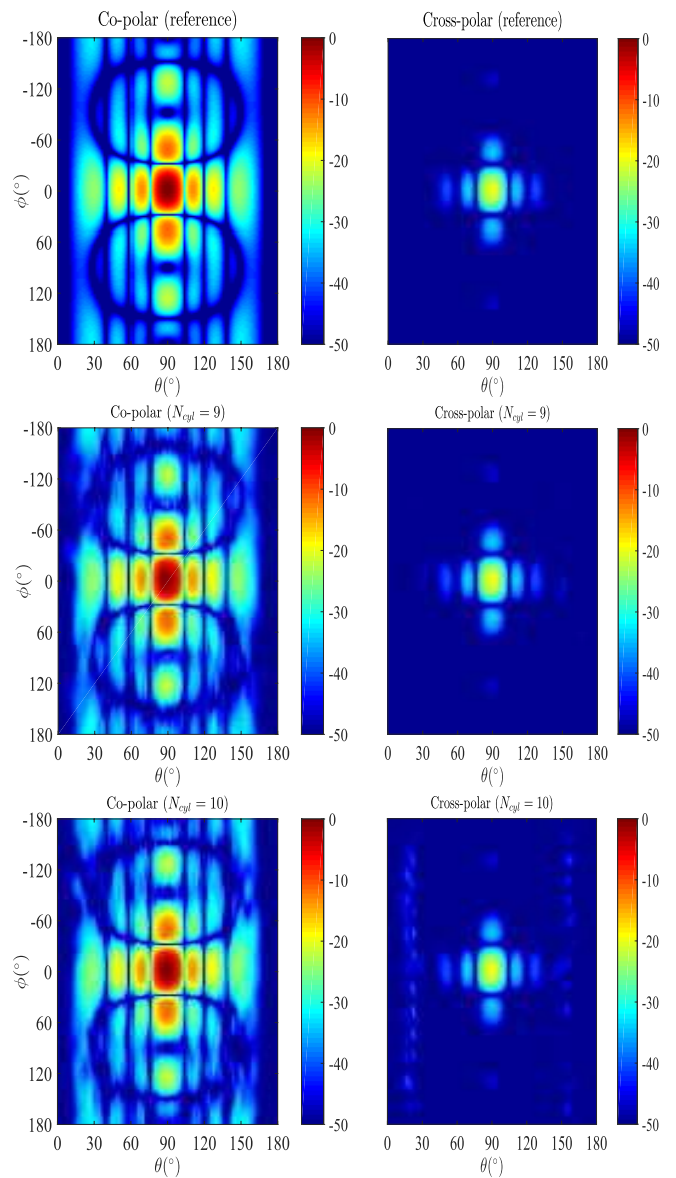


Fig. 7. 2-D co- and cross-polar FF components (dB) for SNR = 30 dB.

the threshold of about 40 dB, the values of the error function are low.

We present in Fig. 7 the 2-D FF radiation pattern considering SNR = 30 dB for $N_{cyl} = 9$ and $N_{cyl} = 10$. Even for a low SNR NF data (30 dB), the 2-D radiation pattern calculated using the matrix method is in good agreement with the reference one for the copolarization and cross polarization. These observations are confirmed by comparing the two main cut planes of the copolar and cross polar presented in Fig. 8 of the calculated FF using the matrix method and the reference one.

B. Numerical Study: Closed Circular Section Cylinder and SWE

To evaluate the performance of the matrix method using the spherical modal expansion in the case of the closed cylinder,

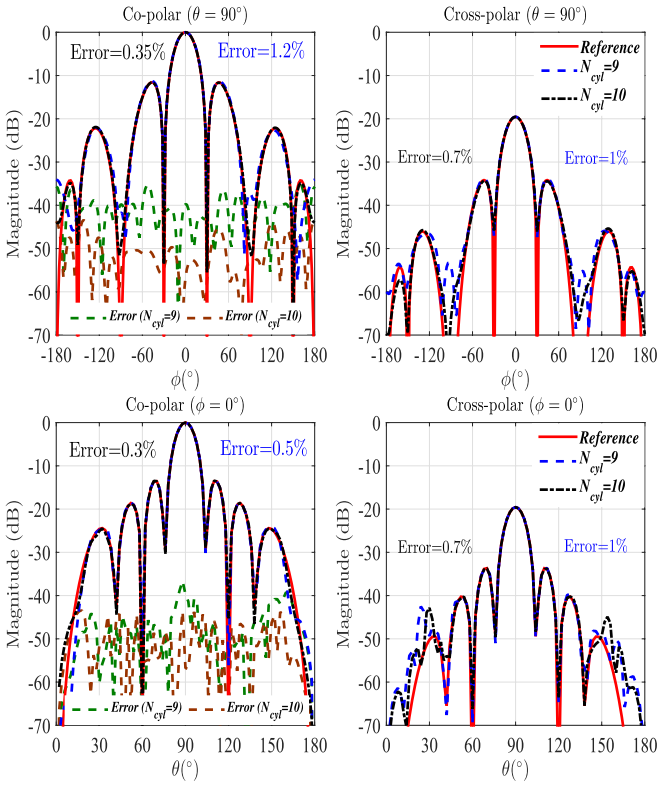
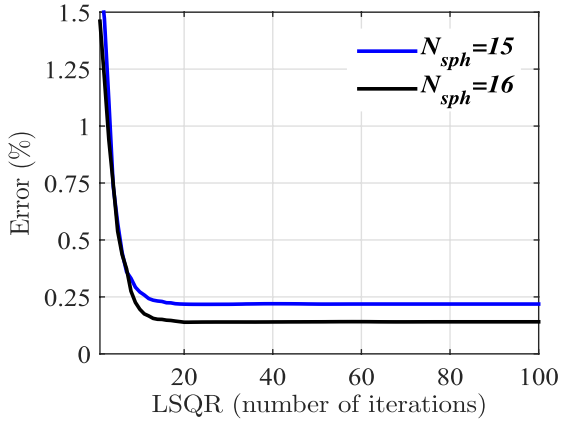


Fig. 8. Co- and cross-polar FF components (dB) for SNR = 30 dB.


 Fig. 9. Error function versus the number of iterations and the considered number of spherical modes N_{sph} .

we use the same AUT (two layers of 4×8 dipoles). The field generated by the AUT is collected over a closed cylinder of $z_{max} = -z_{min} = \rho_{meas} = 5\lambda$ with $\Delta z = \rho_{meas} \tan(\Delta\theta)$. The two disks used to close the cylinder are explored respecting the sampling criterion $\Delta\rho = \rho_{meas} \tan(\Delta\theta)$ (see Fig. 2). In the azimuth direction ϕ , the sampling step is calculated from $\Delta\phi = \pi/N_{sph}$.

The convergence of the matrix method using $N_{sph} = \text{int}(kr_{min}) + 4 = 15$ and $N_{sph} = \text{int}(kr_{min}) + 5 = 16$ (two different numbers of samples) is presented in Fig. 9. It is seen that the error values are low and the convergence is reached rapidly.

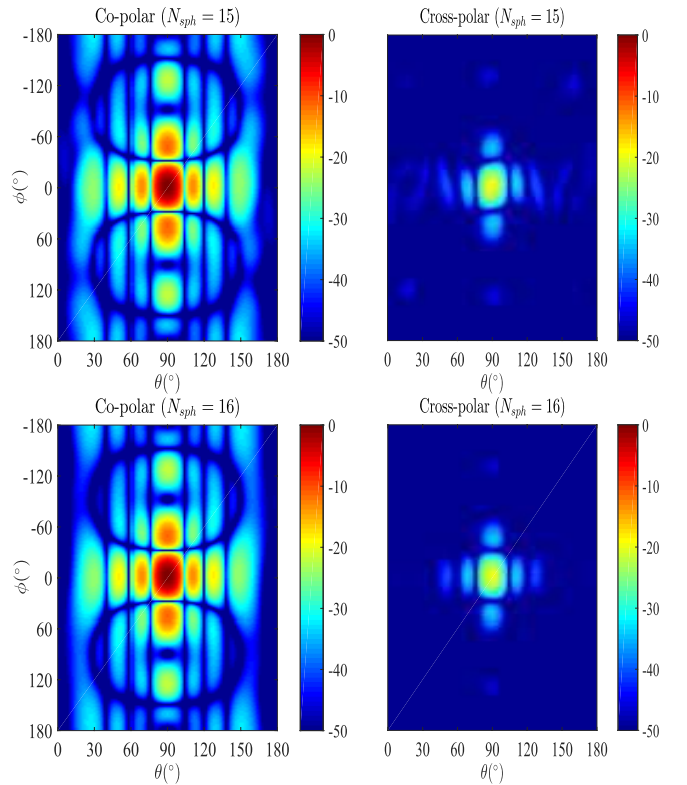


Fig. 10. 2-D co- and cross-polar FF components (dB).

To emphasize the influence of the considered number of modes in reconstructing the FF of the AUT, the 2-D FF radiation patterns and the main cut planes of the AUT calculated using $N_{sph} = 15$ and $N_{sph} = 16$ are, respectively, presented in Figs. 10 and 11. As it can be seen, the copolarization and cross polarization of the FF results are in good agreement (reference FF in Fig. 7). The error function between the reference copolar FF and the calculated one is also evaluated in Fig. 11 for the main cut planes. In the case of $N_{sph} = 15$, the matrix method presents an error = 1% for $\theta = 90^\circ$ and error = 0.5% for $\phi = 0^\circ$. For $N_{sph} = 16$, the error = 0.1% for $\theta = 90^\circ$ and error = 0.15% for $\phi = 0^\circ$.

Furthermore, we have quantified the effects of the SNR and the considered number of spherical modes over the matrix method accuracy. The NF data are calculated over the closed cylinder considering $N_{sph} = 15$ and $N_{sph} = 16$. Then, a white Gaussian noise is added to reduce the data SNR. Fig. 12 shows the error between the reference FF and the calculated one as a function of the SNR of the NF data. The error function takes high values for low SNR data. However, beyond the threshold of 40 dB, the error function stays stable and low.

The 2-D error distributions considering SNR = 30 dB or SNR = 40 dB for $N_{sph} = 15$ and $N_{sph} = 16$ are presented in Fig. 13. As it is seen, the errors appear essentially in the minima of the radiation pattern and in the regions of weak fields. However, the errors decrease significantly considering an SNR = 40 dB and $N_{sph} = 16$.

In Section III-C, we will examine the application of the matrix method in the case of the square section cylinder

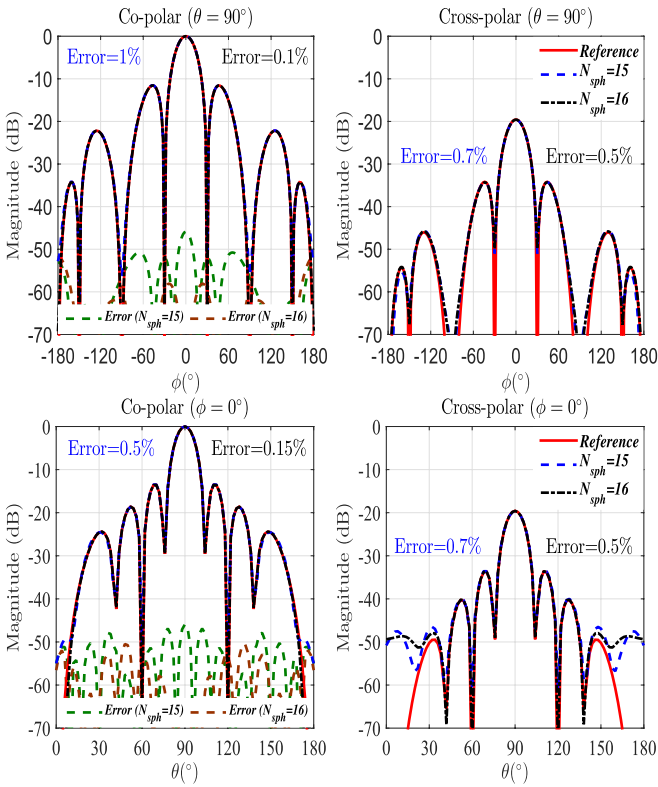


Fig. 11. Co- and cross-polar FF components (dB).

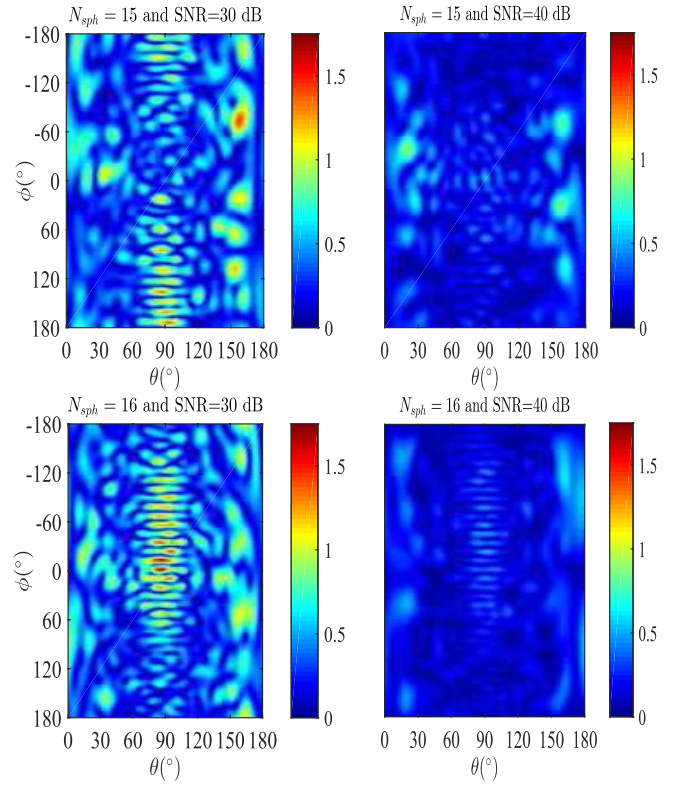
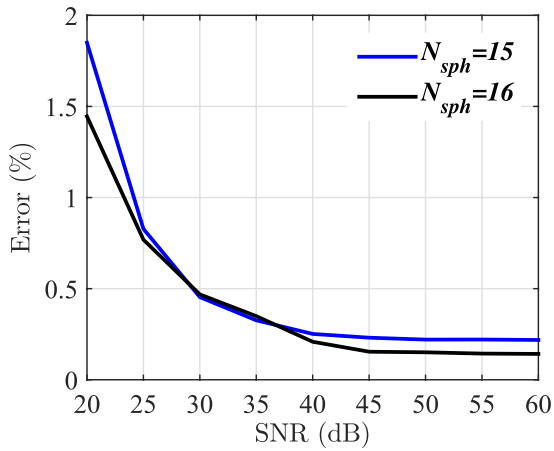


Fig. 13. 2-D error (%) distribution for copolar FF component (dB).

Fig. 12. Error function versus SNR and the considered number of spherical modes N_{sph} .

surrounding the AUT with real data using the Supelec planar NF range.

C. Experimental Study: Square Section Cylinder and CWE

The AUT used to validate the matrix method based on the NF measured over square section cylinder is presented in Fig. 14. The base station antenna and the measurement probe are placed inside the planar NF measurement system of dimensions $4 \text{ m} \times 5 \text{ m} \times 3 \text{ m}$. For our validation measurement, we have used the wideband probe “En-Probe EFS-105” [19]. This probe is composed of a small dipole (dimensions 6.6

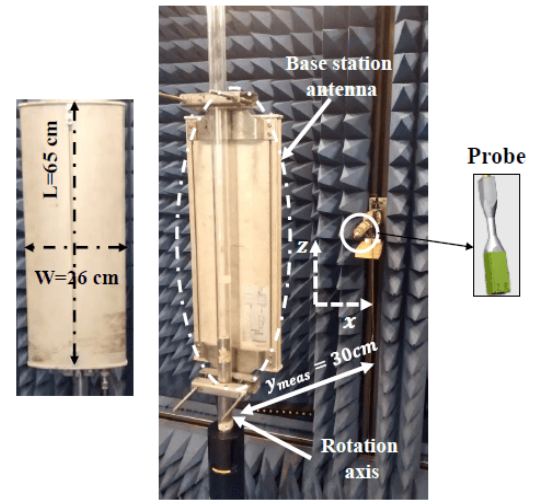


Fig. 14. Base station antenna measured in the planar NF facility.

$\text{mm} \times 6.6 \text{ mm}$) connected to the base unit through a fiber optic link and the base unit is connected to the measuring Vector network analyzer (VNA) through a short cable. This probe is dedicated to NF measurement covering the frequency band (5 MHz–3 GHz). The antenna NF measurement is carried out at the frequency 1.9 GHz. The NF probe used in this measurement is supposed to behave like an ideal electric dipole, which means that no probe correction procedure is needed.

To evaluate the performance of the matrix method, we compare the radiation pattern of the AUT resulting from the

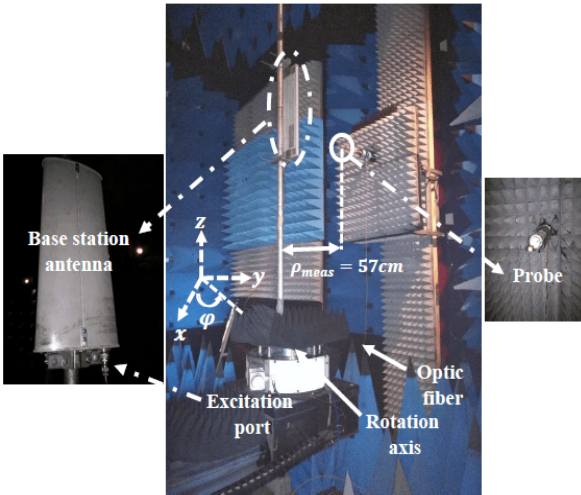


Fig. 15. Base station antenna measured in the cylindrical NF facility.

application of the classical NFFFT in cylindrical coordinates and the one calculated using the proposed method.

The reference FF radiation pattern is issued from the classical cylindrical NFFFT by measuring the tangential components (E_ϕ and E_z) over a cylinder of radius 57 cm and a height of 340 cm with a regular sampling step of $\Delta z \simeq \lambda/3 = 5$ cm and an angular sampling step of $\Delta\phi = 10^\circ$. The photography of the measurement configuration is presented in Fig. 15. This reference measurement is realized using the probe described earlier.

The matrix method validation procedure is described as follows: in the first step, we measure the tangential NF components (E_x and E_z) over a rectangular surface in front of the AUT. The measurement plane is placed at $y_{\text{meas}} = 30$ cm. The height and the width of the rectangular plane are created by moving the probe respecting a sampling step $\Delta x = 6$ cm and $\Delta z = 3$ cm between measurement points over the planar surface ($-30 \text{ cm} \leq x_{\text{meas}} \leq 30$ cm and $-70.5 \text{ cm} \leq z_{\text{meas}} \leq 70.5$ cm). In the second step, we realize the measurement over the three others faces by rotating the AUT considering an angle of $\pi/2$. In the third step, we reorganized the measurement data to calculate the coefficients of the six coupling matrices of (8). In the fourth step, we consider a number of cylindrical modes $N_{\text{cyl}} = 10$ to solve the linear equation systems (8) utilizing the MATLAB LSQR routine. In the last step, we determine the AUT FF radiation pattern using the calculated cylindrical MECs. The assessed FF is compared with the reference one.

The 2-D FF radiation pattern calculated using the cylindrical classical NFFFT and the one evaluated using the matrix method are presented in Fig. 16. As it can be seen, both copolarization and cross polarization evaluated by matrix method present good agreement (the errors are 2% and 9% for copolarization and cross polarization, respectively) in the region corresponding to the critical angle $\theta_{\text{valid}} \leq 50^\circ$ with the reference FF radiation pattern (copolar and cross polar). This region is imposed by the finite planar measurement surface ($-70.5 \text{ cm} \leq z_{\text{meas}} \leq 70.5$ cm).

For a well depth comparison of the 2-D FF radiation pattern, main cut planes of the copolar are shown in Fig. 17. As it

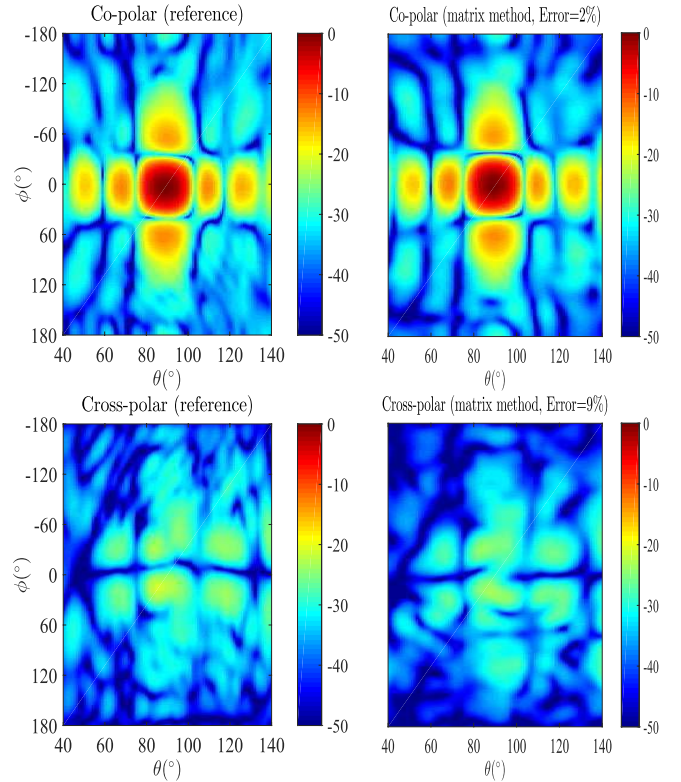


Fig. 16. 2-D co- and cross-polar FF components (dB).

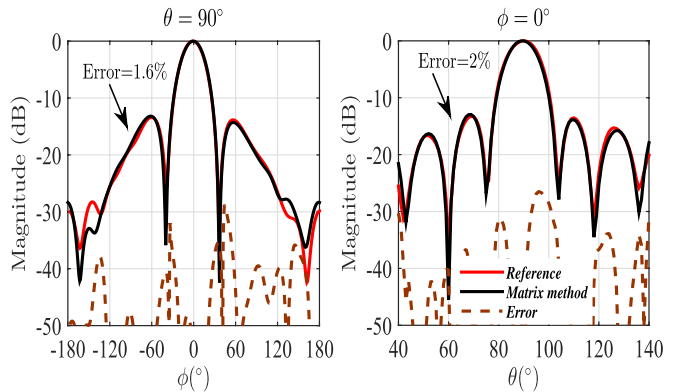


Fig. 17. Copolar FF component (dB).

can be seen, the proposed method presents good agreement with the reference FF in the main cut plane of $\theta = 90^\circ$ (error = 1.6%). However, in the other main cut plane $\phi = 0^\circ$ (error = 2%), the calculated radiation pattern presents a truncation error for $\theta \leq 40^\circ$ and $\theta \geq 140^\circ$.

IV. CONCLUSION

We have developed a matrix method to obtain the MECs from an NF measured over arbitrary surfaces for which classical modal expansion techniques based on FFT are not able to perform. These coefficients are used to assess the radiation pattern of the AUT in all directions.

The proposed method has been tested for two different surfaces. The first surface is the square section cylinder. The measured NF over this surface was exploited with a

cylindrical wave expansion (CWE) in a matrix form to calculate the cylindrical MECs. The second used surface is the closed cylinder with two disks. To calculate the spherical MECs, we had used the spherical modal expansion in a matrix form.

The numerical (dipoles array) and experimental (base station antenna) validation procedures are employed to evaluate the performances of the proposed technique. The obtained results have demonstrated that our technique is able to determine accurately the copolarization and cross polarization typically over a 50-dB dynamic range.

The matrix method allows to extend the usability and to increase the performances of the NF measurement systems for antennas radiation pattern assessment. The proposed method has several advantages, including: reducing the number of measurement samples, decreasing the acquisition time of the NF data, and avoiding the truncation error (closed cylinder). The extension of the proposed matrix method aims at including the probe compensation procedure. This is under development.

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3.4 Caractérisation en champ proche des antennes pulsées

Dans une approche complémentaire à l'analyse du champ proche dans le domaine fréquentiel, je me suis intéressé à l'étude et la caractérisation en champ proche d'antennes dans le régime temporel. J'ai développé ces travaux durant mon post-doctorat au L2S entre 2009 et 2010. Cette ouverture thématique m'a permis de renforcer mes aptitudes expérimentales dans un autre type de caractérisation et de consolider un axe de recherche naissant sur la conception et l'optimisation d'antennes dédiées au radar de sol.

Je me suis penché sur la caractérisation de la réponse impulsionnelle d'antennes ultra-large bande à l'image des travaux [DE JOUGH et collab. \[1997\]](#) [ANDRIEU et collab. \[2005\]](#) [YAROVY et collab. \[2000\]](#). Pour cela, j'ai développé un banc de caractérisation d'antenne en champ proche en coordonnées cartésiennes ainsi que l'algorithme de transformation champ proche - champ lointain dans le domaine temporel [SERHIR et collab. \[2012\]](#) [SERHIR et PICARD \[2013\]](#).

Cette thématique n'est pas étrangère aux activités associées au radar de sol. Il se trouve que la communauté issue des géosciences a été à l'origine de beaucoup de développements sur la transformée de Radon [CHAPMAN \[1978\]](#) [CHAPMAN \[1981\]](#) qui n'est autre que le développement en ondes planes exprimé dans le domaine temporel.

Les travaux de Thorkild B. Hansen et Arthur D. Yaghjian [THORKILD et ARTHUR \[1994\]](#) [HANSEN et YAGHJIAN \[1994\]](#) [HANSEN et YAGHJIAN \[1995\]](#) [HANSEN et YAGHJIAN \[1999\]](#) présentent une description détaillée de la méthodologie de caractérisation en champ proche pour des acquisitions sur un plan en coordonnées cartésiennes. Bucci [BUCCI et collab. \[1998a\]](#) propose l'analyse associée à un échantillonnage polaire et les autres systèmes de coordonnées sont traités dans [HANSEN \[1997\]](#) [CHRISTIANER \[1998\]](#) [SHLIVINSKI et HEYMAN \[1999\]](#) pour des mesures en champ proche cylindriques et sphériques.

En comparaison avec les mesures dans le domaine harmonique, la caractérisation en temporel est adaptée pour distinguer la chronologie des signaux captés par la sonde de mesure [BLECH et collab. \[2010\]](#). Le signal utile issu de l'antenne sous test et les autres signaux qui potentiellement perturbent ou polluent les mesures arrivent à des instants différents qui dépendent des distances parcourues. Les techniques de mesure dans le domaine temporel seraient davantage adaptées à la caractérisation d'antennes dans des environnements non-anéchoïques [KOH et collab. \[2012\]](#) ou pour l'isolation des signaux parasites en compatibilité électromagnétique [RAVELO et collab. \[2011\]](#). Le fenêtrage temporel permet le filtrage des signaux mesurés selon leur ordre d'arrivée.

J'ai été particulièrement séduit par l'analyse présentée par Hansen dans [HANSEN et YAGHJIAN \[1999\]](#) évoquant le *Early-time far-field*. Cette analyse exploite le fenêtrage temporel (time gating) pour filtrer les signaux selon leur ordre d'arrivée pour éliminer les erreurs dues à la troncature de la surface de mesure. Cette technique de filtrage est adaptée pour la caractérisation des antennes faiblement directives avec une surface de mesure d'extension finie. Les développements théoriques des techniques de transformation du champ proche dans le domaine temporel pour les coordonnées cartésiennes sont détaillés dans [HANSEN et YAGHJIAN \[1999\]](#). Néanmoins, l'implémentation expérimentale est rarement commentée dans la littérature.

Ma contribution dans ce sujet est certainement expérimentale. J'ai mis en place une plateforme robotisée avec son logiciel d'acquisition pour caractériser des structures alimentées par des excitations non-sinusoidales [SERHIR et collab. \[2012\]](#). Il s'agissait d'adapter, dimensionner et automatiser l'instrumentation nécessaire pour permettre des caractérisations en champ proche des antennes dans le régime temporel. J'ai exploité ce système de mesure pour l'acquisition dans le régime temporel du champ diffracté par des cibles métalliques placées en espace libre. Nous souhaitons tester le retournement temporel comme approche de localisation basée sur une cartographie de l'énergie. Les résultats de cette analyse sont présentés dans [BENHAMOUCHE et collab. \[2013\]](#) et [BENHAMOUCHE et collab. \[2014\]](#). Cela se déroula dans le cadre de la thèse de M. Benhamouche (2009 – 2012) dirigée par L. Pichon (GeePs) et D. Lesselier (L2S). Nous avons utilisé un moyennage de plusieurs acquisitions pour améliorer le rapport signal à bruit des données mesurées.

J'ai mis en place les outils de transformation champ proche - champ lointain dans le domaine temporel (basé sur le formalisme de Green) et les résultats obtenus (numériques et expérimentaux) sont comparés à ceux issus de la mesure fréquentielle du champ lointain [SERHIR \[2015\]](#) (joint avec ce rapport). Ce travail est une première brique pour les développements futurs qui prendront en charge les systèmes de coordonnées cylindriques et sphériques. Comme première investigation, j'ai commencé une analyse sur l'échantillonnage en champ proche dans le domaine temporel. Mon questionnement concerne le choix des critères à adopter pour assurer une conservation optimale de l'information échantillonnée sur la surface de mesure.

L'utilisation d'un pulse d'excitation nous offre une large plage fréquentielle pour caractériser l'antenne sous test. Néanmoins, le résultat du champ lointain dépend de l'étendue spatiale de la surface de mesure, de

la distance de mesure, de la directivité de l'antenne en fonction de la fréquence, de la fréquence maximale et minimale imposée par le pulse d'excitation.

Dans SERHIR [2015], j'ai mené cette étude sur l'échantillonnage pour établir que la taille de la surface de mesure est un paramètre clé pour le choix du pulse d'excitation. La troncature de la surface de mesure impacte plus ou moins sévèrement le champ lointain selon le comportement de la directivité de l'antenne à caractériser en fonction du spectre fréquentiel du pulse utilisé (amplitude du champ proche sur les frontières de la surface de mesure). Le choix du critère d'échantillonnage est lié à la fréquence maximale utilisée (critère de Nyquist-Shannon) et la distance de mesure est choisie en conformité avec la fréquence minimale de travail. Le champ proche est mesuré dans la zone de Fresnel dont la limite dépend de la fréquence basse choisie.

Publications représentatives sur le champ proche dans le domaine temporel

Les articles ci-dessous apportent les éléments théoriques de la transformation champ proche - champ lointain dans le domaine temporel. Ils appuient les conclusions avancées dans ce rapport.

- [RI.10] M. Serhir, « Transient UWB antenna near-field and far-field assessment from time domain planar near-field characterization : simulation and measurement investigations », *IEEE Transactions on Antennas and Propagation*, vol. 63, no. 11, pp. 4868-4876, 2015
- [RI.9] M. Serhir, « On the near-field sampling and truncation errors in planar time-domain near-field to far-field transformation », *Progress In Electromagnetics Research B*, 62, pp.181-193, 2015

Transient UWB Antenna Near-Field and Far-Field Assessment From Time Domain Planar Near-Field Characterization: Simulation and Measurement Investigations

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Abstract—In this paper, we present the experimental validation of time domain (TD) near-field to near-field (NFNF) and near-field to far-field (NFFF) transformations for ultrawideband (UWB) antenna transient characterization. The used computation schemes for near-field to near- or far-field transformations are based on the Green's function representation of the radiated field where NF and FF are directly calculated in the TD. The first step of the validation process comprises the electromagnetic simulation results dedicated to evaluate the accuracy of NFFF and NFNF transformations. The second step uses near-field measurement data of a Vivaldi antenna to validate the developed computation schemes, whereas the measurement and calibration procedures are fully described. The measured NF data using Supelec planar and cylindrical near-field facilities are also compared with the electromagnetic simulation transient results of the Vivaldi antenna.

Index Terms—Antenna measurement, near-field far-field (NFFF) transformation, time domain (TD).

I. INTRODUCTION

FOR ground-penetrating radar (GPR) applications, we are interested in measuring the free space transient response of ultrawideband (UWB) antennas composing the radar [1], [2]. The duration of the transient response is of crucial importance and the designers try to shorten the antenna free space impulse response while enlarging its frequency bandwidth of matched input impedance ($S_{11} < -10$ dB) [3]. More generally, GPR using UWB nondispersive antennas provides easily interpretable radargrams [4], [5].

UWB antennas transient characterization can be carried out using NF or FF techniques [6], [7]. Based on the Huygens principle, the NF measurement techniques make use of the tangential components of electric or magnetic field collected over a scan surface in the vicinity of the AUT. Then, these measured data are transformed to calculate the electromagnetic field at different distances [8]. The NF measurements can be conducted in a small and low-cost anechoic chamber and by

time-gating the measured data, one can filter out multiple reflections occurring during the measurement [9], [10]. This allows the data accuracy enhancement especially for low-frequency measurement in small anechoic chambers for which the absorption quality is unsatisfactory. The TD techniques permit also the radiation pattern measurement in nonanechoic environment as presented in [11]–[14]. They are well adapted for studying the parasitic electromagnetic radiation emitted from electronic devices for electromagnetic compatibility purposes [15].

Even the theoretical development of the NFNF and NFFF transformations are detailed in [6], [16], and [17], the experimental utilization of TD NF techniques are less popular than the FD techniques. This is due to difficult experimental implementation combined with the drawback related to the computer time requirement, which could make the application of the technique to cases of practical interest unrealistic. The transient NF measurement systems use wideband and small probes with low interaction with the AUT [18]. Moreover, in NFNF and NFFF transformation schemes, the measured E-field and its time derivative are involved in the calculation process. The accuracy of the NFNF and NFFF transformation results depends on the highest point of the chosen interpolation technique to keep working with the minimum number of measured data [19]. In this paper, we use the reconstruction formula to calculate and interpolate the E-field and its time derivative as well for NFNF and NFFF transformation schemes.

Here, we aim at validating NFNF and NFFF transformations using the experimental measurement and the electromagnetic simulation software CST MWS [20]. An UWB Vivaldi antenna designed for GPR is used for this purpose. The Vivaldi TD NF data are collected at different distances from the antenna. First, the NFFF transformation results are presented in order to outline the errors due to the measurement surface truncation. Second, we evaluate the accuracy of the NFNF calculation using different set of measured NF data collected at different distances from the AUT. Third, we present the experimental validation of the NFNF transformation scheme, which we have developed based on the Green's function representation. Our algorithms are applied to data obtained from the measurement of the Vivaldi antenna in Supelec planar near-field measurement facility. Finally, the experimental validation aims also at comparing the simulation and measurement NF results.

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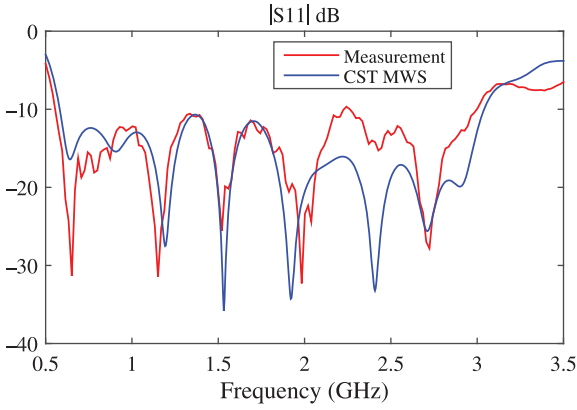


Fig. 1. S_{11} of the Vivaldi antenna used for our transient analysis (optimized for the GPR application). This present the comparison between the S_{11} of the antenna modeled in the CST MWS and the measured one (fabricated antenna).

This paper is organized as follows. In Section II, the planar NFNF and NFFF transformations formulations are described. In Section III, we present comparisons of FF and NF transformation results with the actual fields in the time domain (TD) using the simulation software CST MWS. Then, the measurement facilities used for our validation purpose are described and the NFNF transformation routine is validated using planar and cylindrical experimental setups in Section IV. Finally, concluding remarks are presented summarizing the relevance of our contribution in Section V. All theoretical formulations are expressed in the S.I. rationalized system with $e^{-i\omega t}$ time dependence.

II. PLANAR NEAR-FIELD TRANSFORMATION FOR FORWARD PROPAGATION

Let us consider UWB Vivaldi antenna. This is used to validate NFNF and NFFF transformation algorithms. The antenna has been optimized for the frequency band (0.5–3 GHz) and its S_{11} is presented in Fig. 1. In this paper, we have adopted the NFNF and NFFF transformation algorithms based on the Green's function representation where NF and FF are determined directly in the TD. This choice has been made to clearly distinguish our work from the NFNF or NFFF transformations based on modal expansions (plane wave expansion in the frequency domain) [15].

A. Planar NFNF Transformation

Using the NF tangential components measured over the planar surface at the distance x_{meas} from the AUT, the NFNF transformation algorithm allows calculating the transient field at different distances in the half space $x \geq x_{\text{meas}}$. The NFNF transformation is expressed as [6, eq. (5.6)] [16]

$$\vec{E}(\vec{r}, t) = \frac{-1}{2\pi} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \vec{R} \times \left[\frac{1}{cR^2} \vec{x} \times \frac{\partial \vec{E}_{\text{meas}}(\vec{r}_0, t - R/c)}{\partial t} + \frac{\vec{x} \times \vec{E}_{\text{meas}}(\vec{r}_0, t - R/c)}{R^3} \right] dydz, \text{ for } x \geq x_{\text{meas}}.$$

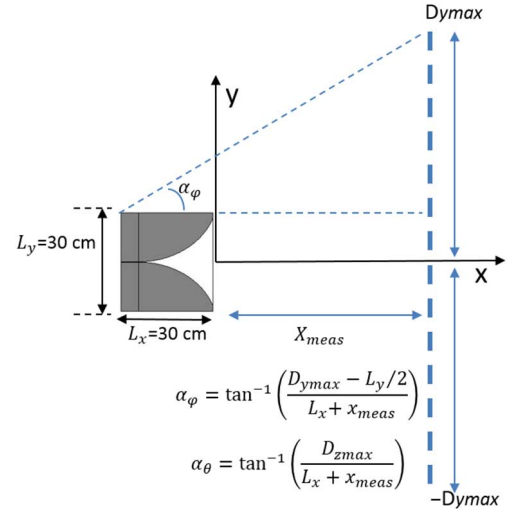


Fig. 2. Near-field measurement configuration at the distance x_{meas} from the AUT. The tangential components E_y and E_z are regularly recorded over the scan surface in the zy plane. The far-field validity area is dependent on the angular domain described by α_θ and α_ϕ .

$$\approx \frac{-1}{2\pi} \sum_{m=-N_y}^{N_y} \sum_{n=-N_z}^{N_z} \vec{R} \times \left[\frac{1}{cR^2} \vec{x} \times \frac{\partial \vec{E}_{\text{meas}}(\vec{r}_{0mn}, t - R/c)}{\partial t} + \frac{\vec{x} \times \vec{E}_{\text{meas}}(\vec{r}_{0mn}, t - R/c)}{R^3} \right] \Delta y_0 \Delta z_0. \quad (1)$$

The calculated field $\vec{E}(\vec{r}, t)$ in the left-hand side of (1) is expressed at the observation point $\vec{r} = x\vec{x} + y\vec{y} + z\vec{z}$ at the instant t as a function of field $\vec{E}_{\text{meas}}(\vec{r}_{0mn}, t - R/c)$ and its time derivative $\partial \vec{E}_{\text{meas}}(\vec{r}_{0mn}, t - R/c) / \partial t$. These are measured at the position $\vec{r}_{0mn} = x_{\text{meas}}\vec{x} + m\Delta y_0\vec{y} + n\Delta z_0\vec{z}$ at the instant $t - R/c$ with $\vec{R} = \vec{r} - \vec{r}_{0mn}$, $R = |\vec{r} - \vec{r}_{0mn}|$ and c is the free space light velocity. As presented in Fig. 2, the measurement surface is defined for $-D_{y\text{max}} \leq y_{\text{meas}} \leq D_{y\text{max}}$ and $-D_{z\text{max}} \leq z_{\text{meas}} \leq D_{z\text{max}}$. The NF is sampled regularly using the Nyquist criterion $\Delta t = \pi / \omega_{\text{max}}$, $\Delta y_0 = \Delta z_0 = \pi c / \omega_{\text{max}}$ with ω_{max} being the maximum angular pulsation of the AUT excitation signal, $N_y = D_{y\text{max}} / \Delta y_0$ and $N_z = D_{z\text{max}} / \Delta z_0$.

From (1), the first step in NFNF transformation is the interpolation of the measured field $\vec{E}_{\text{meas}}(\vec{r}_{0mn}, t)$ and its time derivative to set their values at the instant $t - R/c$. The accuracy of NFNF transformation depends on the chosen interpolation method. In our analysis, we use the reconstruction formula to interpolate and assess the time derivative of the measured field at $t - R/c$.

Let us consider NF data measured over the scan surface at x_{meas} from the AUT. This NF is a band-limited function that can be accurately interpolated using the reconstruction formula. Namely, the field $\vec{E}(\vec{r}_{0mn}, t)$ at the instant t in the time window $t_0 \leq t \leq t_0 + (N_t - 1) \Delta t$ is calculated using the

field samples $\vec{E}(\vec{r}_{0mn}, t_0 + l\Delta t)$ with the cardinal series as follows [6], [17]:

$$\begin{aligned}\vec{E}(\vec{r}_{0mn}, t) &= \sum_{l=0}^{N_t-1} \frac{\sin\left[\pi\left(\frac{t-t_0-l\Delta t}{\Delta t}\right)\right]}{\pi\left(\frac{t-t_0-l\Delta t}{\Delta t}\right)} \vec{E}(\vec{r}_{0mn}, t_0 + l\Delta t) \\ &= \sum_{l=0}^{N_t-1} h(t) \cdot \vec{E}(\vec{r}_{0mn}, t_0 + l\Delta t)\end{aligned}$$

$$\text{with, } h(t) = \sin c\left[\pi\left(\frac{t-t_0-l\Delta t}{\Delta t}\right)\right], \quad \text{for } t_0 \leq t \leq t_0 + (N_t - 1)\Delta t. \quad (2)$$

This means that in the time window $t_0 \leq t \leq t_0 + (N_t - 1)\Delta t$ and provided the NF is sampled respecting the Nyquist criterion $\Delta t = \pi/\omega_{max}$, the time derivative of the E-field is expressed as

$$\begin{aligned}\frac{\partial \vec{E}}{\partial t}(\vec{r}_{0mn}, t) &= \sum_{l=0}^{N_t-1} \frac{\partial h(t)}{\partial t} \vec{E}(\vec{r}_{0mn}, t_0 + l\Delta t) \\ \text{with } \frac{\partial h}{\partial t}(t) &= \frac{1}{t - (t_0 + l\Delta t)} \\ &\quad \times \left[\cos\left(\pi\frac{t - (t_0 + l\Delta t)}{\Delta t}\right) - h(t) \right]. \quad (3)\end{aligned}$$

In our analysis, we use (2) and (3) to calculate the E-field and its time derivative between time-samples (at $t - R/c$) provided the sampling Nyquist criterion is respected. This criterion, which is linked to spatial NF sampling $\Delta y_0 = \Delta z_0 = c\Delta t$, is sufficient to accurately perform the NFFN transformation.

B. Planar NFFF Transformation

The TD NFFF transformation is formulated as [6, eq. (5.57)] [16], [17]

$$\begin{aligned}\vec{E}_{FF}(\theta, \varphi, t) &= \frac{-1}{2\pi c} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \vec{e}_r \\ &\quad \times \left[\vec{x} \times \frac{\partial \vec{E}_{meas}}{\partial t}(\vec{r}_0, t + \vec{e}_r \cdot \vec{r}_0/c) \right] dydz, \text{ for } x \geq x_{meas} \\ &\approx \frac{-1}{2\pi c} \sum_{m=-N_y}^{N_y} \sum_{n=-N_z}^{N_z} \vec{e}_r \\ &\quad \times \left[\vec{x} \times \frac{\partial \vec{E}_{meas}}{\partial t}(\vec{r}_{0mn}, t + \vec{e}_r \cdot \vec{r}_{0mn}/c) \right] \Delta y_0 \Delta z_0. \quad (4)\end{aligned}$$

The antenna radiation pattern at the far-field distance $\vec{E}_{FF}(\theta, \varphi, t)$ in the direction (θ, φ) is written as a function of the time derivative of the tangential measured NF $\partial \vec{E}_{meas}/\partial t$. The spherical coordinates (θ, φ) are associated with the planar coordinates system (x, y, z) , $\vec{e}_r = \vec{r}/|r| = \cos\varphi \sin\theta \vec{x} + \sin\varphi \sin\theta \vec{y} + \cos\theta \vec{z}$ is the FF observation direction and the measured field data \vec{E}_{meas} are collected at $\vec{r}_0(x_{meas}, y_{meas}, z_{meas})$. As expressed in (4), the infinite integration is truncated and we implicitly suppose that the field is weak beyond the measurement surface edges. However, one of

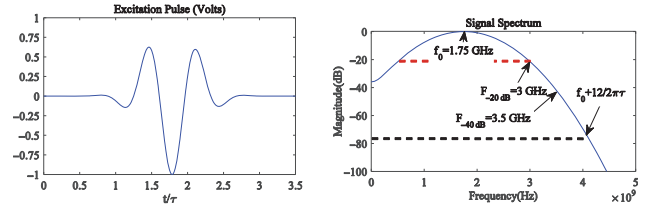


Fig. 3. TD excitation signal applied to the Vivaldi antenna (left). The excitation signal spectrum (right), the maximum of the signal spectrum is at 1.75 GHz.

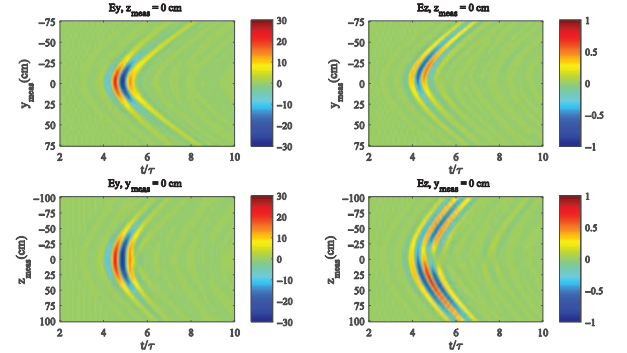


Fig. 4. Vivaldi antenna E_y (V/m) and E_z (V/m) radiated near-field as a function of the time at $x_{meas} = 30$ cm for the plane cuts $z_{meas} = 0$ and $y_{meas} = 0$.

the advantages of TD near-field measurements is that one can obtain exact early-time far-field pattern values even though the field outside the scanned area is strong.

In Sections III and IV, we present the validation results of the NFFN and NFFF transformation using simulation and measurement results and the E-field time derivative calculated using (3).

III. ELECTROMAGNETIC SIMULATION RESULTS

We have developed a routine expressing (4) to study the NFFF transformation results when using different NF data collected at different measurement planes x_{meas} . The Vivaldi antenna has been simulated using the transient simulator of the CST MWS to provide the NF data. The AUT is excited using a cosinusoidal voltage modulated by Gaussian signal to cover the AUT frequency band (0.5–3.5 GHz), $e(t) = -\cos(\omega_0(t - t_0))e^{-4(t-t_0)^2/\tau^2}$ with $\omega_0 = 2\pi f_0$, $f_0 = 1.75$ GHz, $\tau = 0.8 \cdot 10^{-9}$ s, and the delay $t_0 = 1.75\tau$. The excitation signal and its spectrum normalized at $f_0 = 1.75$ GHz are presented in Fig. 3.

The antenna TD NF is collected over the plane $x = x_{meas}$ respecting regularly spaced positions Δy_0 and Δz_0 . Based on [6], the bandwidth of the Gaussian signal is approximately equal to $12/2\pi\tau$. One can sample the radiated field respecting the Nyquist criterion $\Delta y_0 = \Delta z_0 = \lambda_{min}/2$, where $\lambda_{min} = c/f_{max}$, and f_{max} is the maximum frequency expressed as $f_{max} = f_0 + 12/2\pi\tau = 4.14$ GHz, without any loss of information.

Using the CST MWS, we generate the antenna transient-radiated NF over the plane $x_{meas} = 30$ cm. This scan surface is delimited by $D_{zmax} = 100$ cm and $D_{ymax} = 75$ cm as defined in Fig. 2 and the NF in the cut planes $y_{meas} = 0$ and $z_{meas} = 0$ are presented in Fig. 4 for E_y and E_z components. The Vivaldi

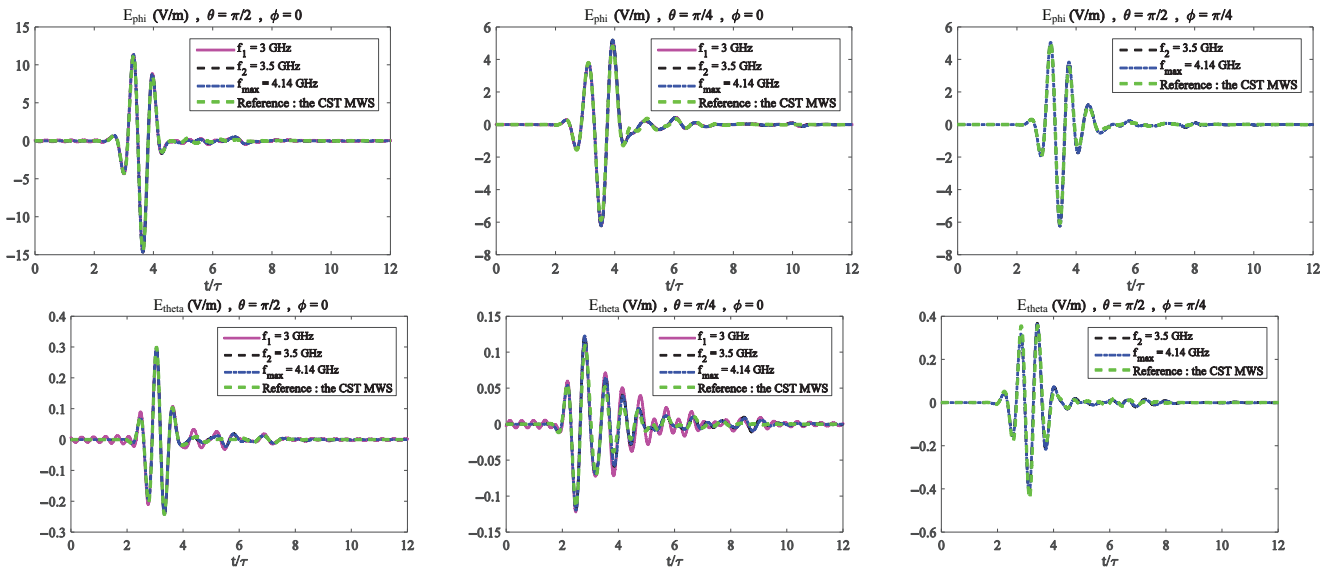


Fig. 5. Transient E_{θ} (V/m) and E_{ϕ} (V/m) far-field radiation pattern comparison considering different sampling frequencies for the spatial directions: $(\theta = \pi/2, \phi = 0)$, $(\theta = \pi/4, \phi = 0)$, and $(\theta = \pi/2, \phi = \pi/4)$. The NF data are collected at $x_{\text{meas}} = 30$ cm.

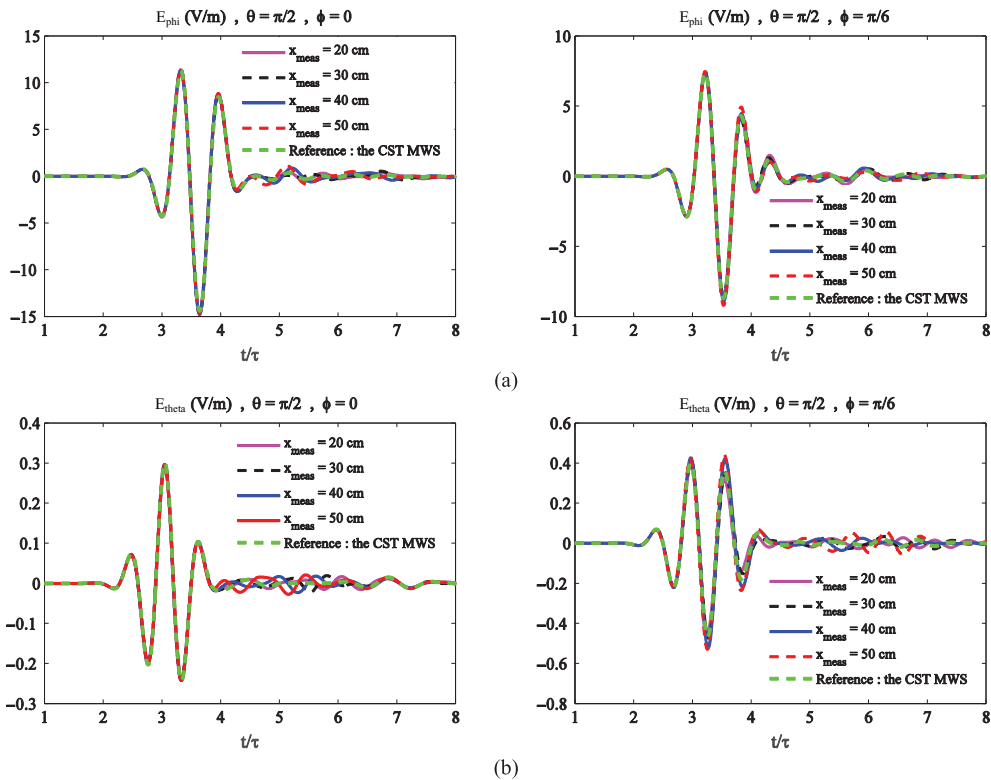


Fig. 6. Transient E_{θ} (V/m) and E_{ϕ} (V/m) far-field radiation pattern comparison considering different NF measurement distances (20, 30, 40, 50 cm) for the spatial directions. (a) $\theta = \pi/2, \phi = 0$ and (b) $\theta = \pi/2, \phi = \pi/6$.

antenna is linearly polarized and the E-field is predominantly radiated through the E_y component and the E_z component is negligible comparatively to E_y .

Since the S_{11} (Fig. 1) of the Vivaldi antenna demonstrates a passband behavior (0.5–3 GHz), we are interested in evaluating the accuracy of the NFFF transformation as a function of the NF sampling frequency. To do so, we compare the actual FF obtained from the CST MWS with the NFFF transformation results using three different NF sampling frequencies $f_1 = 3$ GHz, $f_2 = 3.5$ GHz, $f_{\text{max}} = 4.14$ GHz. In Fig. 5, we

present the FF comparison at the observation directions $(\theta = \pi/2, \phi = 0)$, $(\theta = \pi/4, \phi = 0)$, and $(\theta = \pi/2, \phi = \pi/4)$ for the FF components E_{θ} and E_{ϕ} .

As seen in Fig. 5, good agreements are noticed between the E_{ϕ} (copolar) component of the actual FF and the ones resulting from NFFF transformations. The use of the sampling frequency $f_1 = 3$ GHz is responsible of the aliasing effects clearly seen for the cross-polarization component E_{θ} . In addition, the differences between the curves of Fig. 5 are due to measurement surface truncation. For the far-field observation point $(\theta = \pi/2,$

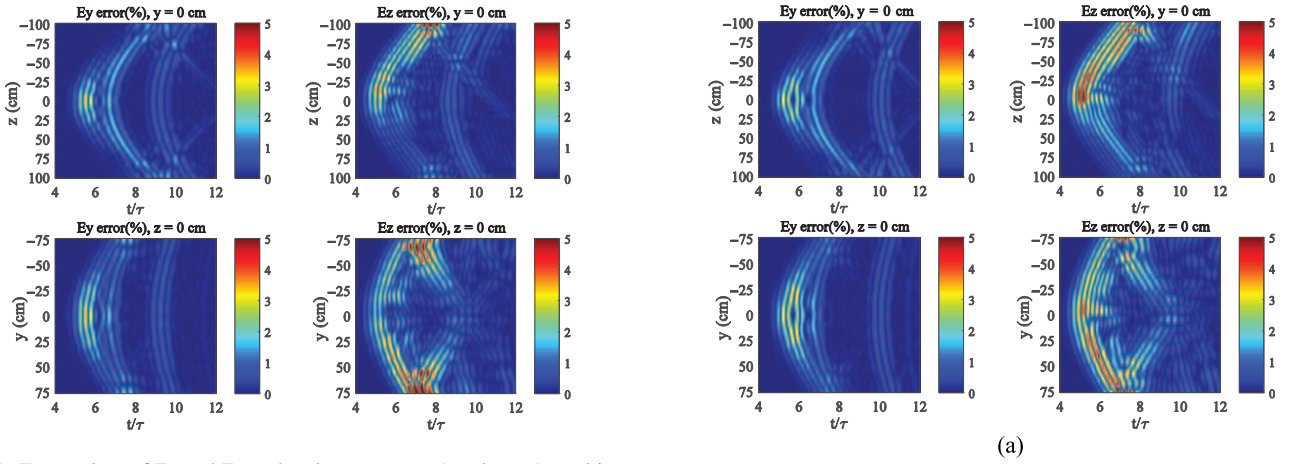


Fig. 7. Error values of E_y and E_z at the plane cuts $y = 0$ and $z = 0$ resulting from the NFN transformation. The NF data is measured at $x_{\text{meas}} = 20$ cm and transformed to calculate the transient field at 50 cm from the AUT.

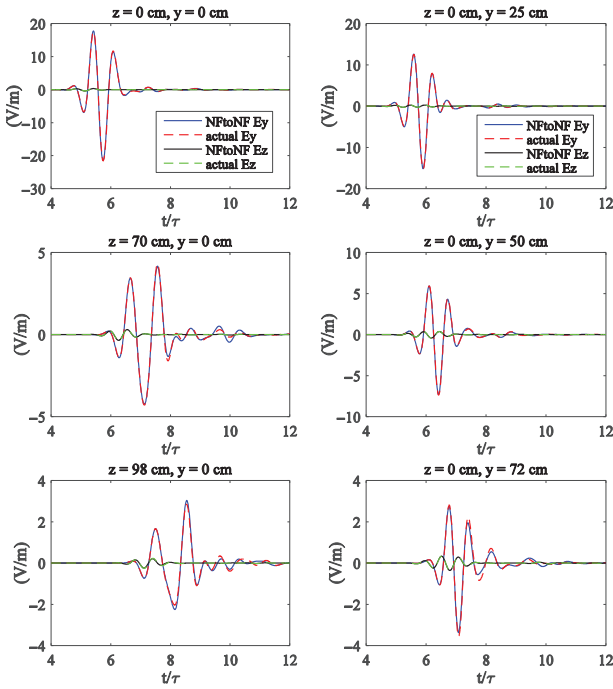


Fig. 8. Actual field comparison with transient E_y (V/m) and E_z (V/m) resulting from the NFN transformation. The NF data is measured at $x_{\text{meas}} = 20$ cm and transformed to calculate the antenna transient response at the spatial positions A($x = 50$ cm, $y = 0$, $z = 0$), B(50 cm, 25 cm, 0), C(50 cm, 0 , 70 cm), D(50 cm, 50 cm, 0), E(50 cm, 0 , 98 cm), and F(50 cm, 72 cm, 0).

$\varphi = 0$), the difference between the actual FF and the calculated one starts at $t = 5.11\tau$ that corresponds to the arrival time when the center of the pulse reached the measurement edge $D_{y\text{max}} = 75$ cm. The error corresponding to the measurement edge $D_{z\text{max}} = 100$ cm is visible around $t = 6.10\tau$.

Using the sampling frequency $f_{\text{max}} = 4.14$ GHz, we compare the NFFF results calculated from NF data collected at different planes $x_{\text{meas}} = 20, 30, 40,$ and 50 cm. These calculated FF are compared in Fig. 6 with the actual one at the directions ($\theta = \pi/2$, $\varphi = \pi/6$) and ($\theta = \pi/3$, $\varphi = 0$). As presented in Fig. 6, good agreements are noticed for $2.5\tau \leq t \leq t_{\text{tr}}$, where t_{tr} depends on the measurement distance x_{meas} and the FF observation point (θ , φ). This effect is known as the truncation error

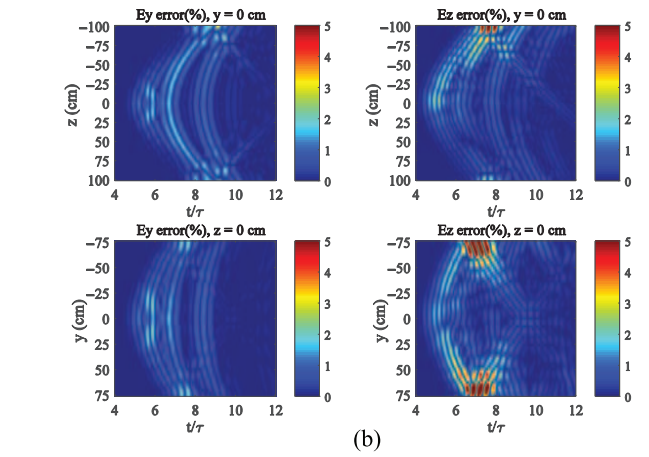


Fig. 9. Error values of E_y and E_z at the plane cuts $y = 0$ and $z = 0$ resulting from the NFN transformation: the NF data is measured at (a) $x_{\text{meas}} = 10$ cm and (b) $x_{\text{meas}} = 30$ cm and transformed to calculate the field at 50 cm from the AUT.

and the FF reliable region is defined by α_θ and α_φ expressed in Fig. 2 depends on the measurement distance.

Based on the presented simulation results, the developed NFFF transformation routine has been validated using the CST MWS. The NFFF transformation results have shown a satisfactory accuracy while using the Nyquist criterion for NF sampling.

The next comparisons aim at validating the TD NFN transformation routine. For this, we provide tangential NF data at $x_{\text{meas}} = 20$ cm for $-75 \text{ cm} \leq y_{\text{meas}} \leq 75$ cm and $-100 \text{ cm} \leq z_{\text{meas}} \leq 100$ cm sampled using $f_{\text{max}} = 4.14$ GHz. Thereafter, we perform the NFN transformation routine to calculate the field at 50 cm from the AUT. The NFN transformation results (E_{NFNtoNF}) are compared with the actual NF (E_{ref}) at the plane cuts ($x = 50$ cm, $y = 0$, $-100 \text{ cm} \leq z \leq 100$ cm) and ($x = 50$ cm, $-75 \text{ cm} \leq y \leq 75$ cm, $z = 0$) using the error expressed as

$$\text{error}(y, z, t) = 100 \frac{|E_{\text{NFNtoNF}}(y, z, t) - E_{\text{ref}}(y, z, t)|}{\max(|E_{\text{ref}}(y, z, t)|)} \quad (5)$$

$t_0 \leq t \leq t_0 + (N_t - 1)\Delta t, -D_{y\text{max}} \leq y \leq D_{y\text{max}}, -D_{z\text{max}} \leq z \leq D_{z\text{max}}$

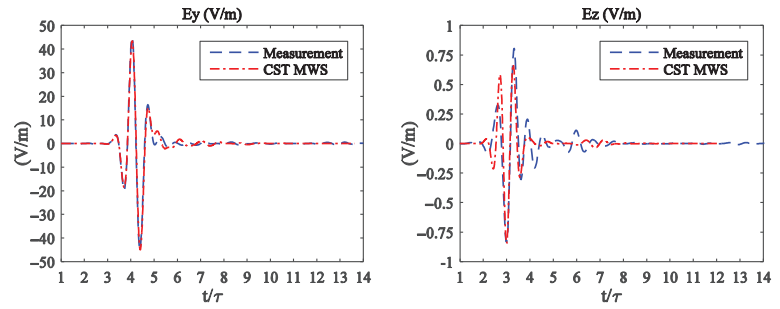
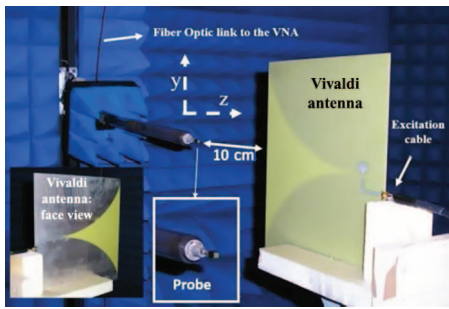


Fig. 10. (Left) Supelec planar measurement setup used for NF antenna measurement. The Vivaldi antenna is placed at 10 cm from the measuring probe (En-probe EFS 105). The probe displacements (translation in y and z directions) are controlled by automated process. (Right) The Vivaldi antenna transient-radiated field [E_y (left) and E_z (right)] in the boresight direction is presented and compared with the CST MWS results after the calibration procedure (normalization).

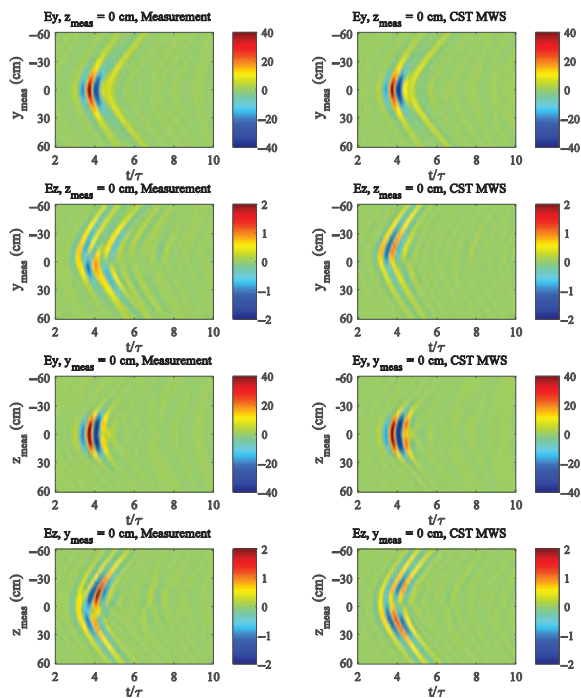


Fig. 11. Comparison between the measured Vivaldi near-field [E_y (V/m) and E_z (V/m)] at the distance $x_{\text{meas}} = 10$ cm from the Vivaldi antenna with the CST MWS results for the plane cuts $y_{\text{meas}} = 0$ and $z_{\text{meas}} = 0$.

The error values are presented in Fig. 7 for the E_y and E_z components. As it is seen, the NFN transformation errors are very low for the E_y components (copolar) and reaches 5% at the edges of the measurement surface ($y = \pm 75$ cm and $z = \pm 100$ cm) for E_z component (cross-polar). These errors are due to NF measurement surface truncation. In Fig. 8, the calculated AUT transient-radiated fields E_y and E_z are compared with the actual ones at different observation points A($x = 50$ cm, $y = 0$, $z = 0$), B(50 cm, 25 cm, 0), C(50 cm, 0, 70 cm), D(50 cm, 50 cm, 0), E(50 cm, 0, 98 cm), and F(50 cm, 72 cm, 0). The calculated NF agrees very well with the actual one even for the observation points E and F situated in the area of the high error values (maximum 5%) of Fig. 7.

To get a deep insight into the origin of the NFN transformation errors, we calculate the field at $x = 50$ cm from the AUT using NF data collected over the scan planes $x_{\text{meas}} = 10$ and 30 cm. These are compared with the actual NF (the CST

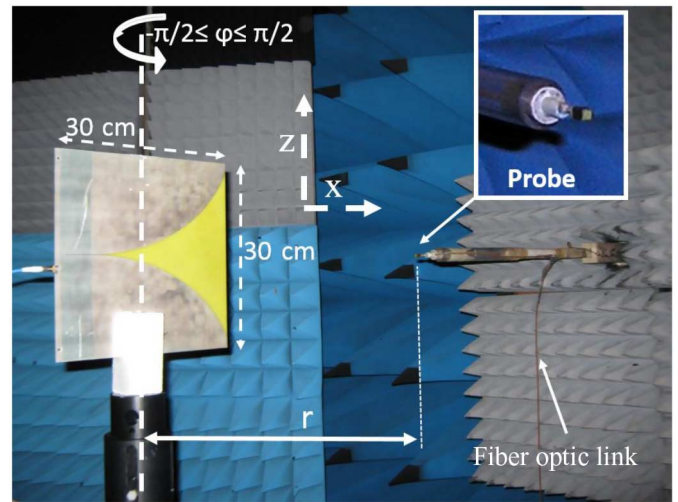


Fig. 12. Supelec cylindrical near-field measurement facility used for our experimental validation. The AUT is rotated following the angle $-\pi/2 \leq \varphi \leq \pi/2$. We use the same probe as the planar near-field facility (En-probe EFS 105).

MWS) at the plane cuts ($x = 50$ cm, $y = 0$, -100 cm $\leq z \leq 100$ cm) and ($x = 50$ cm, -75 cm $\leq y \leq 75$ cm, $z = 0$). The resulting error values are presented in Fig. 9(a) and (b). From Figs. 7 and 9, we can conclude that the calculated results are completely unreliable outside of a certain spectral region. The erroneous ripples are due to the discontinuity of the measured field at the edge of the truncated surface. Hence, the entire calculated pattern is always affected by errors, and it is not possible to define a region where the error is completely zero. However, the concept of the spectral reliable far-field region is usually applied to refer to the region in which the error is not negligible but is low [21].

From Figs. 7 and 9, the highest error value over the principal component E_y stays stable around 3%. In Figs. 7 and 9(a) corresponding to NF data collected at $x_{\text{meas}} = 10$ and 20 cm, the truncation error is higher in the region -50 cm $\leq y \leq 50$ cm and -75 cm $\leq z \leq 75$ cm in comparison with the results of $x_{\text{meas}} = 30$ cm. The truncation error is directly linked to the level of the discontinuity of the measured field at the edge of the truncated surface. At $x_{\text{meas}} = 10$ and 20 cm, this discontinuity is more important than for $x_{\text{meas}} = 30$ cm.

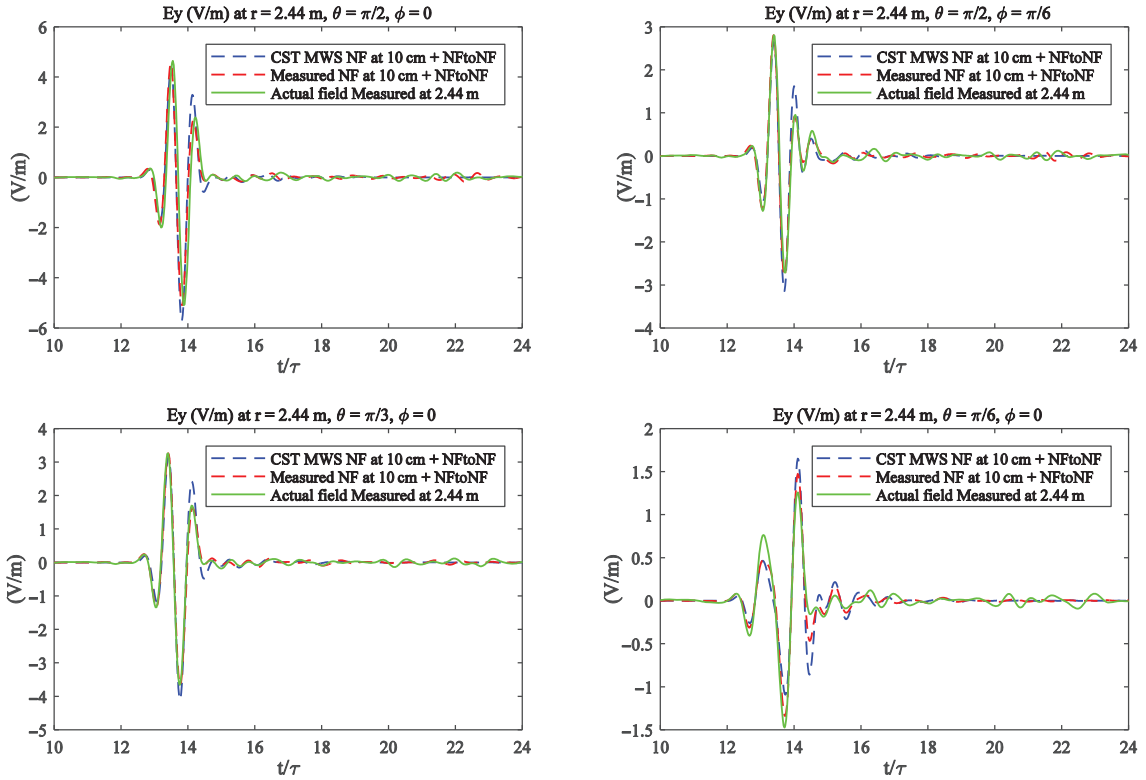


Fig. 13. NFN results at the spatial points A($r = 2.44$ m, $\theta = \pi/2$, $\phi = 0$), B($r = 2.44$ m, $\theta = \pi/2$, $\phi = \pi/6$), C($r = 2.44$ m, $\theta = \pi/3$, $\phi = 0$) and D($r = 2.44$ m, $\theta = \pi/6$, $\phi = 0$) compared with the CST MWS and the directly measured field in the cylindrical facility.

Up to now, we have studied the accuracy of NFFF and NFNF transformation routines using electromagnetic simulation software (the CST MWS). In the next section, we describe the experimental validation of the developed transformation routines.

IV. MEASUREMENT RESULTS

The Vivaldi antenna used in the simulation presented above has been fabricated and Supeclec planar NF range to measure its radiated field is presented in Fig. 10. The Vivaldi antenna and the measurement probe are placed inside the anechoic chamber of dimensions $4 \text{ m} \times 5 \text{ m} \times 3 \text{ m}$. We measure the tangential near-field components E_y and E_z while moving the probe respecting a regular spacing $\Delta y = \Delta z = 3 \text{ cm}$ between measurement points over the planar surface ($-60 \text{ cm} \leq y_{\text{meas}} \leq 60 \text{ cm}$ and $-60 \text{ cm} \leq z_{\text{meas}} \leq 60 \text{ cm}$) at $x_{\text{meas}} = 10 \text{ cm}$ in front of the antenna. For our measurement investigations, we have used the wideband probe “En-Probe EFS-105” [22]. This probe is composed of a small dipole (dimensions $6.6 \text{ mm} \times 6.6 \text{ mm}$) connected to the base unit through a fiber optic link and the base unit is connected to the measuring VNA through a short cable. This probe is dedicated to NF measurement covering the frequency band (5 MHz–3 GHz). The probe performances can be extended to 3.5 GHz with slight measurement sensitivity degradation in the frequency band (3–3.5 GHz).

The antenna NF measurement is carried out in the frequency band (0.3–3.5 GHz) using a Vector Network Analyzer (Agilent ENA 5071b). Thereafter, the measured frequency domain NF complex data (amplitude and phase) are Fourier transformed to

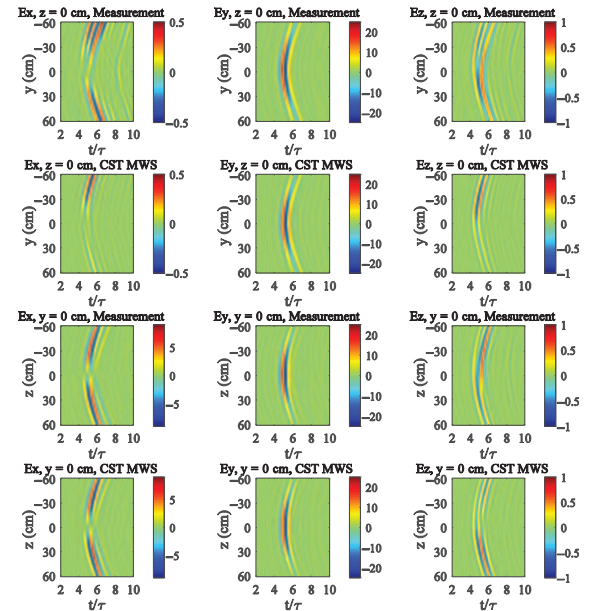


Fig. 14. NFN results at $x = 40 \text{ cm}$ compared with the CST MWS results at the plane cuts $y = 0$ and $z = 0$. NF data (E_y and E_z) have been measured at $x_{\text{meas}} = 10 \text{ cm}$ from the AUT and transformed to reach the planar surface $x = 40 \text{ cm}$. The comparison comprises E_x , E_y , and E_z transient field components.

set the TD NF of the Vivaldi antenna when excited by the pulse given in Fig. 2. In Fig. 10, we compare the measured TD NF at the point ($x_{\text{meas}} = 10 \text{ cm}$, $y_{\text{meas}} = 0$, $z_{\text{meas}} = 0$) with the one obtained from the CST MWS. The E_y component (copolar) fits

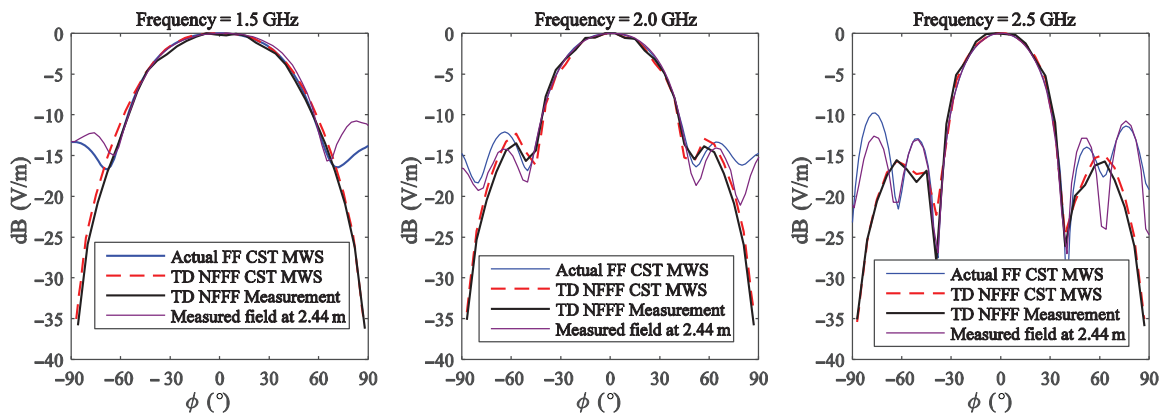


Fig. 15. Frequency domain results comparison. The FF issued from measured NF data ($x_{\text{meas}} = 10$ cm) using the planar facility is compared with the CST MWS directly calculated FF for the frequencies 1.5, 2, and 2.5 GHz for the FF cut plane $\theta = \pi/2$. The NFFF transformation results are Fourier transformed to calculate the frequency domain FF. The FF validity area is about $\pm 48^\circ$.

the CST results well. However, the E_z component (cross-polar) shows some discrepancies.

The NF probe used in this measurement campaign is supposed to behave like a point source. For probe calibration procedure, we have normalized the measured NF magnitude (copolar) in order to reach the CST MWS maximum NF level. This is performed at the principal direction ($x_{\text{meas}} = 10$ cm, $y_{\text{meas}} = z_{\text{meas}} = 0$) and the normalization coefficient is applied to measured NF data (E_y and E_z) over the measurement surface.

In Fig. 11, we compare the measured NF at $x_{\text{meas}} = 10$ cm with the CST MWS NF data after the calibration step. The agreements are very satisfactory for E_y component meanwhile the cross-polarization (E_z) shows some differences compared with the CST results. This is due to many reasons: in part to the antenna fabrication accuracy, measurement positioning errors (alignment), cross-polarization of the probe is not well defined, probe sensitivity for low-field levels.

Using the NFNFF routine, we transform the NF data measured at plane $x_{\text{meas}} = 10$ cm to calculate the field at the spatial positions A ($r = 2.44$ m, $\theta = \pi/2$, $\varphi = 0$), B ($r = 2.44$ m, $\theta = \pi/2$, $\varphi = \pi/6$), C ($r = 2.44$ m, $\theta = \pi/3$, $\varphi = 0$), and D ($r = 2.44$ m, $\theta = \pi/6$, $\varphi = 0$). The reference antenna radiation patterns used for these comparisons have been measured at Supelec cylindrical NF measurement facility placed in an anechoic chamber of dimensions $6 \text{ m} \times 7 \text{ m} \times 8 \text{ m}$. This measurement facility is presented in Fig. 12 and the E-field is measured using the same probe as the planar NF range (En-Probe EFS-105) at the plane cuts $\theta = \pi/2$ and $\varphi = 0$. For our comparison, we consider the CST NF data ($x_{\text{meas}} = 10$ cm) which are transformed to reach the spatial positions A, B, C, and D. These comparisons are carried out for the E_y components (copolar) and are presented in Fig. 13. As it is seen, good agreements are noticed between the measurement results. The NFNFF transformation yields satisfactory results when using NF data measured at 10 cm from the AUT and transformed in order to calculate the transient response at the positions A, B, and C. The differences observed for the position D are due to the NF measurement surface truncation error. The CST MWS NF collected at 10 cm is also transformed to calculate the field at the positions A, B, C, and

D. This has been done in order to verify that our calibration procedure stay correct and no additional calibration step is needed to compensate for the probe spatial response (the probe acts actually as a point source).

The NFNFF transformation allows calculating E_x , E_y , and E_z components in the half space $x \geq x_{\text{meas}}$ in front of the AUT using the tangential components measured at a given distance x_{meas} . Indeed, using Vivaldi antenna transient NF measured in the planar NF facility at $x_{\text{meas}} = 10$ cm from the AUT, we perform NFNFF transformation to set the transient field over the planar surface at $x = 40$ cm ($-60 \text{ cm} \leq y \leq 60$ cm, and $-60 \text{ cm} \leq z < 60$ cm). These are compared with the CST MWS results at the plane cuts ($x = 40$ cm, $y = 0$, $-60 \text{ cm} \leq z \leq 60$ cm) and ($x = 40$ cm, $-60 \text{ cm} \leq y \leq 60$ cm, $z = 0$). As it is seen in Fig. 14, the NFNFF transformation results fit the simulated E_y component (copolar) well. The normal component E_x for the plane cut $y = 0$ agrees well with the simulated results. However, low-level fields (cross-pol) present many discrepancies compared with the CST MWS results. Generally, the results of NFNFF and NFFF transformations present some difficulties to set accurately the low-level cross-polarization field since the cross-polarized NF component is difficult to measure accurately with the EFS-105 probe (Fig. 10).

Finally, using the NF data measured at $x_{\text{meas}} = 10$ cm in the planar NF facility, we perform NFFF transformation to calculate the transient far field over the plane cut ($\theta = \pi/2$, $-\pi/2 \leq \varphi \leq \pi/2$). The NFFF results are Fourier transformed to determine the AUT radiation pattern in the frequency domain. These are compared in Fig. 15 with the actual FF obtained from the CST MWS at the frequencies 1.5, 2, and 2.5 GHz. The comparison also includes the TD NFFF transformation results of the CST MWS NF data collected at $x_{\text{meas}} = 10$ cm from the AUT and the measured field in the cylindrical measurement facility at 2.44 m from the AUT. From Fig. 15, a satisfactory agreement is noticed between the TD NFFF and the CST MWS FF results for the angular region $-48^\circ \leq \varphi \leq 48^\circ$. Beyond this area ($\varphi \leq -48^\circ$ and $\varphi \geq 48^\circ$), the differences between the different curves are due to the measurement surface truncation.

V. CONCLUSION

This paper has presented the experimental validation of TD NFNF and NFFF transformation for antenna transient characterization. Based on the Green's function representation, we have derived NFNF and NFFF transformation calculation schemes which have been validated using electromagnetic simulation software and experimental measurement data. The studied Vivaldi antenna radiated transient field has been measured and near-field to near-/far-field transformations have resulted in a good accuracy compared with the CST MWS simulated transient field. The NFNF transformation can also be used to determine the normal components of the radiated transient field. In addition, the frequency domain comparison has shown a satisfactory accuracy for far-field calculation.

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On the Near-Field Sampling and Truncation Errors in Planar Time-Domain Near-Field to Far-Field Transformation

Mohammed Serhir*

Abstract—This paper studies the effect of three important parameters in planar time-domain (TD) near-field (NF) to far-field (FF) transformation. These parameters are the NF spatial sampling, NF measurement distance and scan surface truncation. The effect of these parameters over the TD FF accuracy are difficult to predict for Ultra Wide Band antennas. In this paper we aim to choose the optimum NF measurement parameters guaranteeing accurate calculation of the time-domain far-field. This allows the optimization of the computation time and memory requirements. Computations using analytic array of elementary dipoles radiation pattern are used to study the impact of each parameter in time-domain near-field antenna measurement. The comparison of the far-field results are presented in time and frequency domains. In particular, it is shown that the choice of the measurement distance and the size of the scan surface decide predominantly on the frequency band of accurate FF calculation. The used formalism in this paper for the NF to FF transformation is based on the Green's function.

1. INTRODUCTION

The Ultra Wide Band (UWB) antennas operating over large frequency bandwidth have attracted a lot of interest over wide telecommunication applications [1–5]. The measurement of the radiation pattern of these antennas has to be carried in a wide frequency band, which is time consuming. For these antennas, the time-domain (TD) techniques seem to be more adapted than frequency-domain (FD) techniques. Indeed, using a short pulse, one can measure the antenna transient response covering the frequency band of interest. This measurement is fulfilled using near-field (NF) or far-field (FF) techniques [6].

In TD FF measurements we record the antenna under test (AUT) transient response when it is excited by a specific pulse. The excitation pulse has to cover the frequency band of interest and supply the AUT with enough power that allows the measurement with acceptable signal to noise ratio at far-field distances. In TD NF techniques the tangential components of radiated field are collected in the antenna vicinity over a scan surface (planar, cylindrical or spherical). Then, the measured data are transformed to calculate the asymptotic behavior of the antenna far-field. For NF measurement, we use pulse generators providing a trade-off between the output voltage magnitude and a short pulse rise-time for wide frequency band characterization.

For accuracy improvement, the post processing tools such as TD gating technique are used to filter out the multiple reflections occurring during the measurement [7]. The TD techniques allow the radiation pattern measurement in non-anechoic environment as presented in [8]. These are well adapted for the parasitic electromagnetic radiation emitted from electronic devices for electromagnetic compatibility purpose [9]. Also, using TD techniques one can measure the radiation pattern of radar structures fed by non-sinusoidal excitation signals. Particularly, when the feeding system is integrated in the radiating antennas for radar applications.

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The main difficulty related to the practical use of NF TD techniques concerns the ability to manage the quantity of measured data. The required computation time and memory storage became problematic for large measurement surface and long-time window (long transient response). The quantity of NF data depends on the used time and spatial sampling criteria. These are determined based on the AUT excitation signal maximum frequency as presented in [10]. Authors in [11–13] present a different NF sampling strategy that aims at representing the NF measurement data using plane polar samples allowing a minimum redundancy in NF sampling. In the FD, Wang in [14] presents a detailed analysis on the theory and practices for planar near-field measurement. The sampling strategies, the NF probe characteristics, the filtering of the NF data and the evanescent modes effect have been addressed. Authors of [15–18] have presented some experimental setups for NF or FF antenna measurement. These papers did not present the sampling strategies used in the NF or FF measurements.

In planar TD near-field antenna measurement, the sampling approach is based on the assumption that the time signal is band-limited to a maximum angular frequency $\omega_{\max} = 2\pi c/\lambda_{\min}$. This leads to the NF sampling criterion $\Delta y = \Delta z = \lambda_{\min}/2$ and the AUT excitation pulse defines approximately the frequency band of interest (minimum and maximum frequencies). This assumption gives a reduced view of the reality of TD NF measurement practices. Considering the spatial sampling criterion based on the highest frequency of the AUT excitation signal ($\lambda_{\min}/2$), as presented in [10], does not guarantee the accuracy of the TD far-field calculation. Indeed, the AUT *directivity* depends on the frequency and using the same measurement surface for time-domain FF calculation leads to a frequency dependent truncation error. Also, the NF measurement has to be carried in the radiating NF zone (Fresnel zone). This Fresnel zone distance depends on the AUT dimensions and the operating frequency. For NF TD technique, the measurement distance has to be adapted to avoid the reactive NF zone for low frequencies. These three parameters are responsible of the TD FF errors.

In this paper we propose to determine the AUT TD FF using different sampling criteria and for several NF measurement distances. By comparing the far-field results, we show that the TD FF errors are resulting from NF evanescent modes with low measurement distances, under-sampling and scan surface truncation errors. It is difficult to distinguish between these errors in the TD FF directly. The frequency-domain (FD) comparisons are presented to make out the effect of each error.

In the NF to FF transformation scheme we use the reconstruction formula to rigorously interpolate the NF in the time-domain. We propose a simple and rigorous formulation to calculate the E -field time derivative from the measured E -field as complementary development of the analysis presented in [10]. The computation scheme for NF to FF field transformation is validated using analytic radiating array composed of infinitesimal dipoles. The paper is structured as follows. In Section 2, the NF to FF formalism based on the Green's function is described and the different formulation needed for our analysis are commented. In Section 3, we present the transformation results and a comparison of the calculated far-field with the actual one in the time and the frequency domains. In Section 4, recommendations and concluding remarks are presented.

2. TIME-DOMAIN NEAR-FIELD TO FAR-FIELD TRANSFORMATION

In this section, we first recall the equations expressing the NF to FF transformation based on the Green's formulation. The detailed developments can be found in [10]. Then, we present how to calculate rigorously the E -field time derivative using the reconstruction formula. Finally, we present the analytic expressions of infinitesimal dipole radiated field in the TD when it is excited by modulated Gaussian pulse. This infinitesimal dipole is used to generate the TD radiation pattern of our AUT composed of 40 infinitesimal dipoles.

2.1. Time-Domain Near-Field to Far-Field Formalism

In this paper we use the time-domain near-field to far-field transformation algorithm based on the Green's function representation [10]. The antenna NF tangential components E_y and E_z are collected over the plane $x = x_{meas}$. Equation (1) expresses the antenna far-field $\vec{F}(\theta, \phi, t)$ in the direction (θ, ϕ) (spherical coordinates associated with the Cartesian coordinates) as a function of the time derivative of

the tangential components of the measured NF E_{meas} .

$$\vec{F}(\theta, \phi, t) = -\frac{1}{2\pi c} \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \vec{e}_r \times \left[\vec{e}_x \times \frac{\partial \vec{E}_{meas}}{\partial t} (\vec{r}_0, t + \vec{e}_r \cdot \vec{r}_0/c) \right] dy_0 dz_0 \quad (1)$$

with $\vec{F}(\theta, \phi, t) = \lim_{r \rightarrow \infty} r \vec{E}(\vec{r}, t + r/c)$

The measurement surface is defined by $-D_{max} \leq y, z \leq D_{max}$, and (1) is expressed in a truncated form as

$$\vec{F}(\theta, \phi, t) = -\frac{1}{2\pi c} \int_{-D_{max}}^{+D_{max}} \int_{-D_{max}}^{+D_{max}} \vec{e}_r \times \left[\vec{e}_x \times \frac{\partial \vec{E}_{meas}}{\partial t} (\vec{r}_0, t + \vec{e}_r \cdot \vec{r}_0/c) \right] dy_0 dz_0 \quad (2)$$

The tangential components of the measured field vector \vec{E}_{meas} are collected at a regular grid described by $\vec{r}_0 = x_{meas} \vec{e}_x + y \vec{e}_y + z \vec{e}_z$ with $-D_{max} \leq y, z \leq +D_{max}$. Using the sampling theorem [10], we transform (2) to the following double summation

$$\vec{F}(\theta, \phi, t) = -\frac{1}{2\pi c} \sum_{n=-N_y}^{N_y} \sum_{m=-N_z}^{N_z} \vec{e}_r \times \left[\vec{e}_x \times \frac{\partial \vec{E}_{meas}}{\partial t} (\vec{r}_{0mn}, t + \vec{e}_r \cdot \vec{r}_{0mn}/c) \right] \Delta y_0 \Delta z_0 \quad (3)$$

where $\vec{r}_{0mn} = x_{meas} \vec{e}_x + m \Delta y \vec{e}_y + n \Delta z \vec{e}_z$, $\vec{e}_r = \cos(\phi) \sin(\theta) \vec{e}_x + \sin(\phi) \sin(\theta) \vec{e}_y + \cos(\theta) \vec{e}_z$.

We have developed a Matlab routine expressing (3) to calculate the FF using NF data calculated at several measurement distances $x = x_{meas}$ with different sampling criterion $\Delta y, \Delta z$. In (3), the first step in NF to FF transformation is the calculation of the time derivative of the measured NF at the instant $\frac{t + \vec{e}_r \cdot \vec{r}_{0mn}}{c}$. The accuracy of the result depends on the chosen interpolation method. In our analysis, we will use the reconstruction formula to determine the time derivative of the measured field and to interpolate the field at the required instants.

2.2. The Reconstruction Formula

Let us define a function $f(t)$ where f_ω is its Fourier transform. $f(t)$ is a band limited function if for $\omega \geq \omega_{max}$

$$f_\omega = -\frac{1}{2\pi} \int_{-\infty}^{+\infty} f(t) \exp(j\omega t) dt = 0 \quad (4)$$

The reconstruction formula allows the exact interpolation of the band limited function $f(t)$ using the discrete points $f(t_0 + l\Delta t)$. $f(t)$ is expressed for every $t \in [t_0, t_0 + (N_t - 1)\Delta t]$ using the cardinal series as

$$f(t) = \sum_{l=0}^{N_t-1} \text{sinc} \left(\pi \left(\frac{t - (t_0 + l\Delta t)}{\Delta t} \right) \right) f(t_0 + l\Delta t) \quad (5)$$

Using the discrete measurement samples $\vec{E}_{meas}(t_0 + l\Delta t)_{1 \leq l \leq N_t-1}$, one can express accurately $\vec{E}_{meas}(t)$ for every $t \in [t_0, t_0 + (N_t - 1)\Delta t]$. To do this, the Nyquist sampling criterion $\Delta t = \pi/\omega_{max}$ has to be respected. The reconstruction formula is used to interpolate the measured E -field as:

$$\vec{E}_{meas}(\vec{r}_{0mn}, t) = \sum_{l=0}^{N_t-1} h(t) \vec{E}_{meas}(\vec{r}_{0mn}, t_0 + l\Delta t) \quad \text{with} \quad h(t) = \frac{\sin \left(\pi \frac{t - (t_0 + l\Delta t)}{\Delta t} \right)}{\pi \frac{t - (t_0 + l\Delta t)}{\Delta t}} \quad (6)$$

For $t_0 \leq t \leq t_0 + (N_t - 1)\Delta t$, the time derivative of the E -field is expressed rigorously as

$$\frac{\partial \vec{E}_{meas}(\vec{r}_{0mn}, t)}{\partial t} = \sum_{l=0}^{N_t-1} \frac{\partial h(t)}{\partial t} \vec{E}_{meas}(\vec{r}_{0mn}, t_0 + l\Delta t) \quad (7)$$

where $\frac{\partial h(t)}{\partial t} = \frac{1}{t - (t_0 + l\Delta t)} (\cos(\pi \frac{t - t_0}{\Delta t} - l) - h(t))$.

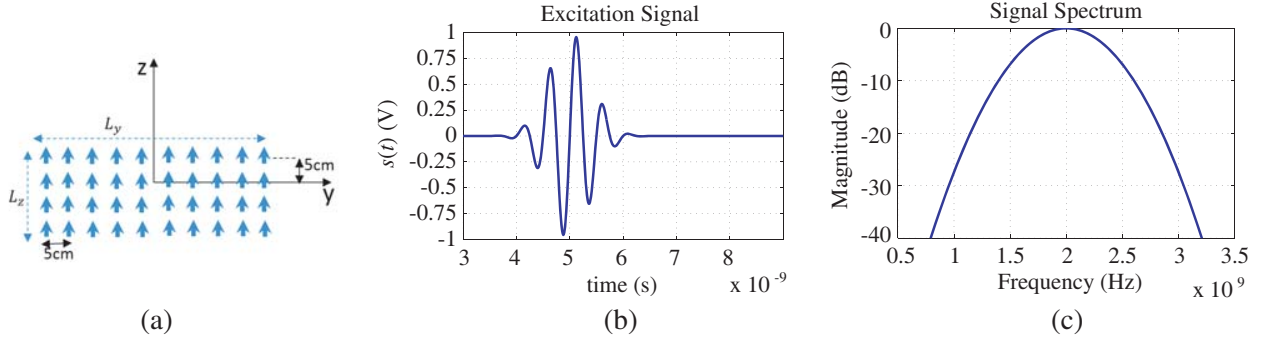


Figure 1. (b) The time-domain excitation signal applied to (a) each of the 40 dipoles composing the AUT and (c) the magnitude of the excitation signal spectrum. The maximum of the spectrum is at 2 GHz.

If the sampling Nyquist criterion is respected, (6) and (7) interpolate the E -field and its time derivative. Here, we consider the antenna under test (AUT) composed of infinitesimal dipoles distributed over the plane $x = 0$ as shown in Fig. 1. Each dipole is excited by a pulse and the TD NF is collected over the plane $x = x_{meas}$ at regularly spaced positions. In the next paragraph, we express the analytical expression of the field radiated by an elemental dipole.

2.3. Transient Radiation of an Infinitesimal Dipole

Let us consider a z -polarized elemental dipole placed at the origin of the coordinate system. This dipole is excited by $e(t)$, a sinusoidal current modulated by a Gaussian pulse written as

$$e(t) = e_0 \sin(\beta t) \exp(-\alpha t^2) \quad \text{with} \quad \alpha = 2/\tau^2 \quad (8)$$

The transient radiated fields E_r and E_θ are written in the spherical coordinates system

$$\begin{aligned} E_r(r, \theta, \phi, t) &= \frac{2\eta}{4\pi r^2} \cos(\theta) \left[\frac{\partial e(t-r/c)}{\partial t} + \frac{c}{r} e(t-r/c) \right] \\ E_\theta(r, \theta, \phi, t) &= \frac{\eta}{4\pi r} \sin(\theta) \left[\frac{\partial^2 e(t-r/c)}{c \partial t^2} + \frac{\partial e(t-r/c)}{r \partial t} + \frac{e(t-r/c)}{r^2} \right] \end{aligned} \quad (9)$$

with

$$\begin{aligned} \frac{\partial e(t)}{\partial t} &= e_0 [\beta \cos(\beta t) - 2\alpha t \sin(\beta t)] \exp(-\alpha t^2) \\ \frac{\partial^2 e(t)}{\partial t^2} &= e_0 [(4\alpha^2 t^2 - \beta^2 - 2\alpha) \sin(\beta t) - 4\alpha \beta t \cos(\beta t)] \exp(-\alpha t^2) \end{aligned} \quad (10)$$

We have used (8)–(10) to generate the tangential components E_z and E_y radiated by 40 elementary dipoles placed at the plane $x = 0$. The distance between dipoles is 5 cm, $\beta = 4\pi 10^9$ rad/s and $\tau = 0.8 \cdot 10^{-9}$ s. The dipole distribution is described in Fig. 1. The measurement surface is limited by D_{max} as shown in Fig. 2 and the TD radiation pattern in the cut planes $y_{meas} = 0$ and $z_{meas} = 0$ are presented in Fig. 3.

3. NUMERICAL RESULTS

In this section we are interested in evaluating the accuracy of the time-domain (TD) FF as a function of the spatial sampling criterion Δy and Δz . The antenna under test (AUT) is composed of 40 elementary dipoles excited by a sinusoidal current modulated by a Gaussian pulse. The NF measured at the distance $x_{meas} = 50$ cm $= 5\lambda_{3\text{GHz}}$ ($x_{meas} = 5\lambda_{1\text{GHz}/3}$) is presented in the TD in Fig. 3 and in the FD in Fig. 4.

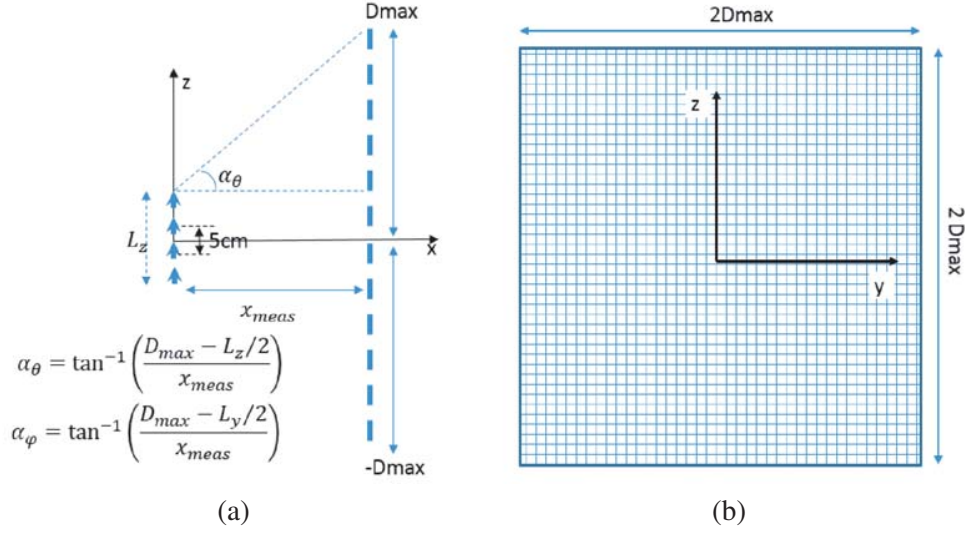


Figure 2. Near-field measurement grid at the distance x_{meas} from the AUT. The tangential components E_y and E_z are regularly recorded over a square surface in the zy plane.

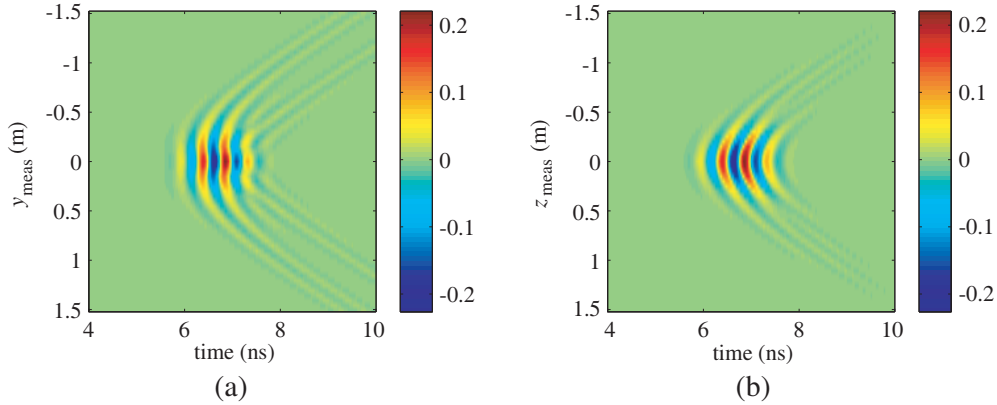


Figure 3. The E_z (V/m) near-field radiated by the AUT (40 dipoles) as a function of the time. (a) The E_z (V/m) at the plane cut $z_{meas} = 0$. (b) The E_z (V/m) at the plane cut $y_{meas} = 0$.

3.1. The Near-Field Measurement Surface Truncation and the Sampling Criterion

3.1.1. The Time-Domain Far-Field Results Comparison

In the planar rectangular frequency-domain measurement technique, the NF is spatially sampled using the Nyquist rule $\Delta y = \Delta z = \lambda/2$ with λ being the working wavelength. As presented in [10] the TD NF is spatially sampled based on the AUT excitation signal maximum frequency $\Delta y = \Delta z = \lambda_{min}/2 = c\pi/\omega_{max}$. For a short pulse (UWB antenna), the maximum frequency can be F_{-3dB} , F_{-10dB} , F_{-20dB} , F_{-30dB} . F_{-XdB} is the frequency for which the magnitude of the excitation pulse spectrum is lower by X value compared with the normalized maximum level (0 dB) of the central frequency. For our case of study, as it is seen in Fig. 1, $F_{-3dB} \approx 2.3$ GHz, $F_{-6dB} \approx 2.4$ GHz, $F_{-10dB} \approx 2.6$ GHz, $F_{-20dB} \approx 2.8$ GHz and $F_{-30dB} \approx 3$ GHz. The maximum magnitude of the excitation signal spectrum corresponds to the central frequency $F_{0dB} = f_0 = 2$ GHz. We propose to express $F_{-3dB} \approx 1.15f_0$, $F_{-6dB} \approx 1.2f_0$, $F_{-10dB} \approx 1.3f_0$, $F_{-20dB} \approx 1.4f_0$, $F_{-30dB} = 1.5f_0$. In order to study the effect of NF spatial sampling criterion over the calculated FF, we use χ to parameterize the spatial sampling $\Delta y_\chi = \Delta z_\chi = \lambda_0/2\chi$, where $\chi = 1.1, 1.2, 1.3, 1.4, 1.5$ and $\lambda_0 = c/f_0$. By comparing the

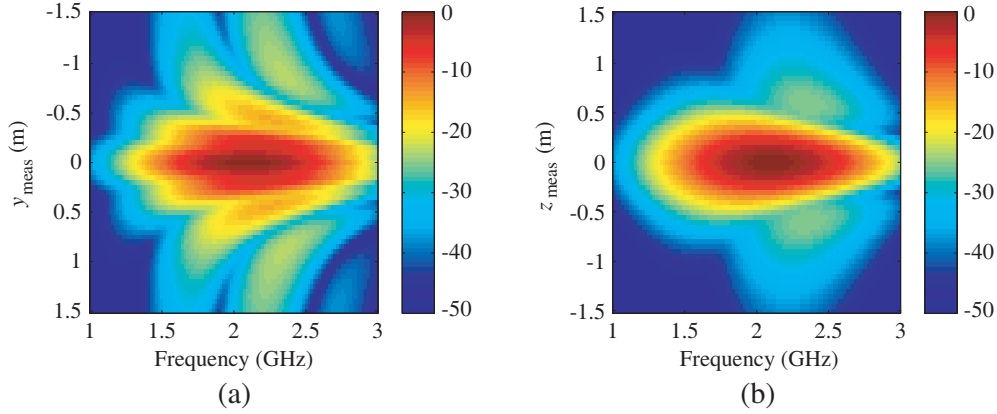


Figure 4. The normalized E_z (dB V/m) NF measured at ($x_{meas} = 50$ cm) radiated by the AUT (40 dipoles) as a function of the frequency. (a) The E_z (dB(V/m)) at the plane cut $z_{meas} = 0$. (b) The E_z (dB(V/m)) at the plane cut $y_{meas} = 0$.

far-field calculated using different χ , we identify the influence of the chosen maximum frequency on the accuracy of the calculated AUT FF. The NF time-axis is sampled using $\Delta t = \frac{1}{3F_{-30\text{dB}}}$ to prevent the time-domain aliasing error.

Over the planar surface at the distance $x_{meas} = 50$ cm from the AUT, we calculate the near-field at regularly spaced points with $\Delta y_\chi = \Delta z_\chi = \lambda_0/2\chi$ and $D_{\max} = 150$ cm ($10\lambda_0$). Then, using (3) and (7), we calculate the FF at the directions ($\theta = \pi/2, \phi = 0$), ($\theta = \pi/3, \phi = 0$), and ($\theta = \pi/2, \phi = \pi/6$) for $\chi = 1.1, 1.2, 1.3, 1.4, 1.5$. The E_z component of the FF is presented in Figs. 5(a)–(c)–(e) and compared with the reference FF determined directly from (9) and (10) for r going to infinity. This actual far-field is labeled Ref in Fig. 5. The differences between the calculated FFs and the actual one are quantified using the error expressed in (11). $E_{\text{NFtoFF}}(\theta, \phi, t)$ is the field calculated using NF to FF transformation and $E_{\text{Ref}}(\theta, \phi, t)$ is the actual far-field.

$$\text{Error}(\theta, \phi, t) = 100 \times \frac{E_{\text{NFtoFF}}(\theta, \phi, t) - E_{\text{Ref}}(\theta, \phi, t)}{\max(E_{\text{Ref}}(\theta, \phi, t))_{t_0 \leq t \leq t_0 + (N_t - 1)\Delta t}} \quad (11)$$

The errors of the curves in Fig. 5(a) are presented in Fig. 5(b) for different χ . The errors of the comparisons presented in Fig. 5(c) are plotted in Fig. 5(d). Fig. 5(f) presents the errors of Fig. 5(e). From Figs. 5(b)–(d)–(f) the early-time far-field introduced in [10] as being the part of the radiated FF without the truncation error is *visually* observed in Fig. 5(e). The measurement surface truncation effect is visible at $t = 7$ ns and $10 \text{ ns} \leq t \leq 13$ ns.

As explained in [10], the effect of the truncation error arrives chronologically after the actual FF signal. This is presented in [10] for a point source. Nevertheless, in Figs. 5(b)–(d)–(f), we notice quantitative differences between the calculated far-fields (FFs) and the actual one even in the time portion of the radiation pattern attributed to early-time far-field ($4 \text{ ns} \leq t \leq 6$ ns).

In order to understand the origin of the errors for $4 \text{ ns} \leq t \leq 6$ ns, we have used a larger measurement surface to calculate the FF ($D_{\max} = 300$ cm) with the same sampling conditions. The obtained results have shown the same error level for $4 \text{ ns} \leq t \leq 6$ ns. These errors are not attributed to the size of the measurement surface. However, we have applied the same analysis (same excitation pulse, same sampling conditions, same measurement distance) for a smaller antenna (4×4) composed of 16 dipoles with 5 cm between dipoles and $D_{\max} = 150$ cm, the error level is smaller for $4 \text{ ns} \leq t \leq 6$ ns. Moreover, this decreasing tendency is confirmed for 4 dipoles (2×2). The errors for $4 \text{ ns} \leq t \leq 6$ ns is directly linked to the size of the AUT.

As presented in Figs. 5(b)–(d)–(f), the error levels depend on the FF observation point (θ, ϕ). The angular FF directions free from truncation error depends on three parameters, namely the measurement distance ($x_{meas} = 50$ cm), the AUT size ($L_y = 45$ cm and $L_z = 15$ cm) and the measurement surface dimensions ($D_{\max} = 150$ cm). Here, the critical angle is defined by $\alpha_\theta = 70$ deg and $\alpha_\phi = 68$ deg as

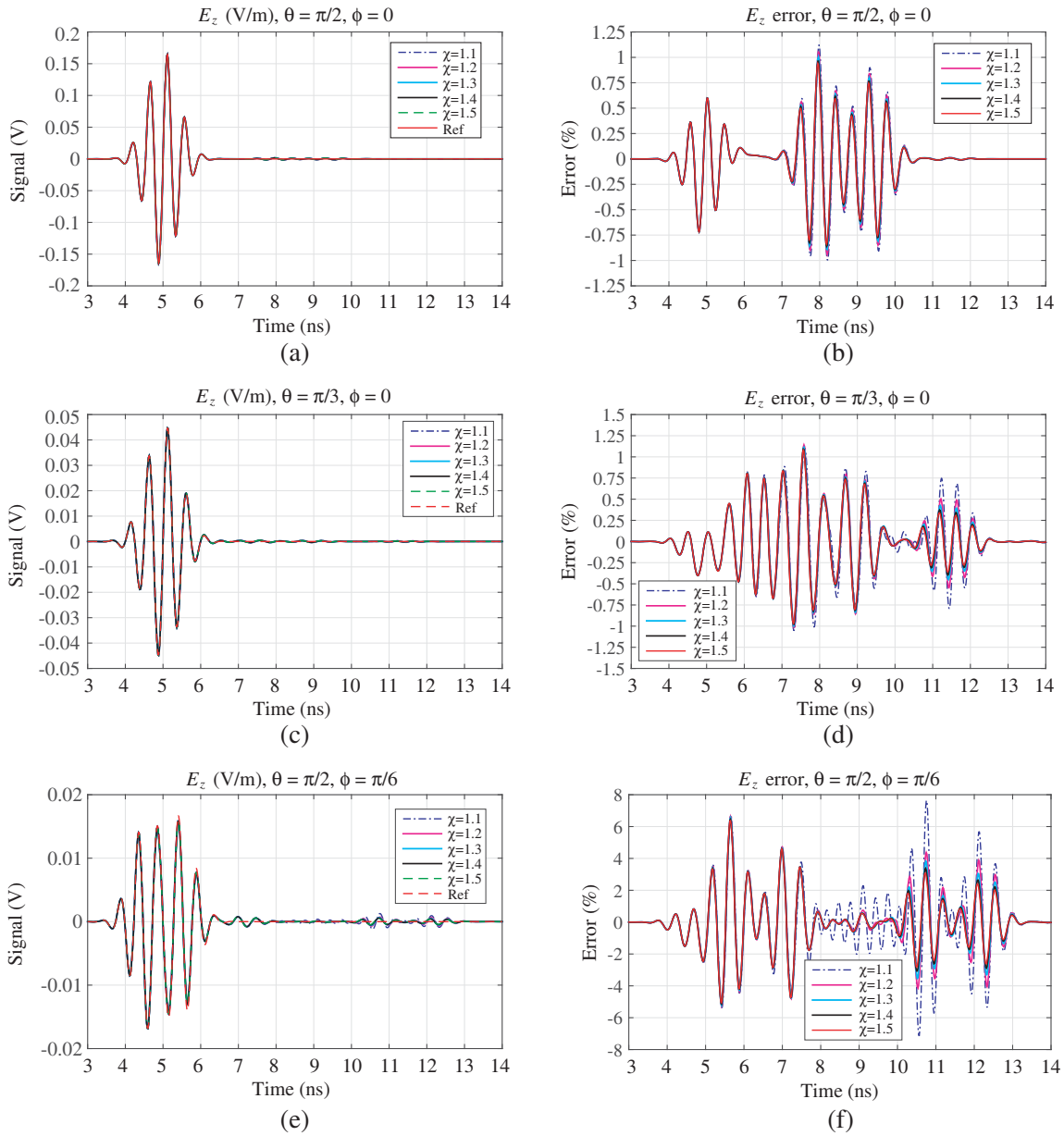


Figure 5. Comparison of the NF to FF transformation results for different sampling criteria $\Delta y = \Delta z = \lambda_0/2\chi$, $D_{\max} = 150$ cm and $x_{meas} = 50$ cm. The TD FF E_z component at (a) $\theta = \pi/2$ and $\phi = 0$, (c) $\theta = \pi/3$ and $\phi = 0$ (e) $\theta = \pi/2$ and $\phi = \pi/6$. The errors of Figs. 5(a)–(c)–(e) are respectively (b) $\theta = \pi/2$ and $\phi = 0$, (d) $\theta = \pi/3$ and $\phi = 0$ (f) $\theta = \pi/2$ and $\phi = \pi/6$.

expressed in Fig. 2. This allows an angular FF area free from truncation errors for $30 \text{ deg} \leq \theta \leq 150 \text{ deg}$ and $-68 \text{ deg} \leq \phi \leq 68 \text{ deg}$. This provides an indication and not the exact angular area, especially when the measured field level in the edge of the measurement surface is of comparable levels with the maximum measured field.

In Fig. 6 we show the FF errors calculated at the principal plane cuts $\theta = \pi/2$ with $-\pi/2 \leq \phi \leq \pi/2$ and $\phi = 0$ with $0 \leq \theta \leq \pi$. The presented errors are for $\chi = 1.2, 1.3$ and 1.5. It is seen from Figs. 6(c) and (f) that the NF sampled using $\chi = 1.2$ is under-sampled since the observable oscillations are due to the aliasing errors. In Figs. 6(a)–(b)–(d)–(e) we have less oscillations. The errors behavior do not

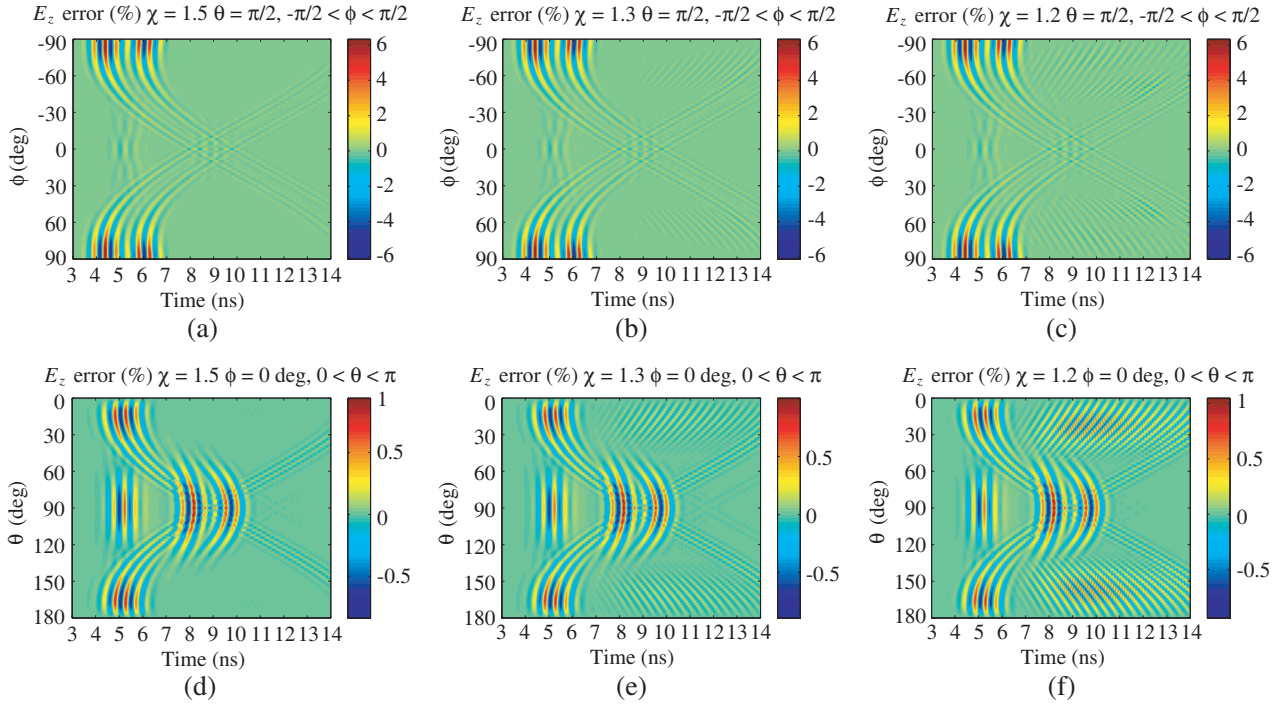


Figure 6. The FF E_z component error as a function of time for the plane cut $\theta = \pi/2$. (a) For $\chi = 1.5$, (b) for $\chi = 1.3$, (c) for $\chi = 1.2$. The E_z error as a function of time for the plane cut $\phi = 0$, (d) for $\chi = 1.5$, (e) for $\chi = 1.3$, (f) for $\chi = 1.2$.

reveal early-time far-field region where the truncation error happens later in time compared with the actual FF. The errors level for $4 \text{ ns} \leq t \leq 6 \text{ ns}$ (actual far-field) is of comparable level with the errors for $7 \text{ ns} \leq t \leq 10 \text{ ns}$. The errors for $0 \leq \theta \leq 30 \text{ deg}$, $150 \text{ deg} \leq \theta \leq 180 \text{ deg}$, $-90 \text{ deg} \leq \phi \leq -68 \text{ deg}$, and $68 \text{ deg} \leq \phi \leq 90 \text{ deg}$ are due to the truncation of the measurement surface as predicted by $\alpha_\theta = 70 \text{ deg}$ and $\alpha_\phi = 68 \text{ deg}$. In addition, the errors for $\chi = 1.5$ and $\chi = 1.3$ have comparable behavior as it is seen from Figs. 6(a)–(b)–(d)–(e). Consequently, comparing the TD FF results for this AUT, the weighting factor $\chi = 1.3$ ($F_{-10 \text{ dB}} \approx 1.3f_0$) allows a comparable accuracy as $\chi = 1.5$ ($F_{-30 \text{ dB}} \approx 1.5f_0$).

3.1.2. The Frequency-Domain Far-Field Results Comparison

Using the NF sampling criterion $\Delta y_\chi = \Delta z_\chi = \lambda_0/2\chi$, $x_{meas} = 50 \text{ cm}$ and $D_{max} = 150 \text{ cm}$ we compare the calculated far-fields in the frequency-domain (FD). These FFs are set by Fourier transforming the previously calculated TD far-fields. The comparisons are carried out at the principal plane cuts $\theta = \pi/2$ and $\phi = 0$ at $F_{-10 \text{ dB}}$ (1.4 GHz and 2.6 GHz), $F_{-20 \text{ dB}}$ (1.2 GHz and 2.8 GHz), $F_{-30 \text{ dB}}$ (1 GHz and 3 GHz) for different χ .

It is seen in Fig. 4 that the AUT NF truncation level depends on the frequency and the field truncation is more visible in the plane cut $z_{meas} = 0$ than the plane cut $y_{meas} = 0$. Hence, the calculated far-field (from NF to FF transformation) is damaged by the truncation error in the cut plane $\theta = \pi/2$ more than the cut plane $\phi = 0$. Let us recall that the angular area of the free error FF is for $30 \text{ deg} \leq \theta \leq 150 \text{ deg}$ and $-68 \text{ deg} \leq \phi \leq 68 \text{ deg}$.

In Fig. 7, we present the FD FFs comparison at the plane cut $\phi = 0$. The FF for high and low frequencies of $F_{-X \text{ dB}}$ fits well the actual far-field when the sampling criterion reaches the Nyquist limit ($\lambda/2$) for $30 \text{ deg} \leq \theta \leq 150 \text{ deg}$. Also, the FF results for $30 \text{ deg} \leq \theta \leq 150 \text{ deg}$ are identical using $\chi = 1.3$ and $\chi = 1.5$ even for the frequency 2.8 GHz. In Fig. 8, we compare the calculated FFs with the actual one in the plane cut $\theta = \pi/2$. As shown, the results are more sensitive to the truncation error compared with the results of Fig. 7. The FFs for $\chi = 1.3$, 1.4 and 1.5 are similar for $-40 \text{ deg} \leq \phi \leq 40 \text{ deg}$.

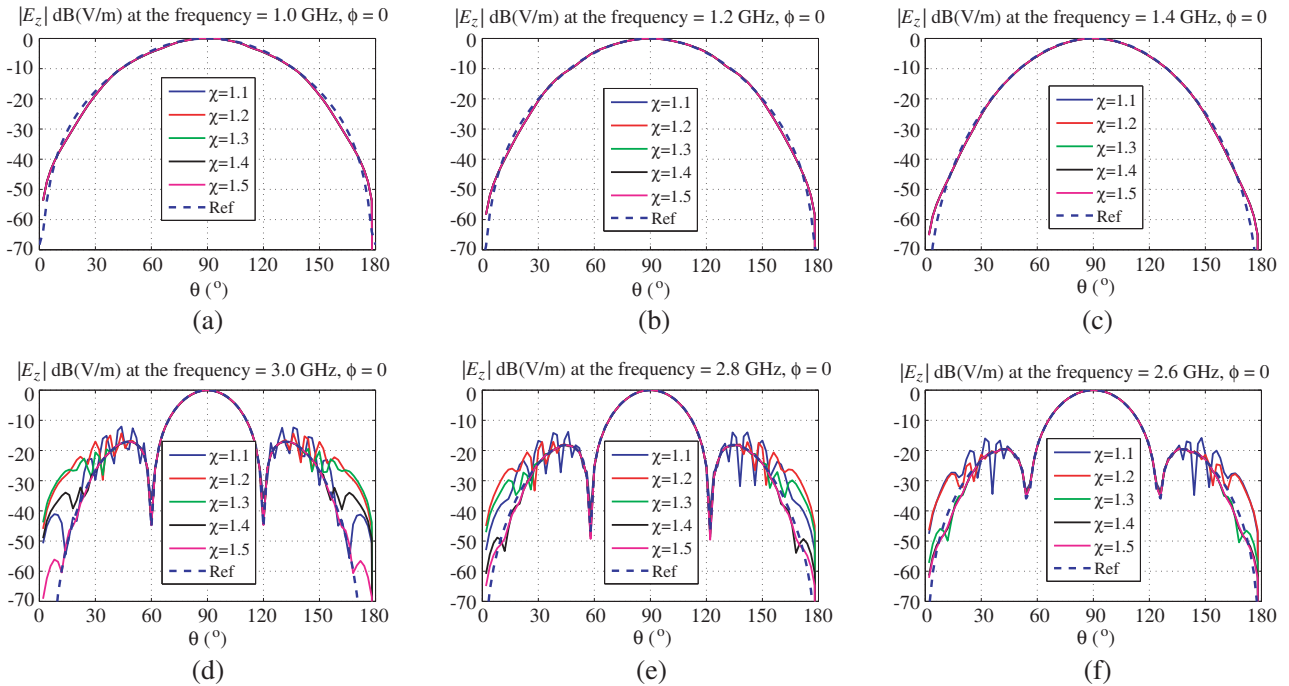


Figure 7. Comparison of far-field E_z -component at the plane cut $\phi = 0$ with $D_{\max} = 150$ cm, $x_{\text{meas}} = 50$ cm and $\Delta y = \Delta z = \lambda_0/2\chi$ for $\chi = 1.1, 1.2, 1.3, 1.4, 1.5$. (a)–(d) For 1 GHz and 3 GHz ($F_{-30\text{dB}}$) respectively. (b)–(e) For 1.2 GHz and 2.8 GHz ($F_{-20\text{dB}}$) respectively. (c)–(f) For 1.4 GHz and 2.6 GHz ($F_{-10\text{dB}}$) respectively.

The measured NF amplitude at the edge of the measurement surface is only -22.6 dB lower than the maximum NF amplitude in the plane cut $z_{\text{meas}} = 0$. Consequently, the FF for $-68 \text{ deg} \leq \phi \leq -40$ and $40 \text{ deg} \leq \phi \leq 68$ presents important errors.

When the truncation level of the AUT NF measurement data changes significantly as a function of the frequency, the measurement surface has to be adapted to satisfy the same truncation level for every frequency. Otherwise, if the measurement surface stays unchanged, the AUT valuable measurement frequency band is chosen as a function of the truncation level of the measured field. In particular, when measuring the NF of the AUT in the time-domain, we can primarily verify by measuring in the principal cut planes $y_{\text{meas}} = 0$ and $z_{\text{meas}} = 0$ the level of the E -field truncation level for every measured frequency by Fourier transforming the time-domain NF. Then, we calculate the NF truncation level by determining the level difference between the maximum E -field and the field level at the measurement surface limit for each frequency. The AUT presents at the cut plane $z_{\text{meas}} = 0$ the truncation levels of (22.58 dB, 26,18 dB), (33.84 dB, 24.40 dB), (28.78 dB, 33.35 dB) and (27 dB, 26.45 dB) for $F_{-30\text{dB}}$ (1 GHz and 3 GHz), $F_{-20\text{dB}}$ (1.2 GHz and 2.8 GHz), $F_{-10\text{dB}}$ (1.4 GHz and 2.6 GHz) and $F_{-3\text{dB}}$ (1.7 GHz and 2.3 GHz), respectively as presented in Fig. 9.

The AUT NF measurement data guarantee truncation levels less than -30 dB in the plane cut $y_{\text{meas}} = 0$. As a consequence, the far-field is accurately calculated at the plane cut $\phi = 0$ for $30 \text{ deg} \leq \theta \leq 150 \text{ deg}$. The NF plane cut $y_{\text{meas}} = 0$ presented in Fig. 9(b) shows truncation levels of -22.5 dB for 1 GHz ($\chi = 1.5$) and 24.4 dB for 2.8 GHz ($\chi = 1.4$) which are responsible of the FF discrepancies in frequency and time domains. From Fig. 9(b), the frequency 2.6 GHz ($\chi = 1.3$) is the highest frequency allowing the lowest truncation error level. For this reason the errors presented in the time-domain in Figs. 5–6–7 for $\chi = 1.3$ were equivalent to the error associated with $\chi = 1.5$. For this AUT and this measurement surface ($D_{\max} = 150$ cm), the maximum frequency to consider for the time-domain NF sampling criterion is 2.6 GHz ($\chi = 1.3$). The FD FF results confirm these conclusions.

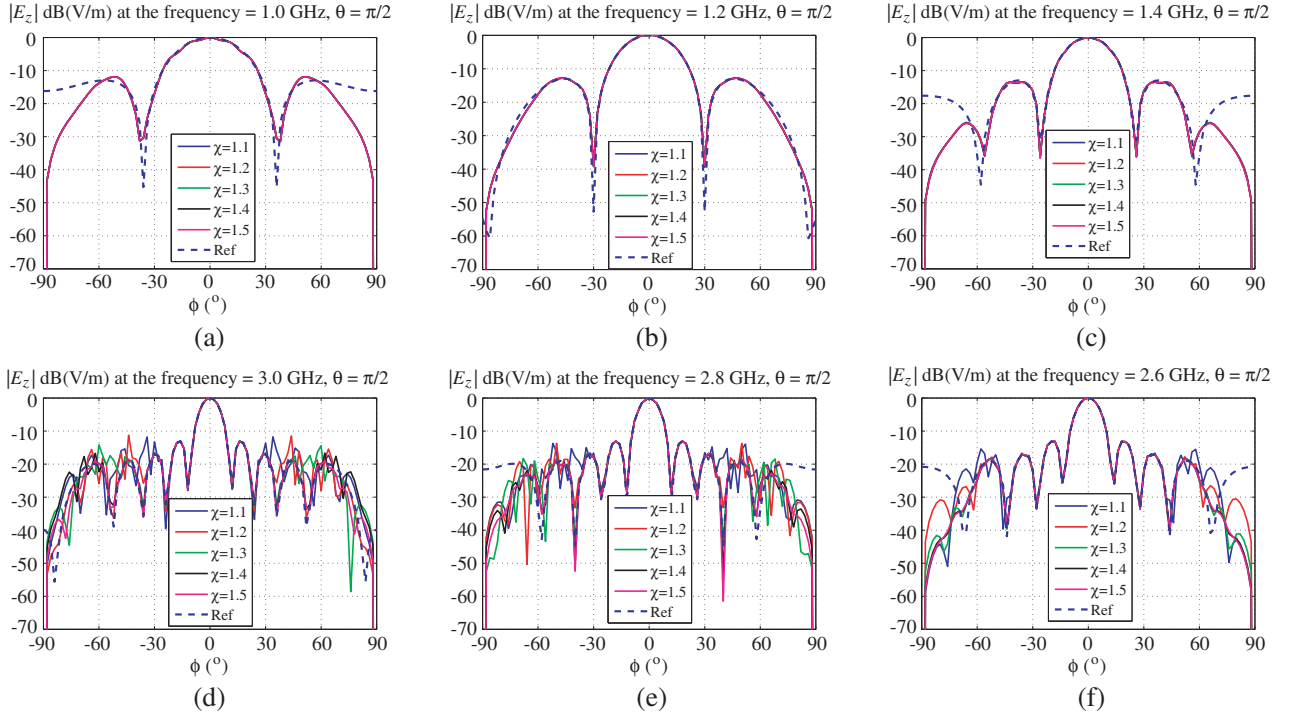


Figure 8. Comparison of far-field E_z -component at the plane cut $\theta = \pi/2$ with $D_{\max} = 150$ cm, $x_{\text{meas}} = 50$ cm and $\Delta y = \Delta z = \lambda_0/2\chi$ for $\chi = 1.1, 1.2, 1.3, 1.4, 1.5$. (a)–(d) For 1 GHz and 3 GHz ($F_{-30\text{dB}}$) respectively. (b)–(e) For 1.2 GHz and 2.8 GHz ($F_{-20\text{dB}}$) respectively. (c)–(f) For 1.4 GHz and 2.6 GHz ($F_{-10\text{dB}}$) respectively.

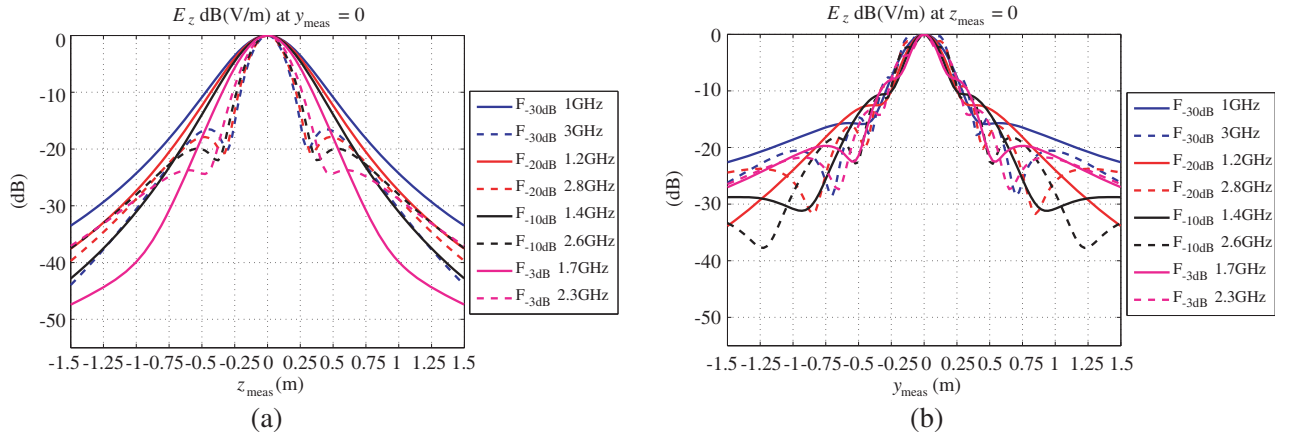


Figure 9. The normalized E_z NF component for $F_{-30\text{dB}}$ (3 GHz, 1 GHz), $F_{-20\text{dB}}$ (2.8 GHz, 1.2 GHz), $F_{-10\text{dB}}$ (2.6 GHz, 1.4 GHz), $F_{-3\text{dB}}$ (2.3 GHz, 1.7 GHz). (a) At the plane cut $y_{\text{meas}} = 0$. (b) At the plane cut $z_{\text{meas}} = 0$.

3.2. The Near-Field Measurement Distance and the Sampling Criterion

3.2.1. The Time-Domain Far-Field Results Comparison

Let us consider the NF measurement surface with $D_{\max} = 150$ cm $= 10\lambda_0$. We are interested in studying the effect of the NF measurement distances x_{meas} where the TD NF data are collected with $\Delta y_\chi = \Delta z_\chi = \lambda_0/3$ ($\chi = 1.5$). The actual TD FF is compared with the ones calculated using the TD NF

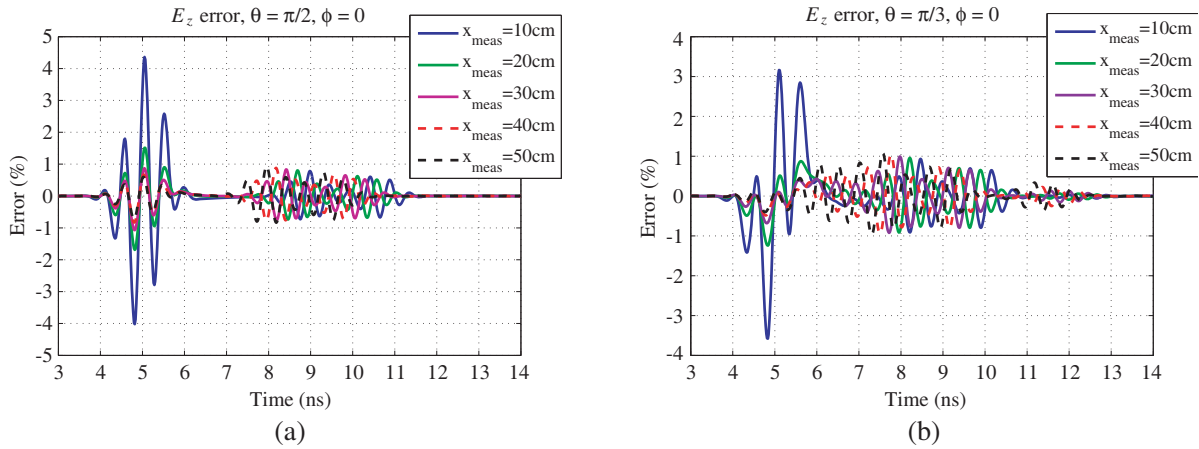


Figure 10. The E_z error as a function of time and as a function of the NF measurement distance (x_{meas}) for the FF observation points (a) $\theta = \pi/2$ and $\phi = 0$, (b) $\theta = \pi/3$ and $\phi = 0$.

to FF transformation for different measurement distances $x_{meas} = 10$ cm ($2\lambda_0/3$), 20 cm ($4\lambda_0/3$), 30 cm ($2\lambda_0$), 40 cm ($8\lambda_0/3$) and 50 cm ($10\lambda_0/3$). The errors resulting from these comparisons are presented in Figs. 10(a) and (b) for the directions ($\theta = \pi/2$, $\phi = 0$) and ($\theta = \pi/3$, $\phi = 0$) respectively. As it is seen in Figs. 10(a)–(b), the error depends on the measurement distance, especially for 4 ns $\leq t \leq 6$ ns. The error values decrease as the measurement distance increases.

Namely, it is difficult to set the limit of the radiating NF zone (Fresnel zone) in the TD NF measurement. The Fresnel zone is defined based on the AUT size and the operating frequency. In TD NF measurement we deal with a frequency band and the NF data measured below the Fresnel zone contain evanescent modes that are probably responsible of the errors for 4 ns $\leq t \leq 6$ ns. Considering smaller antennas, the Fresnel zone is rapidly reached and the errors for 4 ns $\leq t \leq 6$ ns are smaller because less evanescent NF modes are taken into account. In [10] authors have considered a point source and the Fresnel zone is rapidly reached (low evanescent mode) for that case. This explains the early-time far-field free from truncation error as presented in [10].

In addition, from Fig. 10(a), we can isolate the effect of the measurement distance error (4 ns $\leq t \leq 6$ ns) since the truncation error happens later in time and the same error level is observed for every measurement distance for 7 ns $\leq t \leq 11$ ns. If the measurement distance is larger than $2\lambda_0$ (30 cm), the error values for $x_{meas} = 2\lambda_0$ are equivalent to the error values for $x_{meas} = 50$ cm. In contrast with Fig. 10(b), it is difficult to distinguish between the measurement distance error and the truncation error.

3.2.2. The Frequency-Domain Far-Field Results Comparison

We are interested in studying in frequency-domain the effect of the measurement distance x_{meas} . The FF results presented in Fig. 11 are determined by Fourier transforming the time-domain results of the NF to FF transformation of NF data collected at $x_{meas} = 10$ cm, 20 cm, 30 cm, 40 cm and 50 cm with the sampling criterion $\Delta y_\chi = \Delta y_\chi = \lambda_0/3$ ($\chi = 1.5$). The comparisons presented in Fig. 10 show that TD FF errors decrease as the measurement distance increases. The time-domain errors are helpless to identify which frequency is mostly sensitive to the measurement distance. In contrast, the FD comparisons presented in Figs. 11(a)–(d) show that the frequency (1 GHz) is sensitive to the measurement distance with a maximum difference of 1.47 dB at $\phi = 0$ and 1.45 dB at $\theta = \pi/2$. The frequency 1.2 GHz presented in Figs. 11(b)–(e) shows a maximum difference of 0.4 dB. The other frequencies (greater than 1.4 GHz) stay unchanged (error ≤ 0.2 dB) since the NF measurement distance is greater than a wavelength.

As a conclusion, using $D_{max} = 150$ cm and $x_{meas} \geq 2\lambda_0$ the error values are of the same order for $\chi = 1.3$ and $\chi = 1.5$. Hence, $F_{-10\text{dB}} \approx 1.3f_0$ can be considered as the maximum frequency for NF sampling of this AUT. We have tried to isolate the effect of three important measurement parameters in the planar TD NF technique. These parameters are: the measurement distance (reactive NF), the surface truncation (truncation error) and the NF sampling criterion (aliasing error). As it is presented

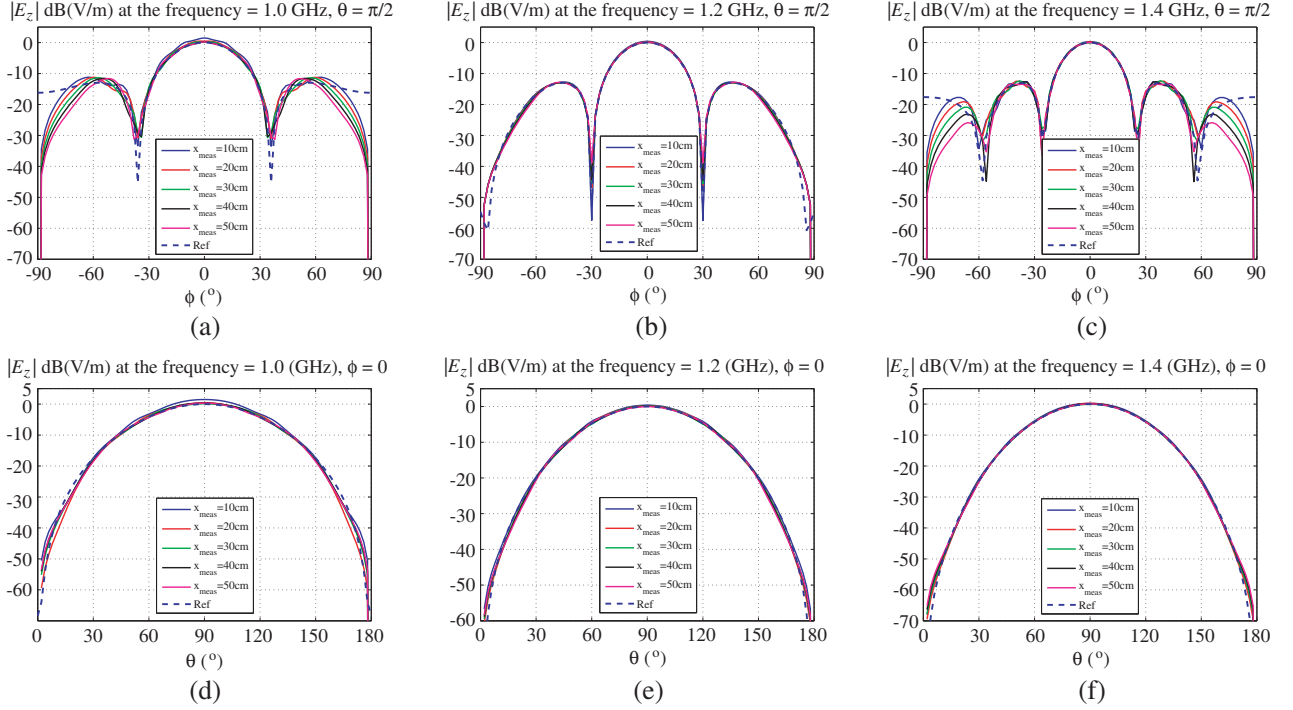


Figure 11. Comparison of the FF radiation pattern for $x_{meas} = 10$ cm, 20 cm, 30 cm, 40 cm and 50 cm at the plane cut $\theta = \pi/2$, (a) at 1 GHz, (b) at 1.2 GHz, (c) at 1.4 GHz. Comparison of the FF radiation pattern at the plane cut $\phi = 0$, (d) at 1 GHz, (e) at 1.2 GHz, (f) at 1.4 GHz.

in Figs. 5–6–7 it is difficult to predict from TD FF errors the consequences over the AUT radiation pattern in the frequency-domain. For this reason, the frequency-domain (FD) comparisons have been presented to make out the effect of each parameter over the AUT FD FF.

4. CONCLUSION

The effect of three parameters in planar time-domain near-field to far-field transformation have been presented. The followed approach aims to optimize the computation time and memory requirements by studying the near-field sampling measurement criterion. For antennas characterization using the time-domain near-field technique, we have shown that multiple conditions have to be met to correctly calculate the far-field. The size of the measurement surface decides predominantly on the frequency band to consider. The NF spatial truncation is responsible for the frequency limitation. The maximum frequency taken into account depends on the antenna excitation pulse and the behaviour of the AUT near-field *directivity* as a function of the frequency. Once the maximum frequency is defined, the sampling criterion is based on the Nyquist rate. Specific care has to be applied in choosing the measurement distance which determine the minimum frequency to be considered. Comparisons in TD and FD have been carried out to confirm these assumptions.

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3.5 Caractérisation d'antennes entre ouverture et collaborations

L'écosystème de CentraleSupélec et l'infrastructure expérimentale dont dispose le laboratoire ont grandement contribué au développement d'échanges académiques avec le laboratoire SONDRRA. Les chercheurs du laboratoire franco-singapourien SONDRRA (Supélec, ONERA, Defense Science and Technology Agency (DSTA, Singapour), National University of Singapore) s'intéressaient à la caractérisation d'antennes ultra-large bande. Cette collaboration (2010 - 2016) avec R. Guinvarc'h (SONDRRA), Hongzhao Ray Fang (NUS), Koen Mouthaan (NUS), M. R. Balakrishnan et Israel Hinoestroza (SONDRRA) est articulée autour de la conception et l'optimisation de réseaux d'antennes spirales à double polarisation circulaire pour des applications RADAR.

Le premier exemple de travaux développés dans le cadre de cette collaboration concernait la conception et la caractérisation d'antennes spirales unidirectionnelles (avec cavité) à double polarisation ainsi que leur système d'excitation **FANG et collab. [2015]** dans la bande fréquentielle [0.6 GHz - 2.4 GHz]. Le deuxième exemple concernait la conception d'un réseau phasé large bande d'antennes spirales connectées **SERHIR et GUINVARC'H [2013]**. Le choix de ces deux exemples est arbitraire car la collaboration avec SONDRRA est un échange quasi-quotidien sur des sujets divers.

Dans **FANG et collab. [2015]**, nous avons proposé une antenne spirale à 4 bras unidirectionnelle (avec cavité) doublement polarisée (circulaire droite et gauche). Le système d'excitation de l'antenne est composé d'un déphaseur contrôlé numériquement (2 bits) qui fournit un déphasage de 90° entre les différents bras de l'antenne. Ce système assure un changement instantané de la polarisation passant de circulaire droite à circulaire gauche. Dans **SERHIR et GUINVARC'H [2013]**, notre objectif était de concevoir un réseau phasé large-bande et unidirectionnel (avec cavité) doublement polarisé. Ce réseau est construit en alternant des antennes spirales à polarisations circulaires droites et gauches. La polarisation du réseau ainsi défini, est tributaire du choix des antennes alimentées par le système d'excitation. Les spirales sont connectées entre elles ce qui permet au courant de circuler vers les structures voisines permettant ainsi un élargissement de la bande passante vers les basses fréquences.

Ma contribution a été essentiellement expérimentale en assurant l'encadrement des doctorants de SONDRRA pendant les phases de mesures. La collaboration avec SONDRRA se poursuit dans le cadre d'un nouveau partenariat (trois prochaines années) autour d'investigations numériques et expérimentales pour l'analyse de la propagation des ondes électromagnétiques dans le milieu urbain en basse fréquence (500 MHz). Pour apporter une preuve de concept, nous procédons à une mise à l'échelle (1/20) pour caractériser la propagation en présence de structures en béton autour de 10 GHz.

Avec l'University of Electronic Science and Technology of China (S. Xiao) et le L2S (D. Lesselier) nous avons développé des travaux qui ont pour finalité l'analyse électromagnétique à la fois numérique et expérimentale de la super-résolution dans des milieux structurés basée sur le retournement temporel. Ce travail s'est déroulé dans le cadre de la thèse de H. Tu (thèse soutenue en 2017) où la visée première était de montrer l'apport d'une structure périodique, de périodicité finie, dans la qualité de localisation de l'élément rayonnant par rapport à la longueur d'onde.

Les phénomènes de résonance (dans le sens de cavité complexe) de la structure périodique considérée apporteraient une meilleure résolution spatiale dans l'identification de l'élément rayonnant **TU et collab. [2016]**. La structure périodique est composée de tiges métalliques dont le diamètre est négligeable devant sa longueur. Ces dernières sont distribuées autour de trois monopôles (légèrement plus longs que les tiges) distants de quelques fractions de la longueur d'onde ($\lambda/30$).

A travers l'analyse expérimentale effectuée en chambre anéchoïque pour des monopôles accordés à la fréquence de 300 MHz, nous avons établi une corrélation entre la fréquence de résonance des tiges et notre capacité à identifier le monopôle émetteur (parmi les trois). En effet, le rayonnement de la structure globale diffère de façon significative selon le choix du monopôle excité. Cette différence est très perceptible à la fréquence 340 MHz qui correspond à la fréquence de résonance des tiges. En absence de la structure périodique nous sommes dans l'incapacité de distinguer le monopôle émetteur car ils rayonnent tous de façon similaire.

En décomposant le champ rayonné sur une sphère par la structure globale (monopôles + structure périodique) en harmoniques sphériques, nous avons constaté que le nombre de modes sphériques présents dans la description du champ rayonné est plus important que pour une structure qui comporte uniquement les trois monopôles. Cette voie est en cours d'exploration dans le cadre de la thèse de Peipei Ran (2017-2020) en collaboration avec D. Lesselier (L2S). Les travaux expérimentaux en cours abordent une autre facette du sujet. Il s'agit d'illuminer et de mesurer le champ diffracté par une structure périodique composée de tiges diélectriques de longueur finies et régulièrement espacées d'une fraction de la longueur d'onde. Nous souhaitons explorer les capacités des réseaux de neurones convolutifs à identifier la (les) tige(s) manquante(s) dans la structure à partir des données expérimentales.

Pour cela, il faut prévoir une phase d'apprentissage où différentes configurations doivent être mise en place pour alimenter l'algorithme. Les apports d'une acquisition multi-statique ou de la diversité spatiale/fréquentielle seront nécessaires pour cette expérience. La validation numérique de cette étude a été faite pour le cas 2D dans [RAN et collab. \[2019\]](#) [RAN et collab. \[2019\]](#). Pour aborder la modélisation numérique du cas 3D nous utiliserons le logiciel CST-MWS pour générer les données des champs diffractés par les différentes configurations de tiges et nous alimenterons l'algorithme de réseaux de neurones convolutifs par une représentation parcimonieuse du champ diffracté basée sur les coefficients du développement en harmoniques sphériques. Ces travaux sont en cours de développement.

La dernière collaboration que je souhaite aborder dans ce rapport englobe les travaux avec le département MIRE de EDF R&D autour de la caractérisation du réseau de distribution basse tension dans la bande fréquentielle utilisée pour la communication par courants porteurs en ligne (CPL) en bande étroite. Une première thèse a été soutenue dans le cadre de cet échange, celle de M. Ait Ou Kharraz (2014 - 2017) qui porte sur la caractérisation de la propagation du signal dans les lignes basse-tension (BT) en rapport avec le déploiement des compteurs intelligents Linky. Les travaux de cette thèse sont liés à l'Institut RISEGRID (Research Institute for Smarter Electric Grids).

La thèse de M. Ait Ou Kharraz avait pour objectif de caractériser le réseau de distribution d'électricité basse tension dans la bande fréquentielle utilisée par les CPL en bande étroite entre 9 kHz et 500 kHz. Pour cela, nous avons adopté une approche ascendante pour décrire le réseau électrique à partir de la théorie des lignes de transmission. Il s'agit de caractériser chaque composant du réseau, ainsi le comportement de l'ensemble résulte de la concaténation des différents composants. Durant la thèse de M. Ait Ou Kharraz, les éléments caractérisés sont les câbles de distribution et de branchement, ainsi que l'installation chez les consommateurs. Les technologies de câble analysées ont été choisies afin de mieux représenter le parc français : câble de distribution souterrain, ligne de distribution aérienne torsadée, câble de branchement triphasé et câble de branchement monophasé. En particulier, les impédances caractéristiques et les constantes de propagation (pertes et déphasages linéiques) des câbles constituent les paramètres physiques ayant été retenus. Plusieurs méthodes de mesures ont été proposées avec une comparaison qualitative des résultats. L'analyse est détaillée dans [AIT OU KHARRAZ et collab. \[2018\]](#) et la méthode retenue exploite le RLC-meter.

Différents appareils domestiques ont été caractérisés en vue d'obtenir leur impédance d'entrée vue par le réseau domestique à l'aide d'un analyseur de réseaux vectoriel. L'impédance d'entrée d'une sélection d'appareils parmi les plus utilisés dans les installations a été mesurée et le câble de l'installation domestique a été caractérisé pour définir son impédance caractéristique et sa constante de propagation. Cela a permis de simuler différentes configurations d'installation et de calculer leurs impédances vues par le réseau BT dans la bande de fréquence 30 kHz - 500 kHz en fonction des appareils connectés et la longueur du câble d'installation. Les résultats sont présentés dans [KHARRAZ et collab. \[2017\]](#)

Cette collaboration se poursuit dans le cadre de la thèse de A. Arbia Haded (2018-2021) et concerne la mise en œuvre d'un logiciel de modélisation de la propagation des signaux CPL dans un réseau électrique. Ce logiciel incorpore les briques développées dans la thèse de M. A. Oukharraz.

Publications représentatives impliquant différentes collaborations institutionnelles

Collaboration avec le laboratoire SONDRRA

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- [RI.7] M. Serhir and R. Guinvarc'h, « A low-profile cavity-backed dual-polarized spiral antenna array », *IEEE Antennas and Wireless Propagation Letters*, vol. 12, pp. 524-527, 2013

Collaboration avec University of Electronic Science and Technology of China

- [RI.14] H. Tu, S. Xiao, D. Lesselier, M. Serhir, « Super-resolution characteristics based on time-reversed single-frequency electromagnetic wave », *Journal of Electromagnetic Waves and Applications*, 30 (13), pp.1670 – 1680, 2016

Collaboration avec le département MIRE de EDF R&D

[RI.21] M. Ait Ou Kharraz, D. Picard, M. Serhir, P. Jensen and C. Lavenu, « Measurement Methods of Outdoor Low Voltage Cable Characteristics for Narrowband Power Line Communication », *IEEE Transactions on Power Delivery*. doi : 10.1109/TPWRD.2018.2876248

3.6 Bilan des contributions et perspectives

IL est difficile de dresser le bilan d'une activité qui a duré plus de dix ans entre 2004, l'année du début de ma thèse de doctorat et 2015 l'année de soutenance de la thèse de M. Farouq. J'ai travaillé sur plusieurs aspects : (i) développements expérimentaux, (ii) transformation champ proche/champ lointain dans les systèmes de coordonnées canoniques, (iii) la modélisation d'antenne, (iv) l'échantillonnage irrégulier et régulier en champ proche, (v) la méthode intégrale des courants équivalents, (vi) le champ proche dans le domaine temporel.

J'estime que le travail effectué sur la méthode matricielle représente une belle synthèse de cet axe de recherche. À travers cette méthode, j'exprime ma contribution dans cette thématique de recherche sur un plan à la fois conceptuel et méthodologique. Nous proposons de désolidariser la géométrie de la surface de mesure du système de coordonnées dans lequel est exprimé le développement modal du champ. L'apport méthodologique réside dans le formalisme qui traduit l'écriture modale du champ en un système d'équations linéaires. La méthode matricielle, par son efficacité nous donne la possibilité d'exploiter une surface de mesure quelconque, pour des raisons de compacité, de simplicité de mise en œuvre ou de suppression d'erreurs de troncature (mesure sur une calotte et développement en ondes planes).

Cette méthode a proposé une rupture dans les pratiques observées dans les techniques de champ proche basées sur le développement modal qui étaient jusqu'ici réservées à des surfaces de mesure canoniques avec un échantillonnage régulier suivant les deux variables de description de ces surfaces. Nous avons développé cette méthode dans la perspective d'équiper le laboratoire d'un robot de mesure capable de déplacer une sonde autour de l'objet rayonnant sans consignes géométriques particulières. La méthode matricielle a été mise en œuvre pour les trois développements modaux : cartésien, cylindrique et sphérique. Nous l'avons exploité pour des surfaces de mesure différentes.

La méthode matricielle offre des perspectives intéressantes. La réduction du nombre de points de mesure en utilisant un maillage adaptatif en est une. Il s'agit, a priori, de repérer les zones de variations rapides du champ dans lesquelles il faut prélever suffisamment de points de mesure. À contrario, dans les zones lentement variables, un nombre réduit d'échantillons seront prélevés. Avec cette stratégie, nous engendrons un échantillonnage irrégulier que l'on peut exploiter avec la méthode matricielle pour déterminer le champ lointain. Sur le moyen terme, j'orienterai mes travaux pour étudier la caractérisation en champ proche sans phase. Je souhaite aussi mettre en œuvre la méthode multipolaire rapide pour tester ses potentialités avec des données irrégulières sans phase.

Sur un autre registre, ma contribution dans la thématique du développement instrumental réside dans la mise en place d'une nouvelle plateforme de caractérisation d'antenne dans le domaine temporel. Ce système de mesure a servi à renforcer notre implantation territoriale dans le tissu socio-économique de la région en collaborant avec Rincet BTP (plusieurs années) sur la caractérisation des radars de sol. Sur le plan national, nous faisons partie du réseau Ruch (Réseau des Utilisateurs de Chambres anéchoïques Hyperfréquence) piloté par J. M. Geffrin de l'Institut Fresnel.

Dans une prochaine étape, je souhaite enrichir mon apport dans la caractérisation d'antennes dans le domaine temporel en généralisant l'utilisation des surfaces arbitraires de mesure du champ proche et en développant des techniques de mesure dans des environnements non-anéchoïques. Les trois collaborations décrites donnent une idée sur le dynamisme de l'activité au laboratoire. Disposer d'une infrastructure matériel et humaine dans son entourage proche est stimulant.

3.7 Références

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Chapitre 4

Imagerie micro-onde appliquée au GPR

« Je laisse Sisyphe au bas de, la montagne! On retrouve toujours son fardeau. Mais Sisyphe enseigne la fidélité supérieure qui nie les dieux et soulève les rochers. Lui aussi juge que tout est bien. Cet univers désormais sans maître ne lui paraît ni stérile ni futile. Chacun des grains de cette pierre, chaque éclat minéral de cette montagne pleine de nuit, à lui seul, forme un monde. La lutte elle-même vers les sommets suffit à remplir un cœur d'homme. Il faut imaginer Sisyphe heureux. »

Albert Camus

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4.1 Introduction et contexte

Le contrôle non destructif à base de radar de sol (Ground Penetrating Radar GPR) est un sujet multi-disciplinaire où sont impliquées de nombreuses communautés scientifiques [DANIELS \[2005\]](#). Ce thème de recherche est basé sur l'étude des interactions onde-matière. Plus particulièrement, la rencontre d'une onde électromagnétique avec un *objet* défini par ses propriétés géométriques et diélectriques engendre un champ diffracté. Ce dernier est utilisé pour détecter/localiser et/ou caractériser cet *objet* de manière aveugle car le champ diffracté porte en lui les caractéristiques de sa source.

Le GPR a été utilisé pour le suivi de l'humidité dans le sol [HUISMAN et collab. \[2003\]](#) [LAMBOT et collab. \[2004\]](#) [BENEDETTO \[2010\]](#), la localisation des canalisations et des structures souterraines [ANNAN \[2003\]](#), la maintenance des infrastructures en génie civil [LAI et collab. \[2018\]](#), les fouilles archéologiques [MELLETT \[1995\]](#) [CONYERS \[2013\]](#), la détection et le diagnostic des systèmes racinaires [MERTENS et collab. \[2016\]](#) [GUO et collab. \[2013a\]](#) [LIU et collab. \[2017\]](#) etc....

Mes travaux de recherche autour de l'imagerie du proche sous-sol ont débuté en 2014 dans le cadre de mes collaborations avec Rincenc BTP (S. Pellevrault) et EDF R&D Laboratoire des matériaux électriques (C. Mirry et E. Amador). Les travaux menés avec Rincenc BTP se focalisaient sur l'optimisation d'antennes ultra-large bande pour la caractérisation des couches supérieures de la chaussée, tandis que EDF R&D s'intéressait à la détection et la localisation des câbles souterrains.

J'ai structuré mes travaux de recherches sur l'imagerie du proche sous-sol autour de trois thèmes : (i) la modélisation électromagnétique et l'analyse du problème direct, (ii) l'imagerie micro-onde comportant inversion et post-traitement des données et (iii) l'instrumentation hyperfréquence qui supporte les expérimentations et les mises en application. Les thèses de [Xiang Liu \(2014 - 2017\)](#) et [Abderrahmane Aboudou-rib \(2017 - 2020\)](#) ont été définies pour apporter des contributions selon ces trois thèmes.

La thèse de Xiang Liu (2014-2017) [LIU \[2017\]](#) s'intitulait « Contrôle non destructif du sol et imagerie d'objets enfouis par des systèmes bi- et multi-statiques : de l'expérience à la modélisation ». Nous nous sommes fixés l'objectif d'étudier les potentialités d'un radar de sol multi-antennes combiné avec un logiciel d'imagerie permettant une reconstruction des objets enfouis. Les acquisitions et les traitements des signaux radar sont effectués dans le domaine fréquentiel dans la bande [0.5 GHz - 3.5 GHz]. Cette thèse a été financée par la fondation Supélec et l'équipe d'encadrement était composée de compétences complémentaires : M. Lambert, L. Pichon, A. Kamani et moi-même.

Mes travaux de recherche durant la période 2014 à 2017 sont orientés pour apporter des contributions dans :

1. La modélisation électromagnétique du problème direct d'un radar en utilisant la méthode du « Galerkin discontinu dans le domaine temporel ». Cette méthode est prometteuse et le travail de modélisation que l'on souhaite faire doit prendre en compte les antennes réelles utilisées lors des mesures.
2. La résolution du problème inverse associé à une configuration multi-statique en développant la méthode d'échantillonnage linéaire « Linear Sampling Method ». Nous avons choisi cette méthode pour sa simplicité de mise en œuvre. Cette méthode doit être éprouvée numériquement et expérimentalement.
3. La mise en œuvre des techniques classiques de migration pour consolider nos connaissances du sujet et faire des comparaisons avec les résultats du « Linear Sampling Method ».

La réflexion expérimentale accompagne toutes les étapes de nos travaux de recherche. Les travaux de modélisation sont toujours appuyés par des mesures en environnement contrôlé.

Dans le cadre de la thèse de Xiang Liu nous avons opté pour une approche multi-statique / multi-fréquence pour atteindre une diversité spatiale et fréquentielle lors des acquisitions des champs diffractés. L'approche multi-statique est jumelée avec la méthode d'inversion qualitative : « Linear Sampling Method » (LSM). L'analyse des résultats de cette méthode d'inversion est faite selon plusieurs configurations d'émetteurs / récepteurs et avec des cibles de formes variées.

Pour les besoins de nos expérimentations, un banc de test a été mis en œuvre pour valider nos analyses du problème direct. Nous avons utilisé la méthode du Galerkin discontinu dans le domaine temporel pour la modélisation électromagnétique des scénarios étudiés. Pour les besoins de nos simulations/mesures, nous avons optimisé plusieurs antennes large bande communément utilisées dans le GPR. Nos mesures sont exclusivement faites dans le domaine fréquentiel via un analyseur de réseau vectoriel.

Le GPR est un thème de recherche multi-disciplinaire. Il rassemble une communauté plurielle comme en témoigne les participants dans l'action COST Action TU1208 (2013 - 2017) (Transport and Urban Development, Civil Engineering Applications of Ground Penetrating Radar) [BENEDETTO et PAJEWSKI \[2015\]](#). Au

sein du COST, quatre groupes de travail (Working Groups) ont été identifiés. Le WG1 est dédié à la conception, optimisation et calibration de radars de sol innovants. Le WG2 est consacré aux travaux du génie civil liés au diagnostic des infrastructures de transport et des bâtiments, à la détection des canalisations souterraines et des cavités, au test des matériaux de construction et l'estimation de la teneur en eau des sols. Le WG3 se focalise sur le développement de techniques d'imagerie et d'inversion et la caractérisation électromagnétique des scénarios complexes. Le WG4 s'oriente vers des domaines complémentaires du génie civil où les approches non-destructives sont exploitées en complément du radar de sol. J'ai participé à cette action COST en tant que *MC Substitute* et mes travaux de recherche s'inséraient dans les WG1 et WG3.

4.2 Analyse du problème direct en contexte GPR

Nous signifions par scénario GPR (Ground Penetrating Radar) un environnement dénué de source où nous considérons deux antennes, jouant respectivement le rôle d'émetteur et de récepteur (cas bi-statique). Il n'y a aucune restriction quant au choix d'explicitation la configuration bi-statique. Le schéma d'analyse reste valable pour les cas mono- et multi-statiques.

L'antenne émettrice (champ incident) est la source d'un champ électromagnétique (harmonique ou impulsionnel) qui interagit avec une cible caractérisée par ses qualités diélectriques et géométriques. Cette cible est placée dans un domaine possédant des propriétés géométriques et diélectriques. L'antenne de réception capture la réponse issue de la scène exposée au champ incident.

Nous distinguons, en conséquence, deux situations différentes. Dans la première, nommée problème direct, les caractéristiques de la scène sont dévoilées et l'on souhaite déterminer la réponse électromagnétique du milieu considéré quand ce dernier est illuminé par un champ incident connu. Dans la deuxième situation, le système étudié est considéré comme une boîte fermée (voilée). Nous connaissons la réponse du système étudié à un champ incident connu, et nous souhaitons remonter aux caractéristiques géométriques/diélectriques des parties prenantes du scénario.

La résolution du problème direct a pour objectif de déterminer la réponse électromagnétique du système quand le domaine d'étude est illuminé par un champ incident connu. Le formalisme de Maxwell est exploité dans son expression harmonique ou temporelle. Le problème électromagnétique direct est résolu à partir d'une discrétisation du problème et une résolution des équations aux dérivées partielles en respectant les conditions aux limites imposées par le cas étudié [CHEW \[1995\]](#).

En régime harmonique, je propose une liste non-exhaustive de quelques méthodes numériques exploitées dans la résolution du problème direct. La **méthode des moments** (MoM) se base sur la résolution des équations de Maxwell sous leur forme intégrale [GIBSON \[2014\]](#) [HARRINGTON \[1987\]](#). Elle se distingue par l'exploitation d'un maillage où seules les zones de discontinuité du champ sont discrétisées et n'est exploitable que dans les problèmes où les fonctions de Green sont calculables. Pour nos travaux de recherche, nous avons utilisé la MoM (cas 2D) pour générer des données synthétiques afin de valider la méthode d'inversion qualitative que nous avons développée. La **méthode des éléments finis** (FEM) [MONK et collab. \[2003\]](#) [JIN \[2015\]](#) est basée sur la résolution des équations aux dérivées partielles sur un maillage non structuré convenable aux géométries complexes. La FEM a été exploitée dans un contexte GPR dans [ARIAS et collab. \[2007\]](#) et [LUBOWIECKA et collab. \[2009\]](#).

Dans le domaine temporel, la **méthode des différences finies** (FDTD) « Finite Difference Time Domain » est souvent utilisée pour modéliser des scénarios GPR. Elle est proposée pour la première fois dans [KANE YEE \[1966\]](#). Cette méthode est caractérisée par sa simplicité, son efficacité et sa stabilité conditionnelle [TAFLOVE et HAGNESS \[2005\]](#). La FDTD conventionnelle utilise un maillage rectangulaire hexaédral peu adapté pour des géométries complexes. Une version avec un maillage non-régulier a été proposée dans [HOLLAND \[1983\]](#).

La FDTD est populaire au sein de la communauté GPR [BOURGEOIS et SMITH \[1996\]](#) [GIANNOPOULOS \[2005\]](#). Antonios Giannopoulos de l'*University of Edinburgh* a proposé « GprMax », un outil FDTD *open source*, qu'il a mis à la disposition de la communauté pour modéliser des scènes GPR. Plusieurs fonctionnalités sont implémentées dans GprMax et permettent de prendre en compte par exemple : la rugosité de l'interface air/sol, un modèle réaliste du sol (Peplinski), hétérogénéité du sol, etc....

La dernière méthode que je souhaite introduire est la **Technique des Intégrations Finies** « Finite Integration Technique » (FIT). Cette méthode est décrite dans [WEILAND \[1977\]](#). Elle est similaire à la FDTD hormis l'exploitation de la formulation intégrale des équations de Maxwell [CLEMENS et WEILAND \[2001\]](#) [MARKLEIN \[1999\]](#). La FIT est aussi avantageuse que la FDTD en terme de simplicité d'implémentation. Elle se distingue par sa flexibilité dans la modélisation des géométries complexes et traite les interfaces entre différents matériaux de façon plus précise. La FIT est utilisée dans le logiciel commercial de simulations électromagnétiques *Computer Simulation Technology, Microwave Studio CST-MWS* pour des modélisations

dans le domaine temporel. Ce logiciel a été largement utilisé durant la thèse de X. Liu

Les méthodes de simulations électromagnétiques sont plurielles. Cependant, il n'existe pas une méthode « universelle » qui garantirait une efficacité absolue : il y a un compromis à trouver entre précision, complexité de mise en œuvre et temps de calcul. Dans l'équipe de Giannopoulos, G. Warren a proposé dans [WARREN et collab. \[2017\]](#) une comparaison entre la FDTD et FIT. Teixeira propose dans [TEIXEIRA \[2008\]](#) une comparaison entre la FDTD et la FEM dans le domaine temporel.

4.3 Contributions dans l'analyse du problème direct : DGTD

Dans le cadre de la thèse de Xiang Liu (2014 – 2017) [LIU \[2017\]](#) nous avons utilisé la méthode de « Galerkin discontinu dans le domaine temporel » (DGTD- Discontinuous Galerkin Time Domain) pour résoudre le problème direct en GPR. Nous avons testé ses potentialités pour modéliser des antennes ultra-large bande en présence d'un bac à sable dans lequel sont enfouies des cibles de différentes sortes. La méthode de Galerkin discontinu propose une discrétisation des équations de Maxwell où le domaine de recherche est maillé de manière non-structurée. La DGTD permet de résoudre les équations sous leur forme variationnelle sur chaque maille à l'image de la méthode des éléments finis. Cependant, elle se distingue par l'indépendance des mailles les unes des autres contrairement à la FEM où le maillage est continu. Dans la DGTD les mailles sont reliées par des termes de « flux numérique ». L'évolution temporelle de l'approche DGTD est régie par les schémas classiques comme Euler, saute-mouton (LeapFrog), Runge-Kutta [BOUBEKEUR et collab. \[2014\]](#) [NTICHI et collab. \[2013\]](#).

Dans le travail de recherche que nous avons mené sur la modélisation du problème direct en contexte de radar de sol, nous avons pris soin de modéliser les antennes non comme des points sources, mais comme des structures ayant des propriétés diélectriques et géométriques. Cela implique directement une réduction du pas de temps selon la condition Courant-Friedrichs-Lewy (CFL). Il s'agit de considérer des scènes relativement grandes contenant des structures de tailles très hétérogènes. Nos résultats de simulations sont comparés aux mesures en environnement contrôlé. Les simulations sont faites dans le domaine temporel et les résultats sont éprouvés dans le domaine temporel et fréquentiel. Dans les simulations des scénarios GPR avec DGTD, le maillage est obtenu avec le logiciel libre GMSH [GEUZAIN et REMACLE \[2009\]](#).

Dans une première contribution [LIU et collab. \[2016\]](#) (communication jointe avec ce rapport), nous avons considéré deux antennes Vivaldi identiques (émettrice et réceptrice) placées au dessus d'un bac à sable 1m x 0.6m x 0.21m. Les A-scan, composés du coefficient de transmission S_{21} , sont calculés sur plusieurs points le long du bac à sable avec le code de DGTD développé par A. Kameni [KAMENI et collab. \[2015\]](#). Le même scénario est modélisé avec le logiciel commercial CST MWS (FIT). Le bac à sable comporte des cibles métalliques et diélectriques et les résultats numériques sont comparés aux mesures effectuées en chambre anéchoïque. Les détails des comparaisons sont produits dans [LIU et collab. \[2016\]](#) où nous commentons les temps de calcul de chaque méthode et présentons les critères de comparaisons utilisés. Les comparaisons avec les mesures confirment la précision et la pertinence de la DGTD pour la modélisation du problème direct en contexte GPR. Cette méthode est légèrement plus précise que CST-MWS et les temps de simulation sont équivalents. L'avantage de la DGTD réside dans l'efficacité de son maillage et son adaptabilité au calcul parallèle.

Publications représentatives sur l'analyse du problème direct en GPR

- [CI.18] X. Liu, A. Kameni, M. Serhir, M. Lambert, L. Pichon, « Discontinuous Galerkin Time-Domain method for 3D modeling of ground penetrating radar scenarios », *Seventeenth Biennial IEEE Conference on Electromagnetic Field Computation*, Nov 2016, Miami, Florida USA
- [CI.28] X. Liu, M. Serhir, A. Kameni, M. Lambert, L. Pichon, « Buried targets detection from synthetic and measured B-scan ground penetrating radar data », *11th European Conference on Antennas and Propagation*, Mar 2017, Paris, France
- [CI.27] X. Liu, M. Serhir, A. Kameni, M. Lambert, L. Pichon, « Ground penetrating radar data imaging via Kirchhoff migration method », *Applied Computational Electromagnetics Society*, Mar 2017, Florence, Italy

Discontinuous Galerkin Time-Domain method for 3D modeling of ground penetrating radar scenarios

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Abstract—In this paper, a Discontinuous Galerkin Time-Domain (DGTD) method is used to model full Ground Penetrating Radar (GPR) scenarios. The bi-static GPR is made of two identical Vivaldi antennas, which are accurately modeled in front of a sand box in order to detect and localize some buried objects. Our approach is validated by comparison with either synthetic data provided by CST-MWS or experimental data acquired in a controlled laboratory experiment in an anechoic chamber. The accuracy of DGTD method is proved. The preliminary imaging result is obtained via Kirchhoff migration method.

Index Terms—Antennas, Microwaves, Numerical Methods.

I. INTRODUCTION

The use of Ground-Penetrating Radar (GPR) has been developed in an extended number of applications for many years. The GPR technique is used in geophysics and civil engineering to provide effective and accurate imaging of underground structures and for buried objects detection. It still needs development as shown by the recent COST Action TU1208 “Civil Engineering Application of Ground Penetrating Radar” [1], [2]. Therefore, it is interesting to explore the GPR performance for estimating the target location.

For the understanding and investigation of the electromagnetic phenomena in Ground Penetrating Radar (GPR) conception application, the computational solution of electromagnetic problem becomes necessary. In order to properly solve the corresponding problem the chosen numerical technique should show its ability to handle complex geometries and different boundary conditions, its efficiency and accuracy of computation.

In decades, many different numerical methods were implemented for GPR simulation, such as Finite Elements Method (FEM) and Finite Difference Time Domain method (FDTD). Each of those methods has several advantages and disadvantages. The most commonly used for the time domain simulation is the FDTD method. Nevertheless the conventional FDTD uses a regular mesh which is a limitation to model complex antenna geometries. The geometries of various scale length lead to fine meshes and result in reducing the stable time step given by the classical Courant Friedrichs Lewy (CFL) condition leading to an increase of the computational cost. For such a complex geometry, using a Cartesian mesh element increases the dispersive error which affects the convergence it is why the FEM methods and its high order mesh elements have been exploited to reduce dissipative error [3]. However FEM use implicit time integration schemes thus its implementation for parallel computing is tedious [4].

The DGTD method is a powerful approach for solving a partial differential equations [4]. It combines FEM and Finite Volume Method (FVM) tools. The discretization of equation in each cell is achieved by FEM approach and the flux terms is used to connect adjacent elements like in FVM method. The DGTD combined with explicit schemes is suitable for parallel computing and for using high order elements. Hence the DGTD seems to be an appropriate method for modeling complex antenna structures located in different environments. It allows to easily discretize GPR scenarios using various dimensions and/or shapes with variable element size. This method has been recently introduced to model antennas [5], [6]. The block diagonal matrices generated by this method facilitate the parallel computing [7]. Its discontinuous aspect is suitable for discontinuous function and simplifies implementation of high order elements constructed thanks to parametrization rules. High accuracy of solution is proved by DGTD simulation with reasonable numerical cost. The GPR data obtained by simulation and measurement are used to localize buried objects via migration technique.

The paper is organized as follows: section II is dedicated to introduce the DGTD formulation. Section III presents the GPR scenarios configuration. Section IV focuses on the discussion of numerical results. Section V proposes the imaging results via Kirchhoff migration (KM) using simulation/measurement data. Finally, a conclusion is drawn to outline the interest of the DGTD method for the GPR application.

II. FORMULATION

The discontinuous Galerkin approach was applied for solving conservative forms of Maxwells’ equations given by:

$$\begin{cases} \int_T \epsilon \partial_t \vec{E} \cdot \vec{\phi} - \int_T \vec{H} \cdot \nabla \times \vec{\phi} - \int_{\partial T} (\vec{n} \times \vec{H})^{num} \cdot \vec{\phi} \\ \hspace{15em} = - \int_T \sigma \vec{E} \cdot \vec{\phi} \\ \int_T \mu \partial_t \vec{H} \cdot \vec{\psi} + \int_T \vec{E} \cdot \nabla \times \vec{\psi} + \int_{\partial T} (\vec{n} \times \vec{E})^{num} \cdot \vec{\psi} = 0 \end{cases} \quad (1)$$

where $\vec{\phi}$ and $\vec{\psi}$ are test functions, and T is mesh element of the domain $\Omega = \cup T$.

Adjacent elements are connected using numerical flux terms $(\vec{n} \times \vec{H})^{num}$ and $(\vec{n} \times \vec{E})^{num}$ written as:

$$\begin{cases} (\vec{n} \times \vec{H})^{num} = \vec{n} \times \frac{\{Z\vec{H}\}}{\{Z\}} - \alpha(\vec{n} \times \frac{(\vec{n} \times [\vec{E}])}{\{Z\}}) \\ (\vec{n} \times \vec{E})^{num} = \vec{n} \times \frac{\{Y\vec{E}\}}{\{Y\}} + \alpha(\vec{n} \times \frac{(\vec{n} \times [\vec{H}])}{\{Y\}}) \end{cases} \quad (2)$$

Where the impedance $Z = 1/Y = \sqrt{\mu/\epsilon}$, $[u] = \frac{u^+ - u^-}{2}$ and $\{u\} = \frac{u^+ + u^-}{2}$. The subscript “-” denotes the values in the current element, and “+” for the adjacent element. In our case, the dissipative upwind flux term ($\alpha = 1$) is used and a 2-stages explicit Runge Kutta scheme with multi-rate time integration is applied to compute the time domain solution of EM problem. This explicit scheme improves the accuracy of the solution by increasing the order of convergence compared to the Leap-Frog (LF) scheme used with the centred flux. Furthermore, it allows an increase of computation efficiency compared with the single-rate methods [8].

III. GPR SCENARIOS AND EXPERIMENTAL SETUPS

A bi-static GPR made of two identical Vivaldi antennas and a sand-filled wooden box compose the GPR scenario (Fig. 1). The front view scheme of GPR scenario is presented in Fig. 1(a), where only the sand-filled part is depicted. E/R denote the transmission/reception Vivaldi antennas operating from 500 MHz to 3.5 GHz with 128 frequency-steps. The wooden box filled with dry sand ($\epsilon_r \simeq 2.53$) is placed over a metal plate in order to eliminate uncontrollable reflections resulting from the support of the box and the soil floor.

The buried objects are two metal bars of the same length and of square/circular cross-section, and a air-filled paper pipe. For the first scenario (Scenario 1), the rectangular metal bar is horizontally placed at the center of the sandbox at a depth of $P_1 = 125$ mm whereas the cylindrical metal bar is placed at a depth of $P_2 = 95$ mm, the lateral distance between the two targets being $E = 50$ mm (Fig. 1(a)). In the second one (Scenario 2), an air-filled paper pipe ($\varnothing = 90$ mm) is buried at the box center and 85 mm deep.

The center of the sandbox is considered as the origin of the axis ($x = 0$), the B-scan displacement range going from -200 mm to 200 mm with a step of 20 mm. Indeed, we have used a step frequency continuous wave strategy to acquire the data using the Agilent E5071B ENA Vector Network Analyzers. Then the results are Fourier transformed to obtain the time domain signals and radargrams.

IV. NUMERICAL RESULT

The same configuration is considered in the time domain simulation procedure with DGTD or CST-MWS in which both the antennas and sandbox are modeled (3D model), the sand being considered as a homogeneous medium with $\epsilon_r = 2.53$. A tetrahedral mesh of 48735/41633 third-order spatial elements is used in DGTD simulations for Scenario 1/2. The number of multi-rate groups is 5. An Sine-Gaussian pulse is used as excitation given by: $e(t) = \exp\left(-\left(\frac{t-\tau}{T}\right)^2\right)$.

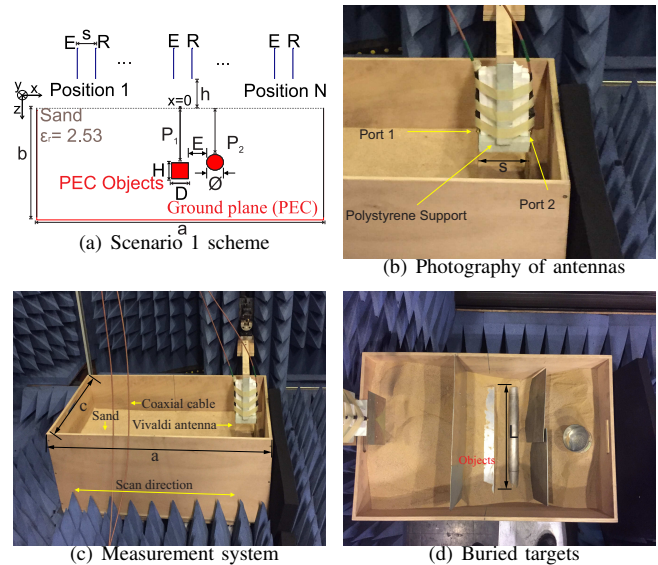


Fig. 1. GPR scenario scheme (a), the dimensions (in mm) being $a = 1000$, $b = 210$, $c = 600$, $H = 50$, $D = 50$, $L = 500$, $P_1 = 125$, $\varnothing = 50$, $P_2 = 95$, $E = 50$, $S = 100$, $h = 2$ and photos of the equipment (b) (c) (d).

$\sin(2\pi f_0 \cdot (t - \tau))$. Its spectrum is normalized at $f_0 = 2.0$ GHz for $\tau = 1.185$ ns and $T = 0.32$ ns, which covers the antenna frequency band [0.5 GHz - 3 GHz]. The frequency response (S-parameters of the antenna) are obtained by FFT (Fast Fourier Transformation) of the time domain simulation results.

The time domain results for the Scenario 1 are presented in Fig. 2 and Fig. 3. The scattered field is the difference between the total and incident fields, where the incident field is obtained by measuring or modeling the sandbox without buried objects. In the radargrams, the solid line contour presents the exact position of buried objects, the dashed line is the bottom metal plate. The E_x presented in Fig. 2(d) and Fig. 3(d) are the x component recorded at the port of receiver antenna (port 2 in Fig. 1(b)) for the measurement/simulation at position $x = 0$. The transmission parameter S_{21} obtain at the same position are presented in Fig. 4.

The results of simulations/measurement are very close. The signature of two buried objects can be clearly observed in either total or scattered field radargrams. The quantitative validations are achieved by comparing the time domain signals (E_x) or transmission parameter (S_{21}).

For the Scenario 2, the radargrams of scattered/total field obtained via measurement and DGTD simulation are presented in Fig. 5. The reflection due to the dielectric object is feeble but visible. The time and frequency domain quantitative results are shown in Fig. 6.

Consideration the presented comparisons between simulations and measurement results for the two GPR scenarios, DGTD method is validated for either metal or dielectric objects.

We introduce the correlation coefficient and the Root Mean Squared Error (RMSE) to analyze the accuracy of the simulation results. The measured electric field E_x for center position ($x = 0$) are set as reference, hence those two

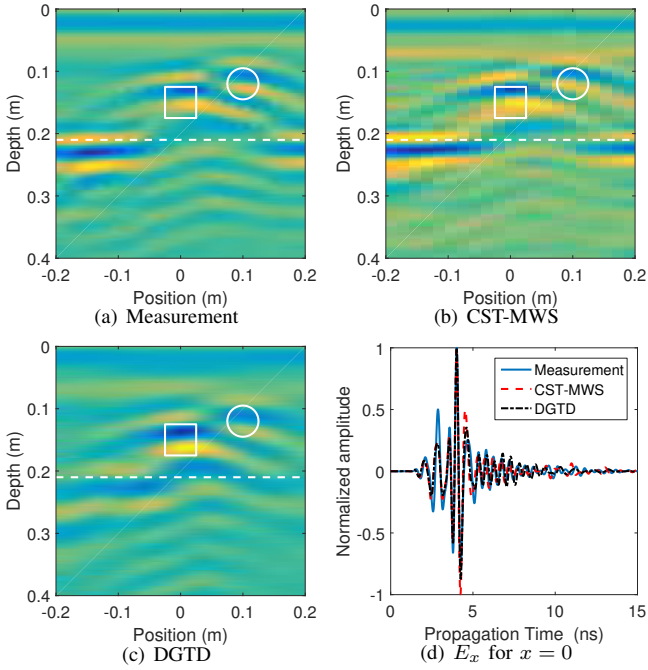


Fig. 2. Time domain results for two buried metal objects. Radagrams of total field: Measurement (a), CST-MWS (b), DGTD (c) and the E_x for the position of $x = 0$ (d).

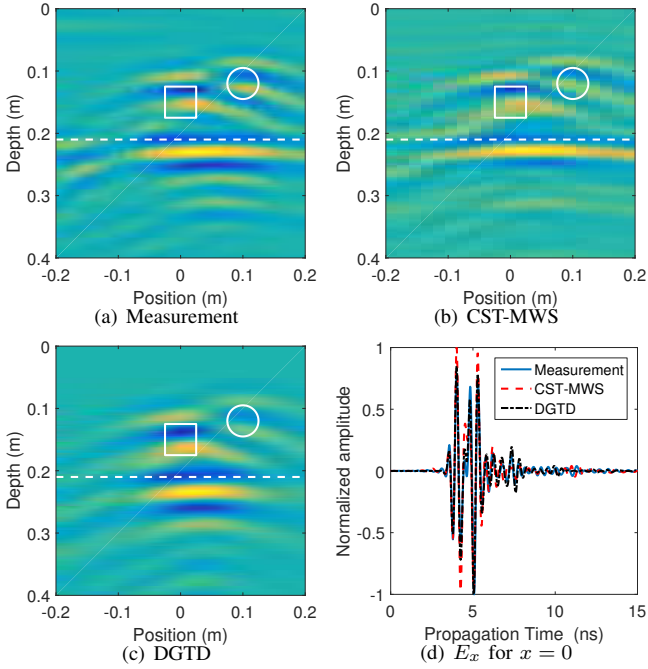


Fig. 3. Time domain results for Scenario 1. Radagrams of scattered field: Measurement (a), CST-MWS (b), DGTD (c) and the E_x for the position of $x = 0$ (d).

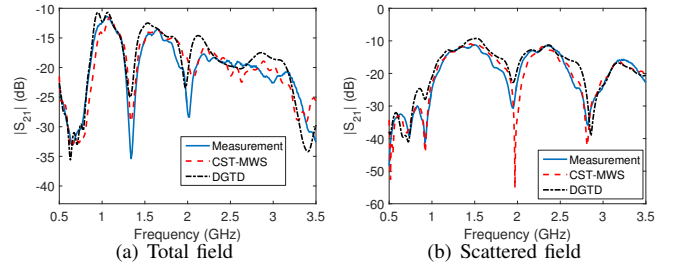


Fig. 4. Transmission parameter S_{21} for Scenario 1: total field (a), scattered field (b)

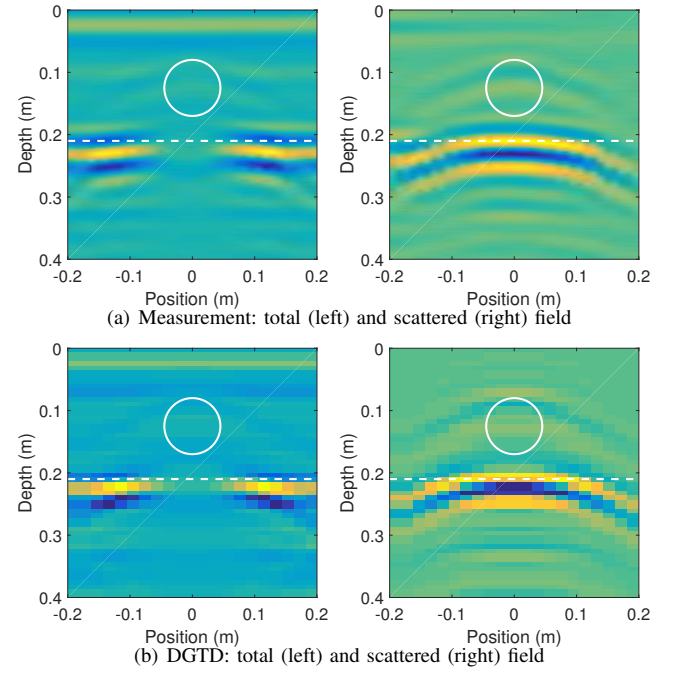


Fig. 5. Time domain results for Scenario 2. Radagrams of total field (left) scattered field (right).

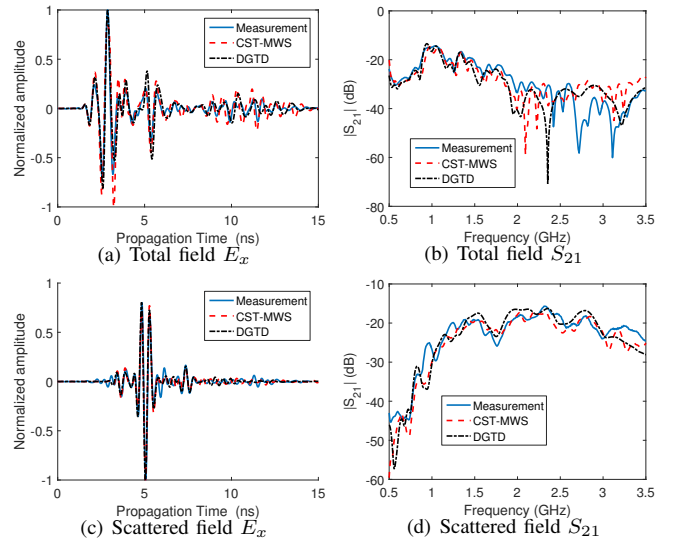


Fig. 6. The quantitative comparison for Scenario 2.

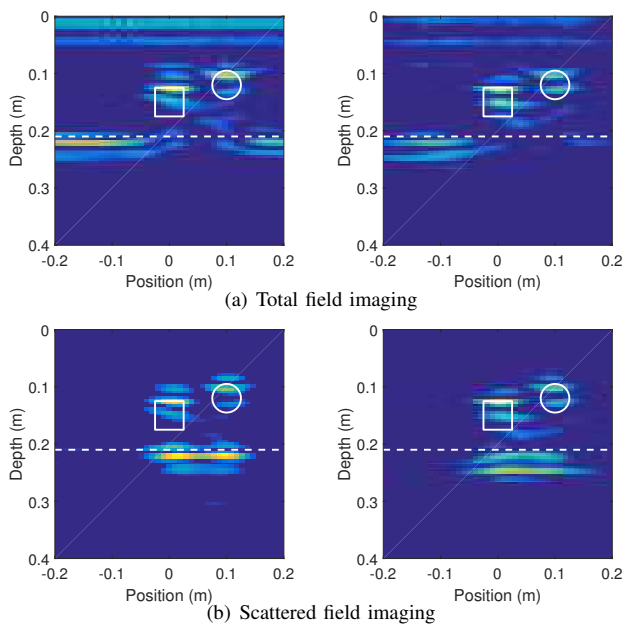


Fig. 7. KM imaging using measurements (left) and DGTD (right) simulation results for two buried metal objects.

parameter are calculated by comparing CST-MWD and DGTD E_x results with measurement values. The results of total field and scattered field for both two scenarios are compared in TABLE I. The improvement of accuracy of the DGTD solutions can be observed for both total and scattered field in any scenario.

The DGTD and CST-MWS simulations are carried out for a duration of 15ns. The DGTD is performed on 32 cores Xeon(R) 2.0 GHz cluster using 16 - 24 cores. The CST-MWS runs on a 8 cores i7 2.5 GHz computer. Their CPU times for one x position simulation of Scenario 1 are $CPU_{DGTD} \approx 1\text{h}40\text{min}$ and $CPU_{CST} \approx 1\text{h}5\text{min}$. Therefore, there is a trade-off between the accuracy of solution and computational time. But the DGTD method will be more efficiency using a powerful cluster/GPU.

Scenario	Total CST	Total GD	Scattered CST	Scattered GD
1	0.901/0.073	0.935/0.031	0.923/0.080	0.964/0.023
2	0.845/0.109	0.908/0.036	0.940/0.040	0.957/0.021

TABLE I

CORRELATION COEFFICIENTS/RMSE BETWEEN SIMULATION RESULTS AND MEASUREMENT FOR TOTAL AND SCATTERED E_x FIELD SIGNALS.

V. IMAGING PROCESSING

After the validation of our simulations and measurement, an imaging processing based on Kirchhoff migration (KM) is proposed to imaging the GPR scenarios with higher resolution than B-scan radargrams. KM is performed by using the Kirchhoff integral representation of a field at a given point as a superposition of waves propagating from adjacent points and times [9]. The KM imaging results for both Scenario 1 and 2 are presented in Fig. 7 and Fig. 8. The localization of objects is achieved by using measurement or DGTD data.

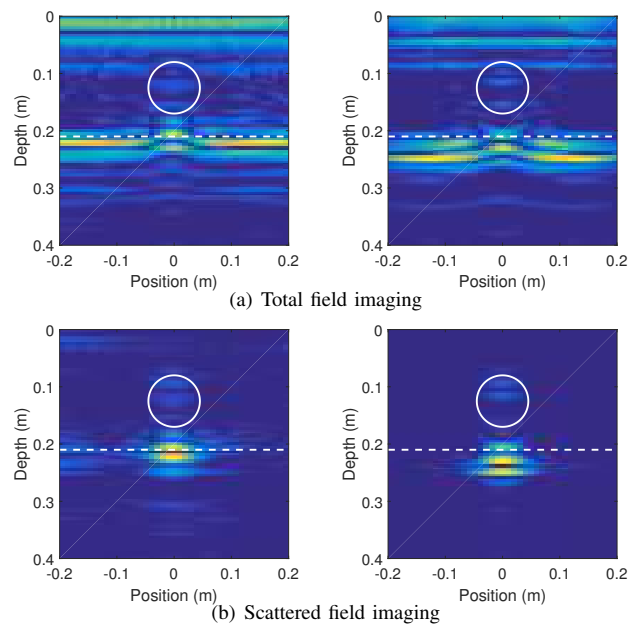


Fig. 8. KM imaging using measurements (left) and DGTD (right) simulation results for one buried dielectric object.

VI. CONCLUSION

In this paper, the numerical modeling of GPR scenarios validates our 3D full-wave simulation with DGTD method. The DGTD method proved its accuracy for modeling a complete electromagnetism scenario with buried metal or dielectric objects. Since the quantitative validation of GPR modeling is achieved, the simulation results could be used for GPR inverse scattering problem with numerical reconstruction algorithms. The DGTD method is proved as an alternative for GPR scenario modeling, and it will be interesting for high accuracy required applications while a powerful cluster/GPU is available. The localization of buried objects is achieved with Kirchhoff migration. Some advanced imaging techniques such as an Enhanced Linear Sampling Method already developed by the authors is currently under extension for this GPR application.

ACKNOWLEDGMENT

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4.4 Analyse du problème inverse en contexte GPR

Dans l'analyse du problème inverse, nous supposons la connaissance du champ diffracté et nous résolvons un problème inverse afin de localiser/identifier/caractériser son origine (source responsable du champ diffracté). Il convient de choisir une méthode d'inversion adaptée pour extraire l'information contenue dans le champ diffracté. Dans le chapitre [CATAPANO et collab. \[2015\]](#) on trouve une large description des techniques d'imagerie qualitatives et quantitatives exploitées dans le contexte GPR. Avant de procéder à la résolution du problème inverse en contexte GPR, les données mesurées sont post-traitées. Pour cela, il est nécessaire de :

1. Filtrer l'effet du "clutter" [SOLIMENE et collab. \[2013\]](#)
2. Caractériser (connaître) la permittivité environnante [GORRITI et SLOB \[2005\]](#)
3. Corriger l'effet de l'antenne émettrice/réceptrice [LAMBOT et ANDRÉ \[2013\]](#).

Ces trois étapes donnent accès au champ diffracté intrinsèquement lié à l'objet que l'on cherche à identifier/localiser/caractériser. Néanmoins, chaque point constitue un axe de recherche avec des variantes qui dépendent du contexte applicatif souhaité : le génie civil [SOLDOVIERI et collab. \[2006\]](#), l'archéologie [CATAPANO et collab. \[2012\]](#), la planétologie [WATTERS et collab. \[2006\]](#) [EYRAUD et collab. \[2018\]](#), etc....

Les méthodes quantitatives visent à résoudre un problème inverse pour reconstruire les caractéristiques géométriques et diélectriques de la source du champ diffracté. Les difficultés de résolution du problème inverse résident dans sa non-linéarité et son caractère mal-posé. Les travaux qui traitent du sujet contournent ces difficultés en ramenant sa résolution à celle d'un problème d'optimisation contraint. En effet, la résolution du problème devient itérative dans l'objectif de minimiser une fonction coût qui évalue la distance entre la solution issue du modèle et le champ diffracté mesuré. Cela nécessite la résolution du problème direct pour chaque itération et les paramètres du système sont modifiés selon différentes stratégies afin de contraindre les solutions pour éviter les minimas secondaires. Parmi les méthodes d'optimisation récurrentes, je cite : méthode de Newton [CUI et collab. \[2001\]](#), méthode du gradient modifié [KLEINMAN et VAN DEN BERG \[1992\]](#), optimisation par essais particuliers (Particle Swarm Optimization) [SALUCCI et collab. \[2016\]](#) ou gradient conjugué [SALUCCI et collab. \[2017\]](#).

Il reste, néanmoins, le problème de la convergence et le caractère mal-posé du problème qui accentue la sensibilité de la solution face au bruit de mesure. Basées sur la résolution successive du problème direct, les méthodes d'inversion non-linéaires sont exploitables dans une petite zone d'investigation et sont souvent limitées au cas 2D.

En revanche, quand on travaille dans le cadre de l'approximation de Born [PASTORINO \[2010\]](#) [CHEW \[1995\]](#), le problème devient linéaire et sa résolution qualitative est efficace numériquement [SOLDOVIERI et SOLIMENE \[2010\]](#). Les techniques de migration [GAZDAG \[1978\]](#) [STOLT \[1978\]](#) appartiennent à cette famille de solutions (problème linéaire). Elles ont été développées pour des problèmes d'inversion des signaux sismiques [YILMAZ \[2001\]](#) et se basent sur la réversibilité du temps. Les techniques de migration sont similaires à la rétro-propagation [HANSEN et JOHANSEN \[2000\]](#), au retournement temporel [FINK \[1992\]](#) et à l'imagerie radar à synthèse d'ouverture (SAR) [CAFFORIO et collab. \[1991\]](#). Si ce type de techniques d'imagerie produit des images focalisées, leurs performances restent néanmoins limitées. Les techniques de migration fonctionnent correctement avec des diffuseurs de petites tailles. Dans le cas où le milieu qui contient la cible est dispersif ou à pertes, ces techniques ne peuvent plus s'appliquer sans une phase de compensation [[YAVUZ et TEIXEIRA \[2009\]](#)].

Dans le cadre de l'approximation de Born, le problème est linéaire mais il reste mal-posé. C'est pour cela que les méthodes de résolutions qualitatives font appel aux techniques de régularisation pour assurer le compromis précision/stabilité de la solution [BERTERO et BOCCACCI \[1998\]](#). Les méthodes d'inversions linéaires permettent d'extraire la position et la forme du diffuseur et cela est suffisant pour une large gamme d'applications GPR.

Dans le cadre de la thèse de Xinag Liu, nous avons mis au point la méthode qualitative basée sur la résolution du problème linéaire dite : d'échantillonnage linéaire « Linear Sampling Method » (LSM) [COLTON et collab. \[2003\]](#) [COLTON et KRESS \[2012\]](#). La méthode du LSM est caractérisée par sa simplicité d'implémentation. Elle a été proposée pour la première fois par D. COLTON et A. KIRSCH dans [COLTON et KIRSCH \[1996\]](#). Cette méthode d'imagerie a été utilisée dans [CATAPANO et collab. \[2007\]](#) pour imager des cibles bidimensionnelles et dans [CATAPANO et collab. \[2008\]](#) pour des cibles 3D y compris en contexte GPR. Dans ces travaux, les sources du champ incident sont considérées idéales et ponctuelles et les résultats des imageries 2D et 3D sont présentées et discutées.

En réalité, les sources du champ incident sont des antennes qui ne peuvent être considérées comme des sources idéales dans le cadre d'une illumination en champ proche. Dans [CATAPANO et collab. \[2011\]](#), les auteurs ont étudié l'influence du diagramme de rayonnement des antennes émettrices sur le résultat de la LSM. Dans ce cas l'équipe de L. Crocco et I. Catapano a proposé des fonctions d'indicateurs plus adaptées ainsi que la définition des sources d'ordre supérieur (high order sources). Cela améliore la performance de la LSM telle que présentée dans [CROCCO et collab. \[2012b\]](#).

La LSM possède des similitudes avec la méthode de classification multiple des signaux, « Multiple Signal Classification » (MUSIC) [DEVANEY \[2000\]](#). Les deux méthodes exploitent des configurations multi-statiques dans le domaine fréquentiel et la reconstruction des objets enfouis est détériorée pour des configurations avec des vues tronquées [CATAPANO et collab. \[2011\]](#).

4.4.1 Contribution dans l'analyse du problème inverse : LSM

Dans le cadre de la thèse de Xiang Liu (2014-2017), nous avons mis en œuvre la Linear Sampling Method (LSM) comme technique d'imagerie qualitative à partir d'une mesure multi-statique. L'algorithme développé a été validé numériquement et expérimentalement. Nous avons utilisé des données synthétiques issues des simulations numériques (MoM) pour étudier les performances de la LSM selon différentes considérations. Nous avons comparé les résultats de reconstruction d'objets qui correspondent à une couverture circonférentielle complète/partielle avec les résultats d'une configuration GPR à aspect limité où des émetteurs/récepteurs sont distribués régulièrement le long d'un segment en face de la scène.

Nous avons explicité ces résultats dans [LIU et collab. \[2016\]](#) (joint à ce rapport). Le nombre d'émetteurs/récepteurs est défini en respectant le nombre de degrés de liberté du problème tel que décrit par [CROCCO et collab. \[2012b\]](#) (L. Crocco était dans l'équipe de O. Bucci il y a quelques années). Nous avons mené des comparaisons pour différents rapports signal à bruit et pour des cibles de formes variées. Nos investigations concernaient aussi les apports d'une diversité fréquentielle en terme de précision de reconstruction.

Se basant sur une acquisition multi-statique du champ diffracté, la LSM est mise en œuvre et validée pour des objets 2D. Nous avons utilisé le principe de Morozov pour déterminer le paramètre de régularisation de la solution (problème mal-posé). Les données synthétiques ont servi à étudier la pertinence du critère multi-fréquentiel à considérer à partir des indicateurs calculés à l'issue des différentes résolutions (mise en série ou en parallèle des indicateurs associés à chaque fréquence [GUZINA et collab. \[2010\]](#)). Cela a été étudié pour les configurations circonférentielles complète/partielle et GPR. Nous avons aussi utilisé les sources d'ordre élevé [CROCCO et collab. \[2012a\]](#) pour reconstruire la forme des diffuseurs à géométries complexes.

Nous avons validé expérimentalement l'algorithme du LSM développé pendant la thèse grâce à la base de données de diffraction de l'Institut Fresnel [BELKEBIR et SAILLARD \[2001\]](#), [GEFFRIN et collab. \[2005\]](#) [GEFFRIN et SABOUROUX \[2009\]](#). Dans cette base, on trouve les données du champ diffracté d'un bestiaire de cibles caractérisées à différentes fréquences selon un schéma multi-statique. Pour les objets à géométrie complexes, nous avons exploité la LSM avec des sources d'ordre élevé pour imager l'objet sous la forme de "U" proposé dans la database de l'Institut Fresnel.

Par ailleurs, nous avons proposé une technique de compensation pour améliorer la performance du LSM lorsque la distribution d'énergie du champ incident n'est pas uniforme au contact de la cible [LIU et collab. \[2016\]](#). Cette situation intervient dans le cas d'une illumination en champ proche. Cette amélioration est apportée grâce à une compensation énergétique qui, en combinaison avec l'exploitation des sources d'ordre élevé, améliore légèrement le résultat des imageries. L'apport de cette amélioration a été quantifié avec l'indicateur de Jaccard et les détails de l'étude sont présentés dans le chapitre 3 de la thèse de X. Liu [LIU \[2017\]](#).

Nous avons considéré par la suite les données de diffraction issues des simulations 3D avec des antennes réelles (vivaldi opérant entre 1 et 5 GHz). Dans cette configuration GPR (9 antennes émettrices / réceptrices distribuées régulièrement le long d'un segment en face la scène GPR), la méthode du LSM doit subir quelques modifications dans sa formulation pour aboutir à une résolution satisfaisante du problème inverse. Il faut irrémédiablement prendre en compte les diagrammes de rayonnement et définir le système à résoudre à partir des paramètres S mesurés au niveau des ports des antennes. Aussi, il faut reformuler l'écriture du LSM en exploitant la fonction dyadique de Green du problème (pour les différentes antennes). Cela signifie qu'il faut utiliser l'expression du champ incident réel issu des antennes dans la zone d'investigation [HAYNES et MOGHADDAM \[2012\]](#). Le développement est détaillé dans le chapitre 3 de la thèse de X.

Liu LIU [2017] (le lien url pour accéder à la thèse est disponible dans les références).

Enfin, nous avons étudié les avantages de la méthode LSM par rapport aux techniques classiques de migration. Nous en avons étudié trois : Kirchhoff, stripmap SAR SOUMEKH [1999] et F-K STOLT [1978]. Nous avons validé ces techniques numériquement et expérimentalement dans LIU et collab. [2017] et LIU et collab. [2017]. Nous avons constaté que les techniques de migration ont des difficultés à localiser correctement des cibles diélectriques (objets pénétrables de taille non-ponctuelle). L'image du signal focalisé affiche un spot légèrement plus bas que la position réelle de la cible. Néanmoins, les objets métalliques sont localisés à leur juste position. La LSM détecte correctement la position des cibles métalliques et diélectriques grâce au caractère multi-statique du champ diffracté. Dans mes activités de recherche autour de la résolution du problème inverse, J'envisage de construire une bibliothèque d'outils destinée à développer différentes applications futures. Ce travail de consolidation des connaissances se poursuit dans le cadre de la thèse de A. Aboudourib sur l'imagerie des systèmes racinaires avec des outils GPR.

Publications représentatives sur l'analyse du problème inverse en GPR

- [CI.26] X. Liu, M. Serhir, M. Lambert, L. Pichon, « Imaging Comparison of GPR Based on Multi-and Bi-static Configurations », *22nd International Workshop on Electromagnetic Non-Destructive Evaluation*, Sep 2017, Saclay, France
- [CI.20] X. Liu, M. Serhir, A. Kameni, M. Lambert, L. Pichon, « Enhanced Linear Sampling Method for GPR Imaging », *International Workshop on New Computational Methods for Inverse Problems*, May 2016, Cachan, France
- [CI.19] X. Liu, M. Serhir, A. Kameni, L. Pichon, M. Lambert, « GPR imaging via multi-frequency Linear Sampling Method », *10th International Symposium on Electric and Magnetic Fields*, Apr 2016, Lyon, France

Enhanced Linear Sampling Method for GPR Imaging

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Abstract. In this paper we consider the electromagnetic inverse scattering problem for GPR application. The reconstruction of the shape of buried dielectric objects is obtained using a non-iterative inverse technique - Linear Sampling Method (LSM). The applicability of LSM for GPR imaging is discussed, its performance with full and limited aperture is explored in a simple 2D free space configuration in a first step. The use of non-ideal sources and the possible influence of the antenna radiation pattern onto the results are considered. An enhanced approach is proposed to improve the robustness of the LSM with non-ideal sources.

1. Introduction

The Ground Penetrating Radar (GPR) technique is constantly used in geophysics and civil engineering to provide effective and high-precision imaging of underground structures and for buried objects detection. It still needs development as shown by the recent (2013) start of the COST Action TU1208 entitled “Civil Engineering Application of Ground Penetrating Radar” [1–3]; Therefore, it is very interesting to improve the GPR performance for estimating the target location, size and shape. The migration techniques are the traditional GPR imaging approaches. They are able to derive relatively high-resolution images but their capability of shape reconstruction is limited. Iterative inverse scattering methods based on the solution of a nonlinear minimization problem are commonly applied because of high accuracy. However, this kind of approaches requires forward solver for an iterative minimization scheme, which is extremely complex and computationally expensive for GPR application. In addition, the iterative scheme may suffer from the presence of false solution (local minima) affecting the reconstruction validity.

Other kind of inverse scattering approaches are qualitative methods. The Linear Sampling Method (LSM) is considered in this framework. LSM can reconstruct the geometrical features of both dielectric and metallic targets with few a priori information [4, 5]. Furthermore, LSM is very effective in terms of computational requirement, it allows of realizing real time reconstructions, which is very attractive in GPR surveys [6, 7]. Several successful examples are available in the literature concerning the GPR imaging via LSM by supposing the antennas as perfect point sources, so that the antenna pattern is not considered.

In this work, the feasibility and reliability of LSM for GPR application are firstly investigated in 2D situation. The synthetic multi-frequency and multi-static data is obtained by 2D Method of Moments (MoM). The feasibility of using LSM for GPR application is studied for a 2D electromagnetic problem. The non-ideal sources and the influence of the antenna radiation

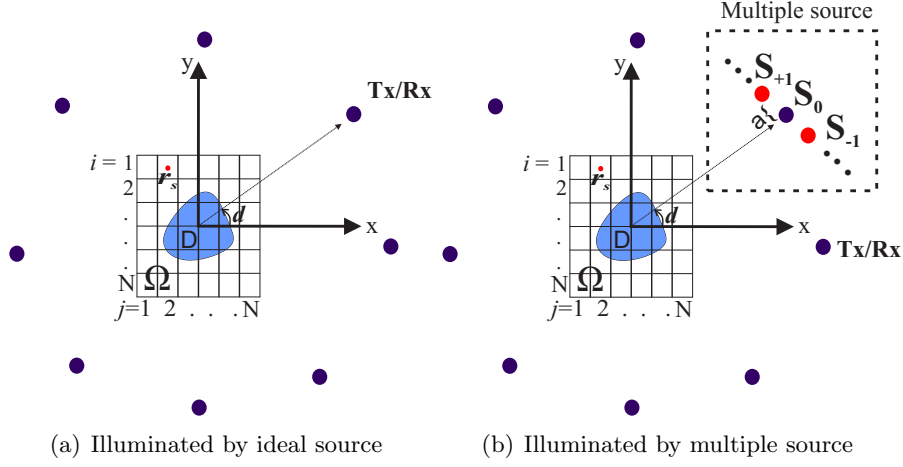


Figure 1. Scattering problem configuration

pattern onto the results are discussed. An enhanced algorithm is proposed to improve the LSM performance with non-ideal sources.

2. The linear sampling method

In this section, we briefly introduce the LSM for resolving the time harmonic inverse electromagnetic obstacle scattering problem.

2.1. Review and formation of LSM

Let us consider a dielectric obstacle $D \in \mathcal{R}^2$ with \mathcal{C}^2 -continuous boundary ∂D as shown in Fig. 1(a). For a given incident electromagnetic wave field u^i injected by a transmitter T_x , a scattering field u^s will be caused by the presence of D . The total field is defined by $u = u^i + u^s$, which should satisfy the Helmholtz system [8].

LSM is based on solving the far field equation for the unknown $g_{\vec{r}_s} \in \mathcal{L}^2(\Omega)$ at each sampling position \vec{r}_s [4, 5]:

$$\int_{\Omega} u_{\infty}(\hat{r}, \vec{d}) g_{\vec{r}_s}(\vec{d}) ds(\vec{d}) = \Phi_{\infty}(\hat{r}, \vec{r}_s) \quad (1)$$

where \hat{r} is unit direction vector, \vec{d} denote the direction of receiver (R_x in Fig. 1(a)), u_{∞} present the far field pattern of scattered field u^s , and Φ_{∞} is the far field of a point source,

$$\Phi_{\infty}(\hat{r}, \vec{r}_s) = \frac{e^{i\pi/4}}{\sqrt{8\pi k}} e^{-ik\hat{r} \cdot \vec{r}_s} \quad (2)$$

The \mathcal{L}^2 -norm of $g_{\vec{r}_s}$ becomes unbounded if the sampling point is outside the scatter D . Therefore the indicator function of LSM can be defined by

$$X(\vec{r}_s) = \|g_{\vec{r}_s}\|_{\mathcal{L}^2(\Omega)}^{-1} \quad (3)$$

We now approximate equation 1 using the trapezoidal rule to approximate the integral such that

$$\mathbf{A} \cdot \mathbf{g}_{\vec{r}_s} = \mathbf{b}_{\vec{r}_s} \quad (4)$$

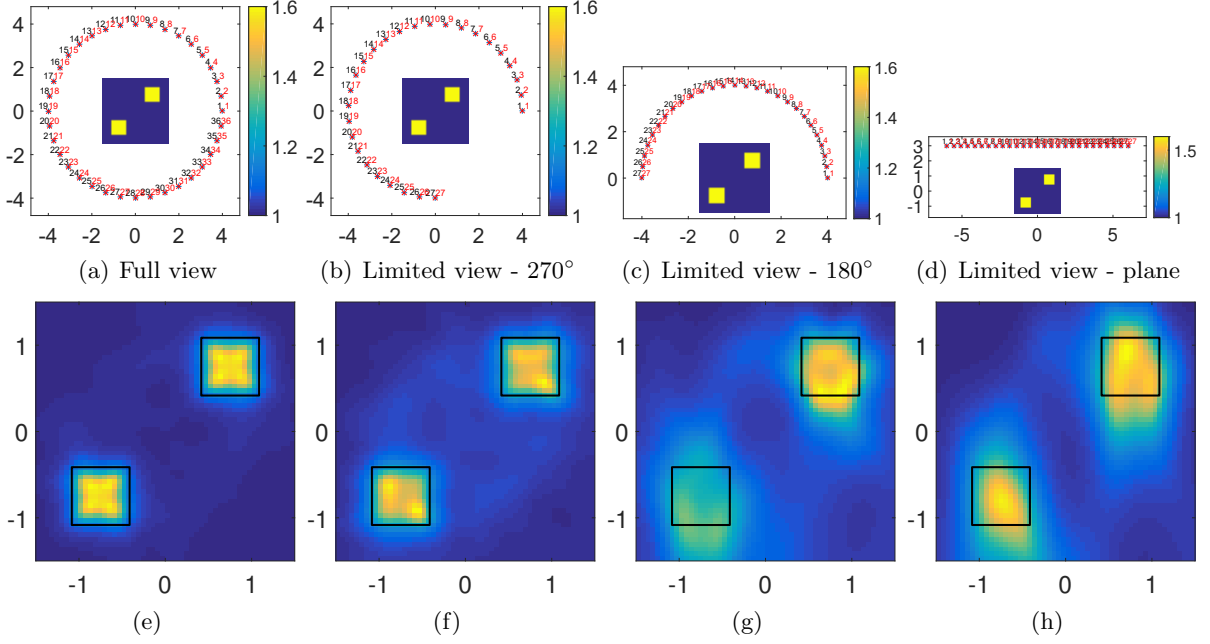


Figure 2. Configuration of inverse scattering problem (a) full view $M = N = 36$, (b) 270° view $M = N = 27$, (c) 180° view $M = N = 27$, (d) plane view $M = N = 27$, and (e) (f) (g) (h) the respective imaging results of four different configurations

where the \mathbf{A} is the $M \times N$ multi-static matrix obtained by GPR measurement or numerical calculation, $\mathbf{b}_{\vec{r}_s}$ is given by $\frac{N}{2\pi} \phi(\vec{d}_l, \vec{r}_s)$ using equation 2, $1 \leq l \leq N$, M and N are respectively the number of transmitters and receivers.

The equation 4 can be actually resolved using Tikhonov regularization via the singular value decomposition of \mathbf{A} [9]. Note $\mathbf{A} = \mathbf{U}\mathbf{S}\mathbf{V}^*$, and let σ denote its singular value, the $\mathbf{g}_{\vec{r}_s}$ can be finally given by

$$\|\mathbf{g}_{\vec{r}_s}\|^2 = \sum_{p=1}^N \left(\frac{\sigma_p}{\alpha_{\vec{r}_s} + \sigma_p^2} \right)^2 |(\mathbf{U}^* \mathbf{b}_{\vec{r}_s})_p|^2 \quad (5)$$

with regularization parameter $\alpha_{\vec{r}_s}$ determined by Morozov principle for each sampling point [9].

2.2. Numerical results of LSM

A 2D example with two dielectric squares is firstly considered in Fig 2(a). The blue area is the grid region Ω (3×3 m), the emitters/receptors are placed over a circle with radius = 4 m and $M = N$. The edge length of target is 0.6 m and its relative permittivity is 1.6, the coordinates of their centers are respectively (0.75 m, 0.75 m) and (-0.75 m, -0.75 m). The emitters are ideal electric line sources. Gaussian noise is added to the data acquired by Method of Moments (MoM) for SNR = 10 dB over 21 operating frequencies from 240 MHz to 360 MHz. The imaging result in Fig 2(e) provide extremely good reconstruction capability of LSM, while a large number of antennas is considered with full view configuration. The LSM performance deteriorate as long as the aperture is reduced (Fig 2(f) and Fig 2(g)). A GPR surveys like configuration (plane view form) is given in Fig 2(d), the imaging result in Fig 2(h) confirms the feasibility of LSM for GPR imaging problem.

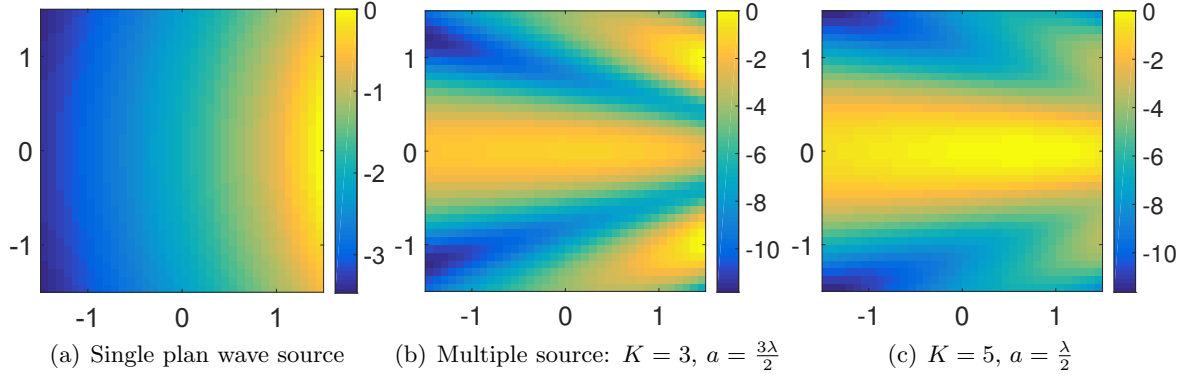


Figure 3. Illumination (normalized amplitude of incident electric field E_{inc} in dB on Ω) of single source and multiple sources

3. The enforced LSM for GPR

In the classic LSM introduced in section 2, the transmitters are supposed to be ideal sources. But in most of the GPR applications, the antenna (transmitter) is placed near the ground, hence the antenna pattern should be considered.

3.1. Synthetic antenna parttern in 2D

In 2D case, A single electric line source is used to illuminate the investigative domain Ω with classic LSM (Fig. 1(a)). The antenna pattern can be simulated by using multiple source as shown in Fig. 1(b). The multiple source is composed by certain elementary ideal sources, which are placed uniformly on a vertical line of \vec{d} , separated by the distance noted a .

Let K denote the number of elementary ideal sources, the normalized amplitude (in dB) of electric fields (TM mode) introduced by a single source and multiple sources onto Ω are shown in Fig. 3. The investigate domain Ω is 3×3 m, the direction of source $\vec{d} = (-1, 0)$, the distance of transmitter to origin is 5m and the signal frequency is 3MHz.

Comparing with the electric field injected by single source (Fig. 3(a)), the 2D synthetic antenna pattern can be obtained by the multiple source illumination (Fig. 3(b), Fig. 3(c)).

3.2. Enforced LSM

In consequence of the antenna pattern, the incident electric field intensity is not proportional to the distance between illuminated position and source. Hence, Ω is not uniformly illuminated while the multiple source rotate around it. A compensation method is proposed to offset the influence of antenna pattern by

$$X_E(\vec{r}_s) = C(\vec{r}_s) \cdot \|g_{\vec{r}_s}\|_{\mathcal{L}^2(\Omega)}^{-1} \quad (6)$$

where X_E denote the enforced LSM indicator function, and the correction function $C(\vec{r}_s)$ is defined by

$$C(\vec{r}_s) = \left(\sum_{p=1}^N \|E_{inc}(p, \vec{r}_s)\| \right)^{-1} \quad (7)$$

One result is shown in Fig. 4. The grid region Ω is 3×3 m, the emitters/receptors are placed over a circle with radius = 5 m and $M = N = 36$. The size of target is $0.5 \text{ m} \times 1 \text{ m}$ and its relative permittivity is 1.6, the coordinates of its center is $(-0.3 \text{ m}, 0.75 \text{ m})$. The emitters are

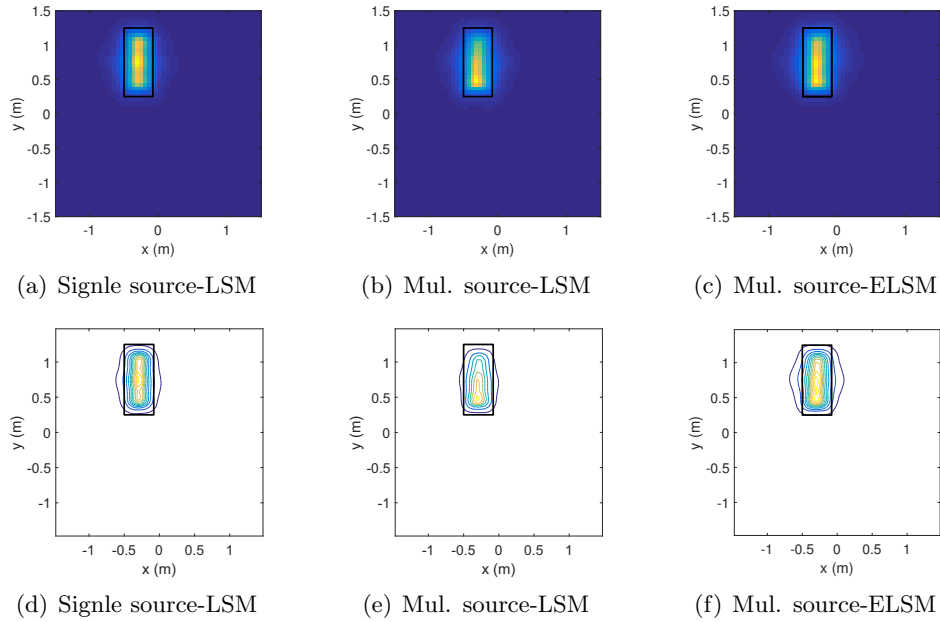


Figure 4. (a) LSM imaging result for single source illumination, (b) LSM imaging result for multiple source illumination, $K = 3$ and $a = \frac{3\lambda}{2}$, (c) Enforced LSM imaging result for multiple source illumination, $K = 3$ and $a = \frac{3\lambda}{2}$, (e) (f) (g) the respective contour plot displays of (a)(b)(c)

ideal electric line sources and multiple source with $K = 3$ and $a = \frac{3\lambda}{2}$. Gaussian noise is added to the data acquired by Method of Moments (MoM) for SNR = 10 dB and operating frequency is 3 MHz.

The deterioration of LSM performance with the presence of synthetic antenna pattern is observed by comparing (Fig. 4(a),4(d)) and (Fig. 4(b),4(e)). The proposed enforced LSM with can prove a better imaging quality by compensating the nonuniform incident field (Fig. 4(c),4(f)).

4. Conclusion

In this framework, the feasibility of using LSM for GPR application is studied for a 2D electromagnetic problem. Preliminary results are presented in a simplified configuration. The use of non-ideal sources and the possible influence of the antenna radiation pattern onto the results is discussed. A compensation method is proposed to correct this influence. The extension of the approach to the 3D case will be dealt with and the confrontation with laboratory controlled experimentation is under investigation.

Acknowledgments

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4.5 Du côté de l'antenne ultra-large-bande pour le GPR

4.5.1 Antenne Bowtie

La conception et l'optimisation d'antennes ultra-large bande dédiées aux radars de sol font partie de mes activités de recherche. Mes premières investigations étaient en lien avec les antennes Bowtie et leur utilisation dans le radar de sol. Ces antennes équipent la majorité des radars du commerce (Mala, GSSI, EDS, etc...). Je vous renvoie vers [PAJEWSKI et collab. \[2015\]](#) pour une description plus exhaustive des antennes utilisées par les constructeurs des radars de sol.

Les antennes Bowtie sont exploitées dans le GPR car leur structure plate permet d'accroître le couplage avec le sol et assurer un transfert optimal de l'énergie électromagnétique. Ce couplage avec le sol dépend de la permittivité de ce dernier. En effet, les antennes Bowtie appartiennent à la famille des dipôles biconiques dont l'impédance d'entrée est corrélée à l'ouverture angulaire des cônes. Afin de modifier l'impédance d'entrée de l'antenne, Lestari dans [LESTARI et collab. \[2004a\]](#) a proposé une structure modulaire que l'on peut modifier en fonction du milieu ausculté tout en conservant une bonne adaptation d'impédance.

La particularité dans le cahier des charges des antennes pulsées dédiées au GPR est l'absence de « ringing » [WIESBECK et collab. \[2009\]](#). Le « ringing » est ce signal caractéristique des antennes dispersives qui sont le siège de réflexions multiples provoquant ainsi un signal résiduel qui rallonge la durée de la réponse impulsionnelle de l'antenne. Pour y remédier, les antennes sont chargées résistivement à leur extrémité [LESTARI et collab. \[2004b\]](#) [WU et KING \[1965\]](#). Cela permet d'élargir leur bande passante vers les basses fréquences même si cela impacte directement leur efficacité.

L'antenne Bowtie, à l'image des antennes dipolaires large bande, est pénalisée par l'apparition des lobes secondaires au-delà d'une certaine fréquence [BALANIS \[2016\]](#). Dans ce cas, l'énergie électromagnétique n'est plus focalisée en face de l'antenne et cela est préjudiciable aux applications GPR. Pour corriger l'apparition des lobes secondaires de l'antenne Bowtie, j'ai utilisé une cavité réfléchissante optimale afin de re-focaliser le diagramme de rayonnement pour les fréquences en question. Cela améliore la directivité de l'antenne pour une meilleure illumination de la scène sous test. Ensuite, j'ai plié l'antenne Bowtie pour aboutir à une Bowtie sous la forme d'un V. Cette idée est inspirée à la fois par les cornets TEM et les dipôles en V [KANGWOOK KIM et SCOTT \[2005\]](#). Les caractéristiques de l'antenne optimisée sont confirmées par des résultats de simulations et de mesures dans [SERHIR et LESSELIER \[2018\]](#) (joint avec ce rapport). J'ai appelé cette antenne « reflector backed folded bowtie antenna ». Cette antenne présente un diagramme de rayonnement unidirectionnel sans lobes secondaires dans la bande fréquentielle [1 GHz - 3 GHz], son gain est compris entre 5 et 12 dBi. J'ai comparé les propriétés de cette antenne Bowtie avec ceux d'un cornet ridgé et d'une antenna vivaldi et j'ai mesuré leur champ rayonné dans le sable à 10 cm de profondeur. Le sol ayant le comportement d'un filtre passe-bas, le champ mesuré dans le sol est maximal à des fréquences plus basses qu'en espace libre.

4.5.2 Cornet Transverse ElectroMagnétique

Dans un deuxième volet, mon intérêt s'est porté sur l'optimisation d'une antenne cornet TEM en partenariat avec Rincent-BTP. Ce concepteur/utilisateur de radar souhaitait optimiser une antenne cornet Transverse Electro-Magnétique (TEM) pour un radar bi-statique off-ground destiné à la caractérisation des couches supérieures de la chaussée (radar à 40 cm du sol). Le prototype final retenu du cornet TEM a la forme d'une flamme similaire à celle que l'on retrouve dans [TAN et collab. \[2012\]](#). L'antenne présente une bande passante allant de 0.3 GHz à 4 GHz.

Le choix d'un cornet TEM est motivé par sa large bande passante, sa directivité, son caractère peu dispersif et une polarisation linéaire [SHLAGER et collab. \[1996\]](#) [LEE et SMITH \[2004\]](#) [THEODOROU et collab. \[1981\]](#). Même si les notions que je viens de citer sont liées à une caractérisation fréquentielle, les caractéristiques importantes en GPR sont : une réponse impulsionnelle courte et la capacité de l'antenne à rayonner un pulse propre sans « ringing » [SCHEERS et collab. \[2000\]](#) [FARR et BAUM \[1998\]](#).

Les plaques métalliques du cornet TEM sont imprimées sur un substrat FR4. Leur forme géométrique évolue progressivement pour assurer une transition graduelle de l'impédance caractéristique du mode TEM. L'impédance caractéristique évolue de 50 Ω au niveau de l'alimentation à 377 Ω au niveau de l'ouverture. Aux extrémités des parties métalliques de l'antenne, j'ai placé deux plaques résistives pour atténuer les réflexions résiduelles au niveau de l'ouverture [KANDA \[1982\]](#). Cela améliore l'adaptation en basse fréquence (au détriment de l'efficacité) et raccourcit la réponse temporelle de l'antenne.

En dernier lieu, j'ai placé, une structure « mille-feuille » à l'intérieur du cornet TEM pour améliorer sa directivité et par conséquent réduire sa sensibilité aux perturbations extérieures. La structure « mille-feuille » est constituée de plaques de substrat FR4 régulièrement espacées qui jouent le rôle d'une lentille. Cette idée est similaire à celle présentée dans [YAROVY et collab. \[2002\]](#) [AURAND \[1997\]](#) lors de la présentation du cornet TEM rempli d'un matériau diélectrique. Cela permet de réduire la taille de l'antenne et de raccourcir sa réponse impulsionnelle. J'ai valorisé le travail effectué sur l'antenne TEM avec Rincent-BTP dans le cadre d'un brevet [SERHIR M. \[2016\]](#) (lien vers le brevet en bibliographie).

Après ce partenariat, Rincent-BTP utilise désormais dans son nouveau radar de sol un analyseur de réseau. Les mesures sont faites en fréquentiel et le traitement temporel exploite une transformée de Fourier des signaux mesurés. Le rapport signal à bruit est nettement meilleur que lors de l'utilisation des signaux dans le domaine temporel. D'ailleurs, les travaux de [SEYFRIED et collab. \[2014\]](#) et [KAISER et SCHAFER \[1980\]](#) proposent des techniques de fenêtrage (alternative au Hanning/Hamming) à utiliser avant d'appliquer la transformée de Fourier inverse.

Publications représentatives des travaux sur les antennes pour le GPR

[RI.20] M. Serhir, D. Lesselier, « Wideband reflector-backed folded bowtie antenna for ground penetrating radar », *IEEE Transactions on Antennas and Propagation*, vol. 66, no. 3, pp. 1056-1063, 2018

Ce document décrit en détail les caractéristiques de l'antenne Bowtie pliée, les résultats des mesures du diagramme de rayonnement dans le sable et les comparaisons entre les simulations et les mesures. Ce document est joint avec ce rapport.

[B.1] M. Serhir, S. Pellevraut, « Antenne pour radar, structure d'antenne et radar », *Europe FR3031396*, 2016

Le descriptif détaillé du brevet est accessible via le lien de bibliographie. Les caractéristiques géométriques de l'antenne et de la structure « mille-feuille » sont décrites.

Wideband Reflector-Backed Folded Bowtie Antenna for Ground Penetrating Radar

Mohammed Serhir¹ and Dominique Lesselier, *Senior Member, IEEE*

Abstract—A loaded ultrawideband unidirectional folded triangular bowtie antenna is proposed for ground penetrating radar (GPR) applications covering the (0.5–3 GHz) frequency range. The behavior of the proposed structure has been modeled and optimized using full-wave numerical simulations and measurements, in far-field and near-field settings in both time and frequency domains in free space and sandy medium. In particular, an impedance bandwidth of 100% with standing-wave ratio ≤ 2 , a broadside gain between 5 and 12 dBi, for total antenna size of $1.4\lambda_m \times \lambda_m$, λ_m free space wavelength at the center frequency, are achieved as confirmed by laboratory-controlled measurements in free space and in GPR settings.

Index Terms—Antenna design, broadband time- and frequency-domain measurements, folded bowtie antenna (FBA), full-wave modeling, ground penetrating radar (GPR).

I. INTRODUCTION

GROUND penetrating radar (GPR) and more generally short-range radar systems are increasingly developed in view of localization and identification of utilities and other targets buried in soil as well as through-wall detection under various guises, refer to [1] and [2] for a recognized nonexhaustive illustration. Their performances in the field are mostly linked with the ultrawideband (UWB) transmitting and receiving antennas that are employed. In particular, the central frequency and the bandwidth of the generated and collected impulse signals are key factors of whatever image quality that can be achieved by such radar systems in GPR situations. As is well known, the low operation frequencies are needed in order to evidence deeply buried targets and the high ones are tailored to yield high resolution for the detection of shallowly buried targets. In addition, the constraints over the GPR antennas are in the time domain, i.e., a short impulse response and absence of ringing [3], [4]. Consequently, UWB GPR antennas have to satisfy over a broad frequency band to the following constraints: reflection coefficient smaller than -10 dB,

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linear phase characteristics (low dispersion), low boresight gain variation over the frequency band, focused radiated field over a narrow solid angle, stable antenna near-field (NF) foot print extent versus frequency [5].

Here, we propose an optimized bowtie antenna for GPR on or near ground. Bowtie antennas have omni-directional radiation pattern and are linearly polarized [6]. This symmetric antenna necessitates a balun to ensure a symmetric excitation and a symmetric radiation pattern; indeed, bowtie antennas (belonging to the ample family of dipole antennas) have the peculiarity of an input port impedance dependent upon the flare angle of the bowtie. To transform this antenna into a unidirectional one, a cavity can be used as in [7]–[9] yet with cavities that are large in comparison with the bowtie antenna sizes. Thus, these backing antenna solutions are not low profile and have not been studied to assess NF characteristics, which, however, is a key element for the GPR community.

For GPR [1], the back antenna radiated energy is not of interest for ground illumination. Classically, absorbers are put inside the cavity to cancel or at least to attenuate the damaging effect of antenna-cavity coupling [10]. Here, we propose an optimized reflector geometry that can constructively bring “back-radiation” to “front-radiation” with minimum antenna mismatch due to antenna-reflector coupling, and keep S_{11} less than -10 dB for a frequency range of same width as the one of the no reflector antenna started from (ideally).

Like with a dipole, the radiation pattern of a bowtie antenna is split, since sidelobes arise when the size of the antenna reaches 1.5 wavelength (λ) at the highest frequencies [6] of the operating bandwidth. The maximum radiated field then is not anymore observed in the boresight direction. So, one will be exploiting the antenna backing reflector in order to correct this phenomenon as much as possible by refocusing the radiation pattern at the high frequencies. In addition, besides using the reflector, bending the bowtie antenna will enhance the directivity over a broad frequency band. This idea is inspired by TEM horn [11] and V-dipole antennas.

The above results into the final prototype termed as reflector-backed folded bowtie antenna (RB FBA) in the following. It achieves a stable radiation pattern in comparison with the RB planar bowtie antenna (RB PBA), and its final design ensures an impedance bandwidth of 100% ($S_{11} \leq -10$ dB), and a boresight gain between 5 and 12 dBi within the (1–3 GHz) frequency band of interest.

Notice that for the bandwidth evaluation, a restrictive definition is being put to the fore. The bandwidth is stated as the frequency interval, wherein $S_{11} \leq -10$ dB and the maximum

radiated field is at the boresight direction (in front of the antenna). Even if S_{11} can be lower than -10 dB for a larger bandwidth, the frequency interval where the maximum energy is radiated in the boresight direction is the only one considered. This definition is directly linked with GPR applications since indeed energy must be focused within a localized area in front of the antenna.

The present contribution is organized as follows. In Section II, the FBA S_{11} and far-field characteristics are introduced in comparison with those of a PBA. In Section III, the S_{11} and far-field radiation pattern properties of the FBA are presented with due attention to performance enhancement enabled by the proposed backing reflectors and the resistive loads added to control the antenna transient response ringing. In Section IV, the free space transient and frequency NF (footprint) characteristics of the presented antenna models are discussed and compared with Vivaldi and dual ridged horn antennas. Radiated NF of the antennas at 10 cm depth in a sandy medium is focused onto here. This NF information, as underlined already before, is indeed the most useful for the GPR antenna community. In conclusion, the performances of the proposed antenna for GPR applications are summed up.

II. FOLDED BOWTIE ANTENNA

The design of a PBA optimized for GPR applications in the (0.5–3 GHz) frequency band is considered now. The PBA geometrical details are found in Fig. 1(a). The antenna is printed over a 1.6 mm thick FR-4 substrate ($\epsilon_r = 4.2$, $\tan \delta = 0.01$). The antenna is fed symmetrically by the balun. Two coaxial cables are soldered together in order to get a 100Ω input impedance [12], [13]. To realize a symmetric excitation, they are plugged into a wideband 180° hybrid coupler to excite the two parts of the bowtie antenna with a phase shift of 180° . This planar antenna (antenna + coaxial cables) is optimized using the transient solver of CST microwave studio [14].

To carry out the measurements—those and all others discussed herein have been performed in the anechoic chamber of CentraleSupélec [15]—the two-port vector network analyzer Agilent ENA E5071 has been used to measure the S_{11m} and S_{21m} , port 1 being plugged to the first antenna coaxial cable and port 2 to the second one. As seen in Fig. 2, the antenna reflection coefficient calculated as $S_{11} = S_{11m} - S_{21m}$ is lower than -10 dB in the chosen frequency band (0.5–3 GHz) save between 1.35 and 1.75 GHz, where S_{11} is about -7 dB.

As seen in Fig. 3, the measured far-field radiation patterns are displayed at principal plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. The far-field pattern of the PBA [Fig. 3(a)] shows a split at 1.6 GHz due to sidelobes appearance. In addition, the radiation pattern of Fig. 3(a) for the plane cut $\theta = 90^\circ$ is symmetric along the ϕ -axis, and the antenna front-radiation is identical with the back one ($\phi = 0^\circ$ and $\phi = 180^\circ$). The PBA directivity drops around 1.6 GHz and the radiated energy in front of the antenna ($\theta = 90^\circ$) is not predominant. This split is clearly visible for the plane cut $\phi = 0^\circ$. As a conclusion and based on our definition of the bandwidth, this antenna provides 91% bandwidth covering the frequency band (0.5–1.35 GHz).

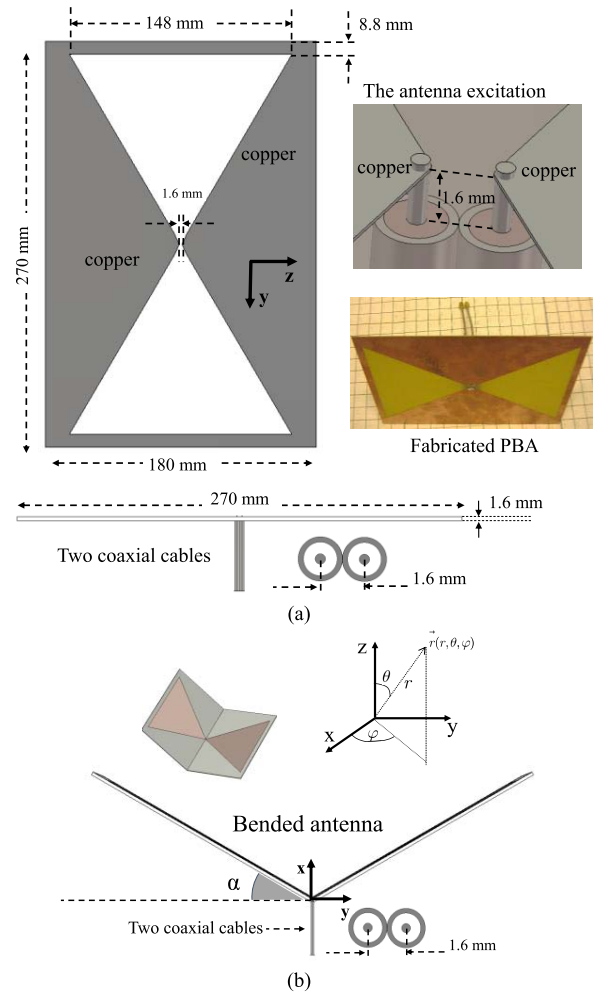


Fig. 1. (a) Sketch of the antenna geometry and its symmetric excitation. (b) Bending antenna principle and α definition.

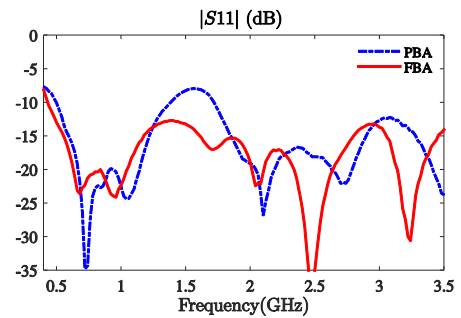


Fig. 2. Comparison between measured reflection coefficient S_{11} as a function of frequency for PBA and FBA antennas.

In order to extend the antenna bandwidth, to bend it as shown in Fig. 1(b), enables to enlarge the frequency range where the front radiation pattern of the antenna is predominant. The aim is to correct the radiation pattern split that happens for frequencies beyond 1.6 GHz. In Fig. 4(a), the S_{11} results of bended bowtie antenna for angles $\alpha = 0^\circ, 10^\circ, 20^\circ, 30^\circ, 40^\circ$, α bending angle as defined in Fig. 1(b), are displayed. For these simulations, one used the discrete port provided

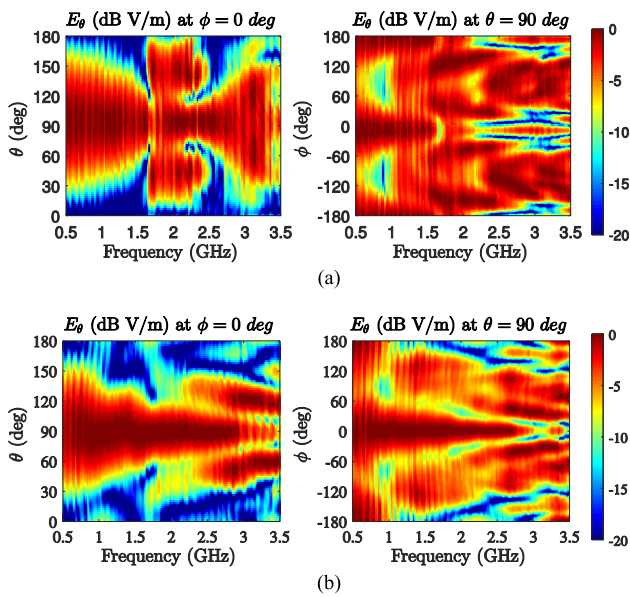


Fig. 3. Co-polar far-field measured radiation patterns in plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. (a) PBA. (b) FBA.

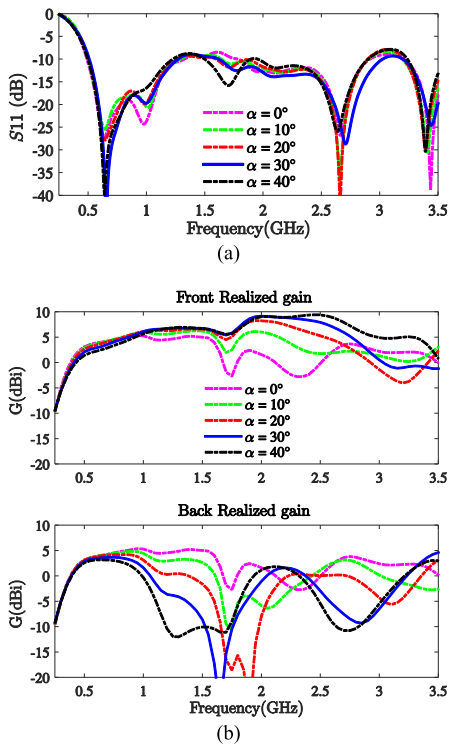


Fig. 4. (a) CST simulation results of S_{11} parameter as a function of α . (b) CST simulation results of realized gain as a function of the frequency for different α values (top) in front of the antenna and (bottom) in the back of the antenna.

by CST (input impedance 100Ω). It is seen that the angle α does not significantly change the S_{11} results. Yet, the S_{11} curve associated with $\alpha = 30^\circ$ exhibits slightly better matching within the frequency bandwidth (0.5–3 GHz).

Gains achieved in the front and in the back of the antenna are shown as a function of angle α in Fig. 4(b). The antenna

back gain decreases as α increases, i.e., bending the antenna contributes to the radiated power being focused into the boresight direction. Increasing α improves the gain at higher frequencies, while degrading it for the lower part of the bandwidth. For $\alpha = 0^\circ$, the PBA radiates the same power in front and in back as seen in Fig. 4(b). One chose the solution corresponding to $\alpha = 30^\circ$ as a good compromise between gain and antenna impedance matching. This antenna acronym is FBA. The measured S_{11} is compared with PBA one in Fig. 2, showing $S_{11} \leq -10$ dB for frequencies between 0.5 and 3 GHz.

In Fig. 3(b), measured radiated far-field of the FBA (once fabricated) are displayed in principal plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. With $\alpha = 30^\circ$, the antenna directivity is enhanced in comparison with $\alpha = 0^\circ$ [Fig. 3(a)] for plane cut $\phi = 0^\circ$. The radiation pattern in front of the FBA ($-60^\circ \leq \phi \leq +60^\circ$) is more directive than with the PBA. The frequency band where the field is maximally radiated in the boresight direction is extended to 2.5 GHz and $S_{11} \leq -10$ dB in frequency band (0.5–3 GHz), as seen in Fig. 2. Consequently, the resulting bandwidth is 133% for the FBA from 0.5 to 2.5 GHz.

III. REFLECTOR-BACKED FOLDED BOWTIE ANTENNA

The far-field performances with backing reflectors are studied in this section. As shown in Fig. 5(a) and (b), the PBA and FBA are backed with copper reflectors. Their shapes have been finalized based on an optimization process using CST MWS simulation software. The measured reflection coefficients S_{11} of both fabricated RB antennas (RB PBA and RB FBA) are displayed in Fig. 5(c). As it can be seen, both antennas yield $S_{11} \leq -10$ dB in the frequency band (1–3 GHz). The RB FBA shows a small oscillation around -10 dB between 1 and 1.3 GHz. The mechanical robustness of the antennas is provided by wood pieces placed at the lateral parts of the antennas.

To evaluate the RB antennas bandwidth, measured far-fields are displayed in Fig. 6 at principal plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. From the 2-D plots shown in Fig. 6, it is noticeable that the back radiation is reduced and the far-field is maximally radiated within the solid angle defined by $45^\circ \leq \theta \leq 135^\circ$ and $-60^\circ \leq \phi \leq 60^\circ$. The directivity of both antennas is increased especially at plane cut $\theta = 90^\circ$. For the plane cut $\phi = 0^\circ$, the RB PBA directivity in the boresight direction drops at 2.5 GHz, as shown in Fig. 6(a), in comparison with the one of the RB FBAs, which is slightly weak at 3 GHz. Based on Fig. 6(a), the field is not predominantly radiated in the boresight direction (sidelobes of comparable level) for the RB PBA beyond 2 GHz. Consequently, the RB PBA has 66% bandwidth (1–2 GHz).

The radiation pattern sidelobes of the RB FBA become predominant beyond 3 GHz. This antenna has 100% bandwidth (1–3 GHz). Measured boresight gains of the antennas are shown in Fig. 7. It is seen that bending the bowtie antenna corrects the gain drop at 1.6 GHz. In addition, the role of the optimized backing reflectors is highlighted since the constructive coupling between the reflectors and the antennas

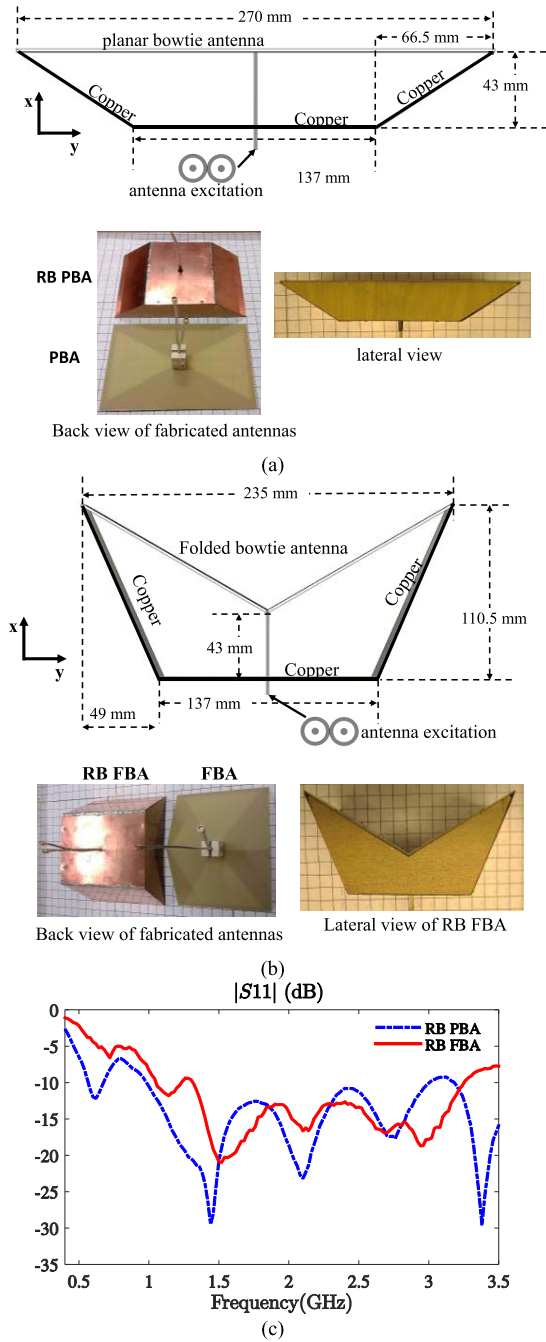


Fig. 5. (a) RB PBA geometric details and photography of the fabricated antennas with/without the reflector. (b) RB FBA geometric description and photography of realized antennas with/without the backing reflector. (c) Comparison of measured S_{11} of RB FBA and RB PBA as a function of frequency.

achieves a significant gain increase (more than 3 dB) between 1.5 and 2.25 GHz. The RB FBA features a realized gain between 5 and 12 dBi.

To complete the analysis presented in this paper, resistive loads have been added to shorten the transient response of RB PBA and RB FBA. To do so, four lumped resistors of 150Ω are soldered at the ends of the bowtie as shown in the photography in Fig. 8(a). The value of the lumped

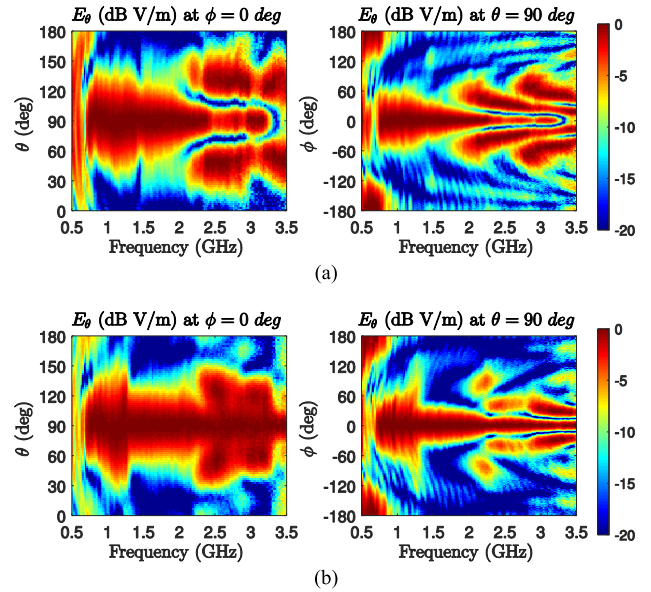


Fig. 6. Co-polar far-field measured radiation patterns in plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. (a) RB PBA. (b) RB FBA.

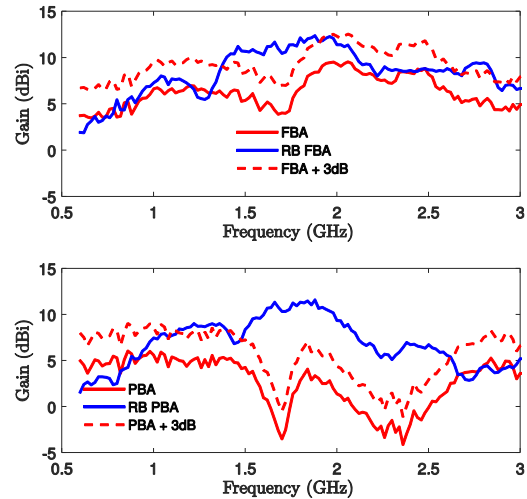


Fig. 7. Measurement results of realized gains of PBA, FBA, RB PBA, and RB FBA in the boresight direction as a function of frequency.

resistors has been optimized using CST MWS. The chosen resistor value has to preserve the matching of the antenna. In Fig. 8(b), we present the S_{11} of loaded antennas. For both RB PBA and RB FBA, the reflections due to low frequencies have been attenuated and $S_{11} \leq -10$ dB for the frequency band (1–3 GHz), thanks to the lumped resistors. In Fig. 9, the measurement of the loaded antennas far-field radiation pattern is presented. In comparison with Fig. 6, it is noticed that the lumped resistors did not disturb the radiation pattern behavior in comparison with unloaded antennas.

IV. FOOTPRINT AND NEAR-FIELD CHARACTERISTICS OF THE PROPOSED ANTENNAS

In this section, the free space NF radiated by loaded RB PBA and RB FBA is investigated and compared with the

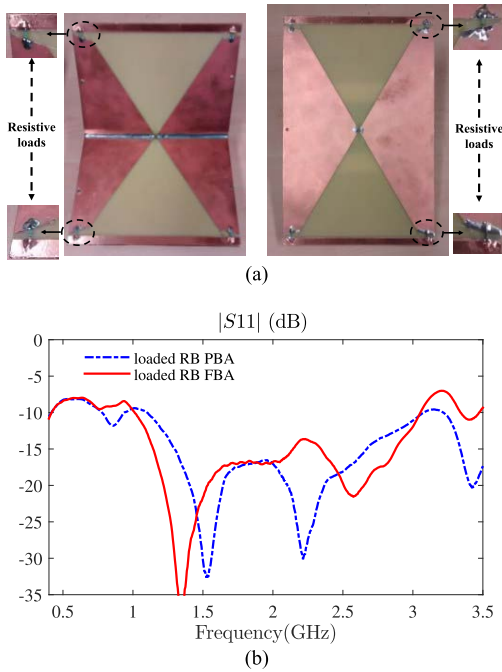


Fig. 8. (a) Photography of loaded RB PBA and RB FBA. (b) Comparison of measured S_{11} of loaded RB PBA and RB FBA as a function of frequency.

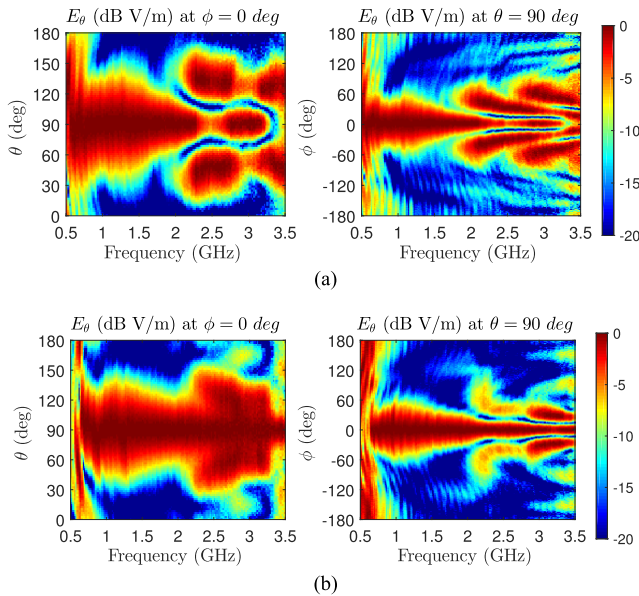


Fig. 9. Co-polar far-field measured radiation patterns in plane cuts $\theta = 90^\circ$ and $\phi = 0^\circ$ as a function of frequency. (a) Loaded RB PBA, (b) Loaded RB FBA.

NF radiated by Vivaldi (300 mm \times 300 mm) [17] and dual ridged horn antennas (SH600 of Microwave Vision company), two well-known UWB antennas. We aim at comparing the transient response of these antennas when excited with the second derivative of a Gaussian pulse. The excitation signal is a Mexican hat signal expressed as

$$e(t) = -\left(4\alpha^2 \left(\frac{t-t_0}{\tau}\right)^2 - 2\alpha\right) \exp\left(-\alpha \left(\frac{t-t_0}{\tau}\right)^2\right) \quad (1)$$

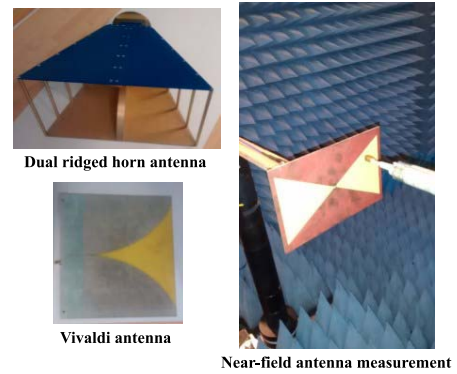


Fig. 10. Dual ridged horn antenna, Vivaldi antenna, and PBA measurement configuration to illustrate the free space NF antenna testing.

for $\tau = 2.10 \cdot 10^{-10}$ s, $\alpha = 0.85/\tau^2$ and $t_0 = 15\tau$ which covers the frequency band (0.5–3 GHz), where the spectrum maximum is at $f = 1.5$ GHz.

The Vivaldi, dual ridged horn, loaded RB PBA, and loaded RB FBA have been tested in the CentareleSupélec planar NF range to measure its radiated NF as illustrated in Fig. 10. The antenna under test and the scanning probe are placed inside the anechoic chamber of dimensions 4 m \times 5 m \times 3 m. We measure the co-polar NF components E_z while moving the probe respecting a regular spacing $\Delta y = \Delta z = 2.5$ cm over the planar surface $-50 \text{ cm} \leq y_{\text{meas}}, z_{\text{meas}} \leq 50 \text{ cm}$ at the measurement distance $x_{\text{meas}} = 10$ cm in front of the antenna. We have used the wideband probe “En-Probe EFS-105” [16]. This probe is dedicated to NF measurement covering the frequency band (3–5 GHz).

For these antennas (SH600, Vivaldi, RB PBA, RB FBA), NF measurements are carried out in the frequency band (0.5–3 GHz) using Agilent ENA 5071b vector network analyzer. Then the measured frequency-domain NF complex data (amplitude and phase) are Fourier transformed to set the transient response of the tested antennas when excited by the pulse expressed in (1). In Fig. 11, we show the measured time-domain NF at the plane cuts $y_{\text{meas}} = 0$ and $z_{\text{meas}} = 0$ of the four antennas. The loaded RB FBA presents a directive transient response in comparison with the RB PBA, Horn, and the Vivaldi antennas. The time-domain-radiated NF in the boresight direction is of comparable duration for the presented antennas as it is seen in Fig. 12, where the excitation pulse and its corresponding spectrum are also plotted. The loaded RB PBA still presents a residual ringing in its transient response. However, the directivity of the loaded RB FBA is confirmed in Fig. 13, where the footprint of the four antennas are determined based on their free space time-domain NF measurement data [18], [19]. This antenna is a good candidate (UWB, short transient response, and directivity) for GPR application using the Mexican hat excitation pulse expressed in (1).

Predicting the behavior of the loaded RB FBA and PBA antennas in their actual operating environment is essential. Here, to get some insight on that behavior, they are placed above a dry sand box (real-part dielectric permittivity of sand

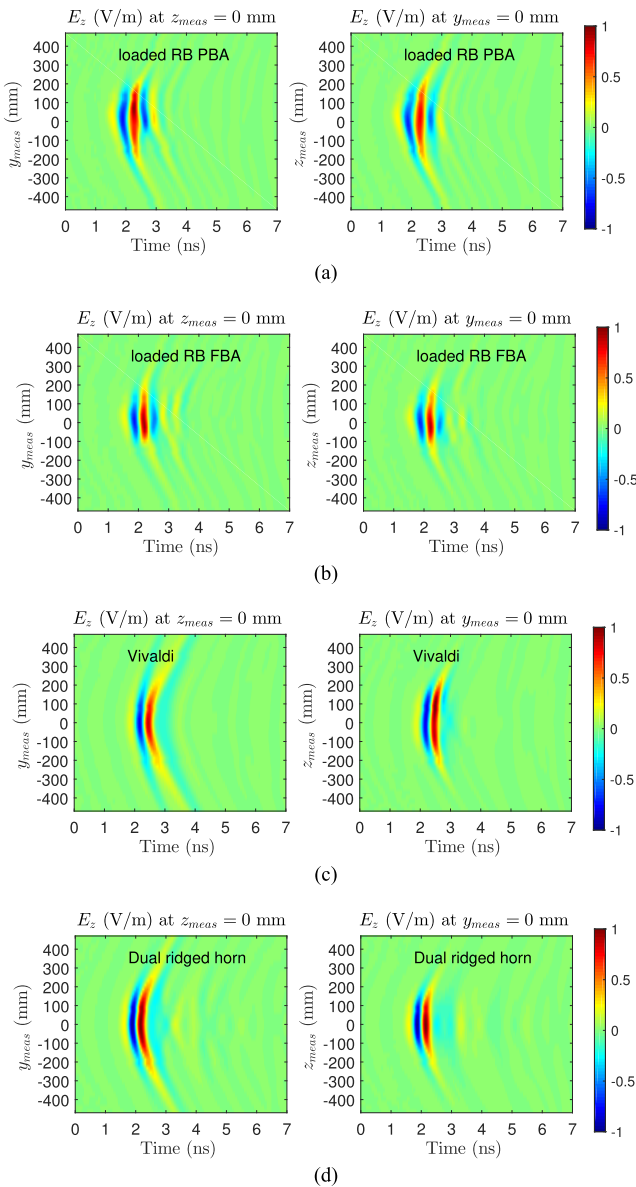


Fig. 11. Normalized transient radiated NFs at 100 mm in front of the antenna under test in the free space at plane cuts $y_{meas} = 0$ mm and $z_{meas} = 0$ mm for (a) loaded RB PBA, (b) loaded RB FBA, (c) Vivaldi antenna, and (d) dual ridged horn antenna.

about 2.5, negligible imaginary part) as an example of such an environment.

A small probe (En-Probe [16]) is buried 100 mm deep inside the sand box. The antenna under test is displaced over a polystyrene (foam) interface of 50 mm thickness that ensures constant distance between antenna aperture and the sand box surface. The probe is linked to the network analyzer via an optical fiber, which guarantees perturbation-free measurement [17].

In order to cancel the reflection caused by the air–sand contrast at the bottom side of the sand box, pyramidal absorbers are employed as seen in Fig. 14. In order to realize the symmetric excitation of the antennas, two measurement steps are needed. In the first step, we measure the antenna radiated

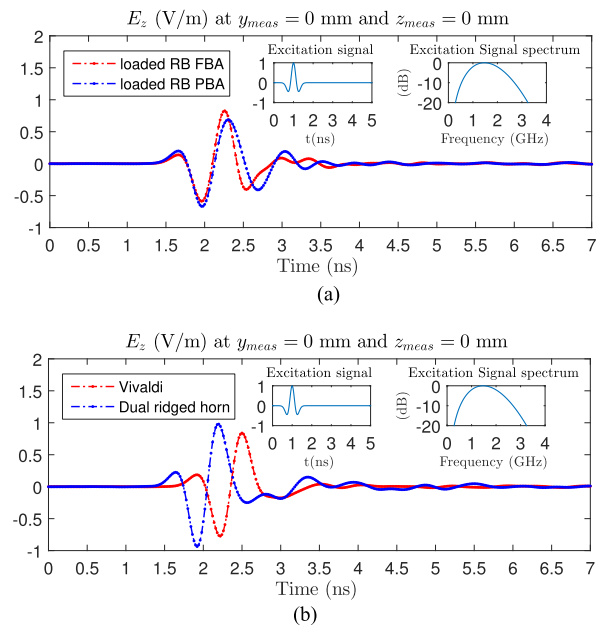


Fig. 12. Normalized measured transient E_z components of the free space radiated NF at 100 mm in front of the antennas at the position $y_{meas} = 0$ mm and $z_{meas} = 0$ mm. (a) Loaded RB FBA and Loaded RB V-FBA. (b) Vivaldi and dual ridged horn antennas.

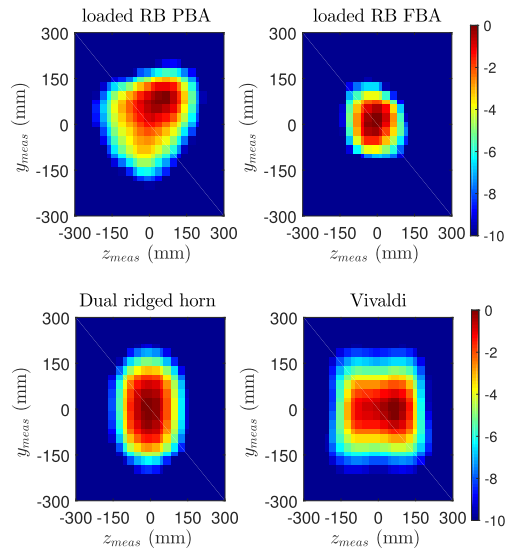


Fig. 13. Antenna footprints for loaded RB PBA, loaded RB FBA, Vivaldi antenna, dual ridged horn antennas.

field when the first port is fed by the VNA and the second one is 50Ω loaded, and in the second step, the field is measured with the second port fed while the first one is 50Ω loaded. By subtracting the results of these two-step measurement results, we achieve the differential excitation of the antennas.

By displacing the tested antennas over the scanning grid (Fig. 14), defined by $-350 \leq y_{meas} \leq 350$ mm and $-200 \leq z_{meas} \leq 200$ mm, with steps $\Delta z = \Delta y = 20$ mm, the radiation field over the frequency band (0.5–3.5 GHz)

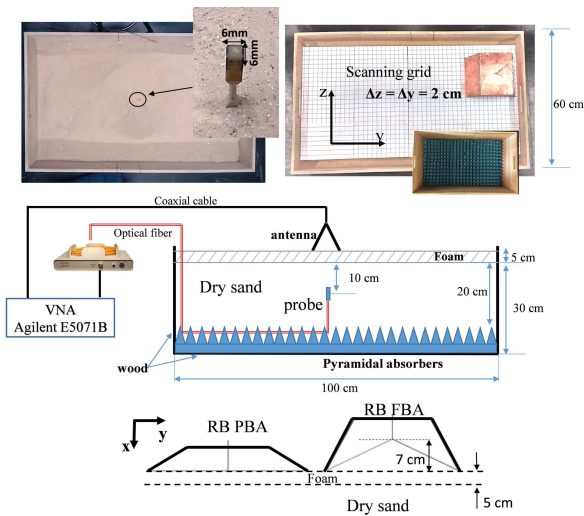


Fig. 14. Over sand box NF measurement procedure description.

100 mm deep inside the sand box is gotten as a function of the frequency. In Fig. 15, the measured E_z field is shown for cut planes $y_{\text{meas}} = 0$ mm and $z_{\text{meas}} = 0$ mm as a function of frequency. It is seen from Fig. 15(a) that the dominant frequency band of the loaded RB PBA is (0.75–1.75 GHz). This means that the loaded RB PBA is radiating significant E_z field at 100 mm depth over this frequency band. The loaded RB FBA E_z field presented in Fig. 15(b) shows the frequency dominant band (0.75–2.25 GHz), which is well suited for the excitation pulse defined in (1). Indeed, the loaded RB FBA guarantees a smooth and maximally radiated field in front of the antenna over the dominant frequency band (0.75–2.25 GHz).

To compute the impulse response of the antennas, it is sufficient to run an inverse Fourier transform, letting them excited by a Mexican hat signal covering the frequency band (0.5–3 GHz). The transient radiated E_z fields at plane cuts $y_{\text{meas}} = 0$ and $z_{\text{meas}} = 0$ are shown in Fig. 15(c) and (d).

The loaded RB FBA [Fig. 15(d)] presents a focused transient radiation pattern in comparison with the loaded RB PBA [Fig. 15(c)], especially at $z_{\text{meas}} = 0$ plane cut, and the transient response of the RB FBA in the boresight direction is shorter than the one of RB PBA as seen in Fig. 16.

Further remarks are of interest. The antenna central part comprising the excitation port of the loaded RB PBA is at 5 cm from the sand surface. For the loaded RB FBA, the excitation port is at 12 cm from the sand surface. This is illustrated in Fig. 14. Bending the bowtie antenna can be helpful to prevent the antenna mismatch caused by the strong coupling between excitation port and soil in GPR context, especially with wet soil. For this reason, the transient response at the boresight direction of loaded RB PBA starts before the loaded RB FBA one as shown in Fig. 16 since the distance between the antenna port and the measuring probe is slightly different (7 cm difference). However, the duration of the radiated signals is shorter for the bended antenna.

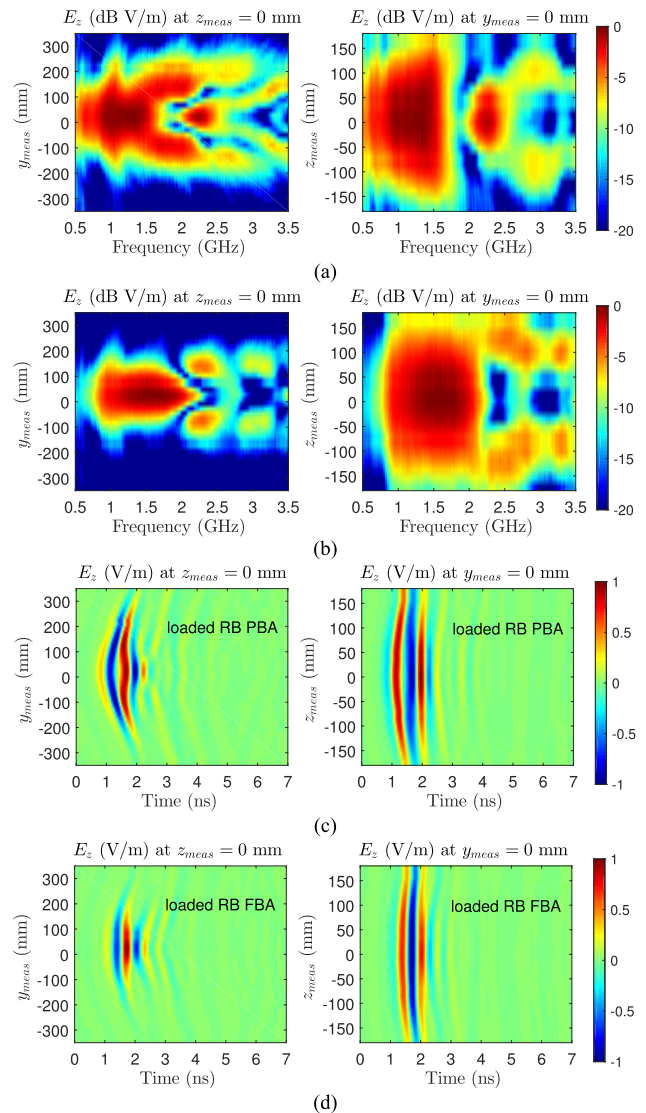


Fig. 15. Normalized measured E_z components of the radiated NF at 100 mm depth in the sand box at plane cuts $y_{\text{meas}} = 0$ mm and $z_{\text{meas}} = 0$ mm as a function of frequency for (a) loaded RB PBA and (b) loaded RB FBA. Normalized transient radiated NFs at 100 mm in the sand at plane cuts $y_{\text{meas}} = 0$ mm and $z_{\text{meas}} = 0$ mm for (c) loaded RB PBA and (d) loaded RB FBA.

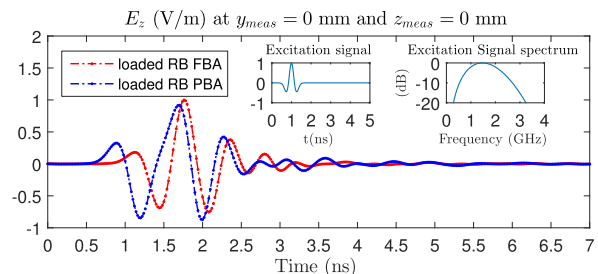


Fig. 16. Normalized measured E_z components of the radiated NF at 100 mm depth in the sand box in front of the antennas ($y_{\text{meas}} = 0$ mm and $z_{\text{meas}} = 0$ mm) as a function of frequency.

V. CONCLUSION

In this paper, an UWB loaded RB FBA dedicated for GPR applications has been proposed. It has been compared

with PBA with and without low-profile backing reflectors. Both simulations and experiments from prototypes have been conducted. Performances are compared in the far-field region for the radiation pattern and realized gain. The optimized resistively loaded RB folded antenna (RB FBA) allows a large bandwidth and is well adapted to GPR applications as comforted by measuring the field distribution at 100 mm depth in a sandy environment.

Overall, two ideas have been followed and validated by simulations and experiments: bending the bowtie antenna enlarges the frequency band where the main radiation energy is in the boresight direction, using an optimized reflector keeps the antenna matching and allows the directive near- and far-field radiation patterns. The final antenna prototype (loaded RB FBA) has been compared with other classically UWB antennas used in GPR applications. We compared the characteristic of this antenna with a ridged horn antenna (SH600) and the presented solution is cheaper, not cumbersome and present a comparable performances. The proposed antenna performances have been compared with a Vivaldi antenna also.

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4.6 Mesures sur site : à la recherche de la jonction perdue

Dans un volet applicatif de la thèse de X. Liu (2014 - 2017) et dans le cadre du partenariat avec EDF R&D, nous avons proposé de tester l'approche GPR pour la détection de la jonction qui relie deux câbles électriques souterrains. Ce travail a été mené en collaboration avec Céline Mîry et Emmanuel Amador dans le site de Moret-sur-Loing de EDF R&D Laboratoire des matériaux électriques. Les câbles se trouvent à une profondeur de 80 cm de la surface du sol et la jonction correspond à cette excroissance de 50% du diamètre du câble. On souhaite détecter/localiser cette variation du diamètre à partir du champ diffracté mesuré via un radar de sol. Ce dernier est composé d'une paire d'antennes Bowtie adaptées dans la bande fréquentielle [0.5 GHz - 3 GHz].

Nous avons étudié le problème direct en considérant un milieu homogène (sable). Dans l'étude numérique nous avons simulé la scène composée d'un câble muni d'une jonction en son centre que nous avons placé dans un volume de sable. Les antennes émettrices et réceptrices ainsi que le câble et sa jonction sont rigoureusement modélisés.

Nous avons modélisé ce scénario avec CST-MWS (transient mode) et nous avons comparé les résultats de migration des champs diffractés (Kirchhoff) selon la nature du sable (humide/sec) et la polarisation des antennes (parallèle/perpendiculaire) par rapport au câble. Cette étude numérique avait pour objectif d'apporter une preuve de concept de la détectabilité de la jonction par une approche GPR. Les conditions de la modélisation numérique sont certes idéales (sable homogène), néanmoins elles peuvent nous renseigner sur le choix de la polarisation des antennes à exploiter lors des mesures.

Sur le site de EDF R&D, nous avons mesuré les signaux radar afin de localiser la jonction avec un seul B-scan (mesure le long du câble). En effet, nous connaissons *a priori* la position du câble. L'environnement de mesure était inhomogène en profondeur et rugueux en surface. Cela n'a pas été pris en compte dans nos simulations électromagnétiques. Nous avons prélevé quelques échantillons de sol pour caractériser la permittivité du milieu de propagation. Pour accéder au champ diffracté nous avons utilisé deux méthodes de filtrage du "clutter". La première méthode consiste à soustraire la valeur moyenne/médiane du champ total mesuré. La deuxième méthode utilisée a été proposée dans [POTIN et collab. \[2006\]](#). Elle repose sur la conception d'un filtre numérique passe-haut. Les composantes spectrales du "clutter" sont associées aux basses fréquences et la cible est caractérisée par les fréquences hautes du spectre.

Le champ diffracté mesuré et filtré indique clairement la localisation du câble et de la jonction. Cependant, l'application de la migration de Kirchhoff en exploitant les données mesurées n'est pas satisfaisante, le bruit environnant est relativement fort. L'utilisation d'antennes polarisées parallèlement au câble assure une détection plus facile. Pour les besoins des mesures nous avons utilisé un analyseur de réseau vectoriel (Agilent ENA E5071) dans la bande fréquentielle 0.1GHz - 3.1 GHz sur 1201 points. Ce travail est commenté en détail dans [LIU et collab. \[2018\]](#) (joint à ce rapport).

Publication représentative

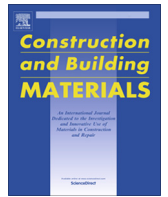
Ce document apporte un complément d'information sur les détails de la modélisation numérique et présente les résultats des mesures sur le site d'EDF R&D dans le site de Moret-sur-Loing, Laboratoire des matériaux électriques.

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Virtual Special Issue

Ground-Penetrating Radar and Complementary Non-Destructive Testing Techniques in Civil Engineering

Detectability of junctions of underground electrical cables with a ground penetrating radar: Electromagnetic simulation and experimental measurements



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HIGHLIGHTS

- Optimized fractal V-folded bowtie antenna for high frequency GPR.
- GPR scene modeling using actual antennas as transmitting and receiving components.
- Proof of concept of the detectability of junction of buried power cable using GPR technique.
- Measurement at the actual environment at EDF test site for the experimental validation of the analysis.

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ABSTRACT

For electricity distribution companies, being able to accurately detect the position of buried power cables using nondestructive methods is a crucial issue. The most important issue is the localization of the junction part linking the cables to each other where most maintenance operations are carried out. In this work we have conducted a feasibility study to confirm the relevance of high frequency Ground Penetrating Radar (GPR) to detect these buried junctions in their environment despite the clutter. The cables are buried in an inhomogeneous Earth medium at about one meter deep. Consequently, we have optimized a folded bowtie antenna to fit the frequency band [0.5 GHz–3 GHz]. Then a rigorous model of the cable is considered in the numerical study using the optimized antenna in a bi-static configuration. A full wave electromagnetic simulation software (CST MWS) has been used to study the detectability of the junction within dry and wet homogeneous sands then a classical Kirchhoff migration technique has been used to process the data and localize the junction. Finally, our analysis has been validated by measurements at a test site (real environment) to show the efficiency of the proposed analysis. The measured data are also post-processed using a digital filtering technique for clutter reduction and visibility enhancement of the buried cable.

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1. Introduction

Ground penetrating radar (GPR) is a well-known non-invasive system used for identification and localization of buried cables, pipes and targets [1]. It is based on radars using electromagnetic waves to illuminate the ground using specific antennas. Echoes are collected and analyzed in order to image the underground. The wave propagates through it and when it reaches an electro-

magnetic contrast, the scattered field collected by the receiving antenna contains dielectric and geometric information about this contrast. Thanks to such information content of the signal, GPR is used in civil engineering [2], transport applications and road diagnostics [3–5], and soil moisture mapping [6–10]. GPR techniques are also used on masonry building to locate metal reinforcement [11,12] or for mapping of archeological sites [13]. Using GPR, rebar position and depth can be evaluated with good accuracy [14–18] however, their diameters are difficult to estimate [17].

In the following, we will focus our work on the detection and localization of the junction that connects two or more buried

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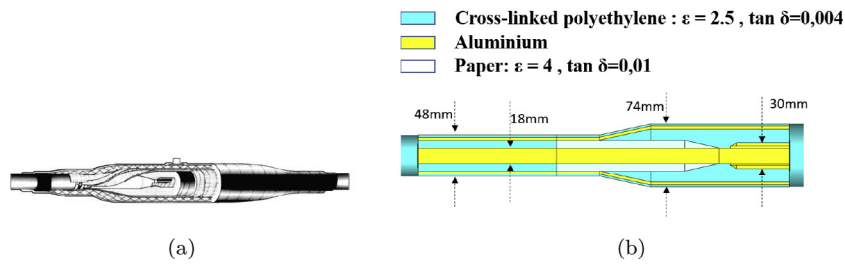


Fig. 1. (a) A photograph of the junction of underground electrical cables, (b) The junction detailed description for the electromagnetic modeling purpose, the junction length is 40 cm.

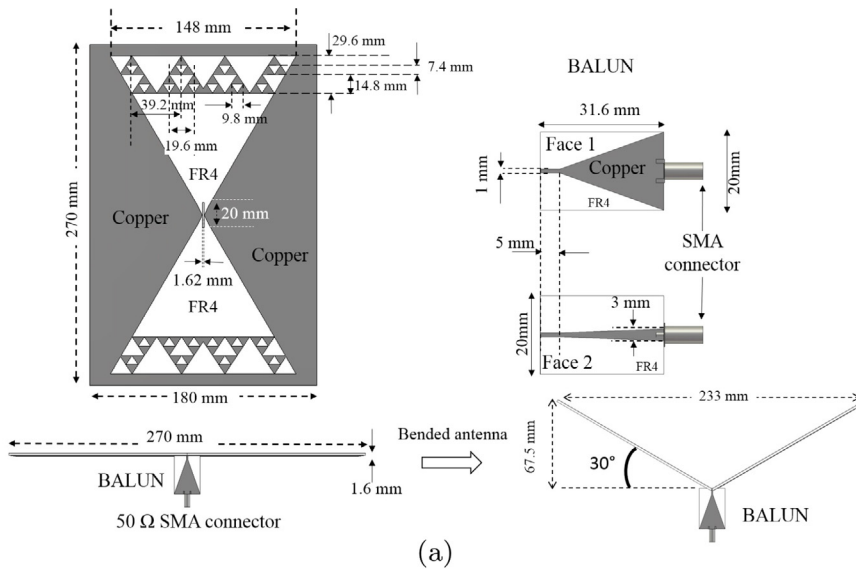


Fig. 2. (a) V-Folded Bowtie Antenna (VFBA) description, (b) the cavity backed VFBA description.

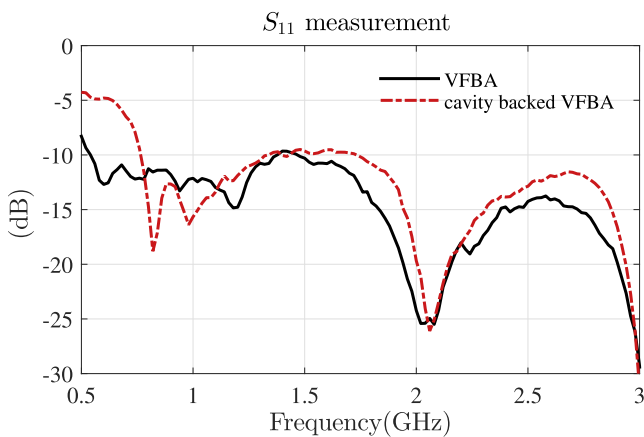


Fig. 3. The measured reflection coefficients (S_{11}) of the cavity backed VFBA and the VFBA.

power cables using a GPR system and the collected B-scan along the latter. The junction is represented as the part of the power cable with a diameter outgrowth of 50% compared with the one of the cable. Such a diameter variation modifies the scattered field enough to distinguish the junction from the cable. The power cables are usually buried at 1 m depth in sandy environments ($\epsilon_r \sim 4$) and an example of real buried junction is presented in Fig. 1 where geometric details are provided. This is one of the smallest junction used by Electricité De France Company (EDF). This work, aims at showing that a wide band GPR system covering the frequency band [0.5 GHz–3 GHz] can allow us to reach the needed resolution (3 cm) to distinguish the contrast between the scattered fields due to the cable and the junction.

Thanks to the help of a full wave software (CST MWS) and its Finite Integration Technique a folded bowtie antenna operating at the frequency band [0.5 GHz, 3 GHz] has been optimized and various scenarios have been studied either in terms of embedding

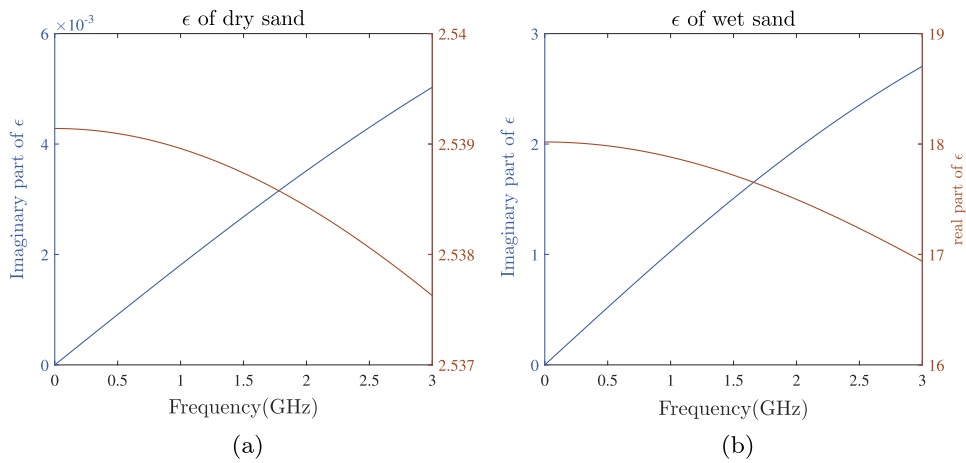


Fig. 4. Permittivity of the sand used for the electromagnetic simulation, (a) dry sand, (b) wet sand.

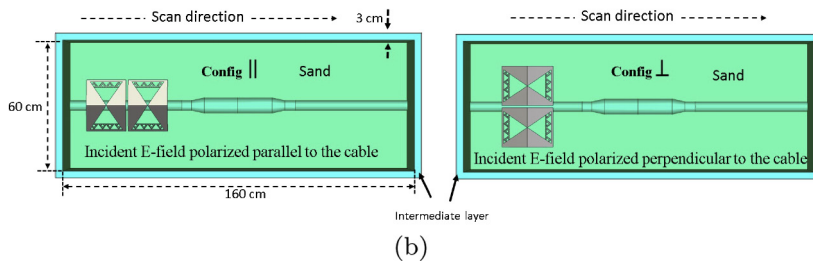
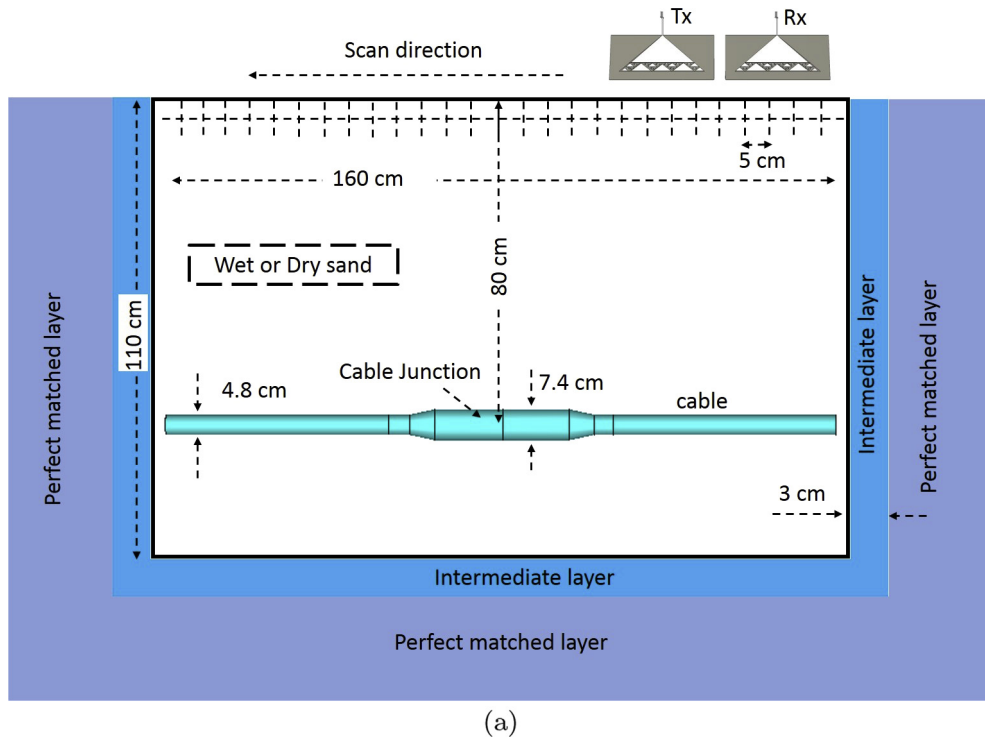


Fig. 5. (a) The simulation configuration with the buried cable, the sandy environment and the transmitting and receiving antennas, (b) description of the up-view of the simulation configuration associated with the polarization of the transmitting and receiving antennas.

Earth medium (dry/wet sand) or in terms of polarization (parallel/perpendicular). The geometric and electric description of the power cable embedded in a homogeneous dry or wet sandbox respectively are considered in the numerical model and the optimized antenna is used as transmitting and receiving structures. In fact, the authors of [19–26] have underlined the impor-

tance of the incident field modeling on the GPR imaging results. The corresponding received signal (S_{21}) in a bi-static configuration along the buried scanned cable is then collected and treated using various filters in order to remove the clutter. Finally, the measurements are realized inside a controlled area of “EDF - R&D” company with the laboratory of electrical material at Moret sur Loing site

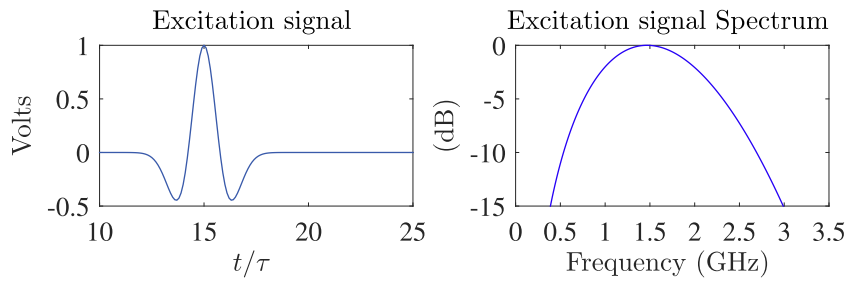


Fig. 6. The used excitation signal and its frequency spectrum.

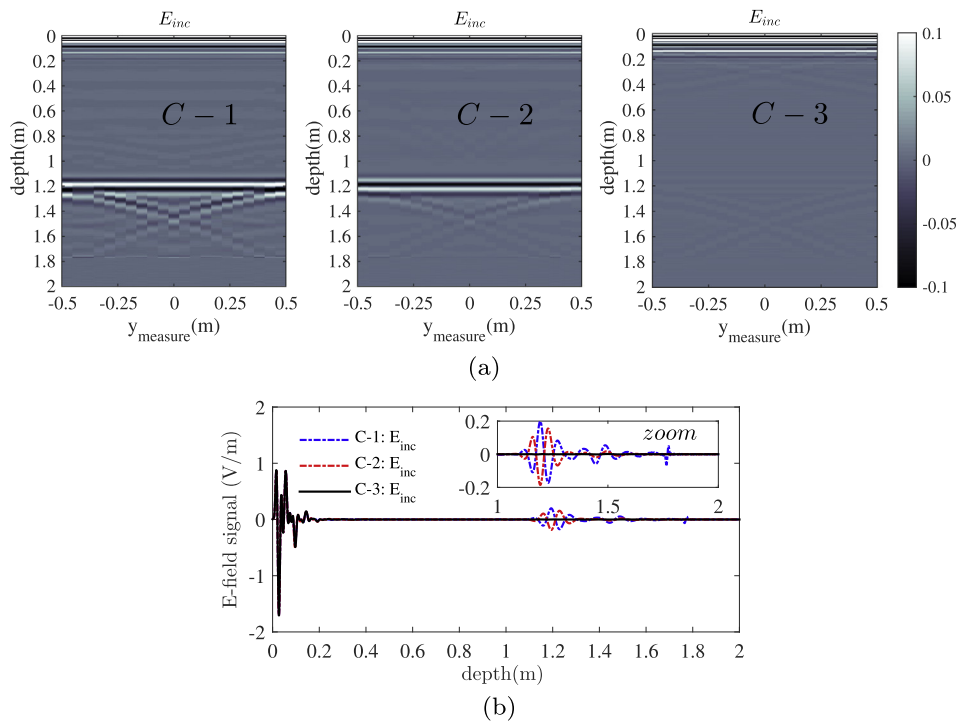


Fig. 7. (a) The background modeling (E_{inc}) for C-1, C-2, C-3 configurations (b) the A-scan of the central position $y_{meas} = 0$ for C-1, C-2 and C-3 configurations.

where multiple cables are buried and where the presented GPR study has been validated. The paper is organized as follows: Section 2 is dedicated to describing the designed antenna. Section 3 proposes the full-wave simulation results for the junction detectability. Section 4 focuses on the on-site measurement results conducted at Moret sur Loing. Finally, a conclusion is drawn to outline the interest of the conducted study for utilities detection and localization in Section 5.

2. The GPR antenna structure

Ultra wideband antennas are widely used for GPR applications [27,28]. These antennas can be TEM horn [29], biconical antenna [30], the bowtie, the lumped element loaded antennas [31]. Here an optimized fractal folded bow-tie antenna for on-ground GPR has been chosen. This antenna is used in the vicinity of the soil (proximal antenna). Such an antenna belongs to the dipole antennas family, which is efficient and easy to design since the input port impedance depends on the flare angle of the bowtie. We have optimized the antenna with its symmetric excitation (BALUN) for the frequency band [0.5 GHz–3 GHz]. The geometrical description of the antenna is given in Fig. 2 where the BALUN is also described. The antenna has been optimized in free space using the transient solver provided by CST MWS and the measured reflection coeffi-

cient (S_{11}) of the realized antenna is presented in Fig. 3. In order to prevent the back-radiation of the antenna we need to transform it to unidirectional one. We can use a cavity as specified in many papers among which [32,33] where a large cavity in comparison with the bowtie antenna size has been used. However, such a solution is not adapted to the GPR application since the size of the cavity has to be comparable with the size of the antenna as presented in [34]. Nevertheless, the use of the cavity is inevitable because of the electromagnetic energy radiated from the back of the bow-tie antenna. Without a shielding around the antenna, the interferences with the external parasitic electromagnetic perturbations reduce significantly the signal to noise ratio of the measured data. In fact, the metallic cavity brings the protection to the GPR antennas and without a specific precaution, the coupling between the antenna and the shielding cavity will inevitably deteriorate the antenna matching.

In this paper we propose, besides using the cavity filled with absorbers, to bend the fractal bowtie antenna (as shown in Fig. 2)) to enhance the directivity of the antenna in the frequency band of interest [0.5 GHz–3 GHz]. The idea of bending the antenna is inspired from TEM antennas as depicted in [35].

The 30° bended antenna is labeled V-Folded Bowtie Antenna (VFBA). As shown in Fig. 3, the VFBA S_{11} is presented and compared with the S_{11} of the VFBA with the shielding cavity filled with absor-

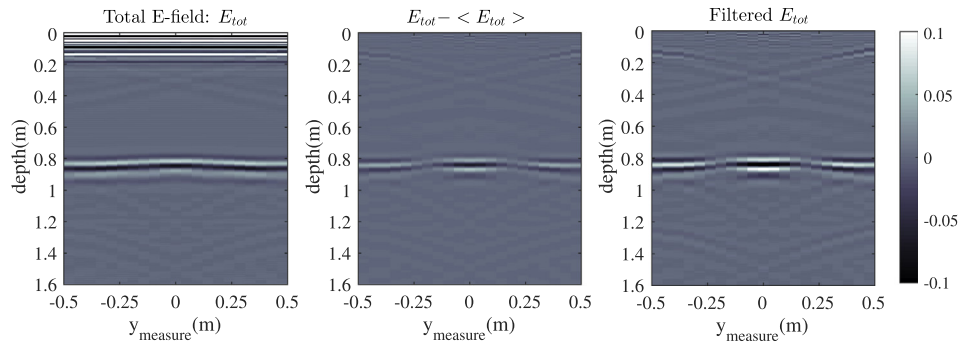


Fig. 8. B-scan radargrams for E_{tot} , $E_{tot} - \langle E_{tot} \rangle$, filtered E_{tot} , for parallel configuration with wet sand.

bers. The cavity backed VFBA presents $S_{11} < -10\text{dB}$ for the frequency band [0.75 GHz–3 GHz] and the cavity filled with absorbers did not disturb significantly the VFBA reflection coefficient (S_{11}) over the frequency band [0.75 GHz–3 GHz].

3. The electromagnetic modelling of the GPR scene

From now, the optimized VFBA will be used as transmitter and receiver to illuminate the scene and collect the scattered field due to a buried cable in dry or wet sands. The permittivity of wet sand is defined for 20% humidity. The permittivity values used for our simulation are predefined in the CST MWS library and their real and imaginary part are given in Fig. 4 whereas the GPR scene modeled in our study is described in Fig. 5. Keeping the simulation time reasonably short (about 4 h per antennas position) the cavity filled with absorbers for the antennas as shown in Fig. 5 has not been considered. For the simulation of the GPR scene, we have used the transient solver of CST MWS with hexahedral meshing based on the spatial criterion $dx = dy = dz = \lambda/12$ where λ is calculated for the frequency 1.5 GHz. The total number of used mesh cells is 29 million for two VFBA placed in front of a sand volume of 1100 mm depth, 660 mm width and 1670 mm length.

The simulation is running on PC with 32 GByte RAM, over Intel Xeon CPU E5-1650 V3 3.5 GHz. The excitation pulse is sent to the antenna via the coaxial connector of the BALUN and the simulation is stopped after 50 ns. For each position along the cable, this simulation takes 3h45min. In order to work with reduced size sandbox volume perfect matched layers (PML) have been used to prevent the echoes produced by the borders of the sand volume. For the case of dry sand scene, PML are placed around the sand volume to simulate an infinite sandy environment. For the case of a wet sand, the PML predefined in the CST software cannot deal with a dispersive environment and this is why an intermediate layer (IL) with a permittivity equal to the frequency-averaged value of the amplitude of the wet sand permittivity has been added. Adding the PML over the IL greatly helps in echoes attenuation. The PML play their role correctly since these have been defined for the IL containing non-dispersive material (Fig. 5). We have optimized the thickness of the IL using CST MWS in order to reduce the echoes produced by the sandbox limitations. The IL thickness optimization is not the main subject of the present study and will not be detailed in the following however in our case a value of 3 cm has allowed a sufficient reduction (cancellation is not reached) of the echoes produced by the sandbox borders. Another solution would have been to use a larger sandy volume and filter out the echoes but with an important increase of the CPU time (more than 12 h calculation per antenna position for the double of the sand volume) with respect to the proposed solution.

The efficiency of the proposed solution for the echoes cancellation due to the sandbox boundaries is demonstrated by comparing

the results of background modeling (the field without the buried cable) in three configurations. In configuration C-1 a wet sand volume placed in free space without perfect matched layers nor intermediate layers is considered, then configuration C-2 is defined as a wet sand with PML around the sand volume then configuration C-3, where a wet sand volume with the PML and IL as presented in Fig. 5 is used. The electromagnetic simulations are carried out in the time domain and the transmitting antenna is excited using the second derivative of the Gaussian signal. The excitation signal is expressed as

$$e(t) = - \left[4\alpha^2 \left(\frac{t-t_0}{\tau} \right)^2 - 2\alpha \right] \exp \left[-\alpha \left(\frac{t-t_0}{\tau} \right)^2 \right] \quad (1)$$

and its spectrum normalized at $f_0 = 1.5\text{GHz}$ are presented in Fig. 6 for $\tau = 2 \times 10^{-10}\text{s}$, $\alpha = 0.85/\tau^2$ and $t_0 = 15\tau$ which covers the antenna frequency band [0.5 GHz–3 GHz].

For each antennas measurement position (A-scan), the receiving antenna collects the field scattered by the sand volume and the transmission coefficient S_{21} is exploited to construct the radargram presented in Fig. 7a. The center of the sandbox is considered as the origin of the axis and the scanning is carried from [–0.5 m, 0.5 m] with a displacement step of 5 cm.

As shown in Fig. 7b, the combination of the PML and the IL (C-3) reduces significantly the echoes resulting from the limitation of the sand volume whereas in configurations C-1 and C-2 those echoes are still visible (mainly the echoes related to the bottom of the sand box around a depth of 1.2 m). For the three configurations, the cross-talk between the transmitting and receiving antennas is seen in Fig. 7b where the A-scan corresponding to the central position $y_{meas} = 0\text{m}$ for C-1, C-2 and C-3 are compared, as seen in the zooming part the C-3 configuration is efficient for the echoes attenuation. In the following, only the combination of the PML and IL will be considered.

Now a numerical study of the influence of the transmitting and receiving antennas polarization on the buried cable detection is performed. In Fig. 5b an upper-view of the antennas disposition above the sandbox is depicted. In the first situation, the transmitting and receiving antennas are placed in order to align the polarization of the incident field with the buried cable and consequently the impinging E-field polarization is parallel to the scan direction along the cable (parallel configuration). In the second situation, the antennas are positioned so that the incident E-field is polarized perpendicularly to the scan direction (perpendicular configuration).

Using CST MWS to model the scene presented in Fig. 5 for a wet sand leads to the B-scan radargrams presented in Fig. 8 for the so-called parallel configuration. The total field (E_{tot}) represents the measured field over the sandbox in the presence of the buried cable. From the E_{tot} B-scan radargram, two clear echoes zones are

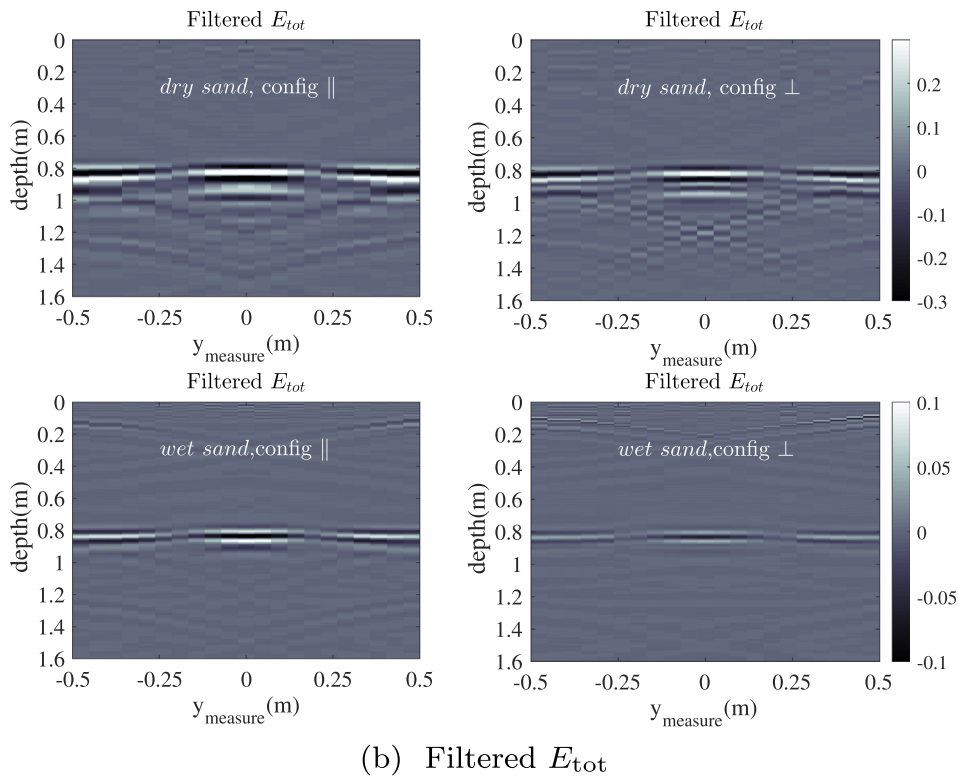
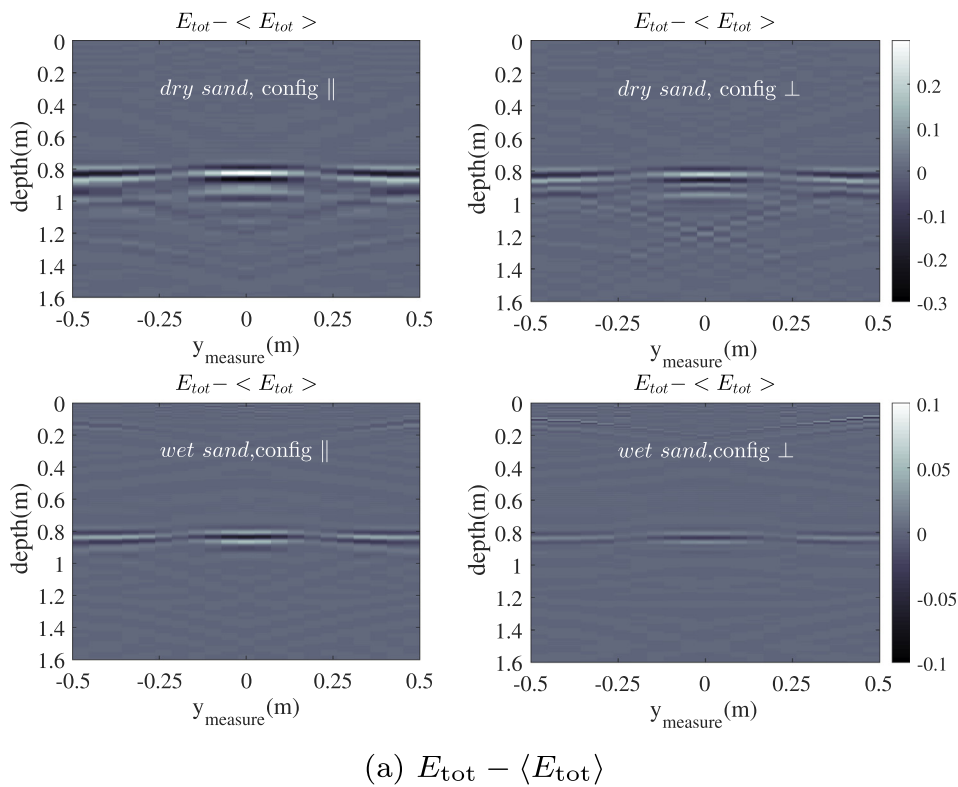


Fig. 9. (a) B-scan radargram of $E_{tot} - \langle E_{tot} \rangle$ for the 4 studied situations: dry/wet (top/bottom), parallel/perpendicular (left/right) polarizations (b) B-scan radargram of Filtered E_{tot} for: dry/wet (top/bottom), parallel/perpendicular (left/right) polarizations.

identified, the first one is due to the antennas cross-talk combined with the air–soil interface whereas the second one is linked to the buried cable position. In order to get access to the scattered field, background removal techniques are needed to isolate the signature

of the buried cable. Among them, two have been used in the following. The first one is one of the procedures described in [1] for clutter reduction by the GPR community, where the radargrams resulting from $E_{tot} - \langle E_{tot} \rangle$ where $\langle E_{tot} \rangle$ means averaged value of

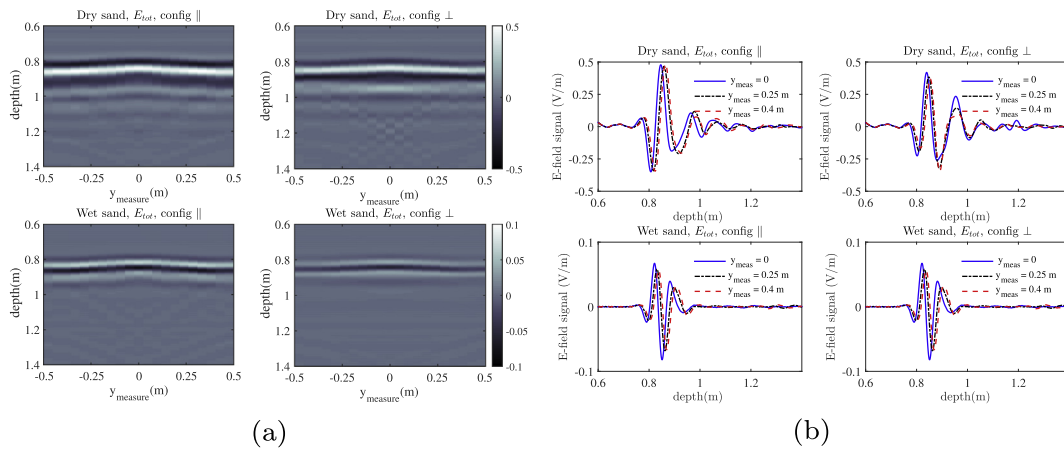


Fig. 10. (a) B-scan radargram of the total field corresponding to the four studied situations. (b) Comparison of A-scan at $y_{\text{meas}} = 0, 0.25, 0.4$ m.

the B-scan radargram of E_{tot} is used. The second one is the digital filtering technique proposed in [36], which is based on the design of a digital filter adapted to the B-scan. The high-pass filter to design aims at eliminating the clutter component in the spectral domain. The filtered radargrams are presented in Fig. 8. These two clutter reduction techniques have shown comparable radargrams.

Four situations corresponding to dry and wet sand for parallel and perpendicular polarizations are studied and the corresponding radargrams are presented either for $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ and for filtered E_{tot} in Fig. 9. In all of them three areas can be identified in the signature of the buried cable. The first and the last areas corresponding to the scanning positions $0.25\text{ m} < y_{\text{meas}} < 0.5\text{ m}$ and $-0.5\text{ m} < y_{\text{meas}} < -0.25\text{ m}$ characterize the echoes due to the cable. The in-between second area shows the echoes due to the central part of the cable (junction) located at $|y_{\text{meas}}| \leq 0.2\text{ m}$. The sloping parts of the cable junction (Fig. 5) produce the echoes situated between $(0.2\text{ m} < y_{\text{meas}} < 0.25\text{ m}$ and $-0.25\text{ m} < y_{\text{meas}} < -0.2\text{ m})$. The magnitude of the echoes gives qualitative information about the position of the junction. In Fig. 9 the cable junction is located at $|y_{\text{meas}}| \leq 0.2\text{ m}$ corresponding to the maximum magnitude of $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ and of the filtered E_{tot} .

As it is seen in Fig. 9, the amplitude of the echoes resulting from dry sand are higher than for wet sand due to the higher propagation losses of the latter. As expected, the parallel antenna polarization produces higher level echoes compared with the perpendicular one. Based on these four studied situations, a common conclusion arises; the outgrowth of the cable diameter (junction position) is characterized by stronger echoes which are localized at $|y_{\text{meas}}| \leq 0.2\text{ m}$ placed at 80 cm depth. When the antenna polarization is parallel to the cable, the impinging E-field is better coupled to the buried target, which increases the scattered echoes amplitude. On the contrary, in the perpendicular polarization, the coupling between the impinging E-field and the buried cable is lower, leading to a lower signature of the cable and its junction. As a conclusion and as expected in wet sand media, the radargram resolution is better and the parallel polarization is more efficient for the junction localization due to the strongest echoes amplitude.

In order to have a better understanding of the behavior of the B-scan radargram at the intermediate area between the junction and the cable around $y_{\text{meas}} = \pm 0.25\text{ m}$ the A-scan signals of E_{tot} at the positions $y_{\text{meas}} = 0\text{ m}$ (junction), $y_{\text{meas}} = 0.25\text{ m}$ (transition part) and $y_{\text{meas}} = 0.4\text{ m}$ (cable) are compared in Fig. 10. Comparing the E-field for the two polarizations at the junction central position ($y_{\text{meas}} = 0\text{ m}$) shows that the signal of parallel polarization results in a stronger signature. The outgrowth of the cable diameter at

the junction position ($y_{\text{meas}} = 0\text{ m}$) justifies the differences between the E-field signals at the positions $y_{\text{meas}} = 0\text{ m}$ and $y_{\text{meas}} = 0.4\text{ m}$. In Fig. 10, the A-scan signals show that the signal amplitude is proportional to the size of the scattering part of the cable. The position $y_{\text{meas}} = 0\text{ m}$ corresponds to the largest diameter (74 mm), $y_{\text{meas}} = 0.4\text{ m}$ corresponds to the smallest diameter (48 mm) and the position $y_{\text{meas}} = 0.25\text{ m}$ is the intermediate region between the junction and the cable where the diameter is between 48 mm and 74 mm.

Finally, for imaging purpose a Kirchhoff migration technique [37–42], has been applied to the data of $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ and filtered E_{tot} and the results are shown in Fig. 11. The buried cable and its junction are clearly identified either for dry sand or wet one and either for parallel polarization or for perpendicular one.

As a conclusion the simulation analysis of the GPR scene composed of a cable buried in a sandbox respecting different polarizations and for two commonly existing environments: dry sand and wet sand with 20% humidity is proposed. The actual fractal V-folded bowtie antennas (VFBA) are used in a bistatic configuration for transmitting and receiving the electromagnetic field in the frequency band [0.5 GHz–3 GHz]. Up to here the electromagnetic (EM) modeling taking into account the coupling between the antennas and the sandbox was the main interest. The EM modeling has proven that the cable junction is detectable using a GPR technique in the frequency band [0.5 GHz–3 GHz]. The results presented for E_{tot} are not trivial for the localization of the junction, however, using $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ and filtered E_{tot} the junction can be localized. The filtering technique is simply implemented for efficient clutter reduction as presented. In addition, the Kirchhoff migration technique has shown good results for the localization of the junction based on a unique B-scan along the buried cable. In the next part, the results with experimental on-site measurements will be carried out in a real (even if controlled) environment.

4. On-site measurement results

We have carried out a measurement campaign inside a controlled area in “Electricité De France - R&D” company with the laboratory of electrical material at Moret sur Loing site (EDF lab Les Renardières). The photography of this site is depicted in Fig. 12 and shows the measurement area and cables before being buried. The cables are placed at around 1 m depth from the soil surface and the covers seen in the photography are removed before burying the entire scene. We scanned over 4 m along a cable as shown in Fig. 12, where the junction is localized with the wooden stick

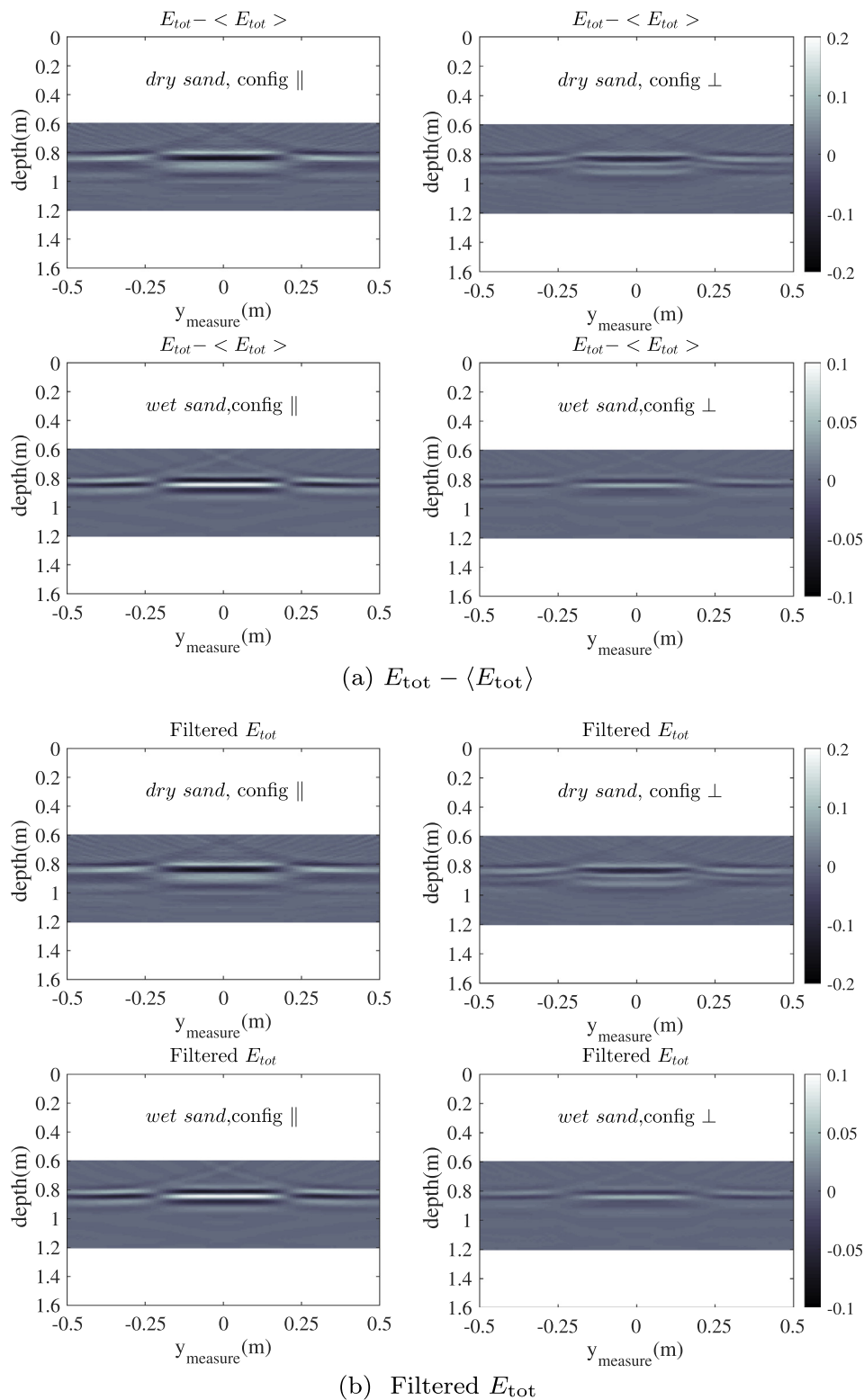


Fig. 11. Kirchhoff migration resulting from 4 situations associated with (a) $E_{tot} - \langle E_{tot} \rangle$ (b) Filtered E_{tot} .

(preserved after burying the cables) seen at the right of the photography.

As seen in Fig. 12, the buried cables are not identical to the simulated one; they are gathered by pair and have a larger diameter (total diameter of about 90 mm) and the junction we are investigating is also larger since it contains the junctions of two cables. The diameter of the buried junction is around 200 mm. The

Frequency-Stepped Continuous Wave (FSCW) GPR technique has been used to measure the scattered field using the cavity backed V-folded fractal bowtie antennas and the vector network analyzer Agilent ENA E5071 in the frequency band [0.1 GHz–3.1 GHz] over 1201 frequency points (Fig. 13). Then the frequency domain data are transformed to extract the B-scan radargram according to the pulse previously depicted in Fig. 6. For each measurement position

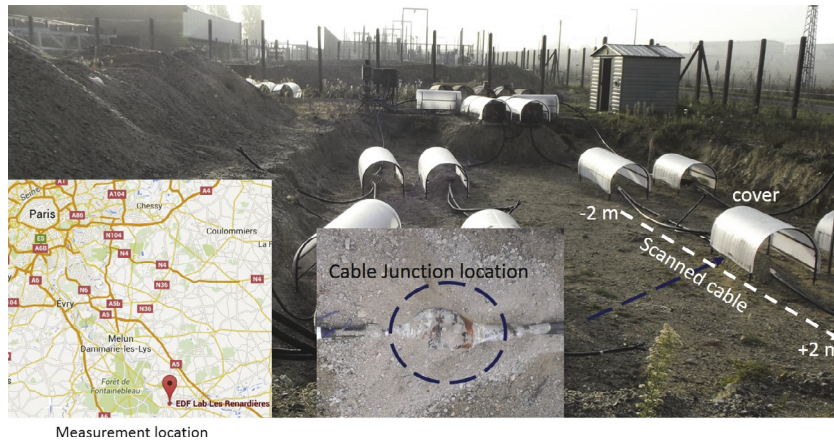


Fig. 12. The measurement area at EDF-lab.

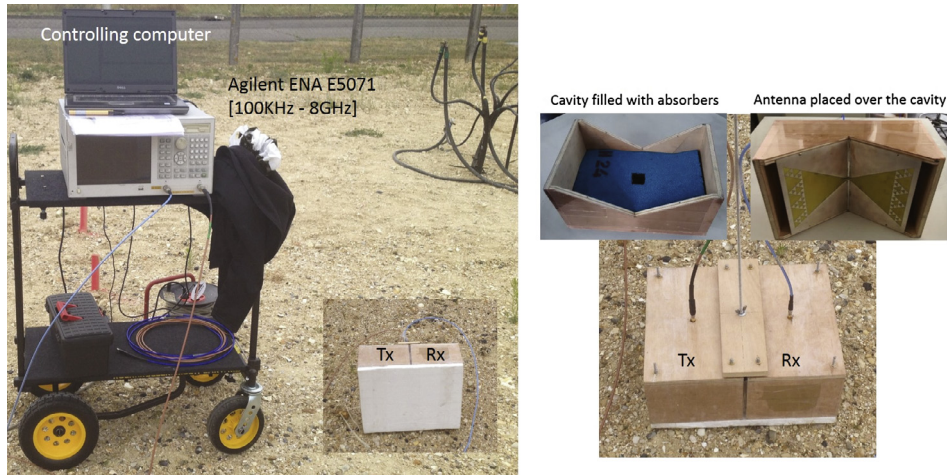


Fig. 13. (left) Vector Network Analyzer and controlling computer, (right) transmitting and receiving antennas cavity backed V-folded bowtie antennas.

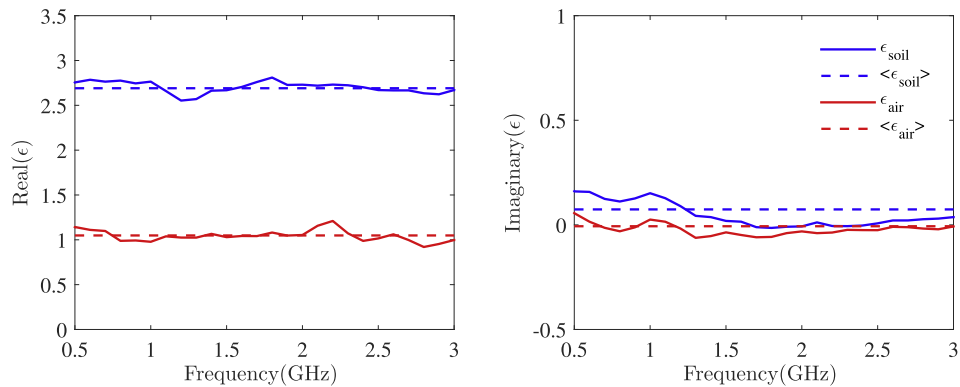


Fig. 14. Measured epsilon of a sand sample collected at the measurement site, (left) real part, (right) imaginary part.

along the cable the controlling computer acquires the transmission parameter S_{21} . The measurement data are collected along the scanning line presented in Fig. 12 and the network analyzer has been calibrated before starting the scan. The antennas are moved manually step by step along 4 m with a displacement step of 5 cm. Unlike the simulation study the surface of the soil is not flat and the soil is not homogeneous.

In order to evaluate the wave velocity in the ground, we have collected a sample of the soil and we measured the permittivity

in the frequency band of interest using a coaxial probe in the laboratory. The results are shown in Fig. 14. The measured real part of the permittivity of the soil is about 2.69 and the imaginary part is about 0.07. Using the same permittivity measurement technique, we have measured the permittivity of the air as a reference. Indeed, the average of the measured real part of the air permittivity is about 1.04 and the imaginary part is about -0.006 . These measurement results are used in the B-scan radargram representation to achieve the time to distance transformation for the cable and junc-

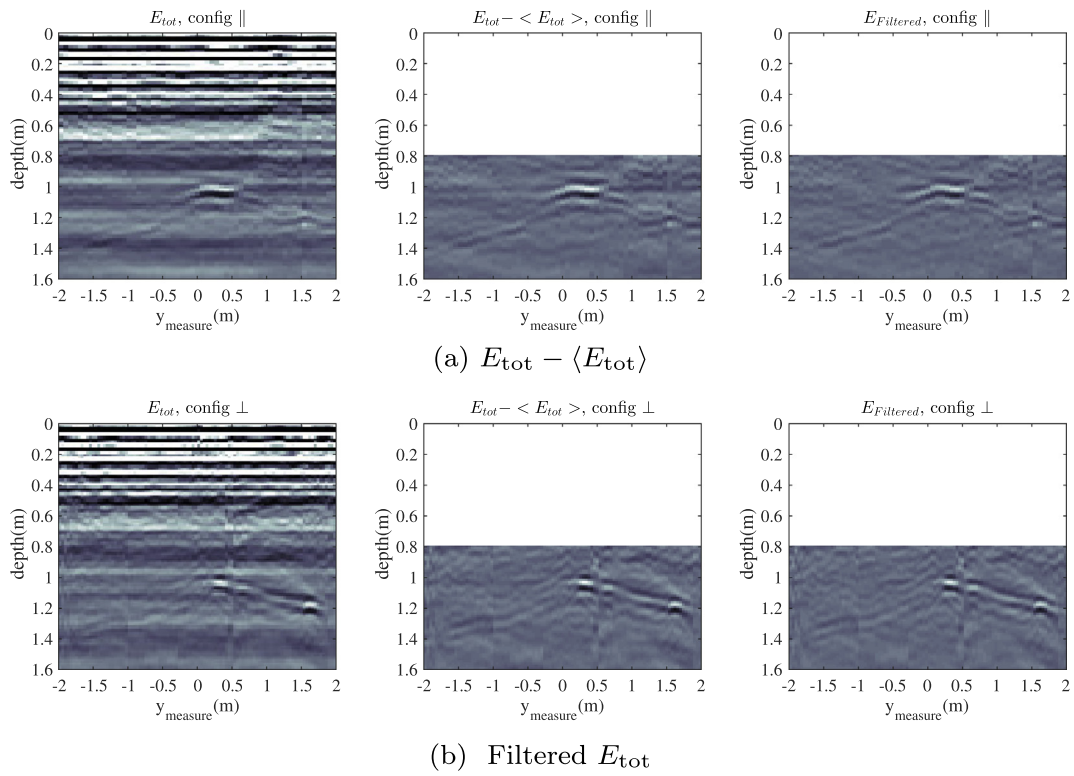


Fig. 15. B-scan radargrams of E_{tot} , $E_{tot} - \langle E_{tot} \rangle$ and filtered E_{tot} corresponding to parallel polarization in (a) and perpendicular polarization in (b).

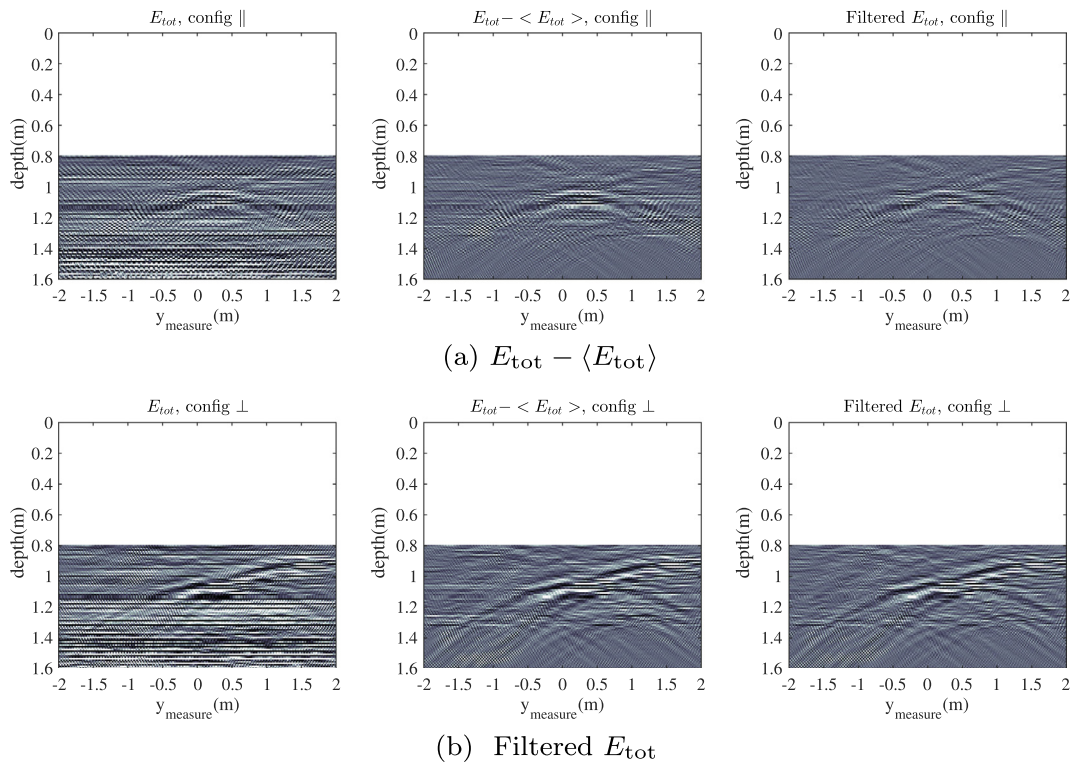


Fig. 16. Kirchhoff migration results of E_{tot} , $E_{tot} - \langle E_{tot} \rangle$ and filtered E_{tot} for parallel (a) and perpendicular (b) configuration.

tion localization. The measurement has been conducted in July with a dry soil.

The B-scan is obtained for the parallel and perpendicular polarizations. The radargrams are presented in Fig. 15. In comparison

with the simulation results, the clutter in the radargrams for E_{tot} is higher mainly caused by the soil heterogeneities. In addition, the antennas response is differently coupled with the non-flat soil. Using the cavity-backed antennas we prevent the external envi-

ronment coupling, however we distort the shortness of the antenna transient response. Based on the radargram of E_{tot} , the cable silhouette is clearly visible between 1 m and 1.4 m depth and the junction corresponds to the brightest zone at 1 m depth at the measurement position 0.5 m. This is applicable for both parallel and perpendicular polarizations.

We have focused our effort on the clutter reduction in the depth range $0.8\text{ m} < \text{depth} < 1.6\text{ m}$. After time gating the measured signal, we have used two techniques for clutter reduction. First the averaging technique to cancel a part of the clutter as presented in Fig. 15, $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ applied for $0.8\text{ m} < \text{depth} < 1.6\text{ m}$ considerably reduced the clutter and the cable is easily identified. In a second time, we have used the digital filtering technique proposed in [36] with the same conclusion. Both techniques help at enhancing the detectability of the buried cable and consequently the junction is localized at 1 m depth at the measurement position $y_{\text{meas}} = 0.5\text{ m}$ as expected. The parallel polarization is more efficient than the perpendicular one which confirms the simulation results.

In Fig. 16, we present the imaging results of the Kirchhoff migration technique resulting from the measurement data. As it is seen, in the parallel configuration the cable and its junction are easily identified using $E_{\text{tot}} - \langle E_{\text{tot}} \rangle$ and filtered E_{tot} . However, in the case of cluttered data, the cable is hardly detectable after the Kirchhoff migration for the perpendicular configuration.

5. Conclusion

In this paper we have shown that the high frequency GPR technique is an efficient non-destructive technique able to accurately detect the location of the buried power cables. Using a unique B-scan measurement along the cable, the junction is clearly detected against the cluttered environment. As a priori information we know the position of the cable. We have considered a complex GPR scene for our electromagnetic modeling study comprising the actual antennas. Comparison between experimental data (anechoic chamber) and synthetic data either computed using CST-MWS or using a laboratory-made software based on Discontinuous Galerkin Time-Domain method are presented in [43,44] and has not been reproduced here.

First, we have optimized the fractal V-folded bowtie antenna for the frequency band [0.5 GHz–3 GHz]. Then, we have presented the scattered field of a buried cable using a rigorous model of the cable provided by Electricity De France R&D. Two homogeneous burying environments have been used (wet and dry sands) which is a simplification of the real soil. However, the measurement campaign carried out in the actual environment have proved that high frequency GPR technique is an efficient tool for the buried cables detection. For the clutter reduction we have exploited two different methods that bring the same enhancement to the measured B-scan. Based on experimental data collected at the measurement site, the cable is localized if its depth is greater than 0.5 m. Indeed, the clutter due to the antenna response ringing and the soil inhomogeneity, engender parasitic oscillation that are hardly filtered. For shallowly buried cable, this technique is not efficient. Also, the polarization of the transmitting and receiving antennas has been investigated. Parallel polarization is recommended for use since the scattered field is higher.

From the antenna point of view, the VFBA is bent so the distance between the excitation part of the antenna (central part) and the soil is about 7 cm, which helps to keep the antenna matching stable even in front of a wet environment. The risk of placing the antenna in front of high conductive medium is that the reflected signal is strong enough to mismatch the antenna excitation. Consequently, the energy accepted by the antenna is reduced.

Acknowledgment

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4.7 GPR pour le diagnostic des systèmes racinaires des arbres

Les travaux sur la détection de la jonction entre deux câbles électriques souterrains ont ouvert une perspective fort intéressante sur l'utilisation du radar de sol pour la caractérisation des systèmes racinaires des arbres.

L'intérêt porté aux racines des arbres est très actuel. Les racines assurent la solidité et la stabilité des arbres sans oublier l'apport vital en eau et nutriments [BRUNNER et GODBOLD \[2007\]](#). Elles contribuent grandement dans le stockage du carbone et de l'excédent de CO₂ présent dans l'atmosphère. La contribution racinaire représente entre 20 et 40% du carbone total stocké dans une forêt [STOVER et collab. \[2007\]](#). En conséquence, une évaluation judicieuse de la biomasse et de la distribution racinaire est un indicateur important sur le rôle que peuvent jouer les arbres dans la lutte contre le réchauffement climatique. L'utilisation du radar de sol pour le contrôle non destructif des racines des arbres est une voie prometteuse comme en témoigne la littérature sur le sujet. Li Guo et son équipe de *Pennsylvania State University* ont proposé un ensemble de travaux dont la finalité est l'étude des potentialités du GPR pour une cartographie efficace des systèmes racinaires des arbres [HIRANO et collab. \[2009\]](#) [GUO et collab. \[2013c\]](#) [WU et collab. \[2014\]](#) [CUI et collab. \[2012\]](#).

Nous avons défini les contours de la thèse de **Abderrahmane Aboudourib (2017- 2020)** pour explorer l'exploitabilité du radar de sol pour cartographier la distribution racinaire des arbres à des fins d'ingénierie civile et d'estimation de biomasse. Cette thèse, intitulée « Imagerie en champ proche de systèmes racinaires : de la modélisation électromagnétique à l'acquisition et au traitement des données en contexte de génie civil et estimation de biomasse », est financée grâce à une allocation de CentraleSupélec. Elle est menée dans le cadre d'une collaboration entre les laboratoires GeePs et L2S (D. Lesselier). Ce travail de recherche doit proposer des hypothèses claires et les conditions suffisantes pour réaliser cette cartographie. Nous avons basé notre réflexion sur les points suivants :

1. Une compréhension de la nature diélectrique des racines vis-à-vis de la permittivité du sol (contraste),
2. L'étude des dispositions spatiales des émetteurs/récepteurs dans des configurations bi- ou multi-statiques et les apports informatifs de cette diversité
3. La modélisation du problème direct en environnement inhomogène
4. Le traitement/inversion des données mesurées pour une interprétation totale ou partielle de la cartographie des racines

Cette thèse s'inscrit dans un cadre applicatif particulier qui nécessite une analyse multi-paramètre du scénario GPR. Les auteurs de [GUO et collab. \[2013b\]](#) proposent une synthèse des éléments à prendre en compte lors de l'analyse du problème direct impliquant des racines. En effet, la modélisation des parties prenantes du problème se décline en trois phases. La première aborde la modélisation de la racine à travers ses propriétés diélectriques complexes. Le modèle de Maxwell-Garnet pour la loi des mélanges est utilisé dans [PAZ et collab. \[2010\]](#) et [STRAUBE et collab. \[2002\]](#) afin de proposer une description du comportement de la permittivité de la racine en fonction de sa teneur gravimétrique en eau.

Dans la deuxième phase, un sol homogène est décrit par les modèles de Topp [TOPP et collab. \[1980\]](#) ou Peplinski [[PEPLINSKI et collab. \[1995\]](#)]. Le second est plus général que le premier car il prend en compte la composition matérielle du sol, sa densité et sa teneur en eau. Le modèle de Peplinski est valable dans l'intervalle fréquentiel [0.3 GHz - 18GHz] où l'évolution fréquentielle de la permittivité est dictée par la relation de Debye. Dans la troisième phase, les hétérogénéités du sol sont décrites et modélisées à partir du travail de Turcotte [TURCOTTE \[1997\]](#) et sa description fractale de l'environnement hétérogène.

Dans la première approche du problème, nous visons une étude paramétrique des facteurs qui limitent la détection des racines en utilisant un radar bi-statique. Nous avons utilisé GprMax [GIANNOPOULOS \[2005\]](#) pour la modélisation du problème direct. Dans [ABOUDOURIB et collab. \[2019\]](#), nous avons focalisé notre attention sur l'impact du diamètre et de la teneur en eau des racines sur leur détectabilité (analyse 2D). En terme de post-traitement, les données de simulation sont focalisées grâce à la migration de Kirchhoff et la décision de la détectabilité est basée sur un seuillage indexé sur l'intensité du champ diffracté.

Nous menons actuellement, dans le cadre de simulations 3D (exploitation de C-scans), les validations numériques (GprMax) et expérimentales (dans un bac à sable) de la méthode du « filtrage adapté » (matched filter) [RICHARDS et collab. \[2010\]](#) pour la reconstruction 3D de l'architecture racinaire à partir d'un C-scan dans un environnement hétérogène. Ces résultats seront présentés dans IWAGPR 2019 au mois de septembre 2019 (joint avec ce rapport).

Dans les prochains mois, nous procéderons à l'étude numérique des configurations multi-statiques combinées avec une technique SAR multi-statique. Cela marquera une réelle contribution sur le plan numérique. L'autre volet qui nous intéressera à partir du mois de septembre 2019 concerne les mesures à

mener sur le terrain avec le radar que nous venons d'acquérir Mala ProEx muni de la technologie HDR avec des antennes 450 MHz et 750 MHz.

Publications représentatives sur la détection racinaire

- [CI.37] A. Aboudourib, M. Serhir, D. Lesselier, « 3D reconstruction of tree roots under heterogeneous soil conditions using Ground Penetrating Radar », *Near Surface Geoscience Conference 2019*, Sep 2019, The Hague, Netherlands.
- [CI.36] A. Aboudourib, M. Serhir, D. Lesselier, « Near-field characterization of root systems, from computational modeling and simulations to controlled laboratory experiments », *PIERS PhotonIcs & Electromagnetics Research Symposium*, Dec 2019, Xiamen, China.
- [CI.35] A. Aboudourib, M. Serhir, D. Lesselier, « Impact of root diameter and water content on tree roots detection using Ground Penetrating Radar », *13th European Conference on Antennas and Propagation*, Apr 2019, Krakow, Poland

3D Reconstruction of Tree Roots under Heterogeneous Soil Conditions using Ground Penetrating Radar

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Abstract—Coarse roots can cause damage to pipes, pavements and building foundations. Hence, reconstructing tree root system in its three dimensions is of paramount importance to understand and reduce interactions between trees and urban infrastructures. Ground Penetrating Radar (GPR) is a promising non-invasive technique for the study of tree roots. In this contribution, a 3D GPR simulation was carried out on two roots having elemental architectures in order to be reconstructed. The simulation was performed with GprMax software under heterogeneous soil conditions. Considering a heterogeneous soil not only provided a realistic soil model, but also enabled us to test the robustness of our post-processing techniques. We mainly used a Singular Value Decomposition (SVD) combined with the matched filter technique.

Index Terms—Ground Penetrating Radar, tree roots reconstruction, forward simulation, heterogeneous soil, Singular Value Decomposition, matched filter.

I. INTRODUCTION

Because tree roots play a variety of roles in plant ecosystems, estimating their size, distribution and 3D structure is of great importance. However, tree root systems are complex since they contain roots that differ in morphology and size. Fine roots are generally small (diameter smaller than 2 mm) with a lifespan up to 6-7 weeks, and have a main function of absorbing water and nutrients; they also play an important role in carbon sequestration and carbon cycling [1]. Coarse roots (diameter larger than 2 mm) with a perennial lifespan, provide the anchorage of the tree with their rigid structure and serve as organs to transport water from deeper soil horizons. They are often called “structural roots”. Actually, coarse roots can generate an average radial pressure of 8×10^5 N/m² [2]. Therefore, they are often negatively impacting man-made infrastructures. So the necessity arises of diagnostic tools to follow their time-spatial evolution.

Moreover, based on numerous investigations, very few tree species’ roots appear to be able to deeply penetrate the soil [3]. Starting from the stem base, coarse roots are ramified in a wide spreading way, with the majority of roots concentrated within the upper 60 cm of soil [4], which makes their detection feasible using a surface-based radar.

Ground Penetrating Radar (GPR) has been found as one of the fast, cost-effective and non-invasive methods for the tree roots systems characterization, as compared to other traditional digging ways such as excavation and uprooting,

which are not only extremely time consuming and labor-intensive, but also likely distorting the real structure of roots.

This contribution suggests a post-processing scheme for 3D reconstruction of tree roots based on GPR data. In fact, using GPR for the reconstruction of tree root systems allows one to make a bridge between real and modeling worlds. Moreover, the topological structure of tree roots and its growth dynamics are indicators of root system’s response to environmental stresses [5].

In this study, we propose to simulate a realistic environment using a geometrically and physically realistic prototype of tree root under heterogeneous soil conditions. To do so, two roots having two elemental forms of common tree roots architectures have been considered. Then, thanks to a chain of post-processing, essentially based on matched filter technique, we have been able to reconstruct the 3D root structures.

This paper is organized as follows: in section II, details of forward simulation including soil and roots modeling are described. In section III, applied data processing procedures are defined and numerical results are presented. In section IV, conclusions are drawn and perspectives for future works are outlined.

II. FORWARD SIMULATION

This section is devoted to presenting the details of 3D FDTD electromagnetic simulations aiming to provide GPR data that are used in our proposed processing procedures. For this, we considered two root models having the basic structures to build any other root architecture.

A. Methodology

All simulations have been conducted using GprMax [6], an open source software, based on the Finite-Difference Time-Domain (FDTD) numerical method. In this study, we used GprMax in the 3D mode and we considered the geometry sketched in Fig. 1. Simulations have been computed using z-polarized dipoles as antennas under a central frequency of 1GHz. The spatial discretization was determined according to the numerical stability and reliability criterion and was set to 3 mm. The excitation current waveform is assumed to be a Ricker pulse (i.e., the first derivative of a Gaussian pulse) as it is widely used for GPR investigations. The time window was set to 8 ns and the separation between transmitter and

receiver to 10 cm. For each B-scan, the displacement step of the antennas above the studied scene and the distance between antennas and soil were set to 2 cm and 5 cm, respectively. In addition, a PML absorbing boundary condition is applied to all domain sides to mitigate the impact of reflective noises introduced by the edge of the computational grid.

To complete the simulation setting, one needs to specify geometric and dielectric properties of roots and soil. To do this, we used some models to characterize realistic roots and soil, which are detailed in parts II.B and II.C, respectively.

B. Root modeling

In this study, circular cylindrical structures have been used to model the simulated roots. As mentioned in the introduction, we have chosen to reconstruct two basic root architectures, both consisting of three cylindrical pieces with a diameter of 3 cm: the first root has a Y shape, and the second one is a sinuous root. Fig. 2 shows these root structures in their simulation environment, buried at a depth of 15 cm.

Regarding the dielectric properties of roots, we used a prototype that has been presented in a previous work [7], which links the root dielectric properties to its volumetric water content. This prototype is based on the Maxwell-Garnett model [8] to characterize the relative permittivity of roots and on the Straube model [9] to characterize their electrical conductivity. For our simulations, we considered a root volumetric water content of 50% for roots, which corresponds to a relative permittivity of $\epsilon_{r\text{root}} = 24$ and an electrical conductivity of $\sigma_{\text{root}} = 0.63$ mS/m.

C. Soil modeling

In order to model a heterogeneous soil for the simulation, we used the semi-empirical model of Peplinski *et al.* [10, 11] combined with the fractal model, both been already implemented in GprMax. The semi-empirical model of Peplinski *et al.* defines the dielectric properties of the soils using a semi-analytic function that relates soil dielectric properties to its sand fraction (S), clay fraction (C), sand density (ρ_s), bulk density (ρ_b) and water volumetric fraction (f_u). The semi-analytic function is approximated using a Debye function to describe the dispersive nature of the soil. In

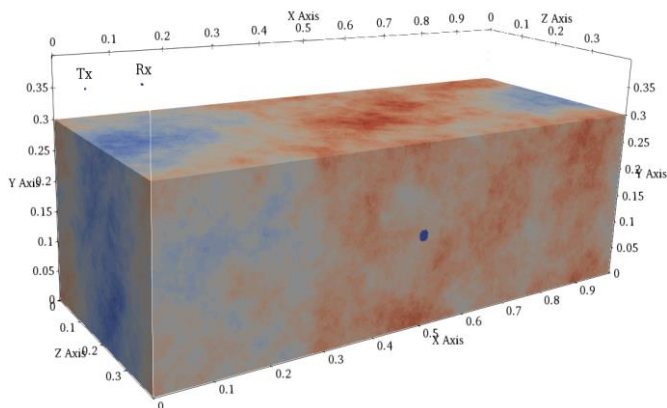


Fig. 1. Scenario of the simulated 3D GprMax environment displayed in Paraview visualization application.

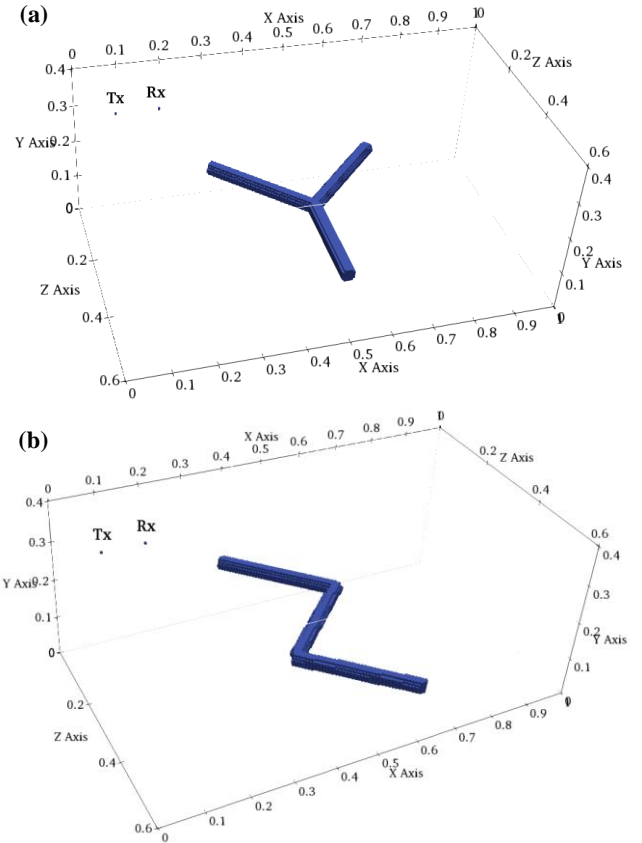


Fig. 2. 3D view of the simulated root structures displayed in Paraview. (a) Y-root, (b) sinuous root.

our simulations we assumed a sandy soil with the following values: $S = 0.7$, $C = 0.3$, $\rho_s = 2.66$ gr/cm³, $\rho_b = 2$ gr/cm³ and $f_u = 1 - 10\%$.

Concerning the geometrical soil distribution, it has been modeled using the fractal correlated noise [12]. We considered that the soil is composed of 20 elements using a fractal dimension β of 1.5 ($0 < \beta \leq 3$). The resulting heterogeneous soil can be seen in Fig. 1 already introduced before. More details about the use of the semi-empirical model of Peplinski *et al.* and fractal model in GprMax can be found in [13].

III. POST-PROCESSING AND NUMERICAL RESULTS

The C-scan simulation results have been computed (antennas sweep: $0 \leq x \leq 100$ cm, $0 \leq z \leq 60$ cm) and post-processed using a MATLAB R2017b program. Several data processing techniques have been applied on radargrams in a stepwise fashion in order to make them more interpretable. Fig. 3 depicts the flowchart of the basic steps as undertaken. In what follows, we give more details about the main techniques.

A. Time-zero correction

Before starting the data processing procedure, a time-zero correction has been carried out. It consists of adjusting the response so that time-zero corresponds to the reflection from the surface of the ground. This operation is critical for accurate

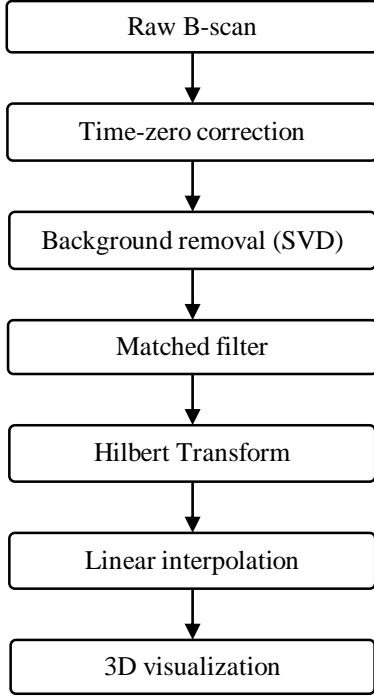


Fig. 3. The flowchart of the post-processing procedure used in this study.

depth determination. Furthermore, to transform the time axis into depth axis, we used the following equation:

$$D = \frac{c}{\sqrt{\epsilon_r}} \frac{T}{2} \quad (1)$$

where D is the depth, T is the two-way pulse travel time, c is the velocity expressed in meters per second ($\text{m}\cdot\text{s}^{-1}$), and ϵ_r is the relative permittivity of the soil. Since the soil in our simulation is built of 20 elements whose dielectric properties were defined using a Debye function, then we took the mean value of the 20 elements. Numerically we get $\epsilon_r = 5.63$.

B. Background removal

Aiming to reduce the clutter and remove the antennas direct coupling from radargrams, we used the Singular Value Decomposition (SVD)-based clutter removal technique [14]. The SVD is a factorization of a rectangular matrix X (in our case a B-scan radargram) into orthogonal matrices, which can be expressed as follows:

$$X = USV^T = \sum_{i=1}^K \lambda_i u_i v_i^T \quad (2)$$

where $U = [u_1, u_2 \dots u_K]$ and $V = [v_1, v_2 \dots v_K]$ (u_i and v_i are vectors) are the orthogonal matrices of X containing the eigenvectors of XX^T and $X^T X$, respectively. S is a diagonal matrix which contains eigenvalues (i.e. the singular values of X) where $X_{i,i} = \lambda_i$ distributed in decreasing order along its main diagonal: $\{\lambda_1 \geq \lambda_2 \geq \dots \geq \lambda_K\}$.

The idea of the SVD-based clutter removal approach is to reconstruct the original radargram X using only some eigenvalues while the rest are set to zero. Large eigenvalues are associated to strongly correlated features such as direct antennas coupling and ground bounce, while low eigenvalues correspond to weakly correlated features namely targets response and high-frequency clutter. In our simulations, excluding the three largest singular values was good enough to get images with reduced clutter.

C. Matched filter

The matched filter is a filter designed to maximize the Signal-to-Noise Ratio (SNR) at the receiver antenna of the GPR system. Actually, when a matched filter is applied to a received signal, the SNR is maximized at a point in time corresponding to the delay to the target [15]. If we call $s(t)$ the transmitted waveform (in this study a Ricker pulse), defined over the time interval $0 \leq t \leq \tau$, the filter that maximizes the SNR takes the form

$$h(t) = a s^*(-t) \quad -\tau \leq t \leq 0 \quad (3)$$

The impulse response of the matched filter is the complex conjugate of the time reversed version of the transmit waveform, scaled by an arbitrary constant a , which is commonly set to 1.

Therefore, the matched filter $h(t)$ is convolved with the received signal $r(t)$ to yield the output $y(t)$:

$$y(t) = r(t) * h(t) \quad (4)$$

$$y(t) = \int_{-\infty}^{+\infty} r(\alpha) h(t-\alpha) d\alpha = \int_{-\infty}^{+\infty} r(\alpha) s^*(\alpha-t) d\alpha \quad (5)$$

From $y(t)$ one obtains an image where pixels at the location of target (root) have a higher SNR compared to surrounding pixels.

D. Numerical results

To give an idea about the post-processing effects on B-scans, we display in Fig. 4 radargrams corresponding to two different positions over the Y-root (at the starting and the end plans of the Y structure).

In Fig. 4a, only time-zero correction has been applied. Consequently, the distance corresponding to the air thickness has been eliminated and the ground surface is the reference of depth axis. Fig. 4b shows the radargram after SVD-based approach application. As it can be noted, the clutter caused by the heterogeneous soil has been highly reduced as well as the direct coupling of antennas. Hence, the root reflection stands out very clearly.

Fig. 4c shows the GPR image after processing with the matched filter imaging algorithm described in section III.C. This yields a focalization of the basic B-scan image indicating the root depth information. Thereafter, the application of the Hilbert Transform enables to get a single spot.

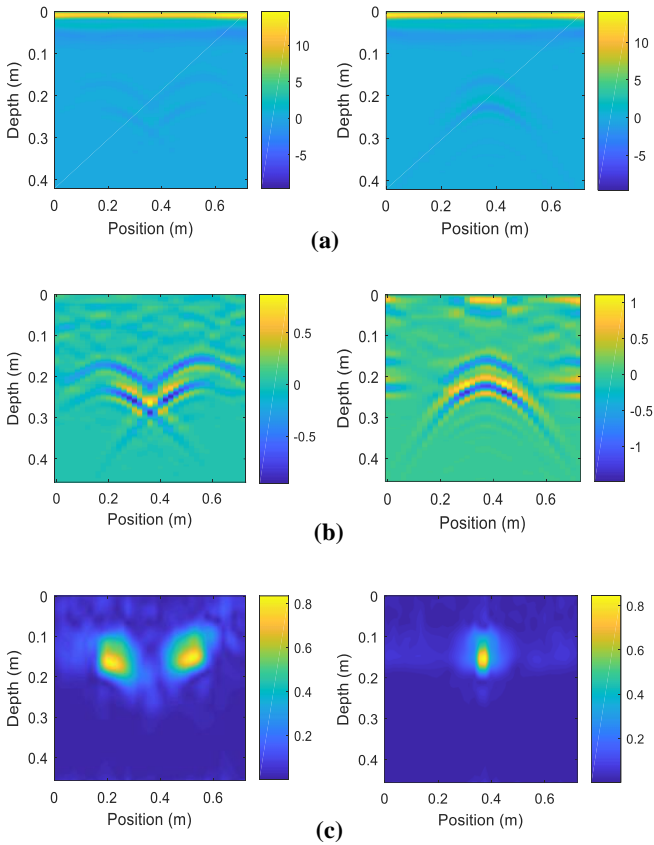


Fig. 4. Effects of post-processing steps: top part (left) and bottom part (right) of the Y-root. (a) The raw radargram with time-zero correction, (b) after background removal, (c) after matched filter, Hilbert Transform and linear interpolation.

Next, by performing linear interpolation on processed radargrams, images are smoothed to prepare a better 3D visualization. Linear interpolation smoothing has been also performed in the parallel root direction (z-direction). This makes the entire root structure smooth and continuous at every point and generates a visually vivid 3D representation.

At the end of the process, we can set a threshold of pixel intensities to adjust the level of heterogeneities to keep in the final representation. Thus, a trade-off must be found between the intensity level of the root and the one of the heterogeneities. In our process, we only kept pixels with normalized intensity ≥ 0.7 .

The difference between radargrams of the top and bottom part of the Y-root can be explained by the difference in polarization. In fact, the bottom part of a Y-root is perpendicular to the line scan of antennas (x-direction), which yields a symmetric and complete hyperbolic response, and, hence, a well-focused spot after matched filter and Hilbert Transform application. Finally, Fig. 5 shows the 3D reconstruction of the simulated roots. We can note the appearance of some parasitic inclusions due to soil heterogeneity. However, we can easily recognize the shape of simulated roots with an acceptable degree of accuracy.

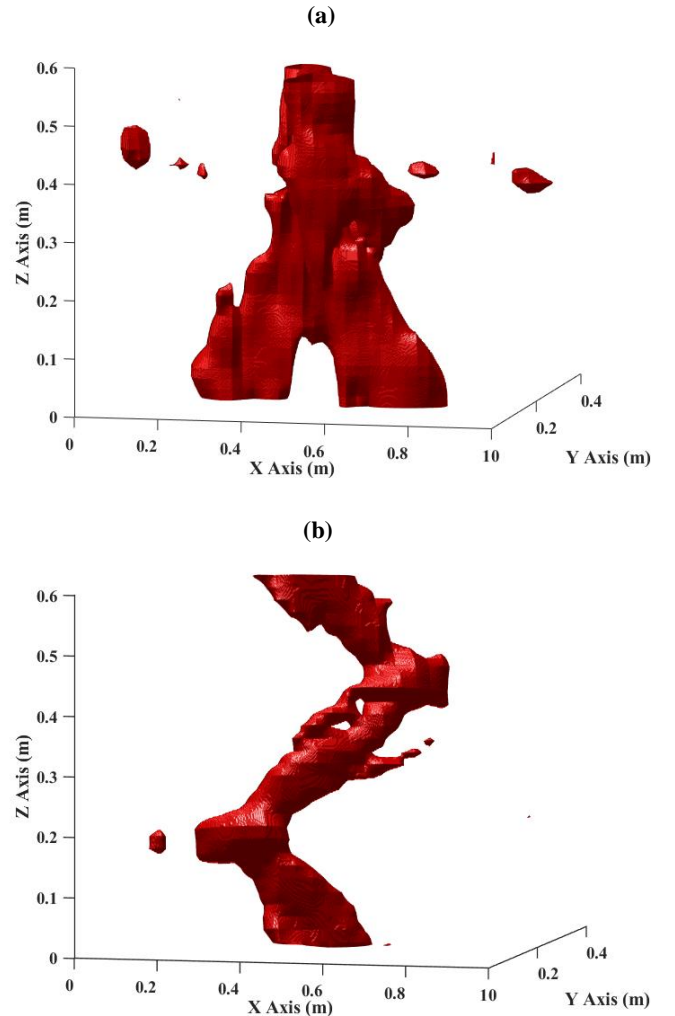


Fig. 5. 3D reconstructions of the simulated root structures. (a) Y-root, (b) sinuous root.

IV. CONCLUSION AND PERSPECTIVES

In this contribution, we proposed a complete approach to reconstruct the 3D structure of two roots. Actually, in the *in situ* environment, these three dimension visualizations can be used to guide researchers to synthesize tree growth dynamics and the environmental effects in the tree development. In our simulations, realistic root characteristics have been considered and heterogeneous soil conditions have been modeled using GprMax features in order to simulate a realistic scene. Regarding the 3D results, thanks to the suggested post-processing chain, the clutter has been meaningfully reduced and a reasonable degree of accuracy and resolution has been achieved.

In the near future, those numerical results will be completed with laboratory measurements using our GPR system previously presented in [7, 16]. In the same context of measurements, we are currently developing a multi-static prototype to test and compare the multi-static SAR technique with Linear Sampling Method previously developed [17].

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4.8 Bilan des contributions et perspectives

Mes activités de recherche sur le GPR gravitent autour des trois volets que sont la résolution du problème direct, le développement des techniques d'imagerie pour résoudre le problème inverse et le développement antennaire pour l'analyse expérimentale des phénomènes. Nous avons abordé la résolution du problème direct avec les outils de simulations numériques basées sur les méthodes : Discontinuous Galerkin Time-Domain (DGTD), Finite Integration Technique (FIT), Finite Difference Time-Domain (FDTD).

Dans le cadre de la thèse de Xiang Liu (2014 - 2017), nous avons utilisé et comparé les résultats de simulation de DGTD et FIT avec les mesures en chambre anéchoïque. Les correspondances sont satisfaisantes. Dans nos simulations, les scénarios GPR modélisés sont identiques à ceux de la mesure. Les antennes (vivaldi) émettrices et réceptrices en configuration bi-statique sont rigoureusement décrites (détails géométriques et diélectriques) au-dessus d'un bac à sable (homogène) dans lequel sont enfouies des cibles diélectriques et métalliques. Nous avons exploité les paramètres S_{21} dans la construction du radargramme issu du déplacement régulier des antennes le long d'une ligne (B-scan) pour balayer la zone où les cibles sont enfouies. Par cette analyse, nous avons montré la pertinence de la DGTD pour modéliser une scène complexe comprenant des antennes en interaction immédiate avec un milieu comprenant des objets enfouis. La DGTD est légèrement plus précise que la FIT pour des temps de simulations équivalents.

Dans le cadre de la thèse de Abderrahmane Aboudourib (2017 - 2020), nous avons utilisé la FDTD pour l'étude du problème direct. Plus particulièrement, nous exploitons GprMax car il prend en charge des modèles réalistes et inhomogènes de sol (Peplinski). Le modèle fractal proposé par GprMax permet la description de l'hétérogénéité du sol et de la rugosité de l'interface air/sol. Nous avons étudié des résultats des simulations 2D et 3D avec des sources d'excitation dipolaires pour qualifier les facteurs qui limitent la détection des racines.

Nous avons étudié la Linear Sampling Method (LSM) comme technique d'imagerie qualitative pour une approche multi-statique. Nous avons analysé les résultats de la LSM à partir de différentes configurations d'observation. Une technique de compensation a été proposée pour améliorer la performance du LSM lorsque le champ incident n'est pas uniforme dans la zone d'investigation (à cause de la directivité de l'antenne et l'illumination en champ proche). La LSM a été validée numériquement et expérimentalement y compris pour le cas des sources d'ordre supérieur. Lors du traitement du LSM dans un cas d'imagerie 3D, nous avons proposé une formulation qui exploite les paramètres S mesurés au niveau des ports des antennes. En dernier, nous avons mis en avant les avantages de la LSM par rapport aux techniques classiques de migration pour la localisation précise des cibles diélectriques.

Actuellement nous développons des travaux sur la méthode du filtrage adapté (matched filter) pour la reconstruction 3D d'objet enfouis. Cette méthode s'apparente aux techniques de migrations et nous l'utilisons comme approche d'imagerie pour la reconstruction 3D des systèmes racinaires à partir d'un C-scan. Cette approche est en cours de valorisation.

Nos travaux s'inscrivent dans un contexte applicatif avec des partenaires industriels. Nous avons développés des solutions antennaires adaptées au contexte GPR et nous avons mené des mesures sur le terrain pour localiser des câbles souterrains. Nous avons adopté l'approche harmonique dans nos mesures car les dynamiques obtenues sont meilleures.

Les perspectives des travaux présentés dans ce chapitre sont nombreuses. Elles concernent les différents volets que sont : l'analyse du problème direct, l'imagerie pour la résolution du problème inverse et l'expérimentation sur le terrain. En matière d'analyse du problème direct, nous souhaitons approfondir le traitement des scénarios GPR avec la méthode FDTD. Son efficacité numérique et sa popularité au sein de la communauté la rendent incontournable. Dans ce travail de modélisation, il faut s'orienter vers la généralisation de la prise en compte de la fonction de transfert des antennes dans les études du problème direct (paramètre S pour l'analyse fréquentielle). Cela passe par une connaissance *a priori* des antennes utilisées dans les scénarios simulés. Cette approche a un coût. Les temps de simulation deviennent exorbitants si les antennes des radars sont rigoureusement maillées et décrites dans le simulateur. L'autre solution consiste à définir un modèle comportemental simplifié (valable dans un secteur angulaire limité) de l'antenne dans des conditions d'utilisation bien définies. Ce modèle est utile dans la phase d'identification des résultats lors de l'analyse du problème direct et en même temps indispensable lors de la correction des champs diffractés mesurés pour accéder aux champs intrinsèquement liés aux cibles étudiées. Cette approche s'apparente fortement aux techniques de correction de sondes dans les mesures en champ proche.

Pour le problème inverse, il est nécessaire d'aborder l'extension de la méthode LSM au cas 3D. La LSM formulée via les paramètres S pourrait être mise en oeuvre expérimentalement pour permettre l'imagerie

quantitative. Cela nécessite une accélération du traitement du problème direct mais donnerait accès à plus d'informations sur le/les objets enfouis. Parallèlement, nous souhaitons étudier des algorithmes SAR multi-statique en comparaison avec des techniques de migration 3D. L'approche temporelle dans la résolution du problème inverse est intuitive et rapide. Nous pouvons proposer une combinaison de plusieurs techniques. Il s'agit d'exploiter les techniques de migration 3D pour délimiter la zone d'investigation et ensuite adopter une approche fréquentielle quantitative.

En ce qui concerne la mesure, nous devons mettre au point un système Multiple Input Multiple Output (MIMO) pour réaliser des acquisitions multi-statiques. Ceci ouvrira des perspectives de validations expérimentales avec des données mesurées en environnement contrôlé. En conséquence, les temps d'acquisition d'un C-scan seront réduits ce qui nous permettra de faire de l'imagerie 3D.

Enfin, nous souhaitons proposer des techniques de correction de données issues des radars du commerce. Notre objectif est de pouvoir exploiter des données récoltées sur le terrain pour imager les systèmes racinaires. Nous travaillons actuellement sur des techniques de suppression de clutter et nous souhaitons valider nos activités sur le diagnostic racinaire à partir de données réelles (Radar Mala ProEx 450 MHz et 750 MHz).

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Chapitre 5

Conclusions et Perspectives

« Une action n'est bonne que si le principe auquel elle se soumet peut être érigé en loi universelle. »

André Comte-Sponville

Mon parcours de recherche a commencé il y a une quinzaine d'année sur un sujet autour de la modélisation d'antenne par des sources équivalentes (thèse débutée en octobre 2004). Depuis cette date, j'ai travaillé dans divers contextes et dans différents laboratoires. J'ai acquis des compétences dans plusieurs domaines de l'électromagnétisme et je compte dans les 5 années à venir mettre à profit ces compétences et connaissances consolidées pour approfondir mon activité autour de l'imagerie du proche sous-sol. Je suppose avoir atteint une maturité suffisante pour proposer des solutions antennaires compatibles avec les besoins de l'application en vue. La caractérisation et la modélisation d'antennes sont des composantes importantes dans mon parcours de recherche qui s'intègrent parfaitement dans mon choix de poursuivre mes activités futures autour du radar de sol.

Je souhaite définir mon projet de recherche autour du développement de nouveaux dispositifs et techniques pour l'imagerie du proche sous-sol par les ondes électromagnétiques à travers une approche multi-vue et multi-statique dans la bande fréquentielle [200 MHz - 3 GHz]. Ce projet pluridisciplinaire nécessite le développement de plusieurs compétences qui ouvriront potentiellement des collaborations avec d'autres équipes (locales, nationales et internationales) qui évoluent dans le contrôle non-destructif. Nous avons au laboratoire un environnement expérimental riche et un savoir-faire numérique intéressant. Au sein du laboratoire, j'ai travaillé avec plusieurs spécialistes, chacun dans son domaine, et j'envisage de poursuivre cette interaction avec mon écosystème pour développer mon projet de recherche. Les travaux de thèse de X. Liu (2014 - 2017) sont un prélude de mon projet de recherche. A. Aboudourib (2017-2020) développe dans sa thèse une application particulièrement intéressante pour le radar de sol. Durant cette thèse, j'ai réalisé que la confrontation des solutions développées aux mesures sur le terrain était primordiale. En conséquence, les prototypes et dispositifs à mettre au point devront permettre d'effectuer des mesures sur le terrain.

Dans cette partie du rapport je décrirai les perspectives associées à chaque phase de mon parcours de recherche pour faire converger l'ensemble vers mon projet pour les cinq prochaines années.

Dans le cadre de mes travaux sur la modélisation comportementale d'antennes, j'ai montré les potentialités d'une représentation équivalente à base de dipôles et les résultats qu'elle produit dans le domaine fréquentiel. Je souhaite développer une version simplifiée (modèle d'ordre réduit) de la représentation des antennes à base de dipôles équivalents dans le domaine temporel. Ce modèle doit reproduire le rayonnement de l'antenne dans un secteur angulaire limité (footprint) et être construit à partir du champ proche mesuré en face de l'antenne en espace libre. Cette stratégie de modélisation est applicable à l'antenne seule et vise à paramétrer le problème pour un compromis précision - complexité. Cela contribue à la concrétisation de la réflexion sur la possibilité de produire une bibliothèque de modèles simplifiés pour représenter le champ incident des antennes présentes dans les radars du commerce. Il s'agit de caractériser le radar en mesurant le champ proche incident produit à partir duquel on détermine un modèle équivalent capable de rayonner le même champ. Le modèle équivalent profiterait à tous les utilisateurs de la communauté GPR.

En considérant le couplage entre l'antenne du radar et le sol, les pondérations des dipôles équivalents doivent être recalculées. Une première hypothèse consiste à travailler dans des conditions où le sol ne perturbe pas significativement l'adaptation de l'antenne. Dans le cadre de cette hypothèse on peut supposer

dans une première approximation que le modèle équivalent établi en espace libre reste valable en présence du sol. L'établissement d'un modèle équivalent apporte une meilleure connaissance du champ incident dans l'analyse du problème direct et lors de la résolution du problème inverse. Dans le cas où l'antenne doit prendre place dans la structure multi-statique, il faut définir le modèle équivalent en considérant ce nouvel environnement et prendre en ligne de compte le couplage entre antennes. Le banc de caractérisation mis en place dans le domaine temporel sera utilisé pour mesurer le champ proche rayonné par les antennes et le modèle équivalent se doit d'être compatible avec le code FDTD de GprMax. L'utilisation du modèle équivalent a pour seule finalité l'accélération des simulations numériques tout en garantissant une meilleure description du champ incident dans la zone d'investigation.

Le problème de la prise en compte du diagramme de rayonnement des antennes dans la modélisation du scénario GPR se pose quand elles sont directives. Cependant, nous pouvons analyser ce problème autrement. Nous pouvons choisir de travailler avec des solutions antennaires de petites tailles et qui produisent un rayonnement dipolaire et omnidirectionnel. Dans ce cas particulier, l'approximation de l'antenne par un simple dipôle dans l'analyse électromagnétique du problème direct prend tout son sens. Cela accélère significativement l'analyse mais la dynamique de mesure est plus faible qu'avec des antennes directives.

Dans mon travail de recherche sur l'élaboration de méthodes de mesure en champ proche, j'ai développé des compétences en instrumentation hyperfréquence. Cette activité expérimentale a été à l'origine de plusieurs collaborations nationales et internationales. Je souhaite poursuivre cette ouverture en augmentant nos capacités expérimentales. Actuellement, je développe les procédés d'automatisation d'un banc de mesure en champ proche sphérique élévation/azimut récemment acquis. Il est destiné à caractériser des antennes dans les fréquences au delà de 20 GHz pour les applications de téléphonie 5G. Ce travail est orienté dans une perspective de développement partenarial et de consolidation de notre implantation socio-économique historiquement ancrée dans la tradition du laboratoire.

Je souhaite enrichir le travail commencé sur la caractérisation d'antennes ultra large bande dans le domaine temporel. La méthode matricielle a démontré sa pertinence dans le calcul du champ lointain à partir d'une mesure en champ proche sur une surface à géométrie quelconque. Il serait intéressant de l'adapter pour le cas des mesures dans le domaine temporel en exploitant un maillage adaptatif et un échantillonnage irrégulier. À moyen terme, je projette de développer des techniques de mesure en champ proche dans des environnements non-anéchoïques. Pour cela, on peut dans un premier temps combiner les techniques de filtrage modal dans le traitement fréquentiel des données et le fenêtrage temporel pour le traitement dans le domaine temporel du champ rayonné. L'approche temporel en comparaison avec les mesures dans le domaine fréquentiel propose une dynamique de mesure nettement plus faible, néanmoins l'observation chronologique des événements facilite l'interprétation des phénomènes d'interaction avec l'environnement.

A travers les thèses de X. Liu (2014 - 2017) et A. Aboudourib (2017-2020), j'ai exploré les techniques de modélisation électromagnétique du problème direct. Nous l'avons abordé à travers la méthode du Galerkin Discontinu dans le domaine temporel (code DGTD développé par A. Kameni), la méthode des intégrations finies (CST-MWS code commercial) et la méthode des différences finies dans le domaine temporel (code FDTD GprMax open source). J'ai pris conscience, pour ma part, que l'utilisation des codes open source et l'émergence du logiciel libre d'une manière générale allaient devenir indissociables de ma pratique de la recherche. Dans le cadre de la thèse de A. Aboudourib nous avons fait le choix payant de travailler avec GprMax pour deux raisons : il intègre des modèles pertinents de sol et propose un modèle fractal pour décrire un environnement hétérogène. L'utilisation de cet outil ouvert incite les utilisateurs à contribuer en proposant des briques supplémentaires en fonction de leur propre besoin. Je pourrais proposer les fichiers python de description des antennes ultra-large bande optimisées sur l'intervalle fréquentiel [0.5 GHz - 3 GHz] à l'image de l'antenne vivaldi, l'antenne papillon ou l'antenne spirale chargée par des résistances distribuées. La méthode des dipôles couplées (volume-integral equation) est une autre solution pour la modélisation du problème direct. P. Chaumet de l'Institut Fresnel propose l'utilisation de son logiciel basé sur cette méthode. C'est une voie intéressante pour explorer une nouvelle méthode d'analyse dans le GPR.

Les résultats des modélisations numériques s'appuient sur des validations expérimentales. Nous avons utilisé un bac à sable (1m x 0.6m x 0.5m) pour valider nos travaux sur le radar de sol. Ce banc de test est incompatible avec mon ambition de développer des systèmes multi-statiques. Il est nécessaire de mettre en place un banc de test plus grand (2.5m x 1m x 1m) muni d'un système de déplacement robotisé pour pouvoir faire des acquisitions en C-scan et tester nos algorithmes d'inversion 3D. Ce dispositif servirait potentiellement à produire des mesures associées à des cibles standardisées (canoniques) et proposer ces données aux inverseurs pour tester leurs algorithmes à partir des paramètres S_{ij} mesurés sur les ports des antennes. Pour cela, il faut accompagner les mesures d'un développement destiné à (i) la calibration des

données mesurées, (ii) le filtrage du bruit de mesure, (iii) le filtrage de l'effet des antennes, (iv) l'élimination de l'interface air-sol, (v) la prise en compte de la rugosité de l'interface, (vi) la réduction du clutter..

Nos travaux sur la résolution du problème inverse en GPR sont en cours de consolidation des connaissances et du développement d'outils. Nous avons testé les trois techniques de migration : Kirchhoff, strip-map SAR et F-K. Nous avons développé la Linear Sampling Method (LSM) comme technique d'imagerie qualitative adaptée pour une approche multi-statique. Nous développons actuellement des travaux sur la méthode du filtrage adapté (matched filter). Cette famille de solutions qualitatives peut être complétée par la technique SAR multi-statique. Il serait intéressant de développer une étude numérique pour tester l'apport des différentes configurations géométriques et des positions spatiales des émetteurs et des récepteurs.

Il est nécessaire d'aborder l'extension de la méthode LSM au cas 3D et sur le court terme l'adapter pour faire de l'imagerie quantitative. Il serait intéressant de combiner une technique de migrations 3D (approche temporelle) pour localiser rapidement la cible et délimiter la zone d'investigation pour ensuite adopter une approche fréquentielle via la LSM qualitative/quantitative pour déterminer les propriétés diélectriques de la cible. Une collaboration avec M. Lambert (GeePs) dans cette activité serait très constructive. L'approche fréquentielle, à mon sens, est plus efficace dans la résolution du problème inverse.

Les méthodes d'inversion sont utilisées en même temps que les techniques de réduction du « clutter ». Nous en avons testé plusieurs; à savoir le filtrage numérique (analyse spectrale du champ diffracté), la décomposition en valeurs singulières (SVD) et la décomposition en composantes principales (PCA). Je souhaite développer et tester les potentialités de la technique de fenêtrage temporel basée sur l'entropie des signaux mesurés.

Il serait intéressant de poursuivre l'analyse sur la LSM à partir de la formulation qui exploite les paramètres S mesurés au niveau des ports des antennes. Ces travaux et cette écriture n'ont pas été validés expérimentalement. Je suis conscient de la difficulté hyperfréquence qui réside dans la mise au point d'un système multi-statique (MIMO- Multiple Input Multiple Output), mais le challenge mérite qu'on s'y intéresse. En effet, il faut développer des techniques de calibration et de compensation pour corriger les mesures au niveau de chaque antenne que comporte le système multi-statique. La part du développement hyperfréquence est stimulante à plus d'un point. Le prototype pourrait ouvrir des perspectives d'exploitation dans l'imagerie et la vision à travers les murs.

J'ai toujours veillé à inscrire ma recherche dans un contexte applicatif. Cela a été le cas pour le développement d'antennes pour le GPR, pour la détection de la jonction entre les câbles souterrains et c'est aussi le cas pour l'imagerie du système racinaire des arbres. Ce dernier point a une tonalité particulière car j'ai réalisé que les mesures sur le terrain manquaient dans mon parcours. Je travaille actuellement avec A. Saintnoy du Geops (Géoscience Paris-sud) pour mener des mesures sur le terrain avec le radar Mala ProEx muni de la technologie HDR, que je viens d'acquérir, (livraison au mois de septembre 2019) avec des antennes pour les fréquences 450 MHz et 750 MHz.

Dans le travail de thèse de A. Aboudourib, nous avons mené deux campagnes de mesure sur le terrain avec un système bi-statique (deux antennes papillons) et un analyseur de réseau [0.5 GHz - 3 GHz]. Nous avons réalisé que cartographier un système racinaire dans son environnement réel reste très compliqué. Sur le plan numérique, il est évident qu'un contraste prononcé entre la racine et son environnement (hétérogène) est suffisant pour qu'elle soit détectable. Néanmoins sur le terrain, on ne dispose pas d'informations concernant ce contraste ni sur le nombre de racines à imager. Dans les prochains mois et jusqu'à la fin de la thèse de A. Aboudourib (octobre 2020), nous devons proposer une preuve de concept de la viabilité de l'approche GPR appuyée par des données récoltées sur le terrain. Nous développerons une étude paramétrique sur le plan numérique des configurations multi-statiques combinées avec les techniques d'inversion pour analyser différentes structures racinaires. Cela marquera notre contribution sur le plan numérique.

En dernier lieu, ce projet ne saurait voir le jour sans le concours des doctorants. Pour concrétiser les propositions ci-dessus, je souhaite lancer une première thèse sur le développement instrumental pour la réalisation et la matérialisation d'un prototype d'imagerie multi-statique. Les antennes sont déjà disponibles et il faut développer la partie matrice de switch et les parties de correction et de calibration. La seconde thèse sera destinée à adapter la LSM pour faire de l'imagerie quantitative 3D.