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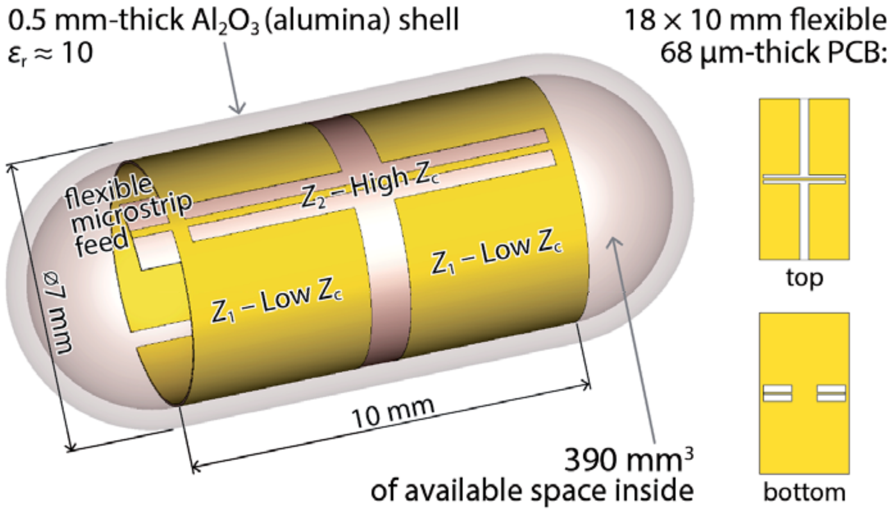

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Our Papers: Negative Group Delay, Refractivity Profiles, and In-Body Antennas

Our first two papers share a common theme: negative group delay. A circuit with negative group delay typically has the property that a time-limited signal (e.g., a pulse) centered at a frequency at which the negative group delay occurs will appear to exit the circuit before it enters the circuit. The first two papers in this issue examine the use of different topologies for realizing the negative group delay functionality.

The paper by B. Ravelo provides a basic overview of negative group delay RF circuits, including those covering the standard categories of filters: low-pass, bandpass, high-pass, and stop-band. The history of negative group delay circuits is reviewed, followed by an introduction to the basic topology and theory. The relationship to filter theory is explained. Methods for synthesizing, analyzing, and characterizing each of the four categories of negative group delay circuits are presented. Examples are given for each, using parallel and series *R*, *L*, and *C* components. Proof-of-concept designs and simulations are provided for the various cases.

The paper by Chung-Tse Michael Wu describes some of the latest developments in passive and active negative group delay circuits using distributed topologies with transmission-line elements. A survey that contrasts the lumped-element and transmission-line approaches to achieving negative group delay is first presented. Negative group delay circuits based on transversal and recursive filters are then described and analyzed. Both active and passive realizations are described. A FET-based distributed active negative group delay circuit is introduced. Simulations were used to validate the proposed design.

Numerical weather prediction models typically provide vertical refractivity profiles from altitudes of 5 m to 10 m up to the stratosphere. Monin-Obukhov-Similarity-based turbulence models provide the refractivity profile in these first 5 m to 10 m above the sea's surface. However, a robust technique is essential for blending the profiles produced by these two techniques without producing nonphysical artifacts. The paper by Robert Marshall, Victor Wiss, Katherine Horgan, William Thornton, and Janet Stapleton presents such a technique. The technique differs from other approaches in two key ways that should make it more physically sound and more robust. A multi-wavelength data-comparison process using the new technique did indicate improved blending.

Our last paper is by the winner of the "Prix Étudiant 2017" (2017 Student Paper Prize) of the French URSI Committee: Denys Nikolayev. He, Maxim Zhadobov, Pavel Karban, and Ronan Sauleau have provided us with a fascinating look at how the radiation performance of conformal antennas for miniature in-body devices can be improved. After a review of the problem of in-body devices and antennas in general, the authors look at the types of antennas that are used for in-body applications and their performance. The various factors that can be optimized to improve the in-body-to-out-of-body operating range are then considered. A method for synthesizing conformal in-body antennas is presented. A design example is presented, going through the whole synthesis, design, fabrication, and testing process. It was shown that the example design could operate at ranges of up to 14 m from various in-body sites.

Our Other Contributions

Stefan Wijnholds' Early Career Representatives Column Presents An Interview With The Recipient Of The 2017 Ursi Koga Gold Medal, Yue Li. His Perspectives On His Career To Date Should Be Of Interest Both To Those At An Early Stage In Their Careers, And To Those Who Work With Early-Career Radio Scientists.

Ethically Speaking, by Randy Haupt and Amy

Shockley, looks at how information can easily be deceptive. The illustration involves a story about coconuts. In reading it, I was reminded of some of the claims I've heard made by politicians – and, unfortunately, of some of the claims I've heard made by scientists and engineers trying to sell research proposals!

In his History Corner, Giuseppe Pelosi brings us a fascinating look at an interesting footnote to electromagnetics (a key part of which is literally contained in a footnote!). The column contains two separate papers dealing with connections between Michael Faraday and Italian physics. The first paper, by Stefano Selleri, examines the case of Francesco Zantedeschi, who claimed priority over Faraday's discovery of magnetic induction. The second paper, by Leonardo Possenti and Stefano Selleri, looks at Leopoldo Nobili, who developed the "astatic galvanometer" that Faraday used in demonstrating magnetic induction. As always, the carefully researched references and marvelous old figures make these most interesting reading.

Ali Farshkaran and Özgür Ergül bring us two problems in the latter's Solution Box column this time. Both involve periodic structures. The first structure is a metamaterial based on split-ring resonators. The second structure is a large array of RF antenna feeds. Both problems involve a large range of scale sizes, and both make use of the Equivalence-Principle Algorithm. Remember: a major intent of the Solution Box is to motivate input from our readers, in the form of either different (and hopefully, better) solutions to the same problems, or new and challenging problems in search of solutions.

In his Telecommunications Health and Safety column, James Lin recalls the interesting history of the microwave signals beamed at the US Embassy in Moscow in the middle to the latter part of the 20th century. His reexamination of what occurred then, in light of what we know now, brings a new perspective to the story.

Asta Pellinen Wannberg presents a Women in Radio Science column by Lee-Anne McKinnell. Dr. McKinnell began her career by modeling the ionosphere. She now is the Managing Director for Space Science within the South African National Space Agency (SANSA). Her story of how she got from that beginning to where she is now provides useful insight into how radio scientists can move from science to management and still make significant contributions to science.

AP-RASC is coming

The URSI Asia-Pacific Radio Science Conference (AP-RASC 2019) will be held March 9-15, 2019, in New Delhi, India. A call for papers appears in this issue. The submission deadline is October 1, 2018. I urge you to submit a paper and plan on attending this important conference.

Ross

On Low-Pass, High-Pass, Bandpass, and Stop-Band NGD RF Passive Circuits

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Abstract

The negative group delay (NGD) function is one of the most intriguing phenomena that can be encountered in area of physics. Since the first experiments in the early 1980s, this abnormal phenomenon remains unfamiliar to conventional engineering. Based on physical, analytical, and experimental investigations, it was found that the NGD phenomenon corresponds to the advanced effect of a pulse signal at baseband frequency. This paper introduces the basic understanding of this extraordinary phenomenon. The RF and microwave-circuit theory dedicated to NGD passive circuits based on the *S*-parameter modeling is established. Two general types of lumped-element-based RF passive topologies, constituted of simple parallel and series mounted impedances, are explored. Low-pass, high-pass, bandpass, and stop-band elementary topologies of passive cells are treated and characterized. The low-pass NGD cells are first-order passive circuits consisting of *RL*-series mounted in parallel and *RC*-parallel networks mounted in series. The different steps of the circuit-parameter calculations as functions of the expected NGD level, insertion loss, and reflection coefficient are explained. To validate the theory, proofs-of-concept (POC) of NGD passive circuits were synthesized, designed, and simulated. As predicted in theory, the simulation results confirmed the feasibility of the low-pass, high-pass, bandpass, and stop-band NGD functions. The NGD circuits respected the basic criteria of RF and microwave circuits with reflection coefficients better than -12 dB. Despite the attenuation, the simulated low-pass NGD cell presented an NGD level of about -1 ns and a cutoff frequency equal to 46 MHz. The high-pass NGD cell presented a group-delay frequency response opposite to the low-pass NGD cell. The bandpass NGD cell exhibited an NGD level of about -2 ns at the center frequency of 0.5 GHz, and a bandwidth of about 92 MHz. Furthermore, the counterintuitive time-advance effect, induced by the NGD phenomenon, proven with various arbitrary-waveform baseband signals, was observed. A pulse advancement of about -1 ns was obtained with the considered low-pass passive cell.

1. Introduction

In the 1970s, an analytical investigation of the propagation of a Gaussian light pulse through an anomalous dispersion medium with negative refractive index was reported by Garrett and McCumber [1]. This theoretical result enabled understanding the negative group delay (NGD) effect in the time domain. After this theoretical conceptualization, the first experiments on NGD were performed in 1982 by Chu and Wong [2]. This experimental demonstration was based on the atomic-vapor medium, and showed the NGD phenomenon at optical wavelengths. However, several physicists remained skeptical about the existence of the NGD phenomenon. The NGD phenomenon was therefore assumed to typically be artifacts of numerical calculations, and not a real physical phenomenon. Consequently, this unfamiliar phenomenon was not sufficiently developed and less interesting to scientists, researchers, and engineers over the 1980s.

Despite this period of low interest, a new era of NGD functionality blew in with the 1990s with the unbelievable demonstrations with basic electronic circuits composed of familiar electronic components such as resistances, inductances, and capacitors. One decade after the first NGD demonstration by Chu and Wong, the first experiments with lumped electronic circuits was performed by the team of Chiao [3-6]. The NGD circuit topology was identified from an analogy between the atomic-vapor medium and the electronic-circuit's transfer function [3, 4]. Theoretical and experimental studies were then conducted with active optical systems [7-9] and electronic systems operating with audio signals [10-12] in order to illustrate the existence of the NGD phenomenon. During this period, one of the NGD effect's basic meanings was explained by electronic experiments considering pulse-shaped voltage signals with millisecond duration. It was visualized that the output pulse's leading and trailing edges occurred before the input pulse had completely penetrated into the NGD circuit. It was emphasized that this counterintuitive NGD phenomenon did not contradict the causality effect [5, 6, 10, 11]. However, at this stage, the complexity and operational speed limitations linked to the operational amplifiers did not motivate the electronic design engineers to exploit the NGD concept.

Some years later, the NGD phenomenon was synthesized with microwave and millimeter-wave circuits that were able to operate at higher frequencies [13-15]. In the same period, the NGD phenomenon was also generated with metamaterial microwave circuits, especially those based on periodic transmission-line structures [16-18]. Based on the microwave design and analyses, it was found that the NGD effect was generally generated with circuits that were typically lossy and resonate [13-24]. The NGD effect appeared around the resonance frequency. However, it was understood from time-domain experiments that this counterintuitive NGD effect was systematically accompanied with significant losses [18, 19]. This combined effect was inherently linked to the absorption around the resonance frequency. Because of the signification loss, there was little interest in the NGD concept during several decades.

To tackle this technical problem, microwave circuit topologies, based on the field-effect transistor (FET) and the low-noise amplifier (LNA), were proposed [25-28]. An analytical methodology enabling identification of simple NGD active microwave topologies was elaborated [26]. Several basic elementary NGD topologies susceptible to generating the NGD function were identified. A generalized NGD circuit theory was introduced. It was emphasized that the NGD function was similar to linear filter gain [29]. Low-pass, high-pass, bandpass, and stop-band NGD cells were identified. It was reported that the low- and high-pass NGD cells were typically first-order linear circuits.

An innovative *RC*-network-based high-pass NGD circuit was introduced in [30]. The bandpass and stop-band elementary cells were based on second-order resonating circuits. The NGD circuit's fundamental properties, such as the NGD level, the NGD cutoff frequencies, the NGD bandwidth, and the NGD figure-of-merit (FoM) were established and described. NGD circuit theory has been made widely available with the testing of several basic and elementary circuits. The feasibility of the theoretical concept has been validated by simulations and experiments. NGD effects have been simultaneously generated with the possibility of gain and loss compensation. Despite these developments in NGD circuit-design methodology, it was emphasized that the NGD effect does not allow any time-advance configuration because of its inherent properties [31].

In addition to NGD design and synthesis, potential applications – notably, *RC*-interconnection propagation delay [32] cancellation – have been forecasted since the 2000s [33, 34]. The feasibility of this delay cancellation was later extended to signal recovery for signal-integrity enhancement, as suggested in [35-37]. Further potential NGD applications for high-performance feed-forward amplifier design were also proposed [38].

Despite this trend of NGD theory becoming widely available, RF and electronic engineers are still not familiar with the NGD function. Further illustration is therefore needed for the NGD function to achieve further understanding in the academic and industrial points of view. In the present paper, we address a simple theoretical approach making it easy to understand the NGD function with lumped passive elementary cells. The aim of the paper is to make the NGD concept familiar to electrical-engineering students, researchers, and those in industry. The present paper is also aimed at making the NGD concept and theory as familiar as possible to the non-specialist. In doing this – based on the analogy between NGD and filter theory [29] – the low-pass, high-pass, bandpass, and stop-band elementary lumped *RC* cells are investigated. This microwave circuit theory is developed from particularly simple low-pass topologies of lumped-element components consisting of an *RL* series network and an *RC* parallel network. These act as first-order circuits that can be investigated in a manner similar to any familiar electrical function. A theoretical approach based on the *S*-parameter analysis is introduced in Section 2. The theory also presents the basic NGD properties and synthesis methodology for the basic NGD cells. By considering the familiar, known low-pass-to-bandpass circuit transformation in filter theory, bandpass NGD cells are also then investigated. The validity of the proposed NGD theory, simulations, and experimental results is presented and explored in Section 3. Both frequency- and time-domain analyses are deployed in order to show the NGD phenomena. Moreover, the ability to tune the NGD function level of the circuit parameters is tested for the deep understanding about the ability to control the NGD function with the lumped-circuit parameters. The conclusion of the paper is then presented in Section 4.

2. Generality of the Passive Topologies Under Study

The present section is focused on a simple way to design low-pass and bandpass NGD circuits, based on the familiar electronic components *R*, *L*, and *C*. After exploring the NGD effect's existence with two passive lumped-element-based topologies, we will investigate the *S*-parameter modeling methodology. The theoretical characterization of the active circuit obtained will then be described. At the end of each step, synthesis expressions enabling one to easily calculate the NGD circuit parameters as a function of the NGD specifications will be formulated.

2.1 Introduction to Basic Topology

The present subsection generally describes the elementary passive topologies of the circuit to be developed in this paper.

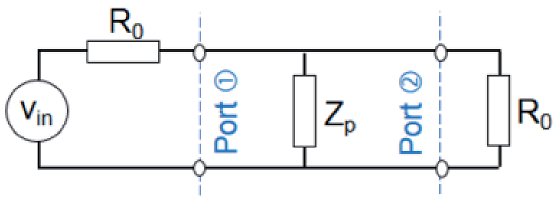

Figure 1. A shunt-impedance Z_p type passive topology.

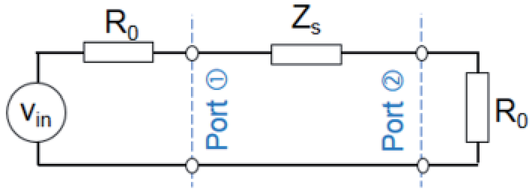

Figure 2. A series-impedance Z_s type passive topology.

Similar to all classical electronic and electrical circuits, the simplest and elementary topologies are based on the two-port circuit as a simple impedance that can be denoted Z. As pointed out in [25, 26, 29], we can start our investigation with a simple two-port circuit composed of a parallel impedance, Z_p, as depicted in Figure 1, or a series impedance, Z_s, as depicted in Figure 2. It should be underlined that in this paper, R_0 represents the reference impedance, equal to 50 Ω.

The present study is based on this two-port circuit S-parameter analysis. For the case of passive topologies, the associated S parameters can be defined by the symmetric relationship

$$[S]=\begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}. \quad (1)$$

2.2 S-Parameter Modeling

The present subsection describes the S-parameter modeling of the passive cells introduced in Figure 1 and Figure 2 of the previous subsections. The constituting element impedances, $Z_p(j\omega)$ and $Z_s(j\omega)$, are assumed to depend on the angular frequency, ω. After the determination of branch currents and applying the definition of the S parameters, we get the following expressions:

For the parallel-impedance-based cell:

$$[S_p(j\omega)]=\begin{bmatrix} \frac{-R_0}{R_0+2Z_p(j\omega)} & \frac{2Z_p(j\omega)}{R_0+2Z_p(j\omega)} \\ \frac{2Z_p(j\omega)}{R_0+2Z_p(j\omega)} & \frac{-R_0}{R_0+2Z_p(j\omega)} \end{bmatrix}, \quad (2)$$

For the series-impedance-based cell:

$$[S_s(j\omega)]=\begin{bmatrix} \frac{Z_s(j\omega)}{2R_0+Z_s(j\omega)} & \frac{2R_0}{2R_0+Z_s(j\omega)} \\ \frac{2R_0}{2R_0+Z_s(j\omega)} & \frac{Z_s(j\omega)}{2R_0+Z_s(j\omega)} \end{bmatrix}. \quad (3)$$

It can be emphasized that the reflection and transmission coefficients of these passive circuits are linked by the relationship

$$S_{11}(j\omega)+S_{21}(j\omega)=1. \quad (4)$$

This means that the circuit must generate significant insertion and reflection losses.

2.3 NGD Circuit Theory

Differently from the classical RF/microwave circuit theory, the unfamiliar NGD circuit theory is based on the sign of the group-delay analysis.

2.3.1 NGD Theoretical Analysis and Similarity to Filter Behavior

By denoting by $j\omega$ the circuit angular-frequency variable, according to circuit and system theory, it can be recalled that the group delay is defined by

$$\tau(\omega)=-\frac{\partial\varphi(\omega)}{\partial\omega}, \quad (5)$$

with

$$\varphi(\omega)=\angle S_{21}(j\omega), \quad (6)$$

Table 1. A comparison of the magnitude and group delay of NGD functions and conventional normal filters.

Characteristics	Normal Filter	NGD Function
Basic specifications	Transfer function gain	Transfer function group delay
Cutoff frequencies ω_c	Root of the equation $\left\|S_{21}\left(j\omega_c\right)\right\| = \left\|S_{21}\left(j\omega\right)\right\|_{\max} / \sqrt{2}$	Root of the equation $\tau(\omega_c) = 0$
Bandwidth	$\left\|S_{21}\left(j\omega_c\right)\right\| \geq \left\|S_{21}\left(j\omega\right)\right\|_{\max} / \sqrt{2}$	$\tau(\omega) \leq 0$
Low-pass with bandwidth defined by $\omega \leq \omega_c$	In the bandwidth, we have $\left\|S_{21}\left(j\omega_c\right)\right\| \geq \left\|S_{21}\left(j\omega\right)\right\|_{\max} / \sqrt{2}$	In the bandwidth, we have $\tau(\omega) \leq 0$ and $\left\|S_{21}(j\omega)\right\|$ should be as flat as possible.
High-pass with bandwidth defined by $\omega \geq \omega_c$		
Bandpass with bandwidth defined by $\omega_{c_1} \leq \omega \leq \omega_{c_2}$		
Stop-band with bandwidth defined by $\omega \leq \omega_{c_1}$ and $\omega \geq \omega_{c_2}$		

being the transmission phase, expressed in radians. This last expression explains that our impedance cannot be a frequency-independent component, such as a simple resistance. In other words, to generate nonzero group delay, a reactive-element-based circuit is necessary.

It is worth noting that the group delay behaves similarly to the filter gain [29]. The NGD bandwidth corresponds to the frequency band where the group delay $\tau(\omega) < 0$. The cutoff angular frequency, ω_c, is analytically the root of the equation $\tau(\omega_c) = 0$. We can identify different types of group-delay cells. We have:

- A low-pass NGD function when the group delay is negative, $\tau(\omega) < 0$, in the lower frequency band, $\omega \leq \omega_c$ or from $\omega \approx 0$ to the cutoff angular frequency, ω_c.
- A high-pass NGD when the group delay is negative, $\tau(\omega) < 0$, in the upper frequency band, $\omega \geq \omega_c$.
- A bandpass NGD when the group delay is negative, $\tau(\omega) < 0$, in the frequency band $\omega_{c_1} \leq \omega \leq \omega_{c_2}$ by supposing $\tau\left(\omega_{c_1}\right) = \tau\left(\omega_{c_2}\right) = 0$. In this case, the NGD central frequency, ω_0, can be verified with

$$\tau\left(\omega_0\right) = \min\left[\tau\left(\omega\right)\right] < 0 . \quad (7)$$

The NGD bandwidth can respectively be written as

$$\Delta\omega = \omega_{c_2} - \omega_{c_1} . \quad (8)$$

- A stop-band NGD when the group delay is negative, $\tau(\omega) < 0$, in the frequency band $\omega \leq \omega_{c_1} \cup \omega_{c_2} \leq \omega$ by supposing $\tau\left(\omega_{c_1}\right) = \tau\left(\omega_{c_2}\right) = 0$.

The performance of the passive NGD circuit can be evaluated with the figure-of-merit (FoM) defined by

$$FoM_{low-pass} = \left|S_{21}\left(\omega \approx 0\right)\right| \tau\left(\omega \approx 0\right) \omega_c \quad (9)$$

$$FoM_{band-pass} = \left|S_{21}\left(\omega \approx \omega_0\right)\right| \tau\left(\omega = \omega_0\right) \Delta\omega . \quad (10)$$

The better is the NGD circuit, the higher is the figure-of-merit.

2.3.2 A Comparison Between NGD Functions and Normal Conventional Filters

Despite the previous definitions of the NGD function types, one may wonder or be confused about the NGD and classical-filter aspects. Table 1 provides a comparative illustration between NGD functions and conventional normal filters. Based on circuit theory, the conventional normal filter's bandwidths are defined where the transmission gain's magnitude satisfies the equation $\left|S_{21}\left(j\omega\right)\right| \geq \left|S_{21}\left(j\omega\right)\right|_{\max} / \sqrt{2}$. As previously described, the NGD functions are mainly based on the sign of the group delay.

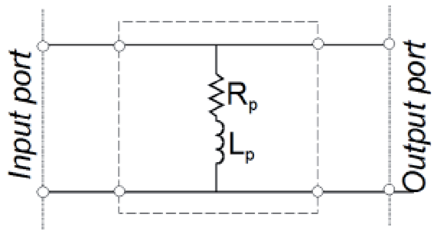

Figure 3. A low-pass NGD cell: *RL*-series shunt network.

2.3.3 Low-Pass NGD Passive Circuit Synthesis Method

Based on RF/microwave engineering, the main circuit parameters are basically the access matching and the insertion loss for passive circuits. Those parameters are still under consideration in the present study. However, in addition, we also deal with the group delay defined in Equation (5). For the low-pass NGD circuit, the NGD characterization described in this section is mainly obtained from the analytical expression of the S parameters and the group delay at very low frequencies, when $\omega \approx 0$. The synthesis method consists of the calculation of the NGD circuit's parameters as a function of the specified NGD level $\tau(\omega \approx 0) = \tau_0 < 0$, which is inversely linked to the NGD bandwidth; the insertion loss; and the reflection coefficients. In other words, the basic relations can be summarized in the following system:

$$\begin{cases} \left|S_{11}(\omega \approx 0)\right| = s_{11} \\ \left|S_{21}(\omega \approx 0)\right| = s_{21} \\ \left|S_{22}(\omega \approx 0)\right| = s_{11} \\ \tau(\omega \approx 0) = \tau_0 \end{cases}. \tag{11}$$

3. Shunt-Impedance-Type Elementary NGD Cells

The present section is focused on the NGD characterization of the family of shunt-impedance-type elementary NGD cell shown in Figure 1. Based on classical circuit theory, the simplest network configurations constituting the shunt impedance Z_p must be first-order circuits. This intuitive analysis leads us to consider *RL* and *RC* networks to build a low-pass NGD cell that will be analyzed in the present section.

3.1 Low-Pass Cell

The analytical identification performed in [26] stated that the only configuration of first-order or low-pass NGD circuits associated with the parallel impedance configuration is an *RL*-series network:

$$Z_p(j\omega) = R_p + j\omega L_p . \tag{12}$$

The low-pass NGD elementary cell is depicted in Figure 3.

3.1.1 Frequency-Dependent S Parameters and Group Delay

For the shunt-impedance-based low-pass NGD cell, the S parameters can be rewritten as follows:

$$\left[S_p(j\omega)\right] = \begin{bmatrix} \dfrac{-R_0}{R_0 + 2\left(R_p + j\omega L_p\right)} & \dfrac{2\left(R_p + j\omega L_p\right)}{R_0 + 2\left(R_p + j\omega L_p\right)} \\ \dfrac{2\left(R_p + j\omega L_p\right)}{R_0 + 2\left(R_p + j\omega L_p\right)} & \dfrac{-R_0}{R_0 + 2\left(R_p + j\omega L_p\right)} \end{bmatrix} \tag{13}$$

From this we can determine the following reflection- and transmission-coefficient magnitudes:

$$\left|S_{p11}(j\omega)\right| = \left|\frac{-R_0}{R_0 + 2\left(R_p + j\omega L_p\right)}\right| \Rightarrow S_{p11}(\omega)$$

$$= S_{p22}(\omega) = \frac{2\sqrt{R_0^2}}{\sqrt{\left(R_0 + 2R_p\right)^2 + \left(L_p\omega\right)^2}} \tag{14}$$

$$\left|S_{p21}(j\omega)\right| = \left|\frac{2\left(R_p + j\omega L_p\right)}{R_0 + 2\left(R_p + j\omega L_p\right)}\right| \Rightarrow S_{p21}(\omega)$$

$$= S_{p12}(\omega) = \frac{2\sqrt{R_p^2 + \left(L_p\omega\right)^2}}{\sqrt{\left(R_0 + 2R_p\right)^2 + \left(L_p\omega\right)^2}} .$$

The associated transmission phase is given by

$$\varphi_p(\omega) = \arctan\left(\frac{L_p\omega}{R_p}\right) - \arctan\left(\frac{2L_p\omega}{R_0 + 2R_p}\right). \tag{15}$$

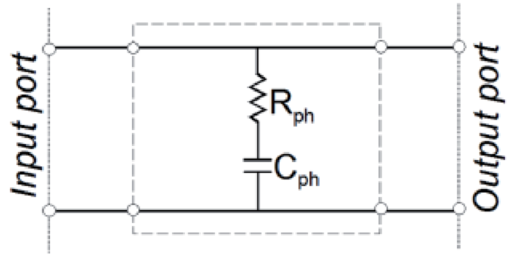

Figure 4. A high-pass NGD cell: *RC*-series shunt network family of the cell shown in Figure 3.

From the group delay defined in Equation (5), we have

$$\tau_p(\omega) = \frac{R_0 L_p\left[2\left(L_p\omega\right)^2 - R_p\left(R_0 + 2R_p\right)\right]}{\left[\left(L_p\omega\right)^2 + R_p^2\right]\left[\left(2L_p\omega\right)^2 + \left(R_0 + 2R_p\right)^2\right]} \quad (16)$$

3.1.2 NGD Characterization

At very low frequencies, the S parameters of the parallel-impedance-based circuit become

$$\left[S_p(\omega \approx 0)\right] = \begin{bmatrix} \dfrac{-R_0}{R_0 + 2R_p} & \dfrac{2R_p}{R_0 + 2R_p} \\ \dfrac{2R_p}{R_0 + 2R_p} & \dfrac{-R_0}{R_0 + 2R_p} \end{bmatrix}. \quad (17)$$

The associated group delay is given by

$$\tau_p(\omega \approx 0) = \frac{-R_0 L_p}{R_p(R_0 + 2R_p)}. \quad (18)$$

It can be understood from the previous expression in Equation (18) that the passive cell introduced in Figure 3 behaves as a low-pass NGD circuit. In addition to the previous characteristics, one of the most important properties of the NGD cell is the NGD cutoff frequency, which can be denoted ω_{c_p}. It is the root of the equation $\tau_p\left(\omega_{c_p}\right) = 0$. The proposed shunt-impedance-based low-pass NGD cell cutoff frequency is given by

$$\omega_{c_p} = \frac{\sqrt{R_p\left(R_0 + 2R_p\right)}}{\sqrt{2}L_p}. \quad (19)$$

The NGD figure-of-merit is expressed as

$$FoM_{low-pass_p} = \frac{-R_0^2}{\left(R_0 + 2R_p\right)\sqrt{2R_p\left(R_0 + 2R_p\right)}}. \quad (20)$$

It can be emphasized that the figure-of-merit is inversely proportional to the resistance R_p.

3.1.3 NGD Synthesis Method

Suppose we are given the desired insertion loss, a, the reflection coefficient, r, and the group delay, $\tau_0 < 0$. Analytically, the synthesis formulas are generated from Equation (18). After some calculations, we obtain the expressions for R_p and L_p associated with the shunt *RL*-series cell. The synthesis relations for this low-pass NGD cell are

$$\begin{cases} R_p(S_{11}) = R_p(S_{22}) = \dfrac{1-r}{2r}R_0 \\ \text{or} \\ R_p(S_{12}) = R_p(S_{21}) = \dfrac{a}{2(1-a)}R_0 \end{cases}, \quad (21)$$

$$\begin{cases} L_p(\tau_0) = -\left(1 + \dfrac{2R_p}{R_0}\right)R_p\tau_0 \\ \text{or} \\ L_p(\omega_c) = \dfrac{\sqrt{R_p\left(R_0 + 2R_p\right)}}{\sqrt{2}\omega_c} \end{cases}. \quad (22)$$

3.2 High-Pass NGD Elementary Cells

The high-pass NGD cell family of the circuit proposed in Figure 3 is depicted in Figure 4.

Similarly to filter theory, the high-pass NGD cells can be easily synthesized from the low-pass cells via the low-pass-to-high-pass transform. As illustrated in Figure 4, the constitute shunt impedance can be expressed as

$$Z_{ph}(j\omega) = R_{ph} + \frac{1}{j\omega C_{ph}}. \quad (23)$$

The present subsection is focused on the NGD characteristics of the bandpass NGD cells associated with the previous cells.

3.2.1 High-Pass NGD Behavior

The S parameters of the series-impedance-based cell shown in Figure 4 become

$$\left[S_{ph}(j\omega)\right]$$

$$= \begin{bmatrix} \dfrac{j\omega R_0 C_{ph}}{2 + j\omega C_{ph}\left(R_0 + 2R_{ph}\right)} & \dfrac{2\left(1 + 2j\omega R_{ph} C_{ph}\right)}{2 + j\omega C_{ph}\left(R_0 + 2R_{ph}\right)} \\ \dfrac{2\left(1 + 2j\omega R_{ph} C_{ph}\right)}{2 + j\omega C_{ph}\left(R_0 + 2R_{ph}\right)} & \dfrac{j\omega R_0 C_{ph}}{2 + j\omega C_{ph}\left(R_0 + 2R_{ph}\right)} \end{bmatrix} \quad (24)$$

Similar to the low-pass-to-high-pass filter transform, the resistance element is not changed: $R_{ph} = R_p$. The associated group delay is given by

$$\tau_{ph}(\omega \approx 0) = \frac{R_0 C_{ph}}{2}. \quad (25)$$

In addition to the previous characteristics, one of the most important properties of the NGD cell is the NGD cutoff frequency, which can be denoted ω_{c_h}. This is the root of the equation $\tau_{ph}\left(\omega_{c_h}\right) = 0$. The proposed shunt-impedance-based high-pass NGD cell cutoff frequency is given by

$$\omega_{c_{ph}} = \frac{\sqrt{2}}{C_{ph}\sqrt{R_p\left(R_0 + 2R_p\right)}}. \quad (26)$$

Moreover, it is also found that for any integer $n > 1$, the group delay can be expressed as

$$\tau_{ph}\left(\omega = n\omega_{c_{ph}}\right)$$

$$= \frac{\left(1 - n^2\right) R_0 R_p C_{ph}\left(R_0 + 2R_p\right)}{\left[R_0 + 2\left(1 + n^2\right)R_p\right]\left[n^2 R_0 + 2\left(1 + n^2\right)R_p\right]}. \quad (27)$$

It can be understood from the expressions in Equations (25), $\tau_{ph}(\omega \approx 0) > 0$, and (27), $\tau_{ph}\left(\omega = n\omega_{c_{ph}}\right) < 0$, which generally means that $\tau_{ph}\left(\omega > \omega_{c_{ph}}\right) < 0$. This means that the passive cell introduced in Figure 8 behaves as a high-pass NGD circuit. The NGD figure-of-merit is expressed as $FoM_{high-pass_p} = FoM_{low-pass_p}$. It can be emphasized that the figure-of-merit is inversely proportional to the resistance R_p.

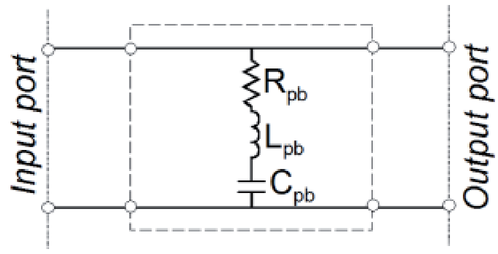

Figure 5. A bandpass NGD cell: *RLC*-series shunt network associated with the elementary cell shown in Figure 3.

3.2.2 Synthesis Method

For the high-pass NGD cell, the resistance parameter $R_{ph} = R_p$ can be synthesized with the relations in Equation (21). However, given the targeted positive group delay $\tau_0 > 0$, the capacitor C_{ph} can be extracted from the high-pass group delay expressed in Equation (25):

$$C_{ph} = \frac{2\tau_0}{R_0}. \quad (28)$$

This high-pass NGD cell capacitor can also be determined from the cutoff frequency expressed in Equation (26):

$$C_{ph} = \frac{\sqrt{2}}{\omega_{c_{ph}}\sqrt{R_p\left(R_0 + 2R_p\right)}}. \quad (29)$$

3.3 Bandpass NGD Elementary Cells

Similarly to filter theory, the bandpass NGD cells can be easily synthesized from low-pass cells via the low-pass-to-band-pass transform. The present subsection is focused on the NGD characteristics of the bandpass NGD cells associated with the previous cells. In this case, to synthesize the associated bandpass NGD cell, we just need to replace the inductance L_p by an LC series network, as shown in Figure 5. Based on filter-circuit theory, during the low-pass-to-bandpass transform, the resistance element is not changed: $R_{pb} = R_p$.

3.3.1 Bandpass NGD Behavior

The proposed cell's resonance frequency, which is also the NGD central frequency, is defined by

$$\omega_0 = \frac{1}{\sqrt{L_{ss}C_s}}. \quad (30)$$

The associated S parameters at the center frequency are expressed as

$$\left[S_{pb}(\omega_0)\right] = \left[S_p(0)\right] = \begin{bmatrix} \frac{-R_0}{R_0+2R_p} & \frac{2R_p}{R_0+2R_p} \\ \frac{2R_p}{R_0+2R_p} & \frac{-R_0}{R_0+2R_p} \end{bmatrix} \quad (31)$$

It can be demonstrated that the associated group delay is

$$\tau_{pb}(\omega_0) = 2\tau_p(0) = \frac{-2R_0L_{pb}}{R_p\left(R_0+2R_p\right)}. \quad (32)$$

This group delay is always negative around the resonance frequency.

3.3.2 Bandpass NGD Characterization and Properties

The NGD center frequency of the passive cell shown in Figure 5 is equal to $\omega = \omega_0$. The NGD bandwidth is defined by

$$\Delta\omega_{c_{pb}} = 2\omega_{c_p} = \frac{\sqrt{2R_p\left(R_0+2R_p\right)}}{L_{pb}}. \quad (33)$$

This means that the NGD cutoff frequencies are defined by

$$\omega_{c_{pb}} = \omega_0 \pm \omega_{c_p} = \frac{1}{\sqrt{L_{pb}C_{pb}}} \pm \frac{\sqrt{R_p(R_0+2R_p)}}{\sqrt{2}L_{pb}} \quad (34)$$

The NGD figure-of-merit is expressed as

$$FoM_{band-pass_p} = 2FoM_{low-pass_p}, \quad (35)$$

which is inversely proportional to the resistance, R_p.

3.3.3 Synthesis Method

The bandpass NGD cell synthesis relationships can be inspired from the low-pass relationships defined in Equations (21) and (22). To achieve the same NGD level, the resistance can be obtained from the same relationship and half of the inductance $L_{ph} = L_p$, which must be considered based on the group delay defined in Equation (32). The associated capacitor can be calculated from the inductance and the expected NGD center angular frequency, which can be denoted ω_0. We therefore have the capacitor-synthesis formula

$$C_{pb} = \frac{1}{\omega_0^2 L_{pb}}. \quad (36)$$

3.4 Stop-Band NGD Elementary Cells

Similarly to filter theory, the stop-band NGD cells can be easily synthesized from the low-pass cells via the low-pass-to-stop-band transform. Figure 6 represents the configuration of the stop-band cell associated with the circuit shown in Figure 3. Similarly to the low-pass-to-stop-band filter transform, the resistance element is not changed, $R_{ps} = R_p$. The present subsection is focused on the NGD characteristics of the bandpass NGD cells associated with the previous cells.

3.4.1 Stop-Band NGD Behavior, Characterization, and Properties

It can be demonstrated that the associated group delay is

$$\tau_{ps}(\omega_0) = R_0C_{ps}. \quad (37)$$

This group delay is always positive around the resonance frequency. The proposed shunt-impedance-based stop-band NGD cell cutoff frequencies are given by

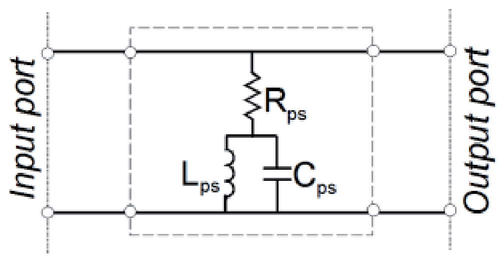

Figure 6. A stop-band NGD cell: R and LC-parallel series shunt network associated with the cell shown in Figure 3.

$$\omega_{c_{ps}} = \frac{1}{\sqrt{L_{ps}C_{ps}}} \pm \frac{\sqrt{2}}{C_{ps}\sqrt{R_{ps}\left(R_0 + 2R_{ps}\right)}} . \quad (38)$$

This means that the NGD bandwidths are

$$\begin{cases} \omega \le \dfrac{1}{\sqrt{L_{ps}C_{ps}}} - \dfrac{\sqrt{2}}{C_{ps}\sqrt{R_{ps}\left(R_0 + 2R_{ps}\right)}} \\ \text{and} \\ \omega \le \dfrac{1}{\sqrt{L_{ps}C_{ps}}} + \dfrac{\sqrt{2}}{C_{ps}\sqrt{R_{ps}\left(R_0 + 2R_{ps}\right)}} \end{cases} \quad (39)$$

For the case of a stop-band cell, the figure-of-merit cannot be defined.

3.4.3 Synthesis Method

The stop-pass NGD cell-synthesis relationships can be inspired from the combination of both low-pass and high-pass relationships. To obtain the attenuation, a, or the reflection coefficient, r, the resistance can be determined from the formula defined in Equation (21). For the given group delay, $\tau_0 > 0$, at the center angular frequency, which can be denoted ω_0 , the capacitor can be calculated with the relationship in Equation (28), $C_{ps} = C_{ph}$. The associated inductance can be calculated from the inductance and the expected NGD center angular frequency, which can be denoted ω_0 . We therefore have the inductance-synthesis formula

$$L_{ps} = \frac{1}{\omega_0^2 C_{ph}} . \quad (40)$$

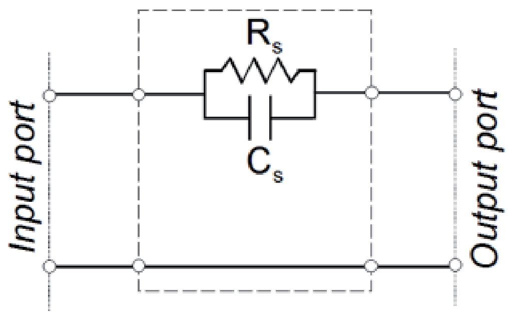

Figure 7. A low-pass NGD cell: *RC*-parallel series network.

4. Series Impedance-Type NGD Passive Cells

Similarly to the study performed in the previous section, the present section is focused on the NGD characterization of the family of series-impedance-type of elementary NGD topology shown in Figure 2. After checking different configurations, the first-order RC-parallel network constitutes the simplest cell to generate the low-pass NGD function.

4.1 Low-Pass NGD Elementary Cell

The analytical identification performed in [29] stated that the only configuration of first-order or low-pass NGD circuits associated with the series-impedance configuration is an RC-parallel network.

$$Z_s(j\omega) = \frac{R_s}{1 + j\omega R_s C_s} . \quad (41)$$

Figure 7 represents the configuration of the elementary low-pass NGD cell.

4.1.1 Frequency-dependent S Parameter and Group Delay

The frequency-dependent S parameter of the proposed impedance-series-type cell can be expressed as

$$\left[S_s(j\omega)\right] = \begin{bmatrix} \dfrac{R_s}{2R_0 + R_s + 2j\omega R_0 R_s C_s} & \dfrac{2R_0\left(j\omega R_s C_s + 1\right)}{2R_0 + R_s + 2j\omega R_0 R_s C_s} \\ \dfrac{2R_0\left(j\omega R_s C_s + 1\right)}{2R_0 + R_s + 2j\omega R_0 R_s C_s} & \dfrac{R_s}{2R_0 + R_s + 2j\omega R_0 R_s C_s} \end{bmatrix} \quad (42)$$

From this we can determine the following reflection-coefficient and insertion-loss magnitudes:

$$\left|S_{s_{22}}(j\omega)\right| = \left|S_{s_{11}}(j\omega)\right| = S_{s_{11}}(\omega)$$

$$= \frac{R_s}{\sqrt{\left(2R_0 + R_s\right)^2 + 4\left(\omega R_0 R_s C_s\right)^2}} \quad (43)$$

$$\left|S_{s_{12}}(j\omega)\right| = \left|S_{s_{21}}(j\omega)\right| = S_{s_{21}}(\omega)$$

$$= \frac{2R_0\sqrt{1+(\omega R_s C_s)^2}}{\sqrt{(2R_0+R_s)^2+4(\omega R_0 R_s C_s)^2}}\ .$$

The associated transmission phase is given by

$$\varphi_s(\omega) = \arctan(\omega R_s C_s) - \arctan\left(\frac{2\omega R_0 R_s C_s}{2R_0+R_s}\right). \quad (44)$$

From Equation (5), we have the group delay

$$\tau_s(\omega) = \frac{R_s^2 C_s\left(2R_0 R_s^2 C_s^2 \omega^2 - 2R_0 - R_s\right)}{\left(1+R_s^2 C_s^2 \omega^2\right)\left[(2R_0+R_s)^2 + 4R_0^2 R_s^2 C_s^2 \omega^2\right]} \quad (45)$$

4.1.2 NGD Characterization

At very low frequencies, the S parameters of the parallel-impedance-based circuit become

$$\left[S_s(\omega\approx 0)\right] = \begin{bmatrix} \frac{R_s}{2R_0+R_s} & \frac{2R_0}{2R_0+R_s} \\ \frac{2R_0}{2R_0+R_s} & \frac{R_s}{2R_0+R_s} \end{bmatrix}. \quad (46)$$

The associated group delay is given by

$$\tau_s(\omega\approx 0) = \frac{-R_s^2 C_s}{R_s+2R_0}. \quad (47)$$

It can be understood from the previous expression in Equation (47) that the passive cell introduced in Figure 3 behaves as a low-pass NGD circuit. In addition to the previous characteristics, one of the most important properties of the NGD cell is the NGD cutoff frequency, which can be denoted ω_{c_s}. It is the root of the equation $\tau_s(\omega_{c_s}) = 0$. The proposed shunt-impedance-based low-pass NGD-cell cutoff frequency is given by

$$\omega_{c_s} = \frac{\sqrt{1+\frac{R_s}{2R_0}}}{R_s C_s}. \quad (48)$$

The NGD figure-of-merit is expressed as

$$FoM_{low-pass_s} = \frac{-R_s}{(2R_0+R_s)\sqrt{R_s+2R_0}}. \quad (49)$$

In this case, it can be emphasized that the figure-of-merit is proportional to the resistance, R_p.

4.1.3 NGD Synthesis Method

The synthesis method consists of calculating R_s and C_s, which can be determined from the system of Equations (11) by considering the proposed series-impedance-type cell. The synthesis formulas are given by

$$\begin{cases} R_s(S_{11}) = R_s(S_{22}) = \frac{2r}{1-r}R_0 \\ \text{or} \\ R_s(S_{12}) = R_s(S_{21}) = \frac{1-a}{2a}R_0 \end{cases}, \quad (50)$$

$$\begin{cases} C_s(\tau_0) = -\frac{R_s+2R_0}{R_s^2}\tau_0 \\ \text{or} \\ C_s(\omega_c) = \frac{\sqrt{1+\frac{R_s}{2R_0}}}{R_s\omega_c} \end{cases}. \quad (51)$$

In the case of reflection-coefficient-based s_p resistance synthesis, the insertion loss is expressed as a function of r by the equation $S_{s_{21}}(\omega\approx 0) = r$. In the case of synthesis from the insertion loss, the reflection coefficient is expressed as a function of a by the equation $S_{s_{11}}(\omega\approx 0) = 1-a$.

4.1.4 Equivalence Between the Parallel-and Series-Impedance Low-Pass NGD Cells

The equivalence between the two parallel- and series-impedance-based passive NGD topologies under study presents an electrical equivalence. By comparison of the transmission-parameter expressions, we have the equivalent relationship between the parallel and series resistances expressed as

$$R_p R_s = R_0^2. \quad (52)$$

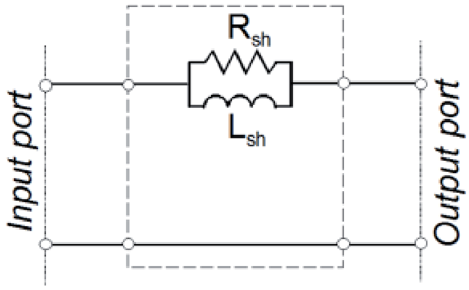

Figure 8. A high-pass NGD cell: *RL*-parallel series network associated with the low-pass cell shown in Figure 7.

In the same manner, the reactive-parameter equivalent relationships can be written as

$$\frac{L_p}{C_s} = R_0^2 . \tag{53}$$

4.2 High-Pass NGD Elementary Cells

The present subsection is focused on the NGD characteristics of the high-pass NGD cell associated with the circuit shown in Figure 7. The corresponding high-pass cell is depicted in Figure 8. The high-pass NGD cell is easily synthesized from the low-pass cell via the low-pass-to-high-pass transform. The resistance element is not changed: $R_{sh} = R_s$.

4.2.1 High-Pass NGD Behavior

The *S* parameters of the series-impedance-based cell shown in Figure 4 become

$$\left[S_{sh}(j\omega)\right] =$$

$$\begin{bmatrix} \dfrac{j\omega R_s L_{sh}}{2R_0R_s + j\omega L_{sh}(2R_0+R_s)} & \dfrac{2R_0(R_s + j\omega L_{sh})}{2R_0R_s + j\omega L_{sh}(2R_0+R_s)} \\ \dfrac{2R_0(R_s + j\omega L_{sh})}{2R_0R_s + j\omega L_{sh}(2R_0+R_s)} & \dfrac{j\omega R_s L_{sh}}{2R_0R_s + j\omega L_{sh}(2R_0+R_s)} \end{bmatrix} \tag{54}$$

The associated group delay is given by

$$\tau_{sh}(\omega \approx 0) = \frac{L_{sh}}{2R_0} . \tag{55}$$

The NGD cutoff frequency, $\omega_{c_{sh}}$, is given by

$$\omega_{c_{sh}} = \frac{R_s\sqrt{2R_0}}{L_{sh}\sqrt{2R_0+R_s}} . \tag{56}$$

Moreover, it is also found that for any integer $n > 1$, the group delay can be expressed as

$$\tau_{sh}\left(\omega = n\omega_{c_{sh}}\right) =$$

$$\frac{\left(1-n^2\right)L_{sh}\left(2R_0+R_s\right)}{\left[2\left(1+n^2\right)R_0+R_s\right]\left[2\left(1+n^2\right)R_0+n^2R_s\right]} . \tag{57}$$

It can be understood from Equation (55), $\tau_{sh}(\omega \approx 0) > 0$, and Equation (57), $\tau_{sh}\left(\omega = n\omega_{c_{sh}}\right) < 0$, which generally means that $\tau_{sh}\left(\omega > \omega_{c_{sh}}\right) < 0$. This means that the passive cell introduced in Figure 8 behaves as a high-pass NGD circuit.

The NGD figure-of-merit is expressed as

$$FoM_{high-pass_s} = FoM_{low-pass_s} , \tag{58}$$

which is proportional to the resistance R_s .

4.2.2 Synthesis Method

The high-pass NGD resistance, $R_{sh} = R_s$, can be synthesized from Equation (50). Given the targeted positive group delay $\tau_0 > 0$, the inductance parameter can be synthesized from the group delay expressed in Equation (55), given by

$$L_{sh} = 2R_0\tau_0 . \tag{59}$$

The high-pass NGD cell inductance can also be alternatively determined from the high-pass cutoff frequency introduced in Equation (57):

$$L_{sh} = \frac{R_s\sqrt{2R_0}}{\omega_{c_{sh}}\sqrt{2R_0+R_s}} . \tag{60}$$

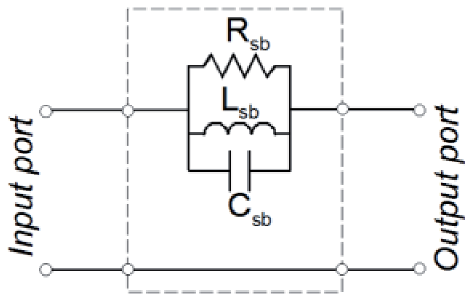

Figure 9. A bandpass NGD cell: *RLC*-parallel series network associated with the low-pass cell shown in Figure 7.

4.3 Bandpass NGD Elementary Cell

In this case, to synthesize the associated bandpass NGD cell we just need to replace the capacitor LC_s by an *LC*-parallel network, and we obtain the elementary cell shown in Figure 9. The resistance element is not changed, $R_{sb} = R_s$.

4.3.1 Bandpass NGD Behavior

The proposed cell's resonance frequency, which is also the NGD central frequency, is defined by

$$\omega_0 = \frac{1}{\sqrt{L_{sb} C_s}} . \tag{61}$$

The NGD bandpass behavior can be understood from the insertion-loss expression. The associated *S* parameters at the center frequency are then expressed as

$$\left[S_{sb}(\omega_0)\right] = \left[S_s(0)\right] = \begin{bmatrix} \dfrac{R_s}{2R_0 + R_s} & \dfrac{2R_0}{2R_0 + R_s} \\ \dfrac{2R_0}{2R_0 + R_s} & \dfrac{R_s}{2R_0 + R_s} \end{bmatrix} \tag{62}$$

The group delay is given by

$$\tau_{sb}(\omega_0) = 2\tau_s(0) = \frac{-4R_s^2 C_{sb}}{R_s + 2R_0} . \tag{63}$$

4.3.2 Bandpass NGD Characterization and Properties

The NGD center frequency of the passive cell shown in Figure 9 is equal to $\omega = \omega_0$. The NGD bandwidth is defined by

$$\Delta\omega_{c_{sb}} = 2\omega_{c_s} = \frac{\sqrt{1 + \dfrac{R_s}{2R_0}}}{R_s C_{sb}} . \tag{64}$$

This means that the NGD cutoff frequencies are defined by

$$\omega_{c_{sb}} = \omega_0 \pm \omega_{c_s} = \frac{1}{\sqrt{L_{sb} C_{sb}}} \pm \frac{\sqrt{1 + \dfrac{R_s}{2R_0}}}{R_s C_{sb}} . \tag{65}$$

This bandpass cell's NGD figure-of-merit is expressed as

$$FoM_{band-pass_s} = 2FoM_{low-pass_s} , \tag{66}$$

which is proportional to the resistance, R_s .

4.3.3 Synthesis Method

Similarly to the previous case of Subsection 3.3.3, the bandpass NGD cell resistance and capacitor parameters can be determined from the associated low-pass cell. The same equation allows determining the resistance. However, two times the low-pass NGD cell capacitor, $C_{sb} = 2C_s$, must be considered, based on the group delay defined in Equation (56). The associated inductance can be calculated from the inductance and the expected NGD center angular frequency, which can be denoted ω_0 . We therefore have the synthesis formula

$$L_{sb} = \frac{1}{\omega_0^2 C_{sb}} . \tag{67}$$

4.4 Stop-Band NGD Elementary Cells

Similarly to Subsection 3.4, the stop-band NGD cell can be easily synthesized from the low-pass cells via the

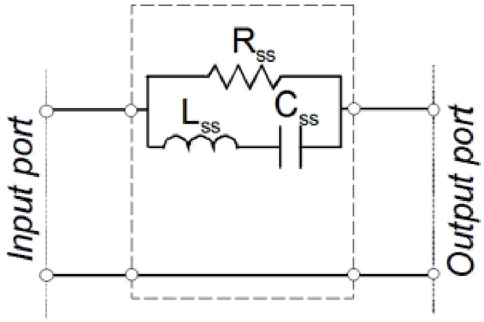

Figure 10. A stop-band NGD cell: *R* and *LC*-series parallel network associated with the low-pass cell shown in Figure 7.

low-pass-to-stop-band transform. Figure 10 represents the configuration of the impedance-series-type stop-band NGD cell. The resistance element is not changed: $R_{ss} = R_s$.

4.4.1 Stop-Band NGD Behavior

The proposed cell's resonance frequency is defined by

$$\omega_0 = \frac{1}{\sqrt{L_{ss} C_{ss}}} . \tag{68}$$

It can be demonstrated that the associated group delay is

$$\tau_{ss}\left(\omega_0\right) = \frac{L_{ss}}{2R_0} . \tag{69}$$

This group delay is always positive around the resonance frequency.

4.4.2 Stop-Band NGD Characterization and Properties

The proposed shunt-impedance-based stop-band NGD cell's cut-off frequencies are given by

$$\omega_{c_{ss}} = \frac{1}{\sqrt{L_{ss} C_{ss}}} \pm \frac{R_s \sqrt{2R_0}}{L_{ss}\sqrt{2R_0 + R_s}} . \tag{70}$$

This means that the NGD bandwidths are

$$\begin{cases} \omega \le \dfrac{1}{\sqrt{L_{ss} C_{ss}}} - \dfrac{R_s \sqrt{2R_0}}{L_{ss}\sqrt{2R_0 + R_s}} \\ \text{and} \\ \omega \ge \dfrac{1}{\sqrt{L_{ss} C_{ss}}} + \dfrac{R_s \sqrt{2R_0}}{L_{ss}\sqrt{2R_0 + R_s}} . \end{cases} \tag{71}$$

For the case of the stop-band cell, the figure-of-merit cannot be defined.

4.4.3 Synthesis Method

By analogy with Section 3.4.3, the stop-pass NGD cell-synthesis relationships can be inspired by the combination of the low-pass and high-pass relationships. The resistance is the same relationship as for the low-pass cell defined in Equation (50). The inductance must be the same as for the high-pass cell defined in Equation (59) or Equation (60), $L_{ss} = L_{sh}$. The associated capacitor can be calculated from the inductance and the expected NGD center angular frequency, which can be denoted ω_0 . We therefore have the capacitor-synthesis formula

$$C_{ss} = \frac{1}{\omega_0^2 L_{sh}} . \tag{72}$$

To validate the overall NGD circuit theory, the design, simulation, and experiments for low-pass, high-pass, bandpass, and band-stop NGD cells built with lumped *R*, *L*, and *C* based elements are presented in the next section.

5. Validation Results

It can be understood from the previous theory that the NGD circuit under study can be assumed to behave like any familiar electronic circuit. The present section is focused on the simulation results of proof-of-concept NGD cells synthesized from the developed theory. The family of two previously explored topologies based on parallel and series impedances is explored. For each family, the low-pass NGD circuits were synthesized, designed, simulated, fabricated, and tested in order to achieve the following arbitrary specifications:

- NGD level $\tau_0 = -1$ ns
- Reflection coefficient $r = -12$ dB.

The associated following circuits were then synthesized:

Table 2. Synthesized NGD circuit parameters.

Type Of NGD Cells	Low-Pass	High-Pass	Bandpass	Stop-Band
Resistance parameter	$R_p = 74.53\ \Omega$ $R_s = 33.54\ \Omega$			
Inductance parameter	$L_p = 2.97$ nH	$L_{sh} = 100$ nF	$L_{pb} = L_p$ $L_{sb} = 0.85$ nH	$L_{ps} = 2.53$ nH $L_{ss} = L_{sh}$
Capacitor parameter	$C_s = 119$ pF	$C_{ph} = 40$ pF	$C_{pb} = 0.34$ pF $C_{sb} = C_s$	$C_{ps} = C_{ph}$ $C_{sh} = 1.01$ pF

- High-pass NGD cell,
- Bandpass NGD cell aimed at operating at a center frequency of $f_0 = 0.5$ GHz,
- Stop-band NGD cell with a center frequency $f_0 = 0.5$ GHz.

The proof-of-concept modeled computed results were compared with simulations run in the *ADS®* environment of the electronic circuit designer and simulator. The proposed *S*-parameter simulations were performed from dc to 1 GHz. The simulation results obtained will be explored in the next paragraphs.

5.1 NGD Cell Proof-of-Concept Syntheses and Designs

After the circuit synthesis from the arbitrary given NGD and specifications, the NGD circuit examples will be described. The NGD functionality can be understood from the simulations. Table 2 summarizes the parameters of the low-pass, high-pass, bandpass, and stop-band NGD cells synthesized given the specifications $\tau_0 = -1$ ns and $r = -12$ dB.

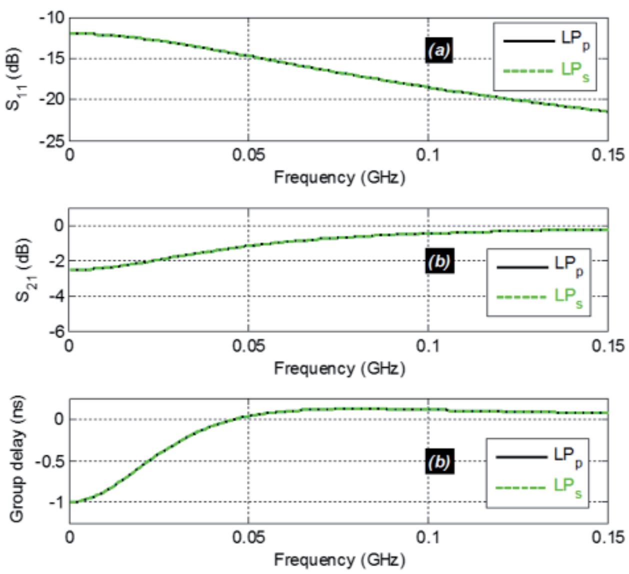

Figure 11. (a) The reflection and (b) transmission coefficients, and (c) group delay of the low-pass NGD circuits shown in Figure 3 and Figure 7.

5.2 Frequency-Domain Results

S-parameter simulations of circuits parameterized as addressed in the first column of Table 2 were performed to illustrate the low-pass, high-pass, bandpass, and stop-band NGD functions.

5.2.1 Low-Pass NGD Results

Figures 11a and 11b respectively display the reflection and transmission coefficients of the low-pass NGD cells shown in Figure 3 and Figure 7 for the shunt- and series-impedance-based topologies. The shunt-impedance-based results are plotted with the solid line, and the series-impedance-based results are plotted with the dashed line. As illustrated by Figure 11c, these passive cells behaved as a low-pass NGD function with the same characteristics. The theoretical prediction and the simulations were in very good agreement, and they confirmed the low-pass NGD function with the proposed passive lumped cells. Both cells generated the same *S*-parameter results. They presented a low-pass NGD level $\tau(0) = -1$ ns and a cutoff frequency of about $f_c(NCD) = 46$ MHz. Over the NGD bandwidth, the insertion loss varied from -2.5 dB to -1.27 dB. In other words, the insertion-loss flatness was about 1.23 dB. The reflection coefficients were better than -12 dB.

5.2.2 High-Pass NGD Results

To simulate the high-pass cells, the parameters of the high-pass cells shown in Figure 4 (for the parallel-impedance-based cell) and in Figure 8 (for the series-impedance-based cell) addressed in the second column of Table 1 were considered. Figures 12a and 12(b) respectively display the reflection- and transmission-coefficient plots. Figure 12c depicts the associated group delays. As expected, both cells generated the same *S*-parameter results. In addition, they behaved as having an NGD high-pass

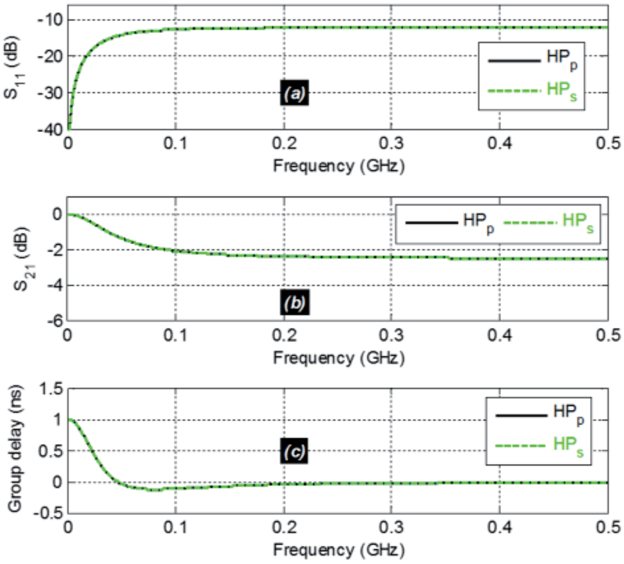

Figure 12. (a) The reflection and (b) transmission coefficients, and (c) the group delay of the high-pass NGD circuits shown in Figure 4 and Figure 8.

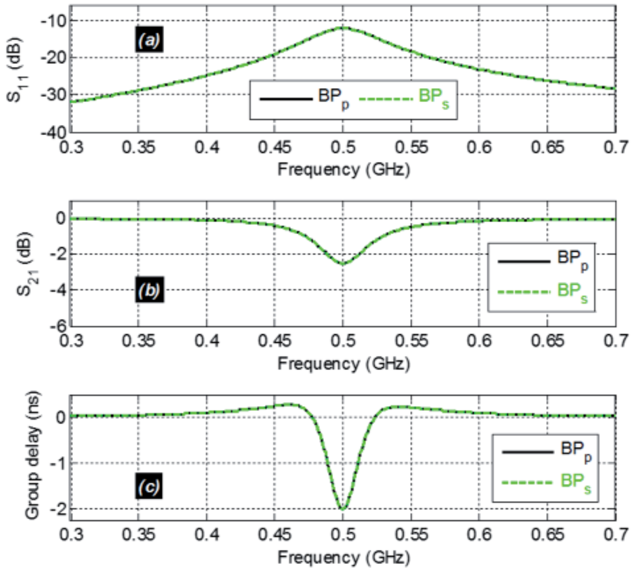

Figure 13. (a) The reflection and (b) transmission coefficients, and (c) the group delay of the bandpass NGD circuits shown in Figure 5 and Figure 9.

function. As theoretically predicted, the high-pass group delay, $\tau_{hp}(f) = -\tau_{lp}(f)$ was the opposite of the low-pass group-delay frequency response. It can be emphasized that the group delay was positive when $f < f_c$ and negative when $f > f_c$. A very good correlation between the theoretical concept of the high-pass NGD function and the simulations was obtained. As seen in Figure 12b, the insertion loss varied from -2.38 dB to -1.25 dB over the NGD bandwidth. The reflection coefficients were better than -12 dB.

5.2.3 Bandpass NGD Results

To illustrate the feasibility of the bandpass NGD function, the parameters of the cells introduced in Figure 5 (for the parallel-impedance-based cell) and in Figure 9 (for the series-impedance-based cell) defined in the third column of Table 1 were considered. The simulated reflection and transmission parameters from 0.3 GHz to 0.7 GHz were respectively plotted in Figures 13a and 13b. Both cells presented similar *S*-parameter responses. More importantly, Figure 13c presents the group-delay frequency responses of the simulated circuits. As predicted in theory, the proposed lumped cells behaved with a bandpass NGD function. The cells generated NGD with level $\tau(f_0) = -2$ ns at the specified center frequency $f_0 = 0.5$ GHz. The NGD bandwidth was about $\Delta f_{NGD} = 92$ MHz. In the NGD bandwidth, the insertion loss varied from -2.5 dB to -1.27 dB. In other words, the insertion loss flatness was about 1.23 dB. The bandpass NGD cell reflection coefficients were better than -12 dB.

5.2.4 Stop-Band NGD Results

The stop-band NGD circuits shown in Figure 6 (for the parallel-impedance-based cell) and Figure 10 (for the series-impedance-based cell) were simulated with the parameters indicated in the last column of Table 1. The simulated reflection and transmission parameters are respectively displayed in Figures 14a and 14b. Both cells presented the same frequency responses. The group-delay responses are plotted in Figure 14c. As expected in theory, the circuit generated the NGD stop-band function. The stop-band NGD circuit group-delay frequency response $\tau_{sb}(f) = -\tau_{bp}(f)$ was the opposite of the bandpass NGD circuit's response. The group delay was positive in the frequency band centered at $f_0 = 0.5$ GHz with a bandwidth of about $\pm\Delta f_{NGD}/2 = 46$ MHz. As seen in Figure 14b, the insertion loss varied from -2.38 dB to -1.25 dB in the NGD bandwidth. The reflection coefficients are better than -12 dB.

5.3 Illustrative Time-Domain Results

In addition to this frequency-domain analysis introduced in the previous subsection, a time-domain investigation was also performed. The results reported in the present subsection are constrained only to the low-pass NGD cells. The proposed time-domain investigation was carried out with the *ADS*® transient analyses in the time parameters $t_{\min} = 0$ and $t_{\max} = 50$ ns with a time step of $\Delta t = 0.2$ ns. Figure 15 shows the configuration of the simulated circuits.

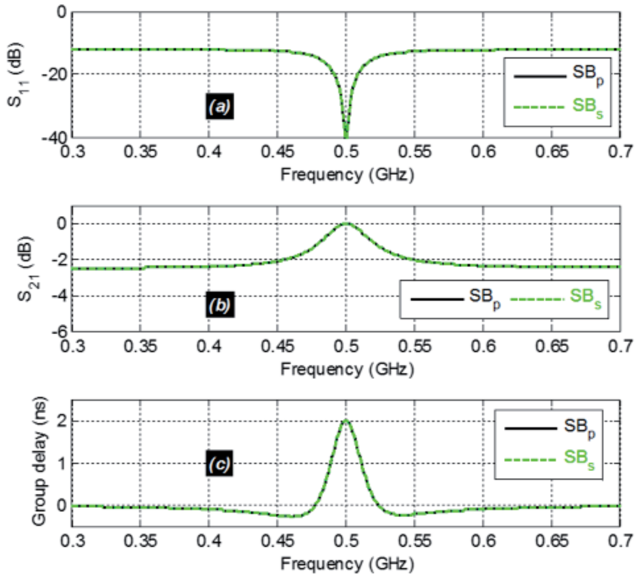

Figure 14. (a) The reflection and (b) transmission coefficients, and (c) the group delay of the stop-band NGD circuits shown in Figure 6 and Figure 10.

Three different input test voltages were attributed to v_{in}: v_1 (a Gaussian signal with a full width of about 40 ns), and arbitrary-waveform signals v_2 and v_3. Figure 16 presents the plots of the comparisons of the transient input and output voltages generated by the parallel-impedance- and series-impedance-based low-pass NGD cells. The input signals were plotted in dotted lines. The shunt-impedance-based cell outputs were plotted with solid lines, and the series-impedance-based cell outputs were plotted with dashed lines. For the three different cases as seen in Figures 16a$_1$, 16a$_2$ and 16a$_3$, the output-voltage signals were significantly attenuated. However, it could be understood from the comparative plots of normalized voltages shown in Figures 16b$_1$, 16b$_2$, and 16b$_3$ that the time advances induced by the NGD phenomena were generated for the different waveform test signals.

The time advance could be assessed with the time shift between the minimum and maximum peaks of the input and output transient signals. This effect can be generated only for the baseband signal presenting a significant spectral frequency band belonging to the NGD bandwidth. Because of the non-flat nature of the group-delay responses shown in Figure 11b over the NGD bandwidth, the time advance could be evaluated from -1 ns to -0.4 ns as a function of the input signal time slope.

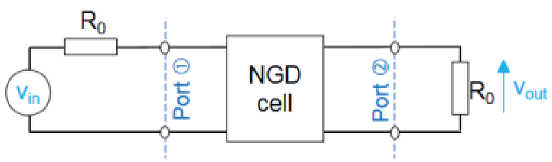

Figure 15. The configuration of the time-domain simulated circuits.

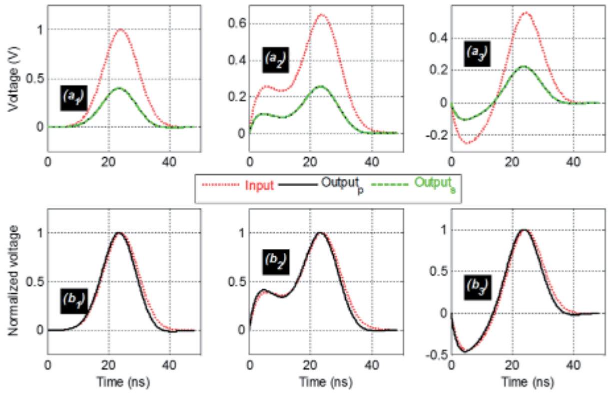

Figure 16. The time-domain simulation results for the low-pass NGD circuit shown in Figure 3 and Figure 7 for three different input signals: (a$_1$) v_1, (a$_2$) v_2, and (a$_3$) v_3, and the associated normalized results: (b$_1$) v_1 (b$_2$) v_2, and (b$_3$) v_3.

6. Conclusion

A basic theory for understand the NGD function has been proposed. The fundamental knowledge enabling familiarization with NGD circuit design was described. More practically, a general RF circuit theory for passive NGD topologies was developed. Two types of topologies, based on the *R*, *L*, and *C* lumped-element parallel- and series-impedances, were investigated. For each topology, the *S*-parameter models were established, and the elementary low-pass NGD cell was identified. The NGD behavior, characterization, properties, and synthesis methods were then explained. The associated high-pass, bandpass, and stop-band cells were also investigated. The theoretical concepts were validated with proof-of-concept designs and simulations. Very good agreement between the theory and simulation results was found. As expected from the theory, the low-pass circuits generated NGD levels of about -1 ns over a bandwidth of about 46 MHz. The bandpass prototypes were synthesized to operate around a NGD center frequency of around 0.5 GHz. The high-pass NGD cell generated a group delay equal to the opposite of the low-pass cell. The bandpass NGD function was obtained with a NGD level of about -2 ns, centered at 0.5 GHz. Finally, transient simulations illustrated the time-advance effect caused by the NGD low-pass phenomenon.

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Recent Developments in Distributed Negative-Group-Delay Circuits

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Abstract

This paper reviews some of the latest developments in active and passive negative-group-delay (NGD) circuits based on distributed topology using transmission-line elements.

First, the concepts of microwave transversal- and recursive-filter-based negative-group-delay circuits are introduced. By using transversal and recursive-filter approaches, the desired amount of negative group delay can be synthesized with an increased bandwidth. The presented prototype, using a transversal-filter-based NGD circuit realized by a multi-section directional coupler, can achieve -1 ns group delay with a 500 MHz bandwidth at a center frequency of 1 GHz with an insertion loss of 20 db to 30 dB. Its active counterpart, realized by a distributed amplifier (DA), can achieve a -1 ns group delay with a 250 MHz bandwidth at 1 GHz and an insertion loss of around 8 dB. Both of these realizations have great input matching, with a reflection coefficient better than -25 dB.

Second, NGD circuits integrated with active transistors and transmission-line elements, such as short stubs, are discussed. The proof-of-concept prototype showed a -0.43 ns group delay with a 1.14 GHz bandwidth at a 5 GHz center frequency and an insertion loss of 2 dB. The fully distributed topology of these NGD circuits shows promising solutions to be utilized in next-generation microwave and millimeter-wave applications.

1. Introduction

The concept of negative group delay (NGD) has recently drawn a lot of attention in the microwave field, and has led to many practical applications. For instance, NGD has been utilized to enhance the efficiency of the feed-forward amplifier, to shorten delay lines, and to realize broadband phase shifters and non-Foster elements [1-4]. Taking advantage of NGD circuits, it is possible to compensate for excessive positive group delays introduced by circuit components and electrical connections in electronic systems. The phenomenon of NGD might yield the output peak of a wave-packet preceding the input peak. This is due to the *RLC* time constants of the circuitry that affect different frequency components of the modulated signal in such a way that some emerge before others when examined in the time domain. The NGD phenomenon does not violate causality, since the initial transient pulse is still limited to the frontal velocity, which will never exceed the speed of light [5].

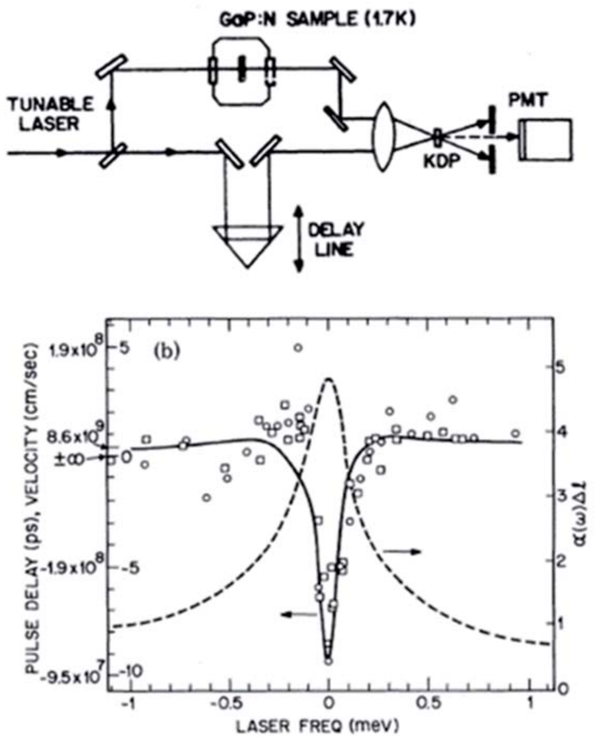

Figure 1. The NGD effect in the optical domain with GaP:N atomic [7].

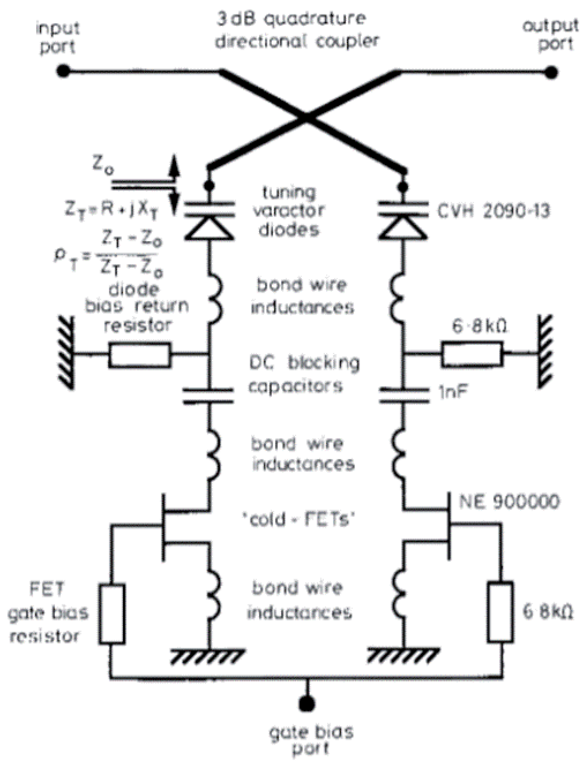

Figure 2. A lumped-element-based tunable NGD circuit [8].

Studies of the NGD phenomenon can be traced back to the 1970s, when analytical investigation and theoretical analysis of Gaussian light propagation in an NGD medium was first proposed [6]. Thereafter, in the 1980s, the first experimentation with the NGD effect was made in the optical domain with GaP:N atomic vapor, as shown in Figure 1 [7]. Later on, in the 1990s, theoretical and experimental investigations in the field of electronic and microwave engineering confirmed that certain passive circuit topologies can generate NGD phenomena [8, 9]. For low-frequency electronic circuits, the NGD phenomenon has been exhibited using simple low-pass and bandpass amplifiers [10-12]. On the other hand, in the microwave regime, because NGD occurs at the frequency band where the absorption or the attenuation is maximum [13], bandstop structures are conventionally used to realize NGD circuits (Figure 2). Based on either series or shunt *RLC* resonators, many active and passive NGD circuits have been proposed and demonstrated [14-18].

Nevertheless, the lumped-element approach may not be very suitable for NGD devices operating at a higher frequency range that goes beyond their self-resonances. In an effort to create a fully distributed topology of NGD circuits, transmission-line-based NGD circuits have recently been proposed and demonstrated to realize the desired negative group delay [19-24]. In this review article, recent developments of distributed-topology NGD circuits will be discussed. It will be shown that these distributed-topology-based methods are able to synthesize the prescribed amount of negative group delay that can be further utilized in electronic circuits.

2. Transversal- and Recursive-Filter-Based NGD Circuits

The first kind of distributed NGD circuits is based on microwave transversal and recursive filters, which can be viewed as a microwave equivalent of a finite-impulse-response (FIR) filter and an infinite-impulse-response (IIR) filter in the digital domain. Transmission lines in these structures are treated as delay elements to provide necessary time delay among the stages. By properly choosing the weighting/coupling coefficients along with the delay elements, the desired amount of NGD can be synthesized at the filters' output port.

2.1 Transversal-Filter-Based NGD Circuits

Figure 3 illustrates a general diagram of a continuous-time-transversal filter. It consists of delay lines (τ), weighting coefficients (b_N), and summing operators. The relationship between the input, $x(t)$, and the output, $y(t)$, in the time domain for a transversal filter with N weighting elements or N sections can be expressed as [25]

$$y(t) = \sum_{n=1}^{N} b_n x\left[t - \sum_{m=1}^{n}\left(\tau_{im} + \tau_{om}\right)\right]. \qquad (1)$$

Its transfer function in the frequency domain can then be obtained by taking the Fourier transform. Furthermore, if we let $\tau_{im} = \tau_{om} = \tau_0$, it follows that the transfer function becomes

$$H(\omega) = \sum_{n=1}^{N} b_n e^{-j2n\omega\tau_0}. \qquad (2)$$

By definition, the group delay of Equation (2) at $\omega = \omega_0$ can be written as follows:

$$\tau_g = -\left.\frac{\partial \angle H(\omega)}{\partial \omega}\right|_{\omega=\omega_0}. \qquad (3)$$

From Equation (2), it can be observed that the phase response of the transfer function can be manipulated by altering the

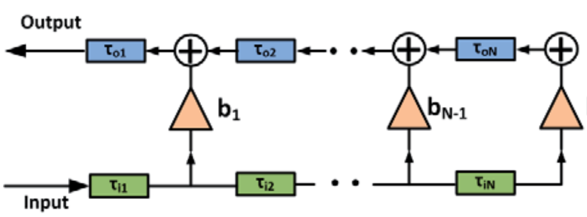

Figure 3. A schematic diagram of a transversal filter [19].

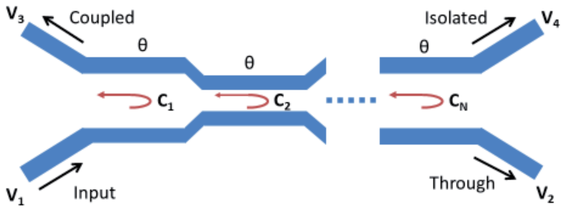

Figure 4. An *N*-section directional coupler [19].

tap coefficients (b_N). In fact, it has been analytically shown that in order for the group delay depicted in Equation (3) to have a maximally flat response, the tap coefficients have the following closed-form solution [19]:

$$b_{N-m} = \frac{\binom{N+\alpha-1}{m}\binom{\alpha-1-m}{N-m-1}}{2\binom{2N-3}{N-1}}, \quad (4)$$

where b_{N-m} refers to the tap coefficients shown in Figure 1, and the group delay, τ_g, of the transfer function, H, depicted in Equation (2) is equal to $(N-\alpha+1)\tau_0$. For a given stage N and negative group delay, which can be determined by the index α, the tap coefficients for a transversal filter exhibiting maximally flat negative group delay can therefore be readily obtained by Equation (4).

2.1.1 Passive Realization

A simple realization of a passive transversal filter in microwave regimes is to use a multi-section directional coupler, as shown in Figure 4. Assuming that the coupling is weak ($C_n \le -10$ dB), and that each section is a quarter-wavelength long ($\theta = \pi/2$) at the center frequency, the total voltage transfer function of the coupled port of an N-section directional coupler can be expressed as [26]

$$\frac{V_3}{V_1} = \sum_{n=1}^{N} (jC_n \sin\theta) e^{-j(2n-1)\theta}, \quad (5)$$

where C_n is the coupling factor for the nth section of the directional coupler, and θ is the electrical length of the coupled lines. It is noted that Equation (5) has a mathematical form similar to Equation (2) by mapping θ to $\omega - \tau_0$.

Table 1. The coupling coefficients for an NGD circuit

	C_1	C_2	C_3	C_4
$N=2$	0.3162	0.2372		
$N=3$	0.2372	0.3162	0.1107	
$N=4$	0.1757	0.3162	0.2012	0.0447

The coupling factors for different sections of directional couplers ($N = 2,3,4$) can be obtained from Equation (4), in which the desired group delay is set to be -1 ns. The maximum coefficients are then normalized to -10 dB as depicted in Equation (6), which is 0.3162 on a linear scale. Table 1 summarizes these coefficients for directional couplers with various stages:

$$C_{N=2} = [0.3162, 0.2372], \quad (6a)$$

$$C_{N=3} = [0.2372, 0.3162, 0.1107], \quad (6b)$$

$$C_{N=4} = [0.1757, 0.3162, 0.2012, 0.0447]. \quad (6c)$$

Figure 5a plots the measured magnitude response of the NGD circuits using multi-section directional couplers. The magnitude of S_{31} decreases more, especially when more sections are used. This may be due to more loss introduced in the actual fabricated prototypes. Figure 5b plots the measured group delay of the fabricated prototypes using printed-circuit-board (PCB) technology with various numbers of stages for directional couplers. Overall, with

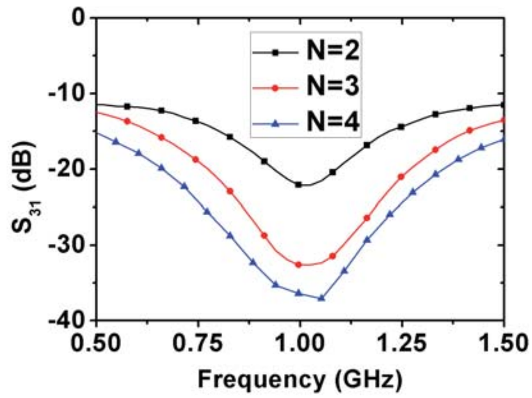

Figure 5a. The measured insertion loss for NGD circuits based on multi-section directional couplers [19].

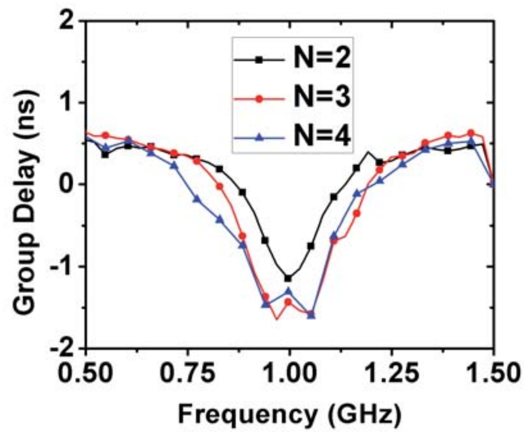

Figure 5b. The group delay for NGD circuits based on multi-section directional couplers [19].

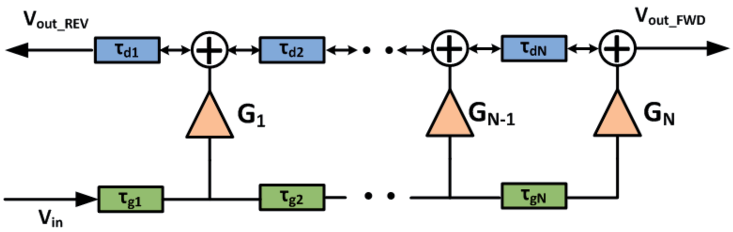

Figure 6. An *N*-stage distributed amplifier.

more sections used in the directional couplers, the bandwidth of the NGD increases. Moreover, as mentioned above, a flat response of NGD can be achieved by using the proper coupling coefficients for each section of the directional coupler.

2.1.2 Active Realization

Although a multi-section directional coupler can provide a very simple way to realize the passive version of a microwave transversal filter, it suffers from tremendous loss in the NGD region, especially when the NGD-bandwidth product is large. This is because the out-of-band gain needs to be large in this case [27]. It turns out that distributed amplifiers (DAs) stand out as an excellent choice for providing amplification for the NGD circuits, as they are essentially one type of transversal filer in active fashion. Figure 6 illustrates the diagram of an *N*-stage distributed amplifier, in which G_k is the gain coefficient of the *k*th tap amplifier, and τ_{gk} and τ_{dk} indicate the time delay of the *k*th section of the gate line and drain line, respectively. Furthermore, if we choose $\tau_{gi} = \tau_{di} = \tau_0$, then the transfer functions of the forward and reverse port with respect to the input can be expressed as [20]

$$H_{FWD}(\omega) = \frac{V_{out\_FWD}}{V_{in}} = \sum_{k=1}^{N} G_k e^{-j\omega(N\tau_0)}, \quad (7a)$$

$$H_{REV}(\omega) = \frac{V_{out\_REV}}{V_{in}} = \sum_{k=1}^{N} G_k e^{-j\omega(2k\tau_0)}. \quad (7b)$$

One can immediately identify that the reverse transfer function, Equation (7b), also has the same mathematical form as Equation (2). It can further be observed that if the coefficients derived in Equation (4) are used for the gain coefficients G_k in Equation (7), an NGD response can be realized at the distributed amplifier's reverse port, while maintaining a flat gain at its forward port, since the G_k s sum up constructively as indicated in Equation (7a). It is worth mentioning that the distributed-amplifier-based NGD circuit can achieve unconditional stability as long as each stage is designed to be unconditionally stable, regardless of the number of stages.

As a proof-of-concept, a distributed amplifier, using commercial off-the-shelf discrete transistors (NEC NE3210S01), was fabricated with PCB technology. As shown in Figure 7, the distributed amplifier contained two transistors with individual gate biases to form a two-stage NGD circuit. The electrical length between each stage was designed to be a quarter-wavelength at the center frequency of 1 GHz. Since the maximum transconductance (g_m) of the transistors we used was 0.075 S, the amounts of g_m in this two-stage distributed-amplifier-based transversal filter were $(0.075\ \text{S}, 0.05625\ \text{S})$.

Figure 8a shows the measured *S* parameters of the fabricated two-stage distributed amplifier. The forward gain showed a flat response of around 10 dB gain, whereas the reverse gain had a minimum of −7.8 dB at the center frequency of 1 GHz where NGD occurred. In addition, the return loss was greater than 30 dB, and the insertion loss was around 0.2 dB at the frequency of interest. The phase and group delays of the distributed amplifier are plotted in Figure 8b, indicating a great agreement between the ideal circuit simulation and measurement. The maximum group delay of the two-stage distributed amplifier was −1 ns at 1 GHz, and the NGD region was 0.9 GH to 1.1 GHz.

2.2 Recursive-Filter-Based NGD Circuits

In addition to transversal filters, microwave recursive filters can also be used to generate NGD of predetermined values in a desired frequency band by using only distributed components [22]. In this case, multi-stage quarter-wave transformers (QWT) are used to generate NGD of

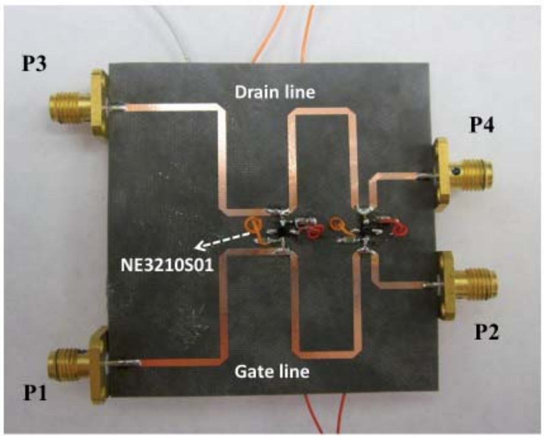

Figure 7. A prototype of a two-stage NGD circuit based on distributed amplifiers [19].

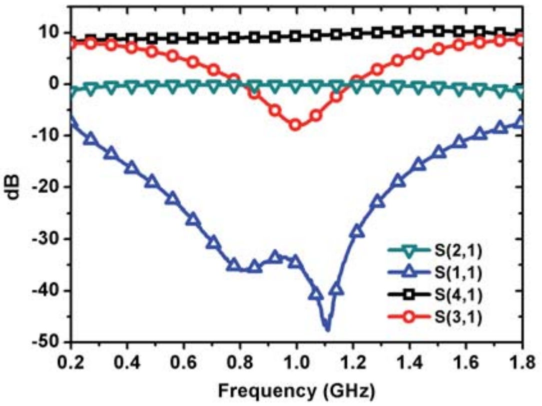

Figure 8a. The measured results of the magnitude for the two-stage NGD-distributed amplifier [19].

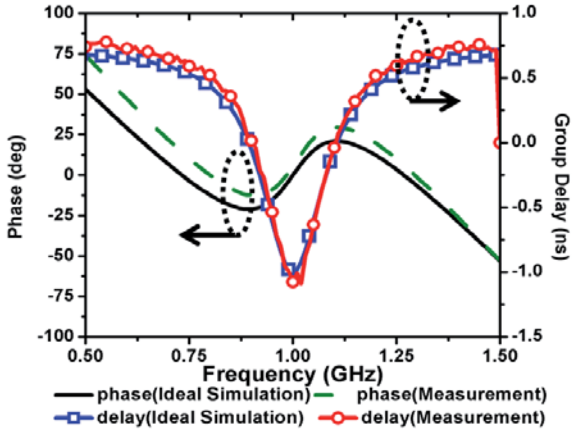

Figure 8b. The group delay and phase response for the two-stage NGD-distributed amplifier [19].

predetermined values, as shown in Figure 9. Essentially, a quarter-wave transformer behaves as a recursive filter, since the reflected waves will bounce back and forth among the impedance interfaces and form an infinite impulse response (IIR). The reflection coefficient, Γ, for the *N*-stage quarter-wave transformer can be written as

$$\Gamma = \frac{V_{in}^{-}}{V_{in}^{+}} = \frac{(A_n - D_n) + j(B_n - C_n)}{(A_n + D_n) + j(B_n + C_n)}, \tag{8}$$

where A_n, B_n, C_n, and D_n represent the transmission-matrix parameters for the entire *N*-stage quarter-wave transformer circuit. The phase of the reflection coefficient is then expressed as

$$\angle\Gamma = \tan^{-1}\frac{(B_n - C_n)}{(A_n - D_n)} - \tan^{-1}\frac{(B_n + C_n)}{(A_n + D_n)}. \tag{9}$$

In order to obtain the prescribed maximally flat group delay, the characteristic impedances of each of the transmission

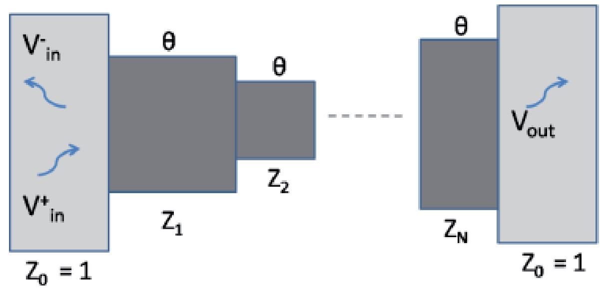

Figure 9. An *N*-stage quarter-wave transformer as a microwave recursive filter [22].

line $[Z_1, Z_2, Z_3, ..., Z_N]$ need to be obtained by solving the following equations:

$$\left|\Gamma\left(\theta = \frac{\pi}{2}\right)\right| = \Gamma_p, \tag{10a}$$

$$-\frac{\partial\angle\Gamma(\theta)}{\partial\theta}\bigg|_{\theta=\frac{\pi}{2}} = \tau_{gp}, \tag{10b}$$

$$-\frac{\partial^{2n-3}\angle\Gamma(\theta)}{\partial\theta}\bigg|_{\theta=\frac{\pi}{2}} = 0, \tag{10c}$$

where $n = 3, 4, ..., (N-1)$, Γ_p is the prescribed magnitude, and τ_{gp} is the prescribed NGD within a unit of the sampling period, *T*. In order to have the maximally flat response, all the higher-order derivatives shown in Equation (10c) are set to zero. Figure 10 shows the simulated results of three different NGDs at 1 GHz for a three-stage quarter-wave transformer using the normalized

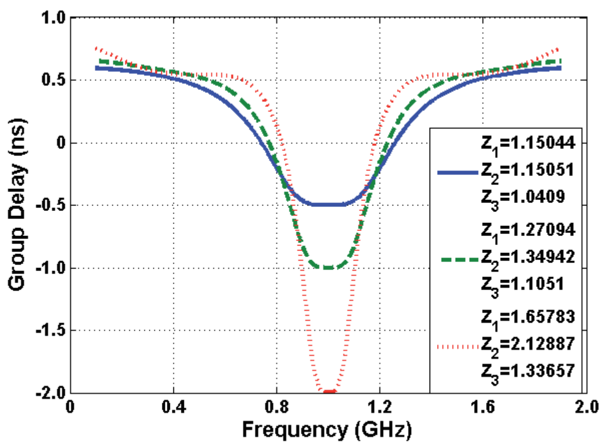

Figure 10. The simulated results for prescribed group delays of −0.5 ns, −1 ns, and −2 ns for a three-stage ($N = 3$) quarter-wave transformer [22].

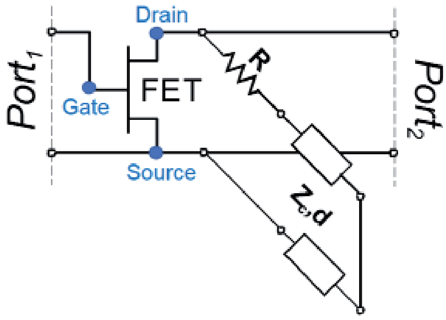

Figure 11. The topology of the NGD distributed active cell (NGDDAC) based on the FET.

characteristic impedances derived from Equations (10). It could be observed that the multi-stage quarter-wave transformer could exhibit NGD with maximally flat response about the center frequency. It was also noted that the bandwidth could be enhanced when more stages were cascaded.

3. FET-Based NGD Distributed Active Circuit with Resistive Short-Circuited Stub

Recently, it was introduced that another simpler active circuit, based on the association of a short-circuited stub, a series resistance, and an FET, was susceptible to exhibiting an NGD function [28, 29]. In contrast to the fully passive circuit proposed in [30, 31], we have the possibility of generating wideband NGD with amplification or positive gain. Unlike transversal-or recursive-filter types of NGD circuits that usually involve multiple stages of gain or coupling cells, this section presents a simple NGD active microwave circuit theory using one active transistor in connection with transmission-line structures.

3.1 Topological Concept

As previously mentioned, the proposed NGD distributed circuit uses an elementary FET. For the sake of analytical simplification, in this paper the FET is assumed to be composed of a voltage-controlled current source (VCCS) with a transconductance, g_m, and a drain-source resistance, R_{ds}.

3.1.1 FET-Based Distributed NGD Active Cell

Figure 11 represents the topology of the NGD distributed active cell (NGDDAC) under study in this section. It acts as a two-port microwave active circuit. The gate-source that constitutes the circuit's input port is assumed to be open but can be matched with a shunt resistance, depending on the targeted applications. The drain-source constituting the output is loaded by a parallel distributed load. The passive part of this circuit is composed of a resistance, R, in series with a short-circuited transmission line (TL) presenting a characteristic impedance, Z_c, and a physical length, d.

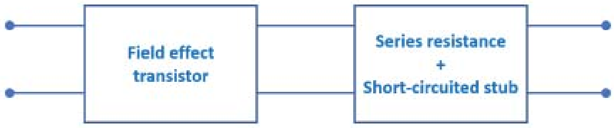

Figure 12. A block-diagram representation of the cell introduced in Figure 11.

The *S*-parameter model of this active cell can be calculated from microwave circuit theory based on the four-port active block represented by the FET in cascade with another four-port passive block represented by the R-TL network. The global configuration of this topology is depicted in Figure 12.

Based on system theory, the total transfer matrix of the system proposed in Figure 12 can be written as

$$\left[M_{total}\right] = \left[M_{FET}\right] \times \left[M_{passive}\right] \tag{11}$$

by denoting, as defined in [28, 29], $\left[M_{FET}\right]$ as the FET transfer matrix, and $\left[M_{passive}\right]$ as the transfer matrix of the passive-network part of the circuit shown in Figure 11.

The equivalent *S* parameters can be determined from the ABCD-to-*S* matrix conversion.

3.1.2. Description of the NGD Circuit Passive Part Constituted by the Distributed Short-Circuited Stub

The short-circuited stub constituting the NGD distributed active cell is assumed to be implemented in microstrip technology, as shown in Figure 13. The

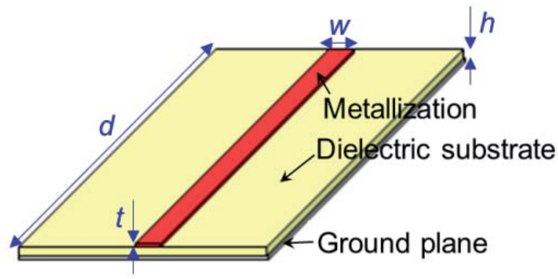

Figure 13. A microstrip-line structure and its physical parameters.

transmission line presents a physical length d, a metallization width w, and a thickness t, and is printed on a dielectric substrate with a height h. Its characteristic impedance is denoted Z_c. According to transmission-line theory, the periodic resonance frequencies are equal to

$$f_0(k) = \frac{kc}{2d\sqrt{\varepsilon_{reff}}} = \frac{\omega_0(k)}{2\pi}, \tag{12}$$

with $k = \{1,2,3,...\}$ being an integer, and where c is the speed of light.

Moreover, the general expression of the stub input impedance is written as

$$Z_{in}(\omega) = jZ_c \tan\left(\frac{\omega d\sqrt{\varepsilon_{reff}}}{c}\right), \tag{13}$$

where j is the complex number $\sqrt{-1}$, ω is the angular frequency, and ε_{reff} is the transmission line's effective permittivity. Based on the theory of Bahl and Trivedi [32], the transmission line's characteristic impedance is given by

$$Z_c = \frac{1}{2\pi}\sqrt{\frac{\mu_0}{\varepsilon_{reff}\varepsilon_0}} \log\left[\frac{h\psi\left(\frac{w}{h}\right)}{w} + \sqrt{1+\frac{4h^2}{w^2}}\right] \tag{14}$$

with vacuum permittivity and permeability ε_0 and μ_0. The real function $\psi(\bullet)$ is defined by

$$\psi(x) \approx 6 + (2\pi - 6)\exp\left(-\frac{13.155}{x^{0.7528}}\right). \tag{15}$$

The $\log(\bullet)$ function under consideration represents the decimal logarithm. These analytical expressions will serve as the initial step during the synthesis of the open-ended stud constituting the NGD distributed active cell.

3.1.3 Proposed NGD Distributed Circuit *S* Parameters and Synthesis Equations

Knowing the FET and the passive network ABCD matrices, the NGD distributed active cell equivalent matrix can be extracted with a matrix product. By denoting $Z_0 = 50\ \Omega$ as the reference impedance, the proposed NGD cell topology presents output reflection and transmission coefficients given by

$$|S_{22}| = \big||S_{21}| - Z_0 g_m\big| / (Z_0 g_m). \tag{16}$$

At the NGD center radian frequency $\omega_0(k)$ previously introduced in Equation (2), the following transmission and output reflection parameters can then be determined from the ABCD-to-S matrix transform:

$$S_{21}[\omega_0(k)] = \frac{-2g_m R_{ds} Z_0 R}{R(R_{ds}+Z_0)+R_{ds}Z_0}, \tag{17}$$

$$S_{22}[\omega_0(k)] = \frac{R(R_{ds}-Z_0)-R_{ds}Z_0}{R(R_{ds}+Z_0)+R_{ds}Z_0} \tag{18}$$

The negative sign of the algebraic value of S_{21} is due to the systematic output-voltage inversion through the FET. It can be underscored that the cell presented in Figure 11 is not matched at its input, because S_{11} is naturally equal to -1. Such an issue can be easily solved by connecting an input shunt resistance between the FET's gate and the source. According to microwave circuit theory, the group delay can be analytically calculated by

$$\tau(\omega) = -\frac{\partial \angle S_{21}(\omega)}{\partial \omega}. \tag{19}$$

At the resonance frequencies $\omega = \omega_0(k)$, it can therefore be demonstrated that the group delay can be expressed as a function of the stub length by the relationship

$$\tau[\omega_0(k)] = \frac{-dR_{ds}Z_0Z_c}{vR[RR_{ds}+Z_0(R+R_{ds})]}, \tag{20}$$

denoting the wave speed $v = c/\sqrt{\varepsilon_{reff}}$. The same group delay can also be expressed with the resonance frequency by the relationship

$$\tau[\omega_0(k)] = \frac{-\pi R_{ds}Z_0Z_c}{\omega_0(1)R[RR_{ds}+Z_0(R+R_{ds})]}. \tag{21}$$

It is noteworthy that this group delay is unconditionally negative for any parameters of the distributed cell.

Similar to filter theory, the proposed NGD cell-synthesis relationships can be established from the expected specifications: the NGD center-frequency resonance, $\omega = \omega_0$; the NGD level; the gain; and the matching level. Moreover, with the distributed configuration introduced in Figure 11, the group delay can be negative at very low frequencies: $\omega_0(k=1)$. The following synthesis relation can thus be derived at base-band frequencies. The stub length can be obtained by inverting Equations (12) and (20). From the desired value of gain, $S_{21}(\omega_0)$, and NGD $\tau(\omega_0)$, the following synthesis relationships are established:

$$R = \frac{|S_{21}(\omega_0)| R_{ds} Z_0}{2g_m R_{ds} Z_0 - |S_{21}(\omega_0)|(Z_0 + R_{ds})}, \quad (22)$$

$$Z_c = \frac{-vR\tau(\omega_0)\left[R_{ds}R + Z_0(R + R_{ds})\right]}{dR_{ds}Z_0}. \quad (23)$$

3.1.4 NGD Properties

First and foremost, it should be pointed out that the maximal transmission gain of the NGD distributed active circuit under study is defined as

$$S_{21max} = 2R_{ds}Z_0 g_m / (R_{ds} + Z_0), \quad (24)$$

which is obtained from Equation (22) by letting the denominator be greater than zero. The NGD function should present a cutoff frequency f_c. For the case of the topology introduced in Figure 1, the periodic NGD frequency band $\left[\omega_c(k), \omega_c(k+1)\right]$ can be defined by the equation

$$\tan\left[\frac{\pi\omega_c(k)}{2\omega_0(1)} - k\pi\right] = \frac{\sqrt{R\left[R(R_{ds}+Z_0) + R_{ds}R\right]}}{Z_c\sqrt{(R_{ds}+Z_0)}}. \quad (25)$$

The NGD bandwidth is substantially given by

$$\Delta\omega_{NGD}\big|_{k=\{0,1,2,\ldots\}} = \omega_c(k+1) - \omega_c(k). \quad (26)$$

The figure-of-merit (FoM) associated with the proposed NGD distributed-active-cell topology is defined as

$$FoM_{NGD}\left[\omega_0(k)\right] = \Delta f \tau\left[\omega_0(k)\right] S_{21}\left[\omega_0(k)\right] \quad (27)$$

As a function of the circuit parameters, this can be rewritten as

$$FoM_{NGD}\left[\omega_0(k)\right] = \frac{-d\sqrt{R_{ds}Z_0}Z_c\Delta\omega_{NGD}/(2\pi)}{\sqrt{2g_m R^3\left[RR_{ds} + Z_0(R+R_{ds})\right]}}. \quad (28)$$

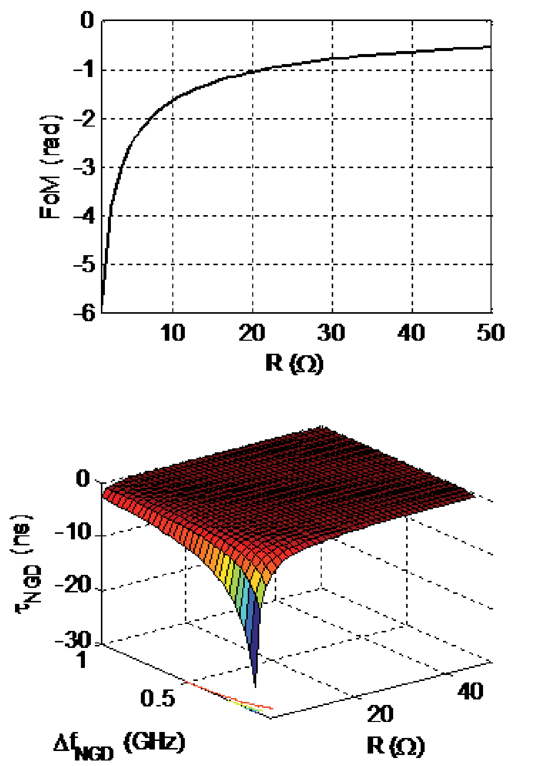

Figure 14. The NGD FoM as a function of R and Δf_{NGD}.

For further understanding of the meaning of the NGD figure-of-merit, a graphical illustration is proposed in Figure 14. It is noted that a higher absolute value of the FOM represents better performance of the NGD circuit. The figure presents the NGD's distributed active cell FoM as a function of *R*, and the level as a function of $(R = 1\,\Omega \to 50\,\Omega,\ \Delta f_{NGD} = 0.1\text{ GHz} \to 1\text{ GHz})$. This illustrative result was computed for the particular case of the FET EC-2612 ($g_m = 98.14$ mS, $R_{ds} = 116.8\,\Omega$).

3.2 Application Example

As a proof of concept (POC), an NGD distributed active cell was synthesized with arbitrary specifications. This proof of concept was designed and simulated in the ADS schematic environment.

3.3 Proof of Concept NGDDAC Synthesis and Design

By applying the synthesis formulas defined in Section 2.3, the circuit parameters R and the short-circuited stub characteristic impedance and length were determined. By considering the desired values of gain, $S_{21dB}(0) = 0$ dB and $\tau(0) = -0.5$ ns, a stub having a quarter wavelength at a frequency of $f_0 = 5$ GHz in series with a resistance $R = 8\,\Omega$ was synthesized and designed. This circuit integrated the bias and dc decoupling capacitors $C = 1$ pF. The schematic of the designed NGD distributed active cell proof of concept is depicted in Figure 15.

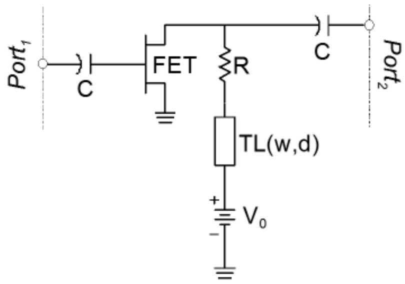

Figure 15. Schematic of the NGD distributed active cell proof of concept: $V_0 = 3$ V, $C = 1$ pF, $w = 1.55$ mm, $d = 16.4$ mm, $R = 8\,\Omega$.

The FET used corresponded to the pHEMT EC-2612 parameters, which are characterized by a transcoductance of $g_m = 98.14$ mS and a drain-source resistance of $R_{ds} = 116.8\,\Omega$. The stub was implemented with a microstrip transmission line printed on FR4 epoxy substrate having a relative permittivity of $\varepsilon_r = 4.4$, a loss tangent of $\tan(\delta) = 0.01$, and a thickness of $h = 0.8$ mm. The Cu conductor metallization thickness was $t = 35$ µm. The stub element, a with characteristic impedance of $Z_c = Z_0 = 50\,\Omega$ was integrated in the NGD circuit.

In order to illustrate the possibility of the NGD level and cut-off frequency, an analytical estimation for $f < f_0/4$ using the approximation $Z_{in} = jZ_c \tan(x) \approx jZ_c x$ was investigated. A comparison between the NGD distributed active cell based on distributed and lumped elements was also performed and is displayed in Figure 16.

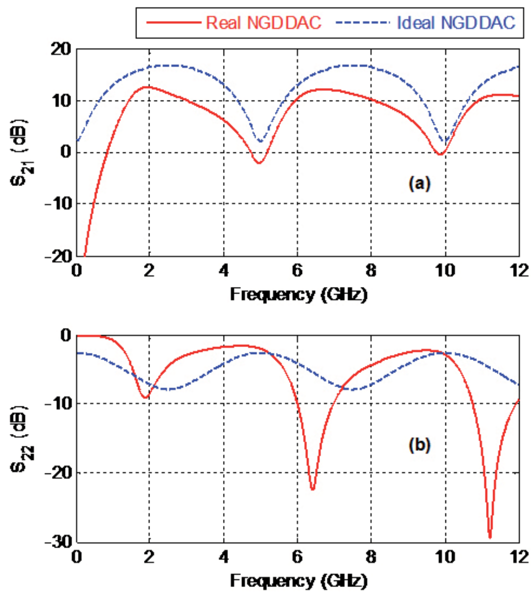

Figure 17. The simulated NGD-distributed active cell ideal and realistic (a) output return losses, and (b) transmission parameters.

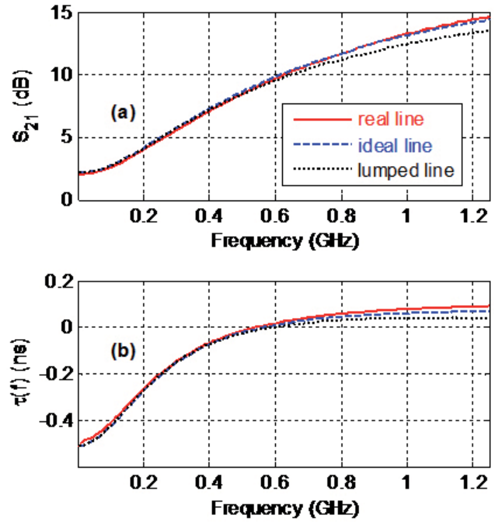

Figure 16. A comparison between the ideal transmission gain and the group delay of the proposed NGD distributed active circuit.

3.4 Validation Results

To check the feasibility of the NGD concept under study, S-parameter simulations of the ideal and realistic circuits were carried out from dc to 12 GHz. A good agreement was obtained between the behaviors of the transmission parameter and the group delay from the circuits built with the realistic and ideal NGD distributed active cell for the responses at very low frequencies, $f \approx 0$ Table 2 summarizes the NGD level, transmission gain, NGD bandwidth, and FoM around the optimal frequency, f_0, using Equation (27). In addition, Table 3 calculates the FOM for the transversal-filter-based NGD circuits discussed in the previous section. It should be noted that the center frequencies (f_0) for Tables 2 and 3 were different, i.e., 5 GHz and 1 GHz, respectively. The two topologies should therefore be compared with each other after taking this factor into account.

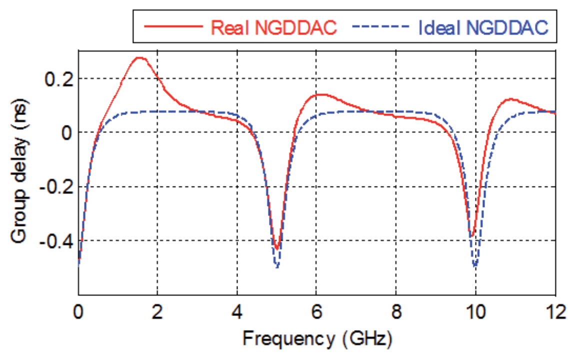

Figure 18. The simulated NGD distributed active cell ideal and realistic group delays.

Table 2. Ideal and realistic NGDDAC characteristics around $f_0 = 5$ GHz.

Parameters	Ideal Circuit	Realistic Circuit
$\tau(f_0)$	−0.5 ns	−0.43 ns
$S_{21}(f_0)$	2.13 dB	−2 dB
$\Delta f(f_0)$	1.11 GHz	1.14 GHz
$\mathrm{FoM}(f_0)$	−0.71	−0.39

Furthermore, comparisons between the output return losses and the transmission parameters are displayed in Figure 17. The group-delay responses are shown in Figure 18. As expected, a 5 GHz periodical multi-band NGD aspect was observed. The main observed differences were due to the FET parasitic elements and the stub transmission-line losses.

Despite this occurrence of an NGD multi-band effect, it was noteworthy that a matching circuit was needed to improve the proposed NGD distributed active cell return losses. Lastly, similar to the resistive amplifier topology, this NGD circuit's noise factor depended on the FET's noise, and could be extracted from the Friss formula.

4. Conclusion

In this paper, some recent developments of NGD circuits – based on microwave transversal and recursive filters, as well as on active transistors loaded with transmission-line elements – were presented and discussed. Unlike the lumped-element approach that is usually limited to its self-resonance, by using the transmission-line-based elements, the fully distributed nature of these NGD circuits make them feasible for further operation at a higher frequency range, such as in the millimeter-wave regions.

5. Acknowledgment

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Table 3. The FOM for transversal-filter-based NGD circuits around $f_0 = 1$ GHz.

Parameters	Passive	Active
$\tau(f_0)$	−1.5 ns	−1 ns
$S_{21}(f_0)$	−35 dB	−7.8 dB
$\Delta f(f_0)$	500 MHz	250 MHz
$\mathrm{FoM}(f_0)$	−0.0133	−0.1018

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using directional couplers.

A Zero-Order-Closure Turbulent Flux-Conservation Technique for Blending Refractivity Profiles in the Marine Atmospheric Boundary Layer

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Abstract

Recent advances in mesoscale numerical weather prediction (NWP) models have supported four-dimensional (4D) radio-frequency (RF) propagation modeling in challenging heterogeneous refractive marine environments. Numerical weather prediction models typically provide a vertical profile of refractivity every 1 km to 3 km horizontally in the domain of interest for each hour in a 48-hour forecast. Due to surface roughness and turbulence constraints, these profiles extend from the stratosphere to within 5 m to 10 m of the sea's surface. Because of strong evaporation at the sea's surface, significant impacts on RF system performance can be induced by refractivity gradients in the first 10 m above sea level (ASL). Historically, the lower-layer refractivity profiles have been calculated using Monin-Obukhov-Similarity- (MOS) based turbulence models. This dual-model approach requires a robust technique for blending on the order of 10^3 profiles per forecast hour without creating non-physical refractivity artifacts. This paper describes a zero-order-closure turbulent-flux technique for blending numerical weather prediction and Monin-Obukhov Similarity refractivity profiles, and presents the results of a multi-wavelength data-comparison process.

1. Introduction

Refraction of RF energy in the atmosphere is determined by the vertical structure of refractivity (N). Atmospheric pressure, atmospheric temperature, and partial pressure due to water vapor contribute to N [1, 2]. This relationship is given by

$$N = (n-1)10^6 \tag{1}$$

$$= \frac{77.6}{T}\left(p + \frac{4810e}{T}\right),$$

where n is the refractive index, T [K] is the atmospheric temperature, p [hPa] is the total atmospheric pressure, and e [hPa] is the partial pressure due to water vapor. $N(z)$ is the vertical profile of N for height z above the surface of the Earth.

The modified refractivity, M, removes the effects of the Earth's curvature from propagation calculations, allowing the use of Cartesian instead of cylindrical coordinates [3]. The vertical profile of M as a function of the height above the Earth's surface is given by

$$M(z) = N(z) + \frac{10^6 z}{a} \tag{2}$$

$$= N(z) + 0.157z,$$

where z [m] is the height above the surface, and a [m] is the mean radius of the Earth (6.378×10^6 m). The vertical gradient of $M(z)$ determines the refractive regime of the marine atmospheric boundary layer (MABL).

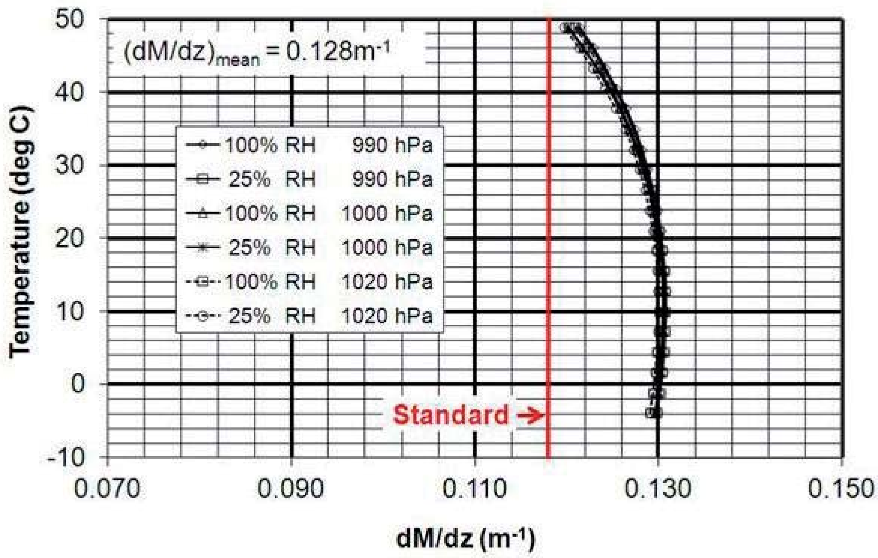

Figure 1. The vertical gradient of modified refractivity in a well-mixed marine atmospheric boundary layer.

The vertical thermodynamic structure of the marine atmospheric boundary layer is traditionally described in terms of potential temperature (θ) and water-vapor mixing ratio (w), since these are conserved variables of an air parcel if it experiences only adiabatic transformation with no net heat transfer as it moves in turbulent eddies throughout the marine atmospheric boundary layer [4]. The blending technique described in this paper is unique, since it is performed on θ, w, and p. The potential temperature in terms of T in Equation (1) is given by

$$\theta = T\left(\frac{p_0}{p}\right)^{\frac{R}{c_p}} . \qquad (3)$$

The reference pressure, p_0, is 1000 hPa. The gas constant for dry air, R, has the value 2.87×10^2 $\mathrm{Jkg^{-1}K^{-1}}$ The specific heat of dry air at constant pressure, c_p, is 1.00467×10^3 $\mathrm{Jkg^{-1}K^{-1}}$.

The water-vapor mixing ratio (w) is given by

$$w = \frac{0.622e}{p-e} \approx \frac{0.622e}{p} . \qquad (4)$$

Inserting Equations (1), (3), and (4) into Equation (2) yields

$$M = \left[\frac{77.6}{\grave{e}\left(\frac{p}{1000}\right)^{0.286}}\right]\left[P + \frac{4810\frac{wp}{0.622}}{\theta\left(\frac{p}{1000}\right)^{0.286}}\right] + 0.157z \qquad (5)$$

The vertical gradient of Equation (5) is given by

$$\frac{dM}{dz} = \frac{dp}{dz}\left(\frac{1.336\times10^7 w}{\theta^2 p^{0.572}} + \frac{399.54}{\theta p^{0.286}}\right)$$

$$-\frac{d\theta}{dz}\left(\frac{6.212\times10^7 wp^{0.428}}{\theta^3} + \frac{559.6p^{0.714}}{\theta^2}\right)$$

$$+0.157\ \mathrm{m}^{-1} . \qquad (6)$$

The marine atmospheric boundary layer is often well mixed, due to mechanical and convective turbulence [5]. In a well-mixed layer, θ and w are conserved, and thus

$$\frac{d\theta}{dz} = \frac{dw}{dz} = 0 .$$

The vertical gradient of M for a well-mixed layer is reduced to

$$\frac{dM}{dz} = \frac{dp}{dz}\left(\frac{1.336\times10^7 w}{\theta^2 p^{0.572}} + \frac{3.995\times10^2}{\theta p^{0.286}}\right) + 0.157 \qquad (7)$$

Figure 1 is a plot of Equation (7) for the ranges of relative humidity, atmospheric pressure, and temperature expected in the marine atmospheric boundary layer. Notice that a well-mixed marine atmospheric boundary layer produces a modified refractivity gradient slightly greater than standard.

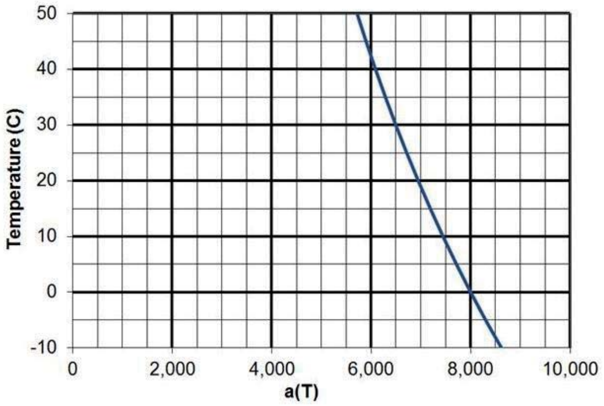

Figure 2. $a(t)$ **from Equation (8) at a pressure of** 10^3 **hPa.**

The mean of the 360 values of the vertical gradient of *M* in Figure 1 is $0.128\ \mathrm{m}^{-1}$. This allows for an instructive description of the vertical gradient of *M* more related to classic atmospheric-boundary-layer thermodynamic structure. Analysis of a wide range of global marine atmospheric-boundary-layer refractivity structures by the Naval Surface Warfare Center, Dahlgren Division, over the last ten years led to the observation that the vertical gradient of *w* is typically two to six times more influential on the vertical gradient of *M* than is the vertical gradient of θ. The radio-frequency engineer is highly knowledgeable about the influence of the dependent variable (*M*) in Equation (7) on RF system performance. The atmospheric-boundary-layer meteorologist is equally knowledgeable regarding the independent variables (θ, w) in Equation (8) and how they relate to classic marine atmospheric-boundary-layer thermodynamic structure. This equation has been successfully employed to exchange valuable knowledge between these two supporting disciplines. The blending technique presented in this paper occurs with $\theta(z)$ and $w(z)$ that leads to a calculation of $M(z)$.

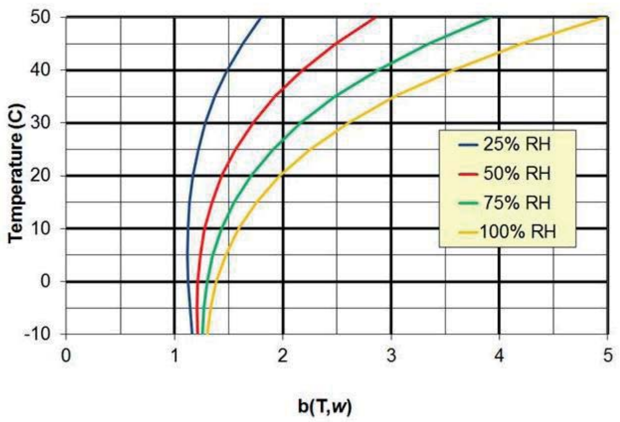

Figure 3. $b(T, w)$ **as a function of relative humidity (%RH) and temperature from Equation (8) at a pressure of** 10^3 **hPa.**

$$\frac{\partial M}{\partial z} \approx 0.128 + a(\theta)\frac{\partial w}{\partial z} - b(\theta, w)\frac{\partial \theta}{\partial z}. \tag{8}$$

From Equation (6), it can be seen that $a(T)$ is weakly dependent on pressure and inversely proportional to the square of potential temperature. Figure 2 is a plot of $a(T)$ for a typical marine atmospheric boundary layer atmospheric pressure of 10^3 hPa for a range of expected temperatures in the marine atmospheric boundary layer. Figure 3 represents $b(T, w)$ over a range of relative humidity and temperature typical in the marine atmospheric boundary layer for an atmospheric pressure of 10^3 hPa.

In recent years, mesoscale numerical weather prediction (NWP) models have been investigated and employed for the purpose of providing prognostic location and time-specific refractivity fields [6-8]. Vertical profiles on horizontally spaced grids of the independent variables in Equation (1) support calculation of vertical profiles of *M* on the same structured grid. Vertical thermodynamic profiles are on discrete levels at each horizontal grid point. Due to roughness length and turbulence modeling requirements, these numerical weather prediction models can only provide data to within 5 m to 10 m of the sea's surface. With the atmosphere at the sea's surface near saturation, the rapid decrease in water vapor with height leads to refractivity structure with significant engineering impacts in this near-surface region, where numerical weather prediction models provide no thermodynamic profile data. Monin-Obukhov-Similarity (MOS) models have traditionally been employed to predict refractivity in the lowest layer, known as the atmospheric surface layer [9].

The complete vertical profile of *M* is thus created by blending two different types of models in a manner that does not produce non-physical gradients in *M*. The mesoscale numerical weather prediction model solves the prognostic equations based on Newton's second law of motion, the equations of state for an ideal gas, the first law of thermodynamics, conservation of mass, and conservation of moisture. The numerical weather prediction model predicts vertical profiles of wind speed, pressure, temperature, moisture, and modified refractivity at discrete vertical levels on a horizontal grid from the surface layer to the stratosphere, and produces vertical profiles every 1 km to 3 km every hour for a 48-hour forecast. The Monin-Obukhov Similarity model assumes that the atmospheric flow is horizontally homogeneous and quasi-stationary; that the turbulent fluxes of momentum, heat, and moisture are constant with height; that molecular exchanges are insignificant compared to turbulent exchanges; that rotational effects can be ignored; and that the influence of surface roughness, boundary-layer height, and geostrophic winds are fully accounted for by the ratio of the surface drag to the air density [10]. The Monin-Obukhov Similarity model parameterizes vertical thermodynamics leading to continuous profiles. For refractivity-profile predictions, the surface-layer model is

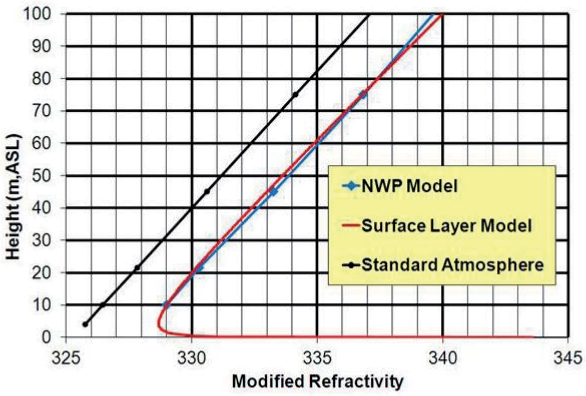

Figure 4. Modified refractivity examples represented by a Monin-Obukhov Similarity surface-layer model in red, and a numerical weather prediction model in blue. The marked data points for the blue numerical weather prediction model curve indicate vertical resolution. The black curve represents a standard atmosphere.

historically referred to as an evaporation-duct model. The problem is demonstrated in Figure 4.

The Monin-Obukhov Similarity surface-layer model in Figure 4 is driven by the wind speed, temperature, and water-vapor content predicted by the numerical weather prediction model at 10 m above sea level in the example in Figure 4. The remaining Monin-Obukhov Similarity surface-layer model input is the sea-surface temperature provided by the numerical weather prediction model at each numerical weather prediction model horizontal grid point. The dilemma, as posed in Figure 4, is to determine at what height to abandon the Monin-Obukhov Similarity surface-layer model and to transfer to the numerical weather prediction model. This must not only be accomplished without producing unrealistic vertical profiles of M, but it must be robustly performed for as many as 56,000 profiles associated with a 48-hour numerical weather prediction forecast.

2. A Zero-Order-Closure Turbulent Flux-Conservation Blending Technique

The use of numerical weather prediction modeling to predict RF system performance is becoming an internationally employed technology. As a result, multiple organizations have developed techniques for blending Monin-Obukhov Similarity and numerical weather prediction refractivity profiles. Prior to the use of numerical weather prediction models, blending techniques were designed for a more limited number of measured profiles of M, where an analyst in the loop was practical.

The Johns Hopkins University Applied Physics Laboratory, Direction general de l'armement (DGA, France), and the Naval Post Graduate School have developed blending techniques for modified refractivity profiles [11-13]. The Defence Science Technology Organisation determines the height of the surface layer in a refractivity-profile-assimilation model by assuming that the wind speed and friction velocity, as well as the potential temperature and temperature scale, are positively correlated at each analysis height [14]

The blending technique described in this paper differs from the techniques described above in two fundamental ways. First, the center of the blending layer is defined as the predicted height of the surface or constant-flux layer. Above this height, the validity of the surface-layer model becomes suspect. Secondly, the blending occurs in thermodynamic space (θ, w, and p), and the resulting blended M curve is calculated from the blended thermodynamic data. Our thinking is that since refractivity is a derived engineering variable, based on the thermodynamic vertical profiles found in natural atmospheric boundary-layer structures, blending is best accomplished at the thermodynamic level. Implicit in this thinking, compared to other models, is that we do not have to assume that ducting always occurs in the surface layer.

The height of the surface layer is defined by the height at which the turbulent fluxes vary by a mathematically tolerable level, commonly taken as 10% [15]. The height of the surface layer is determined by calculating the momentum, heat, and moisture flux at each numerical weather prediction model level, beginning at the lowest, using the Monin-Obukhov Similarity model output driven by the numerical weather prediction at that height. The Monin-Obukhov Similarity accepts sea-surface temperature, and temperature, humidity, and wind speed at each discrete numerical weather prediction height. It is assumed that the air at the surface is saturated, and the wind speed is zero. It is then determined at what increasing numerical weather prediction model height these turbulent fluxes first vary by more than 10% from the surface value, thereby testing one of the major assumptions of the Monin-Obukhov Similarity surface-layer model: that turbulent fluxes are constant with height [10]. Monin-Obukhov Similarity is a zero-order turbulence closure scheme where all resulting equations are fully parameterized [16]. Higher-order closure schemes that employ prognostic equations could be considered, but this Monin-Obukhov Similarity flux-profile approach appears to provide a suitable surface-layer height, leading to a robust engineering product. Two assumptions are made in this determination. It is assumed that the first numerical weather prediction model level is in the surface layer. It is also assumed that the fluxes vary linearly between discrete numerical weather prediction model levels.

The momentum flux is defined as the covariance of the turbulent components of the horizontal and vertical wind speeds:

$$\overline{uw} \equiv \text{vertical momentum flux } = u_*^2 , \qquad (9)$$

$$u_* \equiv \text{friction velocity } = Uk\Big/\left[\ln\left(z/z_0\right)-\psi_m\right], \qquad (10)$$

where

$U \equiv$ mean wind speed at each NWP model height
$k \equiv$ von Karmon constant ≈ 0.4
$z \equiv$ height of each NWP model layer
$z_0 \equiv$ roughness depth
$\psi_m \equiv$ similarity function for momentum.

The heat flux is defined as the covariance of the turbulent components of potential temperature and vertical wind speed.

$$\overline{\theta w} \equiv \text{vertical heat flux } = \theta_* u_*, \qquad (11)$$

$\theta_* \equiv$ turbulent heat scale

$$= \left(\Theta-\Theta_s\right)k\Big/\left[\ln\left(z/z_0\right)-\psi_\theta\right], \qquad (12)$$

where

$\Theta \equiv$ mean potential temperature at each NWP model height
$\Theta_s \equiv$ mean potential temperature at the surface
$\psi_\theta \equiv$ similarity function for heat
$\Theta = T\left(\dfrac{p_0}{p}\right)^{\frac{R}{c_p}}$
$p_0 = 10^3\, hPa$
$p \equiv$ mean pressure at each NWP vertical level
$R \equiv$ gas constant for dray air 2.87×10^2 $\mathrm{Jkg^{-1}K^{-1}}$
$C_p \equiv$ specific heat of dry air at constant pressure $= 1.00467\times10^3$ $\mathrm{Jkg^{-1}K^{-1}}$.

The moisture flux is defined as the covariance of the turbulent components of specific humidity and vertical wind speed.

$$\overline{qw} \equiv \text{vertical moisture flux } = q_* u_*, \qquad (13)$$

$q_* \equiv$ turbulent moisture scale

$$= \left(Q-Q_s\right)k\Big/\left[\ln\left(z/z_0\right)-\psi_q\right], \qquad (14)$$

where

$Q \equiv$ mean specific humidity at each NWP model height
$Q_s \equiv$ mean specific humidity at the surface
$\psi_q \equiv$ similarity function for humidity

$$Q = \frac{0.622e}{p-e} = \frac{w}{w+1}$$

The similarity functions are empirical functions of the ratio of the height in the surface layer (z) to the Monin-Obukhov length (L). L is defined as the height in the surface layer where turbulent kinetic energy (TKE) produced by wind shear is equal to turbulent kinetic energy produced by buoyancy [10]. Using Equations (10), (12), and (14), the turbulent fluxes of momentum, heat, and moisture are calculated at each numerical weather prediction model vertical level, beginning at the lowest.

$$\overline{uw} = u_*^2 = \left\{Uk\Big/\left[\ln\left(z/z_0\right)-\psi_m\right]\right\}^2, \qquad (15)$$

$$\overline{\theta w} = \theta_* u_* \left[\frac{\left(\Theta-\Theta_s\right)k}{\ln\left(z/z_0\right)-\psi_\theta}\right]\left[\frac{Uk}{\ln\left(z/z_0\right)-\psi_m}\right], \qquad (16)$$

$$\overline{qw} = q_* u_* \left[\frac{\left(Q-Q_s\right)k}{\ln\left(z/z_0\right)-\psi_q}\right]\left[\frac{Uk}{\ln\left(z/z_0\right)-\psi_m}\right]. \qquad (17)$$

The first height higher than the first numerical weather prediction model level that corresponds to a 10% change from the nearest surface value for any of the three fluxes is considered to be the point where the Monin-Obukhov Similarity surface-layer profile is gracefully abandoned for the numerical weather prediction profile. The surface-layer model employed in this work was the Navy Atmospheric Vertical Surface Layer Model (NAVSLaM) [17] version 1.0, and the numerical weather prediction model was the Coupled Ocean Atmosphere/Mesoscale Prediction System (COAMPS®) [18]. NAVSLaM was run at 0.25 m vertical resolution. COAMPS® was typically run with a 10 m vertical resolution in the surface layer and a 3 km horizontal resolution. The similarity functions used to calculate the turbulent fluxes at each numerical weather prediction height were those employed by NAVSLaM. L used to evaluate the similarity functions was also provided by NAVSLaM. Fluxes were calculated beginning at the lowest COAMPS® level, and progressing to subsequent higher levels until the fluxes departed from the lowest-level values by 10%.

The technique assumes the surface-layer height is the height at which the COAMPS® derived fluxes vary by 10% from the NAVSLaM (N) fluxes calculated for the constant-flux layer. As an example, examine moisture flux:

$$0.9 \geq \frac{q_C^* u_C^*}{q_N^* u_N^*} \geq 1.1,$$

$q_C^* \equiv$ The required turbulent moisture scale at eachCOAMPS height to provide the specific humidity at that height

$u_C^* \equiv$ The required turbulent momentum scale at each COAMPS height to provide the wind speed at that height

$q_N^* \equiv$ The constant flux layer turbulent moisture scale calculated by NAVSLaM

$u_N^* \equiv$ The constant flux-layer turbulent momentum scale calculated by NAVSLaM

$$q_N^* = \frac{\left(Q_{z,N} - Q_{0,N}\right)k}{\ln\left(\frac{z}{z_0}\right) - \Psi_q}$$

$$u_N^* = \frac{\left(U_{z,N} - U_0\right)k}{\ln\left(\frac{z}{z_0}\right) - \psi_m} = \frac{U_{z,N}k}{\ln\left(\frac{z}{z_0}\right) - \psi_m}$$

$$q_C^* = \frac{\left(Q_{z,C} - Q_{0,C}\right)k}{\ln\left(\frac{z}{z_0}\right) - \psi_q}$$

$$u_C^* = \frac{\left(U_{z,C} - U_0\right)k}{\ln\left(\frac{z}{z_0}\right) - \psi_m} = \frac{U_{z,C}k}{\ln\left(\frac{z}{z_0}\right) - \psi_m}$$

$$\frac{q_C^*}{q_N^*} = \frac{\left(Q_{z,C} - Q_{0,C}\right)}{\left(Q_{z,N} - Q_{0,N}\right)}$$

$$\frac{u_C^*}{u_N^*} = \frac{U_{z,C}}{U_{z,N}}$$

$$0.9 \geq \frac{q_C^* u_C^*}{q_N^* u_N^*} \geq 1.1$$

$$= 0.9 \geq \frac{\left(Q_{z,C} - Q_{0,C}\right)}{\left(Q_{z,N} - Q_{0,N}\right)} \frac{U_{z,C}}{U_{z,N}} \geq 1.1 \qquad (18)$$

Similarly, it can be shown for momentum flux and temperature flux, respectively,

$$0.9 \geq \frac{u_C^* u_C^*}{u_N^* u_N^*} \geq 1.1$$

$$= 0.9 \geq \left(\frac{U_{z,C}}{U_{z,N}}\right)^2 \geq 1.1 \qquad (19)$$

$$0.9 \geq \frac{\theta_C^* u_C^*}{\theta_N^* u_N^*} \geq 1.1$$

$$= 0.9 \geq \frac{\left(\Theta_{z,C} - \Theta_{0,C}\right)}{\left(\Theta_{z,N} - \Theta_{0,N}\right)} \frac{U_{z,C}}{U_{z,N}} \geq 1.1\,. \qquad (20)$$

It thus can be seen from Equations (18) to (20) that the determination of the surface-layer height is in fact calculated by the excursions of the COAMPS® bulk variables from the NAVSLaM bulk variables with height, and these do not depend on the form of the similarity functions.

2.1 Turbulent Flux Calculations

Example heat, momentum, and moisture turbulent flux calculations using Equations (15) to (17) and COAMPS® data are displayed in Figure 5 (COAMPS® is a registered trademark of the Naval Research Laboratory). The COAMPS® data points at 10 m, 22 m, 35 m, and 75 m are indicated. Because all of the required NAVSLaM input parameters were not available from COAMPS® at the surface, it was assumed that the lowest COAMPS® level was in the constant-flux layer, and the turbulent flux at that level was representative of the surface value. Flux values were assumed to vary linearly between discrete COAMPS® levels. The gold vertical bars were the $\pm 10\%$ excursion values. The heat flux varied by 10% from the lowest value at 14.8 m, the momentum flux varied by 10% at 18.2 m, and the moisture flux varied by 10% at 30.9 m. The height of the surface layer for this case was thus assumed to be 14.8 m, as predicted by the heat-flux profile. This height was then defined as the center of the blending zone.

2.2 Blending Zone

Analogous to molecular mean free path, Prandtl [19] proposed a turbulent mixing length over which a parcel of air will maintain identity before mixing with the surrounding fluid. In this blending technique, the Prandtl mixing length

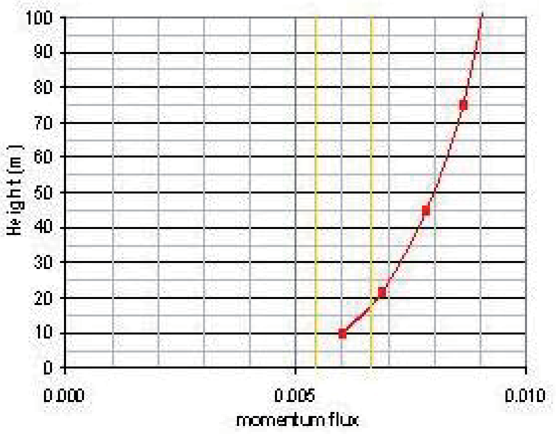

Figure 5a. The momentum flux in Nm^{-2}, calculated from the COAMPS® data depicted in Figure 4 and used in Equation (12). The gold vertical bars indicate the $\pm 10\%$ excursion from the nearest surface value.

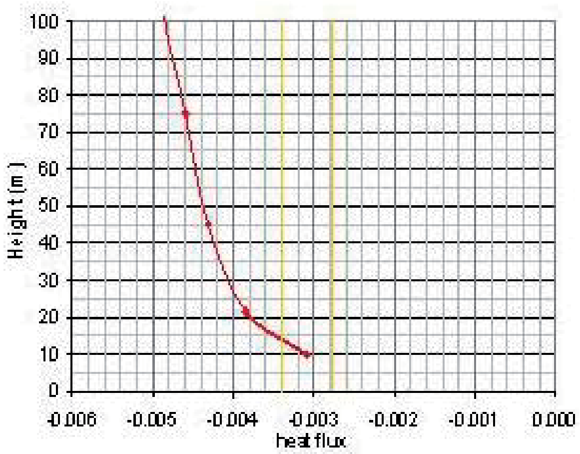

Figure 5b. The heat flux in Wm^{-2}, calculated from the COAMPS® data depicted in Figure 4 and used in Equation (13). The gold vertical bars indicate the $\pm 10\%$ excursion from the nearest surface value.

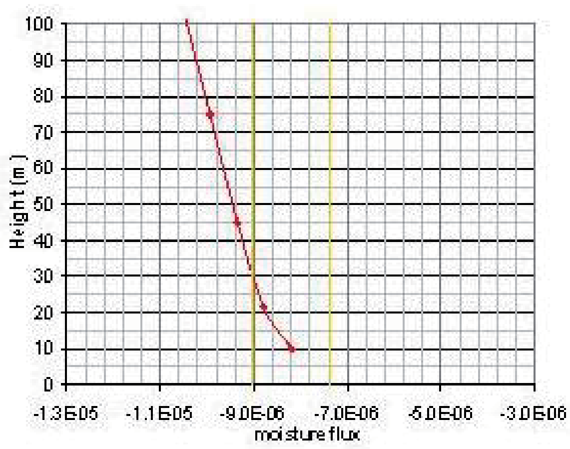

Figure 5c. The moisture flux in $kgsm^{-2}$, calculated from the COAMPS® data depicted in Figure 4 and used in Equation (14). The gold vertical bars indicate the $\pm 10\%$ excursion from the nearest surface value.

defines the depth of the blending zone, and is centered at the blending point determined by the flux-excursion method described in the previous section. There are many parameterizations available in the literature for the Prandtl mixing length. The formulation for the Prandtl mixing length chosen for this blending technique was developed by the European Centre for Medium Range Weather Forecasting (ECMWF) [20], and is displayed in Equation (21):

$$\frac{1}{l_m} = \frac{1}{k z_s} + \frac{1}{\lambda}, \qquad (21)$$

$l_m \equiv$ Prandtl mixing length
$k = 0.4$
$z_s \equiv$ surface-layer height
$\lambda = 150$ m.

For the surface-layer height of 14.8 m in this example, the resulting Prandtl mixing length from Equation (18) was 5.7 m. The resulting blending zone was from 11.9 m to 17.6 m. A $\ln^2(z)$ weighting function was employed every 0.25 m across the blending zone to create a smooth blend of the slopes of the NAVSLaM thermodynamic variables and the slopes of the COAMPS® thermodynamic variables. Linear interpolation was employed between discrete COAMPS® levels. The 0.25 m resolution was chosen as a conservative value for the purpose of resolving evaporation duct heights to within 1.0 m. A $\ln^2(z)$ function was chosen instead of a linear function to minimize the impact of the blend, and to allow the Monin-Obukhov Similarity and numerical weather prediction model data to provide the solution. The weighting function is shown for the current example in Figure 6.

The smoothed thermodynamic slope curves were

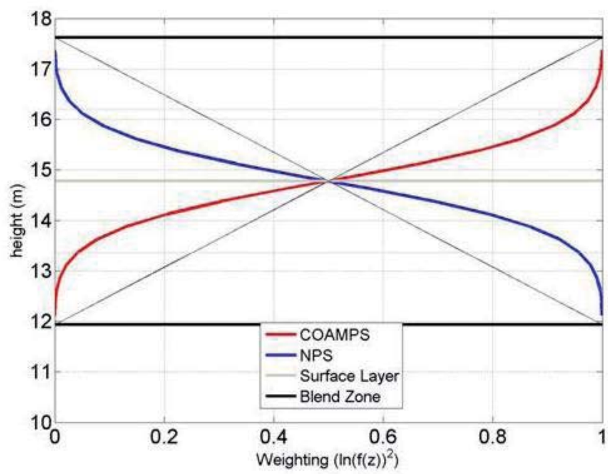

Figure 6. The weighting function for the numerical weather prediction and Monin-Obukhov Similarity models for a surface-layer height of 14.8 m and a blending length of 2.8 m.

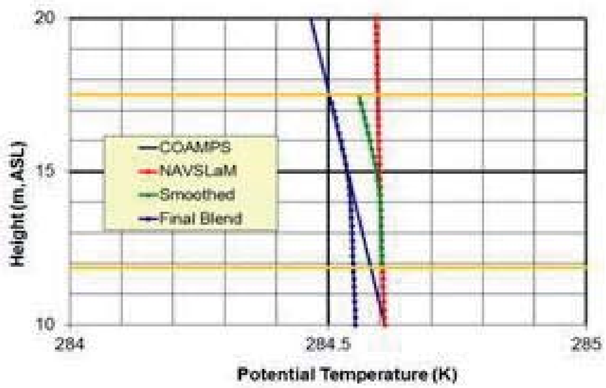

Figure 7a. The blending results for the potential temperature. The horizontal gold lines delineate the blending zone.

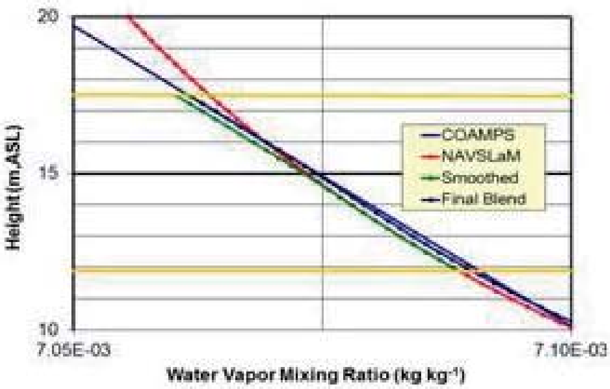

Figure 7b. The blending results for the water-vapor mixing ratio. The horizontal gold lines delineate the blending zone.

vertically integrated to produce vertical profiles of potential temperature and water-vapor mixing ratio. Because the Monin-Obukhov Similarity model and the numerical weather prediction model intercepted at the bulk parameter input level of 10 m, the blending of the thermodynamic curves had to be completed by sliding the blended curves toward the COAMPS® curves so that they matched at the top of the blending layer. This procedure did not affect the vertical gradient of refractivity or the resulting RF propagation. Atmospheric-pressure blending was accomplished in the same manner. This smoothing and blending technique for the current example is shown in Figure 7.

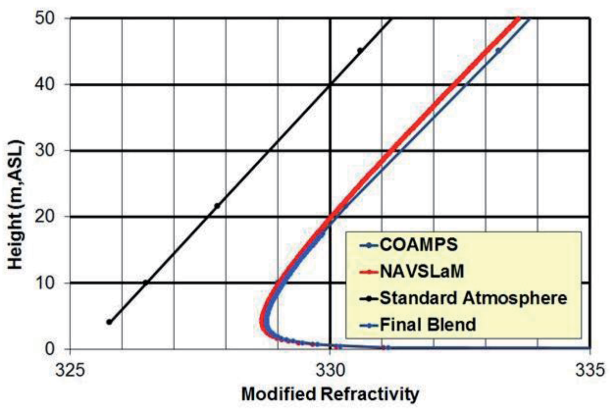

Figure 8. The final blending results for the modified refractivity.

The final blended values of θ and w were converted to T and e through Equations (3) and (4), respectively. The resulting final blended profile of M was calculated from Equation (2). The results for the current example are displayed in Figure 8.

2.3 Data Comparison

The Wallops 2000 Microwave Propagation Measurements Experiment [21] provided a multi-wavelength two-dimensional measured propagation data set for comparison with the blending technique described in the previous sections. Figure 9 delineates a 40 km path off Wallops Island, Virginia, USA where S-, C-, and X-band propagation measurements were made between 1512 UTC and 1605 UTC on April 28, 2000. The multi-wavelength propagation measurements were made from 0 m to 10 m above sea level along this 40 km path. A 3-km-horizontal-resolution COAMPS® grid for 1600 UTC was included the 40 km path. The COAMPS® data were processed by a Barnes objective analysis scheme [22, 23] to produce a meteorological profile along the bearing every 2 km. The blending technique described in the previous sections was applied to each of these 21 profiles. The location of the Naval Postgraduate School meteorological flux buoy is also indicated in Figure 9 [21]. Bulk meteorological parameters from this instrument were used to calculate a single surface-layer refractivity curve to be used with each

Figure 9. The propagation measurement bearing off Wallops Island, Virginia, USA, between 1512 UTC and 1605 UTC on April 28, 2000. The NPS flux meteorological buoy was located along the bearing approximately 13 km offshore.

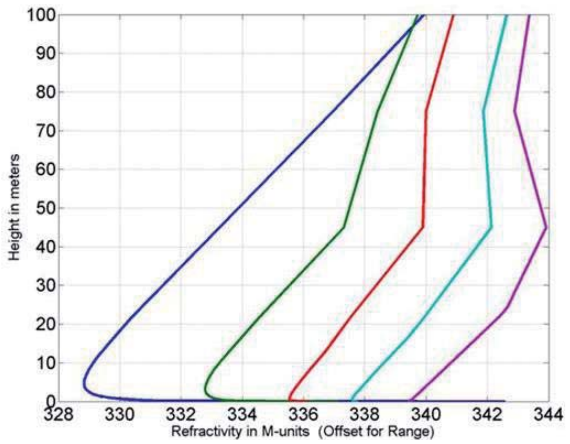

Figure 10. Blended modified refractivity profiles at 0 km, 10 km, 20 km, 30 km, and 40km along the bearing in Figure 9.

helicopter sounding collected along the 40 km path [21].

Figure 10 displays the thinned (for clarity) resulting range-dependent modified refractivity profiles at 0 km, 10 km, 20 km, 30 km, and 40 km, respectively from left to right along the path in Figure 9. The profile at 0 km was that described in the development of the blending technique in the previous sections. Notice the decrease in evaporation duct height in the offshore direction.

Comparisons of the measured S-band one-way propagation factor from 1512 UTC to 1605 UTC with the COAMPS®/NAVSLaM blend one-way propagation-factor prediction are provided in Figure 11. Propagation measurements were made from 0 m to 10 m above sea level along the 40 km path using the Microwave Propagation Measurement System (MPMS) [21]. The propagation modeling was performed by employing the 21 blended profiles along the bearing in Figure 9 with the tropospheric electromagnetic parabolic equation routine (TEMPER) [24]. TEMPER is a parabolic-wave-equation RF propagation model developed at the Johns Hopkins University-Applied Physics Laboratory that considers sea surfaces as well as terrain. At this wavelength, the first multipath null was not observed in this 10-m-deep layer. The difference graphic represents measurement minus model.

A similar analysis for the C-band measurements between 1512 UTC and 1605 UTC on April 28, 2000, is demonstrated in Figure 12. Notice that at this shorter wavelength, the first multipath null is observable in the top left corner. The X-band analysis is provided in Figure 13. Due to the even shorter wavelength, the multipath nulls were more prominent in this 10 m deep layer of one-way

Table 1. S-band comparison statistics.

	Refractivity Source	50th Percentile Absolute Error (dB)	90th Percentile Absolute Error (dB)
All Dates and Times	COAMPS®	0.8	2.4
All Dates and Times	Helicopter	1.3	3.4

Table 2. C-band Comparison Statistics.

	Refractivity Source	50th Percentile Absolute Error (dB)	90th Percentile Absolute Error (dB)
All Dates and Times	COAMPS®	1.2	3.1
All Dates and Times	Helicopter	1.8	4.7

Table 3. X-band Comparison Statistics.

	Refractivity Source	50th Percentile Absolute Error (dB)	90th Percentile Absolute Error (dB)
All Dates and Times	COAMPS®	1.2	5.3
All Dates and Times	Helicopter	2.2	6.8

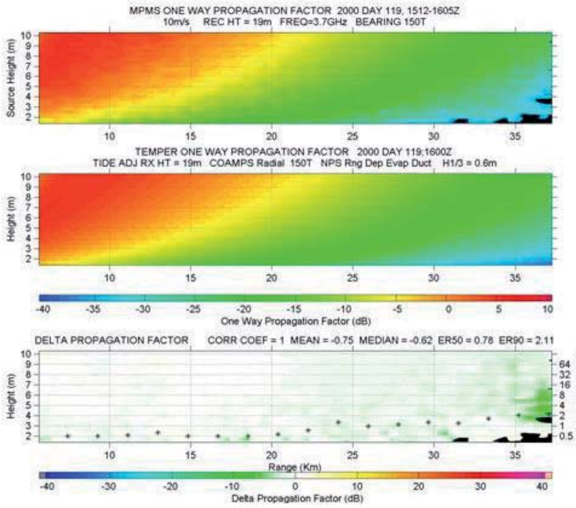

Figure 11. The S-band (3.7 GHz) propagation-factor measurements (upper graphic), the COAMPS®/NPS blend propagation-factor predictions (middle graphic), and the differences (lower graphic), plotted in range and height. The 50th percentile error was 0.8 dB, and the 90th percentile error was 2.1 dB.

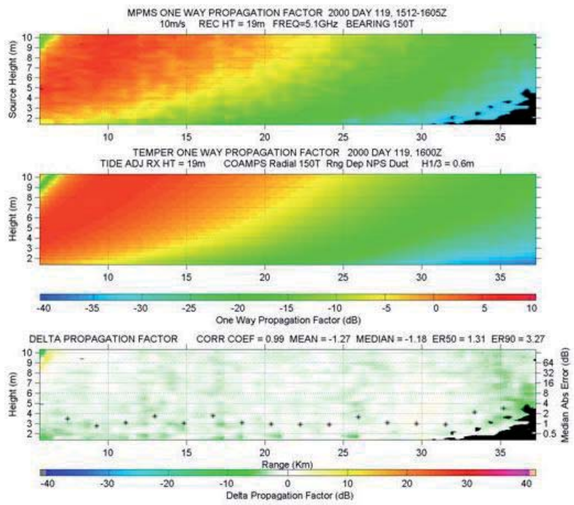

Figure 12. The C-band (5.1 GHz) propagation-factor measurements (upper graphic), the COAMPS®/NPS blend propagation-factor predictions (middle graphic), and the differences (lower graphic), plotted in range and height. The 50th percentile error was 1.3 dB, and the 90th percentile error was 3.3 dB.

propagation factor.

It is important to note that the measurement data was taken over nearly an hour as the transmitter boat sailed along the path in Figure 9. The marine atmospheric boundary layer and the resulting refractivity field were constantly slightly adjusting to the temperature and water vapor in the air flowing over the sea's surface during this period. This data comparison assumed that the propagation factor measured remained constant over this range during the period of collection. It could also be seen that as the multipath nulls showed in the close-range X-band data, the mean absolute error was increased.

A total of six two-dimensional propagation measurements made over three days were used for comparison with similar results. Tables 1 to 3 summarize the 50th and 90th percentile error results for the S, C, and X bands, respectively. They include not only the performance statistics for the COAMPS®/NPS blend, but also the performance statistics for predictions based on range-dependent meteorological soundings made by the Johns Hopkins University Applied Physics Laboratory helicopter [21]. In the case of a COAMPS®/NAVSLaM blend, it is a single COAMPS® model forecast hour. In the case of the helicopter, it is the period over which the meteorological helicopter sampled the vertical structure of the marine atmospheric boundary layer along the MPMS measurement bearing. In some cases, there were multiple helicopter runs for one MPMS propagation measurement. As is standard test and evaluation practice, the fully blended helicopter soundings were composed of helicopter measurements and a surface-layer model. NAVSLaM was driven by bulk inputs from the NPS buoy located approximately 13 km offshore, as in Figure 9. The blending technique was similar to those in the Multi-source Assimilation and Refractivity Interpolator (MARI) [11]. The same evaporation duct defined by data from the NPS buoy was blended with each range-dependent helicopter sounding.

It could be observed from Figures 11 to 13 and Tables 1 to 3 that the blending technique described in this paper produced relatively lower errors compared to measurements. In the median, over all the data-comparison events, the range-dependent COAMPS®/NAVSLaM blend surface-layer refractivity profiles produced slightly better

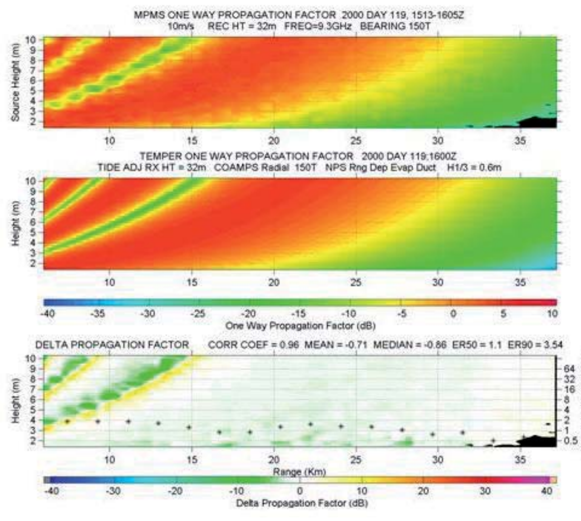

Figure 13. The X-band (9.3 GHz) propagation-factor measurements (upper graphic), the COAMPS®/NPS blend propagation-factor predictions (middle graphic), and the differences (lower graphic), plotted in range and height. The 50th percentile error was 1.1 dB, and the 90th percentile error was 3.5 dB.

results than the constant-evaporation-duct height profiles produced by blending the same surface-layer profile with each helicopter profile. It can also be seen that the measured minus modeled errors tended to increase with decreasing wavelength as the multipath nulls became more prominent.

3. Conclusions

A zero-order-closure turbulent flux-conservation technique for blending refractivity profiles from Monin-Obukhov Similarity and numerical weather prediction models has been described and compared with six two-dimensional multi-wavelength data sets. The technique is unique in that it determines the approximate depth of the surface layer at each numerical weather prediction grid point, and thereby defines the legitimate vertical domain for the surface-layer model at each numerical weather prediction grid point. Blending is accomplished over one Prandtl mixing length with the conserved thermodynamic variables. The blended thermodynamic variable curves are then employed to calculate the final blended modified refractivity profile at each numerical weather prediction grid point. The multi-wavelength data-comparison results indicated that the range-dependent COAMPS®/NAVSLaM surface-layer refractivity profiles in the median over all comparison data showed slight improvement relative to constant evaporation duct heights blended with range-dependent measured soundings.

4. Acknowledgement

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Conformal Antennas for Miniature In-Body Devices: The Quest to Improve Radiation Performance

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Abstract

In-body devices for wireless biotelemetry and neural interfacing could empower many breakthroughs in medicine and clinical research. Despite significant progress in antenna design, improving the radiation performance of miniature in-body devices remains a major challenge because, for the time being, they are able to operate at distances of only up to a few meters. Here, we start by presenting a comprehensive review of existing antenna solutions and operating-range extension methods. We then describe the design and operation of a miniature capsule-conformal microstrip antenna with improved radiation efficiency. We demonstrate that with the proposed methodology, the operating range of deep-body (> 5 cm) devices could reach 10 m to 15 m. This would allow monitoring physiological parameters unaffecting mobility and the quality of life of a user, as well as increasing the energy efficiency of the in-body device.

1. Introduction

Body-implanted electronics came into existence when the first pacemaker was invented in the 1950s [1]. This healthcare innovation exposed the potential and revealed vast opportunities in the field of implantable electronic devices, enabling new ways of diagnostics and treatment, while maintaining a patient's mobility. In particular, Richard Feynman exposed the future of body-implanted electronics in his lecture "There's Plenty of Room at the Bottom" [2]. The first practical realizations of in-body capsule-shaped devices (in-body-capsules), equipped with sensing and transmitting capabilities, appeared in 1957 in *Nature*, in a note from Zworykin and Farrar [3], and a letter from Mackay and Jacobson [4]. Farrar also reported the wireless capsule that senses gastrointestinal motility by means of pressure variation; it operated at frequencies around 1 MHz [5]. Similar devices were reported in the 1960s and 1970s [6-8], all operating in the near field: well below 10 MHz. This substantially limited both the operating range and the throughput of the system [9].

In the 2000s, progress in microelectromechanical systems and microfluidics, along with further miniaturization of electronics and biosensors, empowered a new generation of body-implanted devices. These created numerous applications in medicine, clinical research, wellness, and defense [10-13]. Modern implantable, ingestible and injectable (in-body) devices for biomedical telemetry [14] can continuously monitor and yield physiological parameters (for instance, endoscopic imagery [15], intracorporeal temperature [16], pressure [17], pH [18], glucose levels [19], and so on), while maintaining mobility and improving the quality of life of a patient. Apart from telemetering various diagnostic data, in-body devices can perform the following functions: brain-machine interfacing [20], defibrillation, deep-brain and heart stimulation [21], drug delivery [22], and hyperthermia [23].

Figure 1a. The manufactured prototype of a conformal microstrip antenna by Mahe et al. [63] (©2012 Yann Mahe et al.).

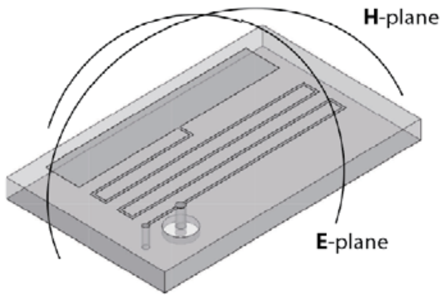

Figure 1b. The design for the antenna shown in Figure 1a (©2012 Yann Mahe et al.).

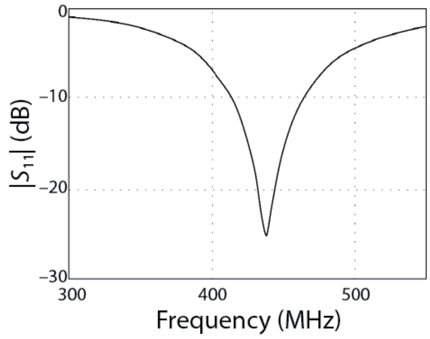

Figure 1c. The simulated reflection coefficient for the antennas shown in Figure 1a (©2012 Yann Mahe et al.).

A modern in-body device typically consists of (1) a biomedical application circuit to acquire physiological data or to provide treatment, (2) a microcontroller to process the data and to give instructions to the device's circuitry, (3) a wireless link containing an antenna, and (4) a power source. The device uses a wireless link to transmit biotelemetry data or to receive operational and treatment instructions [24]. A power source can be a button cell or a miniature wireless powering/recharging circuit [25].

Despite several power-efficient alternatives (e.g., a human-body communication method [26]), far- and midfield (the region where the wavelength is comparable to the distance between an implant and the transmitter [27]) transmissions in the VHF to SHF bands (as defined by the ITU) yield the longest operating range and highest data transfer rates for in-body devices [28].

An antenna operating in the VHF to SHF forms a coupled system of accelerating charges with biological tissues. This system (Tx) radiates to surrounding free space due to non-conserved current and charge in the localized region surrounding the antenna [29]. The antenna's radiation is therefore inseparable from this system, and the operating range is thus the distance from the surface of the body to the antenna of the external equipment (Rx). Assuming the invariable environment (including the fixed position of the in-body device) and given the external receiver properties (antenna gain, receiver sensitivity, and noise level), the operating range depends on the realized gain of the in-body antenna (a function of the reflection coefficient, directivity, and the radiation efficiency) and the input power. In this study, we consider the in-body device to be operating at the maximum authorized input power within the safety limits, and the radiation being omnidirectional. We therefore focus on ways of improving the radiation efficiency and matching stability.

2. In-Body Antenna Types and Performance

Several VHF to SHF antenna designs have been proposed in recent years [30]. The first to appear were standard wire antennas, adjusted to operate in lossy body tissues. For instance, Kwak et al. proposed a helical antenna [31], and Lee et al. reported spiral antennas [32, 33]. Although having satisfactory radiation characteristics (gain $G = -26.2$ dBi in [34] implying a radiation efficiency $\eta < 0.1\%$), wire antennas cannot be cheaply and easily mass produced, and their impedance performance depends strongly on the electromagnetic (EM) properties of surrounding tissues. Printing the antenna on a substrate allows integrating it easily with device circuitry, and partially decoupling the antenna from the surrounding tissues. Planar printed antennas can thus achieve better robustness than wire antennas. Merli et al. proposed a dual-band (MedRadio 403 MHz and ISM 2.45 GHz bands) multilayer spiral antenna [35] that was fully shielded from the circuitry. Liu et al. reported a wideband 2.45 GHz circularly-polarized multilayer antenna [36]. Dissanayake et al. exploited the effects of dialectic loading for miniaturization and impedance-matching purposes, and designed an ultra-wideband (UWB) substrate-integrated slot antenna [37]. Despite having generally better impedance stability than the wire antennas (the radiation efficiency remained $\eta < 0.1\%$), the planar printed antennas are bulky and occupy significant volume (from about 15% to 40%) inside in-body capsules. For this reason, the antenna designs on rigid substrates are used less and less for capsule applications, but remain widely employed for planar devices [38-40].

Conformal designs possess all the advantages of antennas on rigid PCBs, but occupy negligibly small space inside a device (typically less than 5%, Figure 1a). In addition, their electrical size can be larger as they can cover nearly all the surface of the capsule. Various types of conformal designs have been reported. Izdebski et al. designed a meandered dipole antenna conforming to the interior surface of a capsule [41]. The antenna operated at 1.4 GHz with a fractional bandwidth at $|S_{11}| = -10$ dB (hereafter referred to as the "bandwidth" or BW) of about 10%, and a gain $G = -26$ dBi when simulated in the small intestine of the truncated Ansoft human-body model. However, in [42] Rajagopalan et al. used the same antenna to perform link-budget calculations, and reported $G = -32$ dBi. The antenna was sensitive to the surrounding tissue's electromagnetic properties, as well as requiring retuning when circuitry was present. Xu et al. proposed another dipole antenna operating at 403 MHz with an enhanced bandwidth of 36% [43]. However, the gain ($G = -37$ dBi) was significantly reduced compared to the previous antenna. Bao et al. recently developed a tri-band inverted-F antenna [44] with gains of $G = -30,\ 25,\ 23$ dBi for 403 MHz, 915 MHz, and 2.45 GHz, respectively, when simulated inside the stomach of the torso model Gustav (CST *Microwave Studio* [45]).

Much attention has been focused on loop antennas, because – in theory – they may provide higher in-body efficiencies than dipoles and monopoles, due to their predominantly magnetic near field. It allows that the near-field coupling is less to non-magnetic biological tissue (relative magnetic permeability $\mu_r \approx 1$) [46]. However, in practice a magnetic in-body antenna close to the ideal antenna has yet to be proposed, and the radiation efficiency of loop antennas has stayed comparable to counterparts. Nevertheless, some remarkable designs in terms of efficiency-to-bandwidth ratio have been reported. Alrawashdeh et al. designed a range of loop antennas for various applications: from the antennas for knee implants [47, 48] to the multi-band complementary split-ring-resonator-loaded antenna with unprecedented impedance bandwidth [49]. Yun et al. developed the meander-loop antenna conforming to the outer surface of the capsule, thus maximizing the antenna's electrical size [50]. The 434 MHz loop antenna designed by Suzan et al. [51] further improved this approach by extending down to the half-spherical extremity of the capsule. The authors achieved $G = -28$ dBi when simulated in an 80 mm × ⌀180 mm cylindrical phantom with muscle-equivalent EM properties (hereafter, all tissue EM properties are as reported by Gabriel et al. [52-54]).

The main drawbacks of conformal loop antennas are their sensitivity to the capsule circuitry in terms of both impedance and radiation efficiency, and their lack of robustness against the variation of the EM properties of surrounding tissues. For instance, the resonance frequency of the antenna may shift from 403 MHz in muscle to 4 GHz in air [55]. This requires designing antennas with extremely large bandwidths to compensate for the detuning. That means that the antenna losses must increase, and thus the radiation efficiency has to be sacrificed [56-59]. Bao et al. proposed a workaround to increase the robustness to the environment by designing a wideband slot-loop antenna [60]. It operated around $f_0 = 2.45$ GHz with a bandwidth of 114% and a $G = -22$ dBi when simulated in a cuboid (60 mm × 60 mm × 70 mm) muscle-equivalent phantom.

Another type of conformal in-body antenna is the microstrip antenna. These antennas are intrinsically more robust, both to the varying EM properties of surrounding tissues (as f_{res} is defined mostly by the substrate geometry and its dielectric properties [61]), and to the device circuitry (as the antenna contains the ground plane). On the other hand, microstrip antennas usually have much narrower bandwidths [62] compared to the above-mentioned types of antennas, especially considering that substrates should be as thin as possible to minimize the volume occupied by the antenna. Finding a proper miniaturization technique is also a challenge. Cheng et al. used a complementary split-ring resonator (CSRR) to load a 2.45 GHz $\lambda/2$ patch antenna, reducing its length to 10.5 mm. The substrate was 254 µm thick. No data were given on the antenna's radiation performance in-body. Mahe et al. employed a $\lambda/4$ stepped-impedance resonator along with meandering to miniaturize a 434 MHz antenna (Figure 1) [63] (a 400 MHz version was reported in [64]). The substrate was 50 µm thick. The gain was $G = -33$ dBi and the bandwidth was $\approx 10\%$ when simulated in a ⌀200 mm cylindrical phantom with $\varepsilon_r = 49.6$, $\sigma = 0.51\ \mathrm{Sm}^{-1}$. Finally, Psathas et al. proposed a microstrip meandered antenna operating at 402 MHz [65]. The bandwidth was comparable to the previous antenna ($\approx 10\%$), with the gain reaching $G = -30$ dBi when simulated in a homogeneous 100 mm^3 cubic phantom with muscle-equivalent EM properties. The miniaturization employed a shorting pin and meandering techniques, which permitted obtaining dual-band characteristics as well (such as, for instance, in [38]).

Despite significant progress in antenna design, improving the operating range of miniature in-body devices remains a major challenge (for the time being, they are able to operate only up to a few meters). Among the main issues to face are low radiation efficiencies ($\eta < 0.1\%$) and antenna-impedance detuning due to strong coupling to lossy and dispersive biological tissues.

3. Extending Operating Range

The radiation efficiency is constrained by attenuation and reflection losses in tissues [10], as well as by the required miniaturization of antennas, which imposes fundamental limits on the maximum achievable radiation efficiency [66]. The antenna detunes, since its impedance strongly depends on the electromagnetic properties of biological tissues that vary both in terms of permittivity and conductivity [53]. Ingestible antennas are more susceptible to impedance detuning than implantable antennas, as they must operate in a dynamically changing gastrointestinal (GI) tract.

3.1 Choosing Optimal Operating Frequency

The maximum achievable radiation efficiency, η, of an ideal in-body antenna depends on the operating frequency [10]. The optimal frequency range depends on how deep the antenna is in the body, what tissues surround it, and how electrically large the antenna is. Poon et al. [67] showed that the optimal wireless powering of an in-body device by a small current loop can be achieved by operating at GHz frequencies (mid-field region). A number of studies suggest that an optimal frequency exists as well for the energy transfer in the far field from a body to surrounding free space [68-70]. Considering the dispersive properties of body tissues, attenuation and reflection losses, and the efficiency-to-bandwidth limitations of electrically small antennas [58], the optimal frequency range for deep-body implantation ($d > 3$ cm) was found to be in the 10^8 Hz to 10^9 Hz range [68, 69]. The most-suitable unlicensed frequency bands for deep-body applications (e.g. ingestible) are therefore the MedRadio (Medical Device Radiocommunications Service) (401 MHz to 406 MHz, 413 MHz to 419 MHz, 426 MHz to 432 MHz, 438 MHz to 444 MHz, 451 MHz to 457 MHz) [71], ISM (Industrial, Scientific and Medical Service) (433 MHz to 434.8 MHz) [72], and WMTS (Wireless Medical Telemetry Service) 611 MHz [73]. As for the shallow implantation (e.g., subcutaneous), the closer the antenna is to the skin, the less the signal attenuates within the tissue. Higher frequencies are Therefore preferable. The most-efficient bands for subcutaneous applications would be then the ISM 915 MHz, WMTS 1.4 GHz, ISM 2.45 GHz, and up to the ISM 5.8 GHz.

3.2 Improving Radiation Efficiency

Considering that the operating frequency, f_0, is close to optimal, the radiation efficiency is still constrained by the attenuation and reflection losses in tissues. In addition, the required miniaturization of antennas operating below ≈ 1 GHz imposes fundamental limits on the achievable radiation efficiency for electrically small antennas [66].

At least two approaches are currently under development [74]: 1) magnetic antennas (e.g., loops) and 2) dielectrically loaded microstrip antennas. Each has its advantages and drawbacks. A magnetic-type antenna couples less to tissues due to the magnetic component dominating in its near field [46]. As stated above, loop antennas give good radiation performance. However, a spatial variation of tissue EM properties affects the antenna's impedance, and thus can deteriorate matching. Typically, to compensate for the antenna's detuning, an extended bandwidth, BW, is preferred. However, this limits the maximum-achievable radiation efficiency for a given antenna size, as $\mathrm{BW} \propto ka/\eta$ [58]. Bao et al. recently demonstrated a potential workaround of this problem by designing a robust multi-band loop-slot antenna [60].

Another technique is dielectric loading: this increases the robustness of antennas [28, 37], which allows reducing the necessary bandwidth and increasing the electrical size, ka. Merli et al. [75] and Skrivervik [76] studied the effect of implantable device encapsulation on radiation efficiency, and found that it improves with increasing permittivity of encapsulation. Increased radiation efficiency was achieved in [28, 40] by loading microstrip antennas with a thin ceramic superstrate (< 1 mm). A high-permittivity filling of the device – ideally, matched with the tissue's permittivity – helps to further increase η [59]. In this way, a dielectrically loaded microstrip antenna can potentially outperform loop antennas in both robustness and radiation performance.

3.3 On-Body Matching Layers and Repeaters

Merli et al. studied the effects of on-body matching layers that improve power transmission from a body [75]. An ideal matching layer (71 mm thick) can increase the power transmission up to 8 dB. Realistic materials (fiber, neoprene, and silicon have been studied) with reasonable thicknesses (up to 20 mm) can provide up to 2 dB improvement.

Kiourti et al. proposed a dual-band on-body repeater antenna [38]. The antenna received a low-power MedRadio signal (401 MH to 406 MHz) from an in-body implant, and retransmitted it using the 2.45 GHz ISM band. This approach reduced the required implant power by a factor of 100 to achieve the same range without the repeater antenna.

4. Towards Conformal Antennas with Enhanced Radiation Performance

The aforementioned approaches allow one to substantially improve the operating range of in-body devices. Depending on the receiver antenna's sensitivity and link-budget evaluation, a range of 10 m to 15 m is foreseeable in the near future for deep-body-implanted devices ($d \approx 5$ cm) [59]. Along with an on-body repeater, this range can be further extended.

Here, we demonstrate the improved radiation efficiency of a 434 MHz conformal microstrip antenna and a method of achieving it. An antenna fitting within a 17 mm × ⌀7 mm is suitable for a wide range of in-body capsule applications. Its ultra-miniature dimensions, enhanced robustness, and efficiency (compared to counterparts) allow one to employ it for both implantable and ingestible applications (such as, e.g., BodyCAP e-Celsius – intracorporeal temperature telemetry capsule [77]).

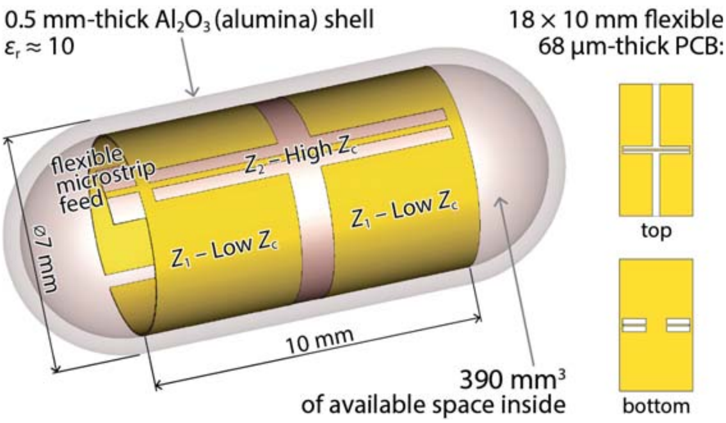

Figure 2. A conformal microstrip in-body antenna inside a 17 mm × ⌀7 mm biocompatible capsule.

4.1 Antenna Miniaturization and Synthesis

Fitting a 434 MHz antenna ($\lambda/2 \approx 34$ cm) within the 17 mm × ⌀7 mm capsule (Figure 2) required miniaturization. A stepped-impedance resonator (SIR) technique [78] showed good performance in terms of miniaturization of microstrip in-body antennas [28, 63]. We used a $\lambda/2$ stepped-impedance resonator with two impedance steps – low Z-to-high Z and high Z-to-low Z – in order to reduce the antenna's size and to match the antenna's input impedance to $Z_{ANT} = 50\ \Omega$.

A $\lambda/2$ stepped-impedance resonator antenna could be synthesized using the transmission-line impedance equation [79, p. 59] for the three-section transmission line with identical first and third elements (Figure 3a):

$$Z_1\left[Z_1 \tan\left(\beta_2 l_2\right) + 2Z_2 \tan\left(\beta_1 l_1\right)\right]$$

$$-Z_2^2 \tan^2\left(\beta_1 l_1\right)\tan\left(\beta_2 l_2\right) = 0\,, \tag{1}$$

where Z_n is the characteristic impedance of each section [Ω], β_n is the phase constant (rad/m), and l_n is the section length [m], with $n = 1,2$. The phase constant, β_n, is defined as

$$\beta_n = \frac{2\pi}{c} f_{res}\sqrt{\varepsilon_{r_n}^{eff}}\,,\quad n = 1,2\,, \tag{2}$$

where $c \approx 3\times10^8\ \mathrm{ms}^{-1}$ is the speed of light in vacuum and $\varepsilon_{r_n}^{eff}$ is the effective relative permittivity of the transmission-line environment. The effective relative permittivity, $\varepsilon_{r_n}^{eff}$, is defined as the relative permittivity of a homogeneous medium that equivalently replaces the heterogeneous medium [79, p. 148].

Equation (1) relates the high Z_1 and low Z_2 line geometries and EM properties of the environment to a resonant frequency f_{res} (when $\mathrm{Im}\left(Z_{ANT}\right) = 0$).

The characteristic impedance of a generic TEM transmission line with negligible losses is given by

$$Z_c = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \approx \sqrt{\frac{L}{C}}, \tag{3}$$

where L is the specific (i.e., per unit length) inductance (Hm^{-1}) and C is the specific capacitance (Fm^{-1}). C and L may be represented through the phase velocity, v_p [80, p. 457]:

$$v_p = \frac{1}{\sqrt{LC}} = \frac{1}{\sqrt{\mu^{eff}\varepsilon^{eff}}}\,, \tag{4}$$

$$\sqrt{L} = \frac{1}{v_p\sqrt{C}} = \sqrt{\frac{\mu^{eff}\varepsilon^{eff}}{C}}\,.$$

Substituting Equation (4) into Equation (3) gives the following expression for the characteristic impedance:

$$Z_c = \frac{\sqrt{\mu_r^{eff}\varepsilon_r^{eff}}}{cC}\,. \tag{5}$$

As neither the capsule environment nor the biological tissues exhibit ferromagnetic properties, we can assume that $\mu_{r_n}^{eff} = 1$. One can therefore evaluate the specific capacitance, C, as

$$C = \frac{2W_e}{V^2}\,, \tag{6}$$

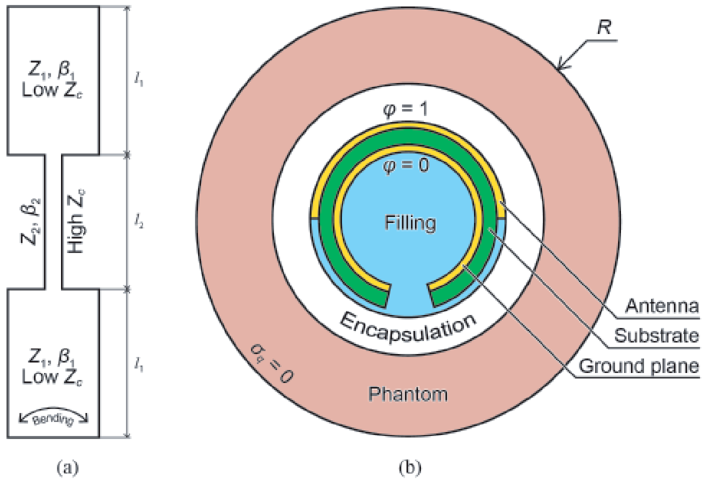

Figure 3. The theoretical model of the antenna: (a) an *H*-type half-wave microstrip stepped impedance resonator, (b) the two-dimensional cross section (not to scale) of the folded antenna inside a capsule within a circular phantom of radius *R*.

where W_e is the total electrical energy stored in the system [J], and V stands for the electrical potential difference [V].

The closed-form solution of Equation (6) is possible only for the simple cases. A numerical solution is required for the given geometry cross section (Figure 3b). The governing partial differential equation (PDE) can be written as $-\nabla \bullet (\delta \nabla \varphi) = \rho$, where ϕ is the sought-for electrostatic potential, and ρ is the charge density. The boundary conditions were defined as the following (Figure 3b): $\phi = 1$ on the cross-section perimeter of the antenna, $\phi = 0$ on the ground plane, and the zero-charge (Neumann) condition was imposed on the outer boundaries of the studied domain as $\varepsilon \partial \varphi / \partial n_0 = 0$. The study domain of radius R circumscribed the cross section of the structure (Figure 3b). The radius R had to be large enough ($R = 4$ cm in our case) in order to accurately estimate the specific capacitance, C.

We used the in-house software *Agros2D* [81] to analyze the formulated problem with the fully *hp*-adaptive finite-element electrostatic formulation of Equation (6). The automatic *hp*-adaptivity was set up to maintain the relative error $\Delta C / C \leq 1\%$. Curvilinear elements were used to accurately represent the circular geometry of the model (Figure 3b).

With this synthesis methodology, the following was necessary to make the most of the design in terms of the radiation efficiency:

1. maximizing the electrical size of the antenna, *ka*, by using the entire cylindrical surface of the capsule;

2. further increasing *ka* by loading the antenna with high permittivity and low loss materials;

3. avoiding meandering the high Z_1 section, therefore preventing the oppositely flowing currents and improving polarization purity;

4. decoupling the antenna from tissues (the results in a narrower bandwidth, as $BW \propto ka/\eta$);

5. averting the impedance detuning due to the varying EM properties of tissues by using a high-permittivity superstrate of a sufficient thickness (determined during optimization routine).

Increasing the substrate thickness of an in-body microstrip antenna results in a higher radiation efficiency, as well [82]. However, the antenna on a thicker substrate occupies more volume inside the capsule. It therefore requires finding a tradeoff between the size and the efficiency. In this study, we used a 50-μm-thick substrate.

4.2 Materials

The conformal antenna was designed for the 50-μm Rogers ULTRALAM® 3850HT substrate (liquid crystal polymer, $\varepsilon_r = 2.9$, $\tan\delta = 0.022$ at 434 MHz). The plating thickness was 9 μm. The capsule shell (acts as the antenna superstrate) was made of Al_2O_3 (alumina, $\varepsilon_r = 10$, $\tan\delta = 10^{-4}$ at 434 MHz) that was biocompatible, inexpensive, and widely available. The shell was 0.5-mm thick, the outer capsule diameter was ⌀7 mm (Figure 2). The capsule with was filled with pure water ($\varepsilon_r = 78.4$ $\tan\delta = 0.022$ at 434 MHz) to further enhance the dielectric loading of the antenna. However, for the industrial implementation, solid high-permittivity (and low-loss) materials may be preferred: for instance, a ceramic-powder-loaded polymer [83] or epoxy [84].

5. Numerical Design and Optimization

We used CST *Microwave Studio® 2016* [45] to model and optimize the antenna. At first, a planar antenna approximation was solved and fine-tuned (in terms of impedance) using the CST time-domain solver (finite integration technique), then validated using the frequency-domain solver (Finite-Element Method (FEM)). Both methods used an automatic adaptive-mesh refinement with

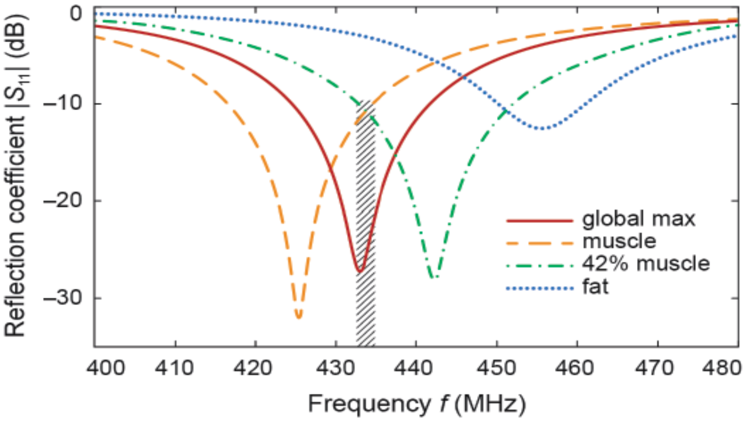

Figure 4a. The antenna's performance in the homogeneous ⌀100 mm spherical phantom: the reflection coefficient, $|S_{11}|$ for muscle-equivalent ($\varepsilon_r = 56.9$, $\sigma = 0.81$ Sm^{-1}), global max ($\varepsilon_r = 77.3$, $\sigma = 2.26$ Sm^{-1}), 42% of muscle ($\varepsilon_r = 23.9$ $\sigma = 0.34$ Sm^{-1}), and global min (fat-equivalent, $\varepsilon_r = 11.6$, $\sigma = 0.08$ Sm^{-1}) EM properties of the phantom.

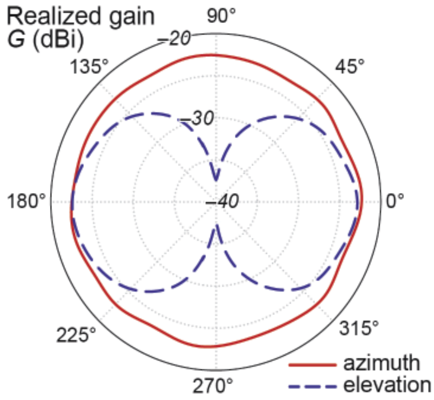

Figure 4b. The antenna's performance in the homogeneous ⌀100 mm spherical phantom: the simulated radiation pattern for muscle-equivalent EM properties of the phantom.

$\Delta\left[\left|S_{11}\left(f_{res}\right)\right|\right] < 1\%$ error tolerance for three consecutive iterations as a convergence criterion. The time-domain models used hardware acceleration (two NVIDIA Tesla K20c GPUs) that significantly reduced the simulation time and, therefore, the fine-tuning routine. The optimized parameters of the planar antenna initialized the conformal antenna analysis and the tuning in a spherical phantom using the FEM solver. The automatic adaptive-mesh refinement settings were set as for the time-domain solver.

The final (conformal) antenna was analyzed and fine-tuned centered inside a ⌀100 mm spherical homogeneous phantom (as in [59]) with EM properties varying from fat (global min, $\varepsilon_r = 11.6$, $\sigma = 0.08$ Sm^{-1}) to global max ($\varepsilon_r = 77.3$, $\sigma = 2.26$ Sm^{-1}) in order to assess the robustness (both tissue EM properties at 434 MHz according to Gabriel et al. [52-54]). A spherical phantom allows accurate estimation of the radiation performance of antennas with undefined position in a body (such as, e.g., an ingestible) as the spherical shape keeps the intrinsic radiation pattern of the antenna. Cuboid and cylindrical shapes may mislead regarding the antenna's performance in terms of directivity.

However, this method does not provide an accurate estimate of the antenna's radiation performance for realistic in-body scenarios. For this purpose, we used an anatomically realistic phantom, "Nelly," implemented in CST *Microwave Studio*® (optimized for a tetrahedral mesh).

5.1 Antenna Performance

To maximize the robustness, we tuned the antenna to resonate below the mid-ISM band, so that $|S_{11}| = -10$ dB at 434.79 MHz (the upper bound of the 434 MHz mid-ISM band, Figure 4a) in the phantom with global max EM properties ($\varepsilon_r = 77.3$, $\sigma = 2.26$ Sm^{-1}). The antenna remained well matched (i.e., $|S_{11}| < -10$ dB) for up to 42% of muscle EM properties ($\varepsilon_r = 23.9$, $\sigma = 0.34$ Sm^{-1}, Figure 4a). This range ($\varepsilon_r \in [23.9,\ 77.3]$ $\sigma \in [0.34,\ 2.26]$Sm^{-1}) covered the majority of biological tissues except fat and bones [52-54]. In fat, the antenna matching was $|S_{11}| = -3.2$ dB at f_0; however, $f_{res} = 456$ MHz. Therefore, $|S_{11}| < -10$ dB in fat (as well as in bones) could be achieved using a matching circuit. The bandwidth was 17 MHz, which was substantially narrower than that of the counterparts, indicating better decoupling of the antenna from lossy tissues. In addition, this bandwidth was sufficient for robust operation of the antenna with 50 Ω feeds in the majority of biological tissues.

Figure 4b shows the radiation pattern of the antenna at 434 MHz inside the ⌀100 mm spherical phantom with muscle-equivalent EM properties. The radiation pattern was dipole-like, which was consistent with the electrically small antenna fundamental limitations on directivity. The realized gain was $G = -22.4$ dBi, and the radiation efficiency reached $\eta = 0.4\%$. The radiation performance of the antenna in realistic phantoms (implantation in fat, stomach, colon, and head have been studied), its robustness to circuitry, and the link-budget evaluations were given in [59]. The antenna transmitted for up to 14 m for all studied implantation sites (except in fat, where in the worst case – the direction of the radiation minimum, $\min[G]$ – the signal-to-noise ratio, SNR, > 0 held up to 7 m).

6. Experimental Characterization

Despite the existence of reliable methods for impedance measurements, the radiation characterization of both electrically and physically small antennas remains challenging. The antenna under test (AuT) couples strongly to a feeding cable that leads to impaired measured data [85]. Currents induced on the cable affect

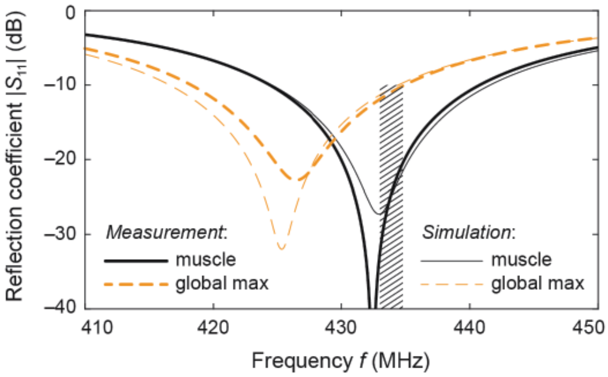

Figure 5a. The antenna's experimental reflection coefficient, $|S_{11}|$, in liquid phantoms with muscle-equivalent, global max, and average GI EM properties.

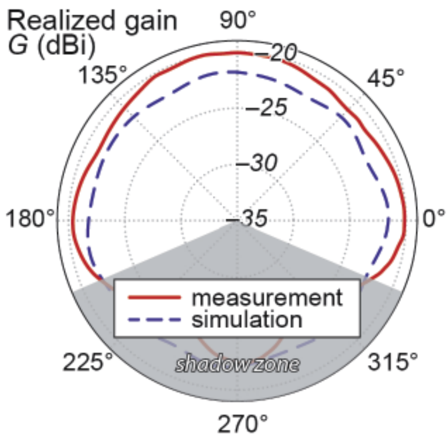

Figure 5c. The experimental far-field characterization: the measured co-polarized radiation pattern of the antenna under test contained in a ⌀100 mm sphere filled with muscle-equivalent liquid phantom.

both the radiation pattern and the matching, confounding the antenna's radiation with the radiation from the cable. Different approaches allow minimizing this effect: baluns [86], differential-feeding techniques [51], electro-optical converters [87], or measurements using a monostatic scattering characterization of an antenna [88]. For in-body antennas, the problem is more complicated since in addition to these sources of error, the antenna has to be measured in a lossy phantom.

6.1 Prototype Manufacturing and Phantoms

We used the laser ablation technique (LPKF ProtoLaser S) to prototype the antenna on a 50-µm Rogers ULTRALAM® 3850HT substrate (Figure 11a) with 9-µm thick copper. The printed antenna was next bent to conform inside a 17 mm × ⌀7 mm alumina tube. This was then soldered to a 100 mm 50 Ω coaxial feed terminated

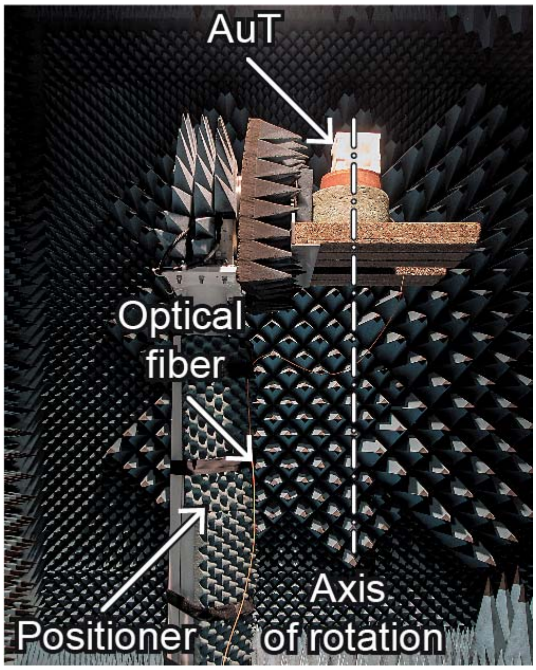

Figure 5b. The experimental far-field characterization: the antenna under test in a phantom mounted on the positioner and fed using an optical fiber.

by an SMA connector. Finally, the capsule was filled with de-ionized water and sealed at both ends. To insulate (with air) the cable from the phantoms, we fitted a ⌀10 mm polyamide tube around the cable, and between the capsule and the SMA connector.

We prepared two liquid phantoms with muscle-equivalent and global-max EM properties to experimentally validate the antenna's performance in terms of impedance, robustness, and radiation. We used a water-sucrose-NaCl recipe to achieve the target EM properties; the concentrations were given in [59]. The EM properties of the phantoms were verified using the SPEAG DAK kit with the DAK-12 probe [89].

6.2 Impedance Characterization

We used an Agilent PNA-X vector network analyzer to measure the antenna's impedance. The reference plane was set at the SMA connector. The cable's effect on the impedance was de-embedded using CST's *Design Studio*. Figure 5a shows the measured reflection coefficients for two liquid phantoms. The experimental results agreed well with the simulations. More-precise manufacturing methods (for instance, photolithography) could further improve the agreement.

6.3 Radiation Performance

To evaluate the radiation performance using the far-field direct-illumination technique (Figure 5b), we used a

Table 1. The designs, capsule sizes, operating frequencies, and radiation performance in the given phantom of conformal capsule antennas as reported in the literature.

Ref.	Type	Size [mm]	f_0 [MHz]	G_{f_0} [dBi]	Phantom: Tissue, Shape, Size [mm]
This	Microstrip	17 × ⌀7	434	–22	Muscle, sphere, ⌀100
[63]	Microstrip	17 × ⌀7	434	–33	$\varepsilon_r = 49.6$ $\sigma = 0.51$, cyl., ⌀200
[65]	Microstrip	24 × ⌀10	402	–30	Muscle, cube, 100^3
[41]	Dipole	26 × ⌀11	1400	–26	Muscle, cuboid, $350^2 \times 200$
[43]	Dipole	24 × ⌀11	402	–37	Skin, cube, 180^3
[44]	IFA	25 × ⌀11	402	–30	Stomach, anatomical
[49]	Loop	25 × ⌀10	402	–25	Mus., ellip. cyl., 180 × 100 × 50
[60]	Slot-loop	25 × ⌀10	402	–22	Muscle, cuboid, $60^2 \times 70$

⌀100 mm spherical glass container filled with a muscle-equivalent liquid phantom. The antenna under test was centered inside of the container. An electro-optical converter (enprobe LFA-3 [90]) fed the AuT so that an optical fiber replaced the RF cable inside an anechoic chamber, thus minimizing the main source of signal impairment. The estimation of the realized gain used the gain-substitution technique, employing a reference antenna of known gain (ETS-Lindgren Model 3164-06 [91], 300 MHz to 6 GHz).

Figure 5c shows the antenna's far field. The maximum realized gain was about –20 dBi for co-polarization, and –33 dBi for cross-polarization (X-pol). The far-field pattern was consistent with the simulated pattern except for the backward radiation, which was distorted due to the shadowing effect of the tower positioner (Figure 5b).

7. Conclusion

Enhancing the operating range of biotelemetry devices up to 10 m to 15 m would allow monitoring physiological parameters without affecting the mobility and quality of life of a user. In addition, power-efficient wireless **neural interfaces could empower many breakthroughs in medicine and clinical research.** Substantial improvement of the range or power efficiency is achievable by combining the approaches reviewed with the radiation-efficiency enhancement of body-implanted antennas.

In particular, conformal microstrip designs showed enhanced radiation performance for in-body applications. As demonstrated by the design, a thin microstrip antenna on a 50 µm-thick substrate, conforming to a miniature 17 mm × ⌀7 mm capsule, could operate up to 14 m from various in-body sites. The miniature dimensions of the antenna, and its enhanced robustness and efficiency (compared to counterparts, Table 1) would allow one to employ it for both implantable and ingestible applications.

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In Memoriam: Jean Van Bladel

It is a privilege for me to pay tribute to Jean Van Bladel, my predecessor as Secretary General, who had a lasting impact on URSI, on science, the scientific community and engineering education and students for over more than half a century.

When in 1968 Professor Jean Van Bladel asked me whether I would join him as an assistant and PhD student, I could not imagine that this would turn out to be a defining moment in my professional life. Having worked closely with Jean Van Bladel at the university, in the framework of the International Union of Radio Science (URSI), and also in the Royal Belgian Academy of Science and Arts, I am very well aware that it is completely impossible to enumerate here all his achievements and awards. So allow me to limit myself to highlight the main points.

After working as an associate professor and full professor at Washington University, St. Louis and the University of Wisconsin, Madison in the USA , Jean Van Bladel returned to Belgium to found, as a full professor, the Laboratory of Electromagnetism and Acoustics at Ghent university. Until his retirement in 1987 he was the driving force behind the growth and the scientific successes of his department. In addition, between 1976 and 1978, Jean Van Bladel served as Dean of the Faculty of Applied Science of the University and, from 1981 until his retirement as professor emeritus in 1987, he was a member of the University Board.

In 1984 Jean Van Bladel was elected as Secretary General of URSI, the International Union of Radioscience. During 10 years until I succeeded him in 1994 he devoted significant time and energy to the reorganization and revitalization of URSI. In recognition of his tireless efforts and tactful diplomacy in tricky international situations he was awarded the title of Honorary President of this International Union of Radioscience.

In 1984 he was elected a member of the Royal Belgian Academy of Science and Arts. Among his contributions to the Academy let me just mention that from 1990 on he served as the delegate of the Belgian Academies to ICSU, the International Council for Science, and in 1995 he was elected president of the Royal Belgian Academy.

Over his distinguished scientific career Jean Van Bladel received many prestigious awards such as Heinrich Hertz Medal and the Antennas and Propagation Society's 1997 Distinguished Achievement Award from the IEEE, an Honorary Doctor's Degree from the University of Liège and both in 1978 and 1984 he was awarded the Francqui Chair at the Free University of Brussels. A complete list of his noteworthy professional service and other honors and awards is too lengthy to delineate here, but let me simply say that it is truly impressive by all international standards.

However I have to mention the vast and outstanding contributions of Jean Van Bladel to science in general and to electromagnetic theory and applications in particular. His books and numerous papers in scientific journals have been important resources to researchers and teachers alike, and they will stand the test of time. Over a long and distinguished career, which lasted for over 60 years, he has exhibited a rare depth of understanding and originality and he, therefore, commands the respect of everyone in the electromagnetics community. Through his scientific contributions and the gratitude of the many students and scientists he advised, helped and inspired his memory will live on for many years to come.

For me personally Jean Van Bladel was a teacher and a mentor who I admired very much. He was in the true, profound sense of the word a gentleman, a man of integrity, a man of culture, gentle and trustworthy, a man for whom duty and honor were all important.

Professor Paul Lagasse
Honorary President of URSI

In Memoriam: Karl Rawer

It is with sadness that we report the passing of Prof. Karl Rawer at his home in March, Germany, two days before his 105th birthday. He died peaceful during his afternoon nap, with the book that he was reading on his lap. Prof. Rawer was one of the pioneers of the exploration of the ionosphere from the ground and from space, and in the understanding of radiowave propagation in the ionospheric medium. He leaves a rich legacy and a large body of work in ionospheric and atmospheric physics, including his book, *Die Ionosphäre*, published in 1952 and translated into many languages. He received many honors. He was the Director of the Frauenhofer Institute for Space Research in Freiburg, Germany, and the project scientist of the first German-US Aeronomy Satellites, AEROS-A and AEROS-B.

Karl Rawer was born on April 19, 1913, in Neunkirchen, Germany. He studied mathematics and physics under Gustav Mie and Gustav Dötsch in Freiburg, and under Arnold Sommerfeld and Jonathan Zenneck in München. In his dissertation in 1939, he used hyperbolic and Epstein functions to solve, for the first time, the problem of radiowave propagation in a stratified medium. From there on, he was hooked on ionospheric research. During World War II, he was charged with developing ionospheric predictions in support of HF communications, working with Johannes Plendl and Walter Dieminger. After the war, he was invited to establish an ionospheric-prediction service in what was then Germany's "French Zone." He developed close ties with his French counterparts, and was the first German after the war to be invited to lecture at the Sorbonne University (prof. associé 1958-1960, and prof. d'échange 1961-1964). After first working under the auspices of the French Service Prévision Ionosphérique de la Marine (SPIM), Rawer in 1956 established the "Ionosphäreninstitut" in Breisach, Germany, and served as its director from 1956 to 1969. This institute gained international reputation in the field of ionospheric radio wave propagation and forecasting; the development of ionosondes in cooperation with research organizations in the USA and France; and joint measuring campaigns in Italy, Greece, Norway, and Africa. Throughout his career, Prof. Rawer was a strong proponent of international science cooperation, recognizing its potential to build bridges after the terrible war years and to foster peace. The International Geophysical Year (IGY) 1957/58 was the perfect opportunity. Representing Germany, he participated in many of the activities. Together with Roy Piggott in the UK, he published the *Handbook for the Scaling of Ionograms*, a precious reference book that is a must at every ionosonde station worldwide. In 1966, Rawer was elected Vice Chair of Commission G of the International Union of Radio Science (URSI), and served as Vice Chair and then Chair until 1972. At the 2017 URSI GASS in Montreal, the first Karl Rawer Gold Medal was awarded, a medal in honor of the work and life of Prof. Karl Rawer.

A new chapter in Prof. Rawer's scientific career began with the first successful launch of the newly developed French "Veronique" rocket from Hammaguir (Algeria) in the Sahara Desert in 1954. With scientific payloads developed under his leadership, ionospheric conditions were observed in situ along the rocket's trajectory, showing for the first time the steep increase in electron density in the D-region. Many more launches followed, and with his expertise in instrument development and mission management, Rawer played a key role in the German involvement in the beginnings of space exploration. When the Committee on Space Research (COSPAR) was founded in 1958 for the promotion of scientific research at an international level, Rawer actively participated. In 1964, he became the Chair of the German National Committee of COSPAR. He vigorously exploited the opportunity that COSPAR offered to establish long-lasting relationships between scientists from West and East across the cold-war borders, but also with researchers in India and in hitherto neglected countries in the Far East and Africa. With a focus on his new area of interest, Rawer, in cooperation with the Fraunhofer Society, founded the Institute für Physikalische Weltraumforschung (IPW) in Freiburg, and served as its director from 1969 to

1979. Under Rawer's leadership, the IPW became a focal point of space exploration in Germany. It was responsible for Germany's second and third satellites, the AEROnomy Satellites AEROS-A and AEROS-B, which were launched in 1972 and 1974, respectively. The EUV spectrometer, impedance probe, and retarding potential analyzer for these missions were developed and tested in-house. Together with the US Atmosphere Explorer C, D, E satellites, the AEROS-A and -B satellites were some of the first to provide a global view of the ionosphere and upper atmosphere, and led to a much improved understanding of the processes that shape this region of geospace.

Data from the AEROS satellites, together with measurements from the worldwide network of ionosondes, were the foundation for the International Reference Ionosphere (IRI), a project that Prof. Rawer initiated under the auspices of COSPAR and URSI. He chaired the IRI Working Group from 1968 to 1976, and continued his active involvement into the mid-1990s. Under his guidance and leadership, the IRI project took off on a path that has led it to become the recognized international standard for the ionosphere that it is today. As participants of the bi-annual IRI Workshops, we will always remember Rawer's scientific rigor and endurance that kept us on our seats and engaged in spirited discussions until late in the day, and his joy and happiness during the social events with a good glass of wine and dancing late into the evening. The Final Discussion session at the end of the workshop was his trademark, with setting out of goals for improvements of the model and finding volunteers to accomplish these tasks.

Since 1955, Prof. Rawer had been affiliated with the Albert-Ludwigs-Universität of Freiburg, lecturing and advising graduate and doctoral students. Even when extremely busy with his many national and international science projects, he always had an open door and ear for his students, and mentored their careers long past their final exams. His research activities in ionospheric modeling continued after his retirement, with support from the German Research Foundation (DFG) and ESA's European Satellite Operations Center (ESOC). In 1993, he published the book *Wave Propagation in the Ionosphere*.

Rawer's greatest love and pride (and maybe one of the secrets of his longevity) was his large family, with three sons, four daughters, 19 grandchildren, and 31 great-grandchildren.

On the occasion of Prof. Rawer's 100th birthday, Bodo Reinisch delivered a laudation on Prof. Rawer's life and accomplishments at the German URSI National Committee meeting: *Adv. Radio Sci.*, **12**, 2014, pp. 221-223 (www.adv-radio-sci.net/12/221/2014/).

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In Memoriam: Thomas J. (Tom) Brazil

Thomas (Tom) Brazil died suddenly on Friday, April 13, 2018. He was Professor Emeritus of Electronic Engineering, University College Dublin (UCD), and the current President of the IEEE Microwave Theory and Techniques Society (IEEE-MTTS). He was a Member of the Royal Irish Academy (RIA) from 2004, serving as its Secretary (2009-2013) and also as Chair of a number of the Academy Committees, including the All-Ireland URSI Committee and the RIA's Communications and Radio Science Committee from 2000 to 2008. He represented Ireland on the URSI Council and as Ireland's Commission D (Electronics and Photonics) Member for very many years. His copious energy, drive and comprehensive insights also led him to contribute to, and be consulted on, science policy both in Ireland and at a European Union (EU) level. He served on a variety of national and international scientific and engineering committees and review bodies, especially in relation to formulating of the EU billion-euro research programs.

Tom retired as Professor of Electronic Engineering at UCD last September, where his reputation as teacher, mentor, and representative was second to none. He was elected by his fellow academics to the Governing Body of UCD for three five-year terms. He also served as a Member of the Senate of the National University of Ireland (NUI).

His humble demeanor belied the tremendous respect in which he was held by his students, the academic fraternity nationally and internationally, and in particular, the global microwave and millimeter-wave research and academic community, especially the IEEE MTT-S and the European Microwave Association (EUMA). He spent four years as an IEEE Worldwide Distinguished Lecturer in Microwave CAD 1999-2003. He was Chair of a number of major international IEEE conferences in microwave engineering, such as the European Microwave Conference in the UK in 2006, and the European Microwave (EUMW) Integrated Circuits Conference held in London in 2016. In fact, only the week before he died, he was working closely with his EuMW 2018 colleagues at a TPC meeting in Madrid (April 6-8, 2018), where he chaired the Subcommittee on Device Modeling topics. Both IEEE MTT-S and EUMA/EUMW will be marking his passing in special ways in the coming months.

Thomas Joseph Brazil was born on September 4, 1952, on the Mall in Birr Co. Offaly in Ireland. Birr is renowned as the home of the Earls of Rosse: in Birr Castle, a great Parsons engineering family, especially of steam-engine fame. However, more than steam engines were invented there. Of interest to URSI astronomers is that William Parsons, Third Earl of Rosse, built the "Leviathan of Parsonstown." This was a historic reflecting telescope of 72 in (1.8 m) aperture. It was the largest telescope in the world from 1845 until the construction of the 100 in (2.5 m) Hooker Telescope in 1917. More recently, as is also well known to URSI astronomers, in 2017, the new LOFAR radio-telescope station IE613, known in Ireland as the I-LOFAR telescope, was constructed and switched on. This westernmost station in the LOFAR network adds considerably to the LOFAR baseline. In 2018, it observed for the first time a billion-year-old red-dwarf flare star called CN Leo, almost 75 trillion kilometers away. Tom was a member of the founding committee of this I-LOFAR telescope, and thoroughly enjoyed his home-to-roots visits to Birr Castle for I-LOFAR planning meetings.

Tom was the eldest of four children of Elizabeth and Walter Brazil, who was then Area Manager in the Irish train and bus semi-state company. In 1960, the family relocated to Dublin. Tom was a stellar student in his local De La Salle Christian Brothers School. From his school days right through life, he was renowned for his wonderful wit and ability – when the serious talking was over and done with – to draw out the funny side of everything. His daughter Lia's favorite story of his childhood was of his constant interest in how things worked, often manifest in very practical ways – including taking the back off the precious family television, and disassembling his sister's dolls on Boxing Day.

After his father died in 1969, Tom worked closely with his mother to ensure his siblings, Marian, Bernice, and Walter, got the love and practical support they needed to flourish.

Tom topped his Electrical Engineering graduation class in UCD in 1973 – first place, first-class honors at the age of 20, – and was in the first cohort of engineering students to do a PhD in UCD. He did so under the direction of the then newly appointed first UCD Professor of Electronic

Engineering, Prof. John O. (Seán) Scanlan. His topic was circuit and solid-state modeling and analysis of the then novel Impact Avalanche Transit Time (IMPATT) devices. He subsidized his meager postgraduate scholarship each year by giving crash courses in numerical analysis to engineering students. This also gave him his first taste for teaching, which became a love of his life. Together with another postgrad student, they would cover the full year's course, including past-examination questions and solutions, in a continuous 12-hour lecturing stint over a day and one-half to a class of 180 students a few weeks before the exam. After graduating in 1977, he took an engineering position as a Research Scientist in Plessey Research (Caswell) in Northampton, UK, working on microwave subsystem development.

In 1979, he moved into his first academic position at the University of Birmingham. This marked the formal beginning of Tom's emergence as a great educator. He had a very strong instinct to share his knowledge and wisdom, and really was a brilliant teacher. His stint in the University of Birmingham was short, as he returned to Dublin to take up a lectureship in electronic engineering in his alma mater, UCD's Electrical Engineering Department, in January 1980. There, with the close collaboration of Prof. Seán Scanlan, he founded the RF and Microwave Research Group. That group thrives to this day, and has a strong international reputation for fundamental and applied research into radio transmitters and receivers. In particular, Tom and his group made notable research contributions in the field of microwave and millimeter-wave nonlinear high-power amplifiers, and in nonlinearity impairment linearization-mitigating concepts and technology. The fruits of his work today already impact across all radio communications infrastructures, from mobile phones and phone masts to Galileo satellite transmitters. The group also produced some dozens of research graduates, PhDs, and Masters in Engineering, and of course a deep track record of research-journal publications, especially in the *IEEE Transactions on Microwave Theory and Techniques*. Those of his graduates who stayed in Ireland helped seed a very advanced technology industry here, the envy of many highly developed economies.

While Tom was very modest about his achievements, on his passing, the renowned microwave engineering academic, industrialist, and entrepreneur, Prof. David Rhodes CBE, FRS, FR Eng, said "[he was] one of the brightest and most intelligent microwave engineers of our time."

Tom succeeded Seán Scanlan in the Chair of Electronic Engineering in 2003, and became Head of the School of Electrical and Electronic Engineering for 14 years. He was proactive in charting a steady course for the school through the choppy waters of rapid and radical change, all the while unwavering in his maintenance of sound academic principles. Not for him were the short-lived academic fashions of the day, or allowing these to ever dilute solid comprehensive engineering-education foundations. Hundreds of UCD graduates have reaped the rewards of this excellent engineering formation in their professional lives, and gone on to contribute to Ireland's immense economic and industrial growth, development, and success.

A sense of care and of duty pervaded his life, from his PhD students to the secondary-school students participating in the annual Irish National Young Scientists Competition, whose projects he loved judging. Nurturing and mentoring intelligence and curiosity, especially in young people, was the cornerstone of his professional life. He believed that good academics help others succeed and go beyond themselves, and he surely did that. He had this generous approach through his life, being available to others and unstintingly giving of himself. In fact, he always worked with his office door open, so that anybody, staff or student, could consult him at any time. Because of the open door, all became aware that when on the phone he would lean back in his chair and put his feet up on his desk. He always enjoyed the wry comments from his colleagues over coffee breaks to which this gave rise. His spirit of service came naturally to him. Even as an undergraduate engineering student himself, he was always ready to help his fellow students. In particular, he was the "go-to" man for anyone who missed a lecture. He had an incredible writing hand, both in neatness and speed, and had the reputation for taking the tidiest, most structured, and most comprehensive lecture notes of all. Even his lecturers sought him out to get the note versions of their own lectures, with the intention of converting their lectures into course textbooks.

Among his peer international community of engineers, he was renowned not only as a leading thinker and a driver of research, but also as someone who readily sought to treat you as a friend. He was that all-rounder combination of teacher and mentor, scientific researcher, contributor to society nationally and internationally, and friend. Most of all, he was a family man: a kind, gentle, and wonderful person, loving husband to Eileen and devoted father to Lia, David, and Tom Junior. He was a gentleman and a scholar. May he rest in peace. As prayed in Ireland's native language: *"ar dheis láimh Dé go raibh a h-anam uasail"* (May his gentle soul be at God's right hand).

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Et Cetera

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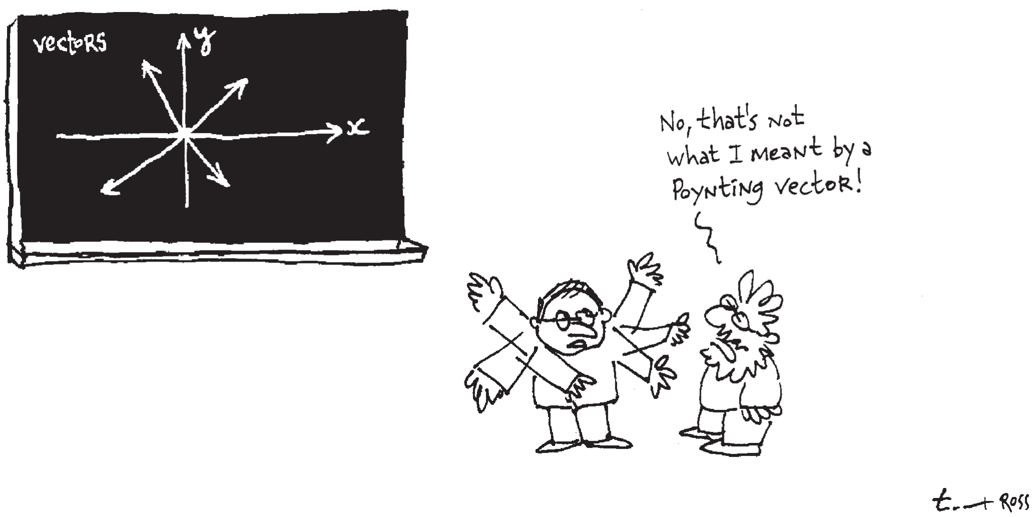

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Foreword

On September 13, 2017, Sir John Meurig Thomas held a lecture in Florence [1]. Thomas, now retired, held the Chair of Chemistry at the Royal Institution, which was created for Michael Faraday. Thomas is renowned for his excellent book Michael Faraday and the Royal Institution: The Genius of Man and Place (Institute of Physics Publishing, 1991). The book was translated in 2006 in Italian as Michael Faraday La storia romantica di un genio (Firenze University Press) by Luigi Dei, now Rector of the University of Florence, and has been released to the public domain on the occasion of this lecture of September 13 (Figure 1).

In the book, it is recalled that from February 21, 1814, to April 7, 1814, Sir Humphry Davy [Penzance, UK, December 17, 1778 - Geneva, CH, May 29, 1829] on the occasion of his continental tour (Figure 2), took Faraday to Florence, where they visited the local museum and saw Galileo's first telescope (Figure 2). In Florence, Davy conducted experiments on the combustion of diamond in oxygen, using the Great Duke great burning glass. As

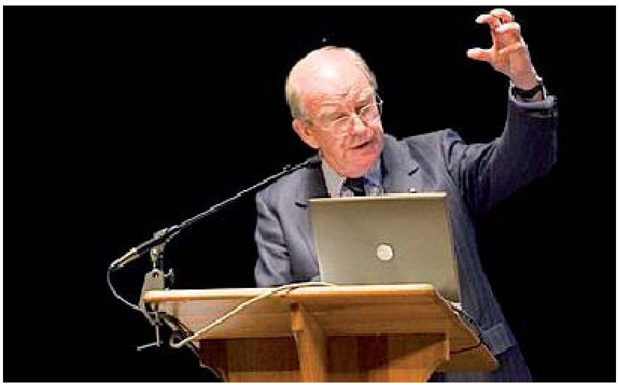

Figure 1. Sir John Meurig Thomas, giving his lecture in Florence in September 2017.

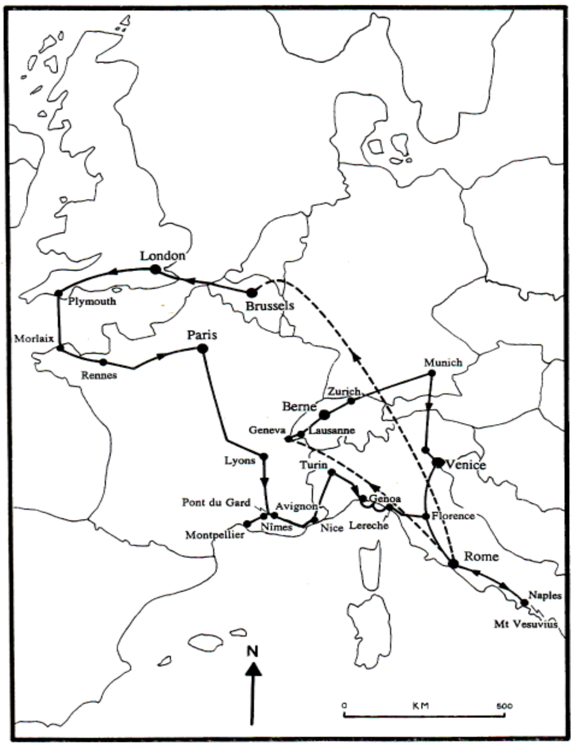

Figure 2a. The continental tour of Davy and Faraday (1813-1815).

Faraday remembered, "After several attempts Sir H. Davy observed the diamond to burn visibly, and when removed from the focus it was found to be in a state of active and rapid combustion. The diamond glowed brilliantly with a scarlet light inclining to purple, and when placed in the dark continued to burn for about four minutes. As Faraday observed, it was "a phenomenon never observed before." In a letter to his mother, he added "Florence, too, was not destitute of its attractions for me, and in the Accademy del Cimento and the Museum attached to it is contained an inexhaustible fund of entertainment and improvement."

Figure 2b. The monumental "Galileo Tribune" (decorated in 1841) at the "Specola" Museum of Florence, where the Scientific Assemblies in Florence used to take place.

I hence sought for further investigations on later connections between Michael Faraday and Italy, and Florence, in particular. In the following, two short papers are published. The first paper is on a minor figure in Italian physics who claimed priority over Faraday's discovery of magnetic induction: Francesco Zantedeschi. The second paper is on the most preeminent physicist in Italy at the time, and the greatest instrumentation builder, Leopoldo Nobili. He was closely connected with Faraday, again concerning magnetic induction, since it was thanks to his "astatic galvanometer" that Faraday managed to prove the phenomenon of magnetic induction.

Reference

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Faraday-Neuman-Lenz Law of Induction or Zantedeschi-Neuman-Lenz Law of Induction?

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A recent, extremely nice and entertaining book [1] (Figure 1a) cited Francesco Zantedeschi [Dolcè, Verona, Italy, August 18, 1797 – Padua, Italy, March 29, 1873] (Figure 1b) as a precursor of Michael Faraday [Southwark, London, UK, September 22 1791 – Hampton Court, Middlesex, UK, August 25, 1867] in discovering the law of magnetic induction in 1829, two years before Faraday himself. This is information that was reported elsewhere [2, 3], as well as on the Internet, most notably on Wikipedia [4]. However, other texts were more cautious [5, 6] or completely avoided citing Zantedeschi [7].

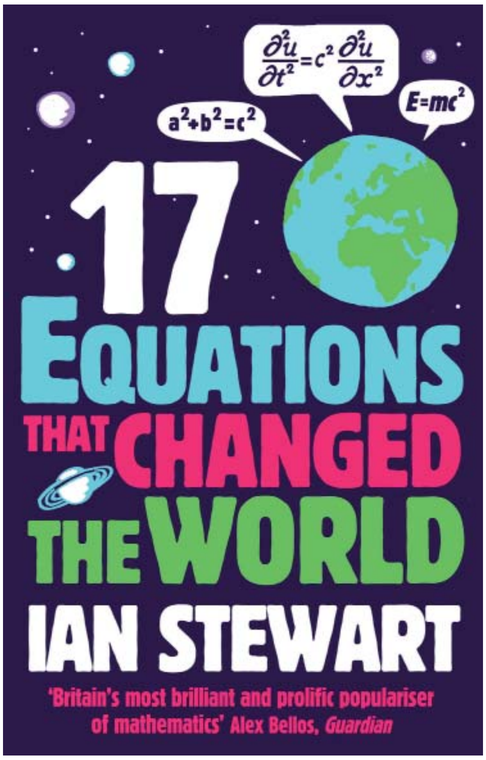

Figure 1a. The cover of the book by Ian Stewart.

Figure 1b. Fancesco Zantedeschi.

It interesting to understand what Zantedeschi really did, and why he is credited. Information was sought in the original papers, which are mostly in Italian and hence of difficult access to our international community.

In 1820, Hans Christian Ørsted [Rudkøbing, Benmark, August 14, 1777 – Copenhagen, Denmark, March 9, 1851] discovered that an electric current could deviate the magnetic needle of a compass. This was a landmark in the study of electricity and magnetism, since it proved a connection between the two. Once the magnetic effects of a current were established, the electric effects of a magnet were highly expected.

In 1929, an obscure Italian Abbot, Francesco Zantedeschi, published a short communication [8] on the chemical effects of magnets. As a *post scriptum*, he added the few lines reported in Figure 2, where he claimed to have obtained such electric effects. He then published a French version of the same communication [9].

PS. Aggiungo in forma di appendice all' esperienza 1.ª e 2.ª della 1.ª parte un altro fatto da me osservato più

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volte in questo mese, il quale non dovrà almeno riuscire discaro, perchè tende quale anello ad unire i diversi fatti elettro-magnetici colla loro sorgente. Ho preso una calamita fatta a ferro di cavallo del peso circa di una libbra francese, che potea sostenere un peso di circa 4 a 5 libbre, ed attorno a ciascun polo ho avvolto strettamente un filo sottilissimo di rame in modo che, collocata la calamita ad una distanza di 15 a 16 piedi parigini, potea sperimentare sulle estremità separate di detti fili. Ora preso un moltiplicatore a due calamite, ho ai capi del filo del medesimo (che è di rame circondato di seta) attaccate due piastrine di rame ben lucide, colle quali, mediante due verghe di legno per non alterare la temperatura, congiunsi i fili che abbiam detto essere in comunicazione coi poli della calamita, ho veduto che l'ago magnetico sviasi dalla naturale sua posizione declinando verso l' oriente il polo al disopra del quale entra l'azione magnetica del polo nord, e verso l' occidente, se questa entra al disotto di esso, non altrimenti di quello che avviene coll' elettrico ordinario. La declinazione era da 8° a 10°. Mi pare che questo fenomeno non si possa ascrivere alla facoltà elettromotrice, perchè il rame trovasi fra due forze eguali e contrarie. E dato anche, come ho esperimentato nei liquidi, che le correnti elettriche, qualunque sia la loro direzione, non sviinsi, come la luce e il calorico raggiante, non dovrebbe il moltiplicatore dare alcun segno, come è chiaro. Pare dunque che tale effetto debba ascriversi al magnetico, e però che il polo nord equivalga al polo zinco d' un apparato voltiano. Io spero che altri esperimentando con moltiplicatori più delicati, come col siderescopio di Lebaillif, potrà ottenere effetti maggiori che udirò quando che sia con piacere.

Pavia, 27 marzo 1829.

Figure 2a. The post scriptum in [8] where Zantedeschi wrote about his experiment on the electric effects of a magnet.

P.S. I add as an appendix to experiences n. 1 and 2 of part 1 another phenomenon I observed several times this month, which should not sound unlikely, because a, as a ring, it unite various electro-magnetic facts with its source. I took a horseshoe magnet, of about 1 French pound of weight, strong so as to lift a weight of about 4 or 5 pounds, and around each pole I tightly winded a very thin copper wire so that, with the magnet 15 to 16 French foot away, I could experiment with the open ended of these wires. Now, taken a two needle galvanometer, I connected to the its wires (which are in copper wrapped in silk) two shiny copper plates, which I moved with wooden sticks, not to alter their temperature, and connected them with the wires which we said winded around the magnet. I saw the galvanometer needle move from its rest position, going east the point of the needle above which the action of the north pole of the magnet enters, going west if the action enters from below, analogously as for ordinary electricity. The deviation was from 8° to 10°. I believe that this phenomenon cannot be due to electromotive force, since the copper is placed between two opposite, equal, forces. Furthermore, as I noticed in liquids, electrical currents, notwithstanding their direction, do not deviate, as it happens for light and caloric rays, the galvanometer should give no sign, as it is clear. It hence looks like that such a phenomenon is to be ascribed to magnetism, such that north pole is equivalent to the zinc pole of a voltaic pile. I hope that further experiments, with more sensible galvanometers, as Lebaillif galvanometer, could show larger effects, of which I will hear news with delight.

Pavia, March 27, 1829

Figure 2b. A translation by the author of the post scriptum in Figure 2a.

Zantedeschi wound coils of copper wire around the poles of a magnet. He then connected these wires with a galvanometer (a two-needle multiplier, as Nobili's astatic galvanometer with a multiplication coil was called at that time), and affirmed seeing a deviation in the magnetic needle (Figure 3). From both texts [8, 9] we can understand that the magnet had tight windings of copper, and that the wires forming the coils extended far from the magnet, so that their free ends were 15 to 16 French feet (about 5 m) from the magnet itself. It was quite unlikely that such an experiment would have led to the described effect. We know now that the electric effect is bound to a *variation* of the magnetic-induction flux in a coil. Zantedeschi's tightly wound coils were not a possible source of such a variation, and the transient nature of the phenomenon was not described at all by Zantedeschi. It is anyway important to stress that something was longed for and awaited. We therefore can think that Zantedeschi might have genuinely believed to have seen an effect. Indeed, Michael Faraday himself had already made some early experiments [10], which failed due to the low sensitivity of his galvanometer.

In 1931, Faraday set up a more-refined experiment. He understood that the magnetic field generated by a coil of currents does indeed induce a current in a close but electrically disconnected coil, *but only at the transients*, that is, when the current in the primary coil passes from zero to a steady-state value and – still extremely important – an opposite current is generated when the primary coil excitation ends. The details of the communication of this discovery are interesting. Other researchers, among whom was Leopoldo Nobili, duplicated these experiments very early, and much excitation ran in the Italian journals on priority. In particular, Giuseppe Gazzeri [Florence, Italy, November 9, 1771 – Florence, Italy, June 22, 1847], in the pages of the *Antologia Fiorentina* [11], explicitly cited Zantedeschi's 1829 work [8] in a note, saying:

> Concerning preliminary studies, we warn the reader that prof. Zantedeschi published in March 1829 (*Biblioteca Italiana* vol. 53, p. 393) a result by him obtained winding a coil around the poles of a magnet, connecting the wires to a galvanometer. His result was the deviation of the needle by 8°-10°. Apparently, it seems the discovery by Faraday, but it cannot be, because in his set-up the currents discovered by Faraday cannot exist [11, p. 174 footnote].

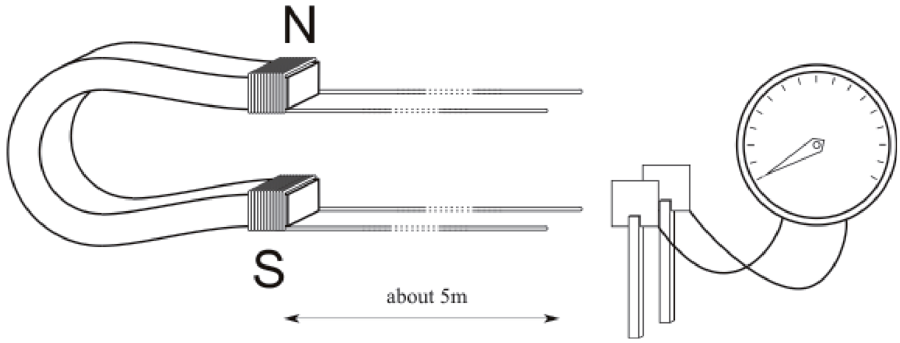

Figure 3. Zantedeschi's experiment, according to his description in [8].

Of course, this enraged Zantedeschi, who bitterly replied on the same *Antologia* pages [11], stating that,

> What is the difference between the Englishman [Faraday] experiments and mine? I coiled the wire on the magnet, while he moved the coils toward the magnet. On anything else and on the fundamental fact [emphasis by Zantedeschi] I see no differences [11, p. 232].

He then claims that the *post scriptum* was too short to explain every detail of his experiment, that he placed and removed an iron bar to the magnet as Nobili did [13, p. 233], and that the effect was transient. Zantedeschi continued claiming for a long time [14] with even more strong words against Faraday:

> When it is said that I connected the coils to a galvanometer "in the usual way" it is unjust. Who, before me, imagined to connect a galvanometer to a coil? Faraday two years later did, and did not mention me. I formally invite Mr. Faraday to break his silence and disrobe himself of the vest of an usurper which he wore up to now on this topic. I wrote letters to him, which never got reply [14, p. 11].

Indeed, Faraday in the end, cites Zantedeschi positively! Zantedeschi, among the other things, experimented with flames, observing a deviation in the flame when a magnet was present. He explained the phenomenon by a diamagnetic effect of hot gasses, and published his observations in an Italian gazette [15]. Faraday had previously affirmed that gasses had no magnetic properties at all. Once Faraday knew about Zantedeschi's results, he experimented again and found that indeed they had magnetic properties [16]. Faraday fully acknowledged Zantedeschi, and added a partial English translation of [15] to his paper.

Yet it is somewhat ironical that, again, Zantedeschi was not the discoverer of the phenomenon he was studying. Another Italian priest, Michele Alberto Bancalari [Chiavari, Genoa, February 20, 1805 – Genoa, August 10, 1864] made the first observation on flame deviations and reported it to the Italian Scientific Assembly in Venice in 1847. No written record by Bancalari stands, only verbal accounts of the sessions [17], which Zantedeschi attended. Yet both Zantedeschi and Faraday correctly cite him.

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Leopoldo Nobili: His Galvanometer and His Connections to the Faraday-Neuman-Lenz Law of Induction

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The discovery of the phenomenon of magnetic induction, or the Faraday-Neumann-Lenz law of induction, was something much expected, after Hans Christian Ørsted [Rudkøbing, Benmark, August 14, 1777 – Copenaghen, Denmark, March 9, 1851] discovered that electric currents had an effect on a magnetic needle. Many researchers, among whom the most notable was Michael Faraday [Southwark, London, UK, September 22, 1791 – Hampton Court, Middlesex, UK, August 25, 1867], actively sought such a phenomenon, yet the instrumentation then available was exceedingly crude [1]. For an electric phenomenon to be apparent, a spark should be observed for voltage, or a current strong enough for visible effects. Some more-sensitive device able to detect feeble currents was necessary. Indeed, Ørsted's discovery made possible the development of such an instrument: the galvanometer. In that same 1820, Johann Schweigger [Erlangen, Germany, April 8, 1779 – Halle, Germany, September 6, 1857] build a first galvanometer by winding a coil of wire around a magnetic needle free to rotate. This was later enhanced by "multiplying" the effect of the current by using several coils of wire.

Figure 1a. Leopoldo Nobili.

Figure 1b. Vincenzo Antinori.

However, it was only with the refinement of such instruments by Leopoldo Nobili [Trassilico, Lucca, Italy, July 5, 1784 – Florence, Italy, August 22, 1835] (Figure 1) that the reliable detection of feeble currents was feasible. Nobili, a preeminent Italian physicist and top scientific instrument maker in Europe, used two oppositely oriented needles on the same suspending wire, and a coil in the shape of a figure eight (1825). The double needle was insensitive to the strong, static, Earth's magnetic field, hence greatly increasing the sensitivity of the instrument [2] (Figure 2). He named his device the *astatic* galvanometer.

In the same period, in 1824 exactly, François Arago [Estagel, Perpignan, France, February 26, 1786 – Paris, France, October 2, 1853] found that a rotating magnetic needle was slowed down by the presence of nonmagnetic metals close to it [3, 4]. He also observed that a rotating copper disk caused a suspended magnet over and close to it to rotate if free, or else was slowed down by the presence of the magnet if this latter was fixed. However, this phenomenon remained inexplicable, and was not immediately connected to electricity (Figure 3, from [5], p. 38).

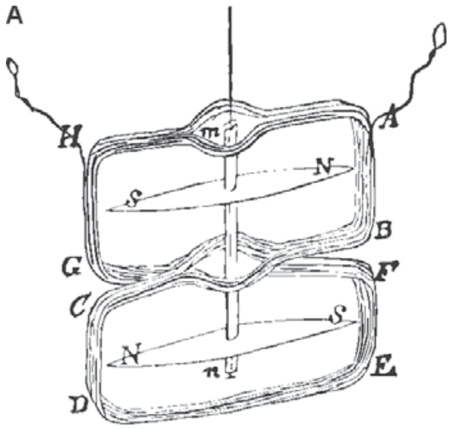


Figure 2a. The layout of the astatic galvanometer, from Nobili's original paper [2].

Eventually, in 1831, Faraday set up an experiment in which Nobili's astatic galvanometer showed a transient current in a coil when another coil connected to a generator, or a permanent magnet, was moved close to or away from it.

The point now is that Faraday devised his experiment in 1831, and read the results on November 24, 1831, at the Royal Society. However, he was quite slow in publishing a paper: a short note appeared only in April 1832 [6]. On the other hand, the news of his highly expected, successful induction experiment spawned at light speed through Europe. The experiment was described by Faraday himself to Jean Nicolas Pierre Hachette [Mézières, Ardennes, France, May 6, 1769 – Paris, France, January 16, 1834], who read a communication in French to the *Académie des Sciences* in Paris on December 26, 1831. Such a French communication, published in the journal *Le Temps* [7], reached Florence, where it was read by Vincenzo Antinori [Florence, Italy, 1792 – Florence, Italy, 1865] (Figure 1) and Leopoldo Nobili, who immediately duplicated the experiment and produced a paper.

Figure 2b. One of the original astatic galvanometers, built by L. Nobili in 1826 and preserved at the "Museo Galileo," the museum of the History of Science, in Florence, which owns several of Nobili's devices.

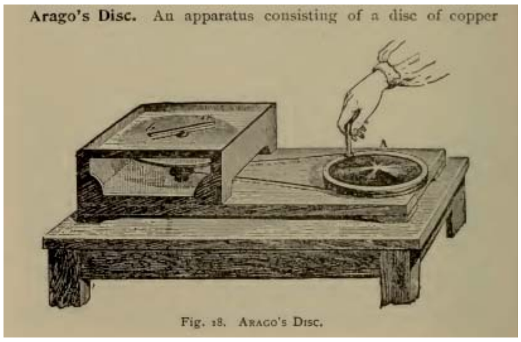
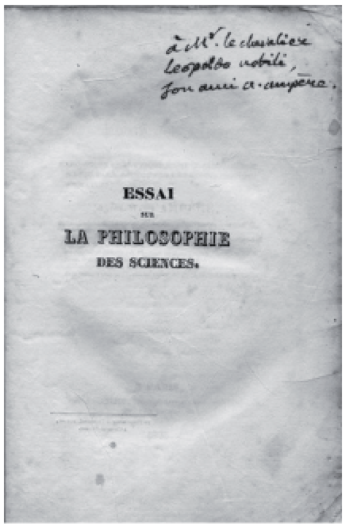


Figure 3a. Arago's disk, from [5].

In Antinori and Nobili's paper, the authors fully acknowledged Faraday's priority. They regretted that Faraday had not yet published his results, and thanked Hachette for having spread the news [8]. Such a paper was dated January 3, 1832, but appeared in an issue of a Journal dated November 1831 (vol. XLIV). Nobili and Antinori quickly wrote a second paper on the subject [9], dedicated to the currents induced on a rotating disk in a magnetic field, hence solving the problem of the rotation observed by Arago [3, 4].

The problem is that Antinori and Nobili's paper was the only item available in the open literature for a few months. As was common practice in that period, it was submitted in French, German, and even English, to the *Philosophical Magazine*, in an annotated translation performed by Faraday himself [10]. It was thus the only well-organized paper on the subject up to Faraday's paper [11], where Faraday himself summarized these events. The priority issue, never claimed by Antinori and Nobili, was nicely settled.



à Mr. le chevalier
Leopoldo nobili,
son ami A. Ampère.

ESSAI

SUR

LA PHILOSOPHIE

DES SCIENCES.

Figure 3b. The first page of the book *Essai sur la Philosophie des Sciences* by A. Ampére (1834), with a friendly dedication of the author to Leopoldo Nobili [private collection, courtesy of Fausto Barbagli, Natural History Museum, University of Florence].

Yet in the *Literary Gazette* there appeared a series of columns publicizing Faraday's discovery [12], and later announcing [13] that Faraday intended to explain the Arago rotation phenomenon on the basis of this new discovery. This in turn upset Giuseppe Gazzeri [Florence, Italy, November 9, 1771 – Florence, Italy, June 22, 1847], a renowned chemist, who attacked the *Literary Gazette* and its journalists (not Faraday!) from the pages of the *Antologia Fiorentina* [14]. Faraday replied in a letter to the *Literary Gazette*, and a discussion on the Arago phenomenon arose. The notes by Faraday in [10] mainly point to his previous work, and say that "I am criticizing Sig. Nobili and Antinori for not understanding my views. It was impossible that I could put forth in a brief letter, matter which, though I have condensed it as much as possible, still occupies seventy *quarto* pages of the Philosophical Transaction." It was true: Nobili and Antinori only had available [7], not [11]. As a result, a misunderstanding was highly possible.

Nobili and Antinori replied, stating that their interpretation was analogous to that made by Ampére of Arago's discoveries [15]. Ampére was the most renowned physicist at the time, and Nobili was Italy's greatest physicist. Faraday bitterly replied, and a long discussion started. This was quite unusual for Faraday, usually extremely polite with other physicists. A possible explanation is that Faraday was worried that Nobili and Ampére, two of the greatest physicists and friends (Figure 3), were interpreting the induction phenomenon he discovered in terms of actions at a distance, whereas Faraday was figuring those lines of force that would have later opened the road to Maxwell. Faraday could have been afraid that a mathematical theory of induction in terms of actions at a distance could be produced, whereas he, self-taught experimenter, could never come up with one. He might then have thought of indirectly attacking Ampére through Nobili [17].

Surprisingly enough, Nobili, close to the end of his life, denied any claim of explaining induction via action at a distance, and espoused Faraday's concepts [18]. However, he claimed that he had developed a similar yet less refined concept earlier, too [19].

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A Coconut Deception

Randy L. Haupt and Amy J. Shockley

Sue (wife/mother of the authors of this article) was strolling along a beach in Mexico when she got hit in the head by a coconut. Did you know that coconuts actually severely injure people, and may even cause death when they hit someone on the head? If you want to find the odds of dying from a falling coconut and search the Web for "death by coconut," you will find that a brewery just released a new beer called Death by Coconut. We suggest instead searching for "how many people die from coconuts." The predominant number that appears is 150 deaths/year due to a coconut falling on someone's head. Wow!

Where did that number come from? Club Direct (British travel insurance) managing director Brent Escott said that "Coconuts kill around 150 people worldwide each year, which makes them about ten times more dangerous than sharks" [1]. He got his number by extrapolating from the data published in a medical study [2]. One Web page estimates that "6 out of every one million people get bonked by a coconut" [3]. They arrived at that number by first estimating the number of coconuts on trees from the number of coconuts harvested in a year, then used the number of people on Earth and the land area where coconuts live. Assuming that 1% of the people die from getting hit by a coconut, then they estimated that 400 people die each year. These numbers should make you think twice before napping under a palm tree.

So, what happened to poor Sue? If you knew the following facts, you would be concerned: "Mature coconut palms may have a height of 24 up to 35 meters and an unhusked coconut may weigh 1 to 4 kg, meaning that blows to the head of a force exceeding 1 metric ton are possible" [2]. Fortunately, the tree that Sue meandered under was a rather small palm. In what is probably the most fortunate part of this situation, her head bumped into the coconut: the coconut never fell from the tree. While startling to head-butt a coconut when walking on the beach, not paying attention, physical injury was thus minimal.

Our coconut story was very misleading. We did not lie, but we did purposely deceive you. The coconut was not on a tall tree, and it did not fall. The statistics we cited from the references were also bogus. In reality, it is difficult to find an instance of one person ever dying from a coconut that fell from a tree. The deaths per year that are frequently cited are based on faulty assumptions, and not real data.

To deceive is to instill a belief in someone that is not actually true. Deception is quite prevalent in today's society, whether driven by true malicious intent, as a mistake, or as a practical joke. The ethical line in deception is intent. If the deceiver is acting with the intent to gain an unjust benefit or cause harm to another, they are acting unethically. Identifying ill-intended deception can be difficult, and requires building upon the skills required to be less gullible. In addition to those skills, you should use your inner compass to gauge if a scenario "feels right." Include information that you know about the individual, and their morals and past actions. Asking probing questions can help identify a deception as well as if it is driven by poor intentions (often times, deceit is driven by peoples' inability to say "I don't know"). Don't hesitate to consult others in order to get varying opinions and views of the situation.

Just because a "fact" is quoted in many different places and by reputable people doesn't make it true. Vladimir Lenin said, "A lie told often enough becomes the truth."

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Maxwell Newsletters Available

The spring and summer 2018 *James Clerk Maxwell Newsletters* of the Clerk Maxwell Foundation are available at www.clerkmaxwellfoundation.org. Among other items, the spring issue contains an article on the discovery of gravitational waves. The summer issue surveys the use of light as a replacement for radio frequencies in Wi-Fi and microcellular communication situations.

Second Announcement

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Asia Pacific Radio Science Conference (AP-RASC) is a triennial flagship event of the International Union of Radio Science (URSI). It is an international conference in the field of radio science, where eminent researchers and experts from all over the world gather to share their knowledge and experience, and also to encourage scientific exchange and fellowship amongst industry colleagues and professionals globally. It is our pleasure to invite participation to AP-RASC 2019 not just from Asia but also from Europe, South and North America, Oceania and Africa. The scientific program will provide an opportunity for participants to exchange new ideas and information on many important issues related to recent advances in radio science. The scope of radio science broadly covered in the AP-RASC 2019 fall under the ten commission of URSI as follows.

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- Commission B: Fields and Waves
- Commission C: Radio-communication Systems and Signal Processing
- Commission D: Electronics and Photonics
- Commission E: Electromagnetic Noise and Interference
- Commission F: Wave Propagation and Remote Sensing
- Commission G: Ionospheric Radio and Propagation
- Commission H: Waves in Plasmas
- Commission J: Radio Astronomy
- Commission K: Electromagnetics in Biology and Medicine

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The following two programs are planned for young scientists:

- Student Paper Competition (SPC)
- Young Scientist Award (YSA)

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- Paper Submission Open
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- Paper Submission deadline
  01 October, 2018
- Acceptance Notification
  01 December, 2018
- Online Registration Open
  01 September, 2018
- Early Bird Registration End
  05 January, 2019

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SOLBOX-09, SOLBOX-10

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1. Introduction

Periodicity is a major property of many structures in electromagnetic applications, such as antenna arrays [1], frequency-selective surfaces, metamaterials [2], and photonic crystals [3]. Array elements (unit cells, antennas, etc.) are periodically arranged in order to enhance the response of a single element or to gain new capabilities due to interactions between them. Naturally, electromagnetic simulations of periodic structures have a long history in the literature. This is still an active area as the structures become more complicated and require more sophisticated (accurate, efficient, stable) tools for precise analysis. When the structure is infinitely large, and particularly when the periodicity is sub-wavelength, there are excellent tools based on Fourier series and finite-difference models, e.g., see [4] and [5]. Similarly, when the elements can be represented point-wise with or without interactions, there are analytical (e.g., using array factor) and semi-analytical (e.g., using lumped elements and/or transmission-line models) techniques. On the other hand, challenges arise when each element is a complex electromagnetic problem itself while it needs to interact with other complex elements. In addition, if the finiteness is an important geometric parameter of a periodic structure, its full-wave models may lead to large-scale problems that can be difficult to solve numerically. In such cases, it is tempting to substitute high-order mathematical models for the elements [6], specifically by isolating their inner dynamics to derive the required equivalence [7] within the decomposed models. A well-known method in this path is called the Equivalence-Principle Algorithm (EPA), which has been used and improved (e.g., hybridized with or used in other methods) by numerous researchers [7-11].

In this issue, two different electromagnetic problems involving periodic arrangements are presented (SOLBOX-09 and SOLBOX-10). In SOLBOX-09, a microwave metamaterial involving 3 × 9 × 9 split-ring resonators (SRRs) is considered. Resonances of SRRs make the problem challenging at certain frequencies. A larger structure is considered in SOLBOX-10, where an array of 20 × 20 radio-frequency antennas needs to be analyzed. The antenna design – namely, a cage-dipole antenna – involves thin lines and very small gaps. For both problems, solutions using an EPA implementation are also presented and compared with brute-force solutions via the Method of Moments (MoM) and the Multilevel Fast Multipole Algorithm (MLFMA).

The contributors are confident that different numerical solutions and analysis methods, probably leading to faster and/or more accurate solutions, can be found. We are looking for alternative solutions to these problems (SOLBOX-09 and SOLBOX-10), as well as to previous problems

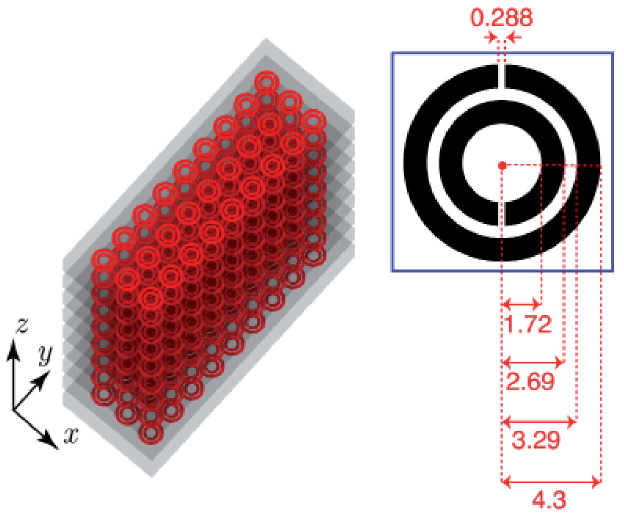

Figure 1. An array of 3 × 9 × 9 SRRs that are arranged periodically. The SRR geometry is the same as that in [12], which was used to construct a composite metamaterial structure. Equivalence boxes that were used in the NEPA-MLFMA solutions are also shown.

(SOLBOX-01 to SOLBOX-08) that can be found in the earlier issues of this column. Please consider submitting your contributions to Ozgur Ergul (ozergul@metu.edu.tr).

2. Problems

2.1 Problem SOLBOX-09

(by Ali Farshkaran and Ö. Ergül)

Figure 1 depicts a 3 × 9 × 9 array of SRRs that are arranged periodically in free space. The periodicities in the *x*, *y*, and *z* directions are 10.508 mm, 10.508 mm, and 18 mm, respectively. The dimensions for a single SRR are shown in detail. This geometry was previously used in [12] to construct a composite metamaterial. The surfaces are assumed to have zero thickness, while they are modeled as perfect electric conductors. The array is excited by a *y*-directed Hertzian dipole located at $x = 48$ mm, if the center of the array is at the origin. It is desired to find the response of the structure (e.g., far-zone scattering, near-zone field distributions, induced current density, etc.) in a range of frequencies from 4.5 GHz to 6.5 GHz. The SRRs are expected to resonate at around 5.25 GHz to 5.5 GHz, at which frequencies the numerical solutions are expected to be challenging.

2.2 Problem SOLBOX-10

(by Ali Farshkaran and Ö. Ergül)

Figure 2 depicts an array of 20 × 20 cage-dipole antennas located in free space. The antenna geometry with 3.2 cm × 6.8 cm size is the same as that in [13], which was optimized and designed for radio-frequency-identification (RFID) applications. The surfaces are assumed to be perfectly conducting with zero thickness. The antennas are arranged periodically on the *x*-*y* plane, with 8 cm and 13 cm center-to-center distances in the *x* and *y* directions, respectively. The overall size of the structure is hence 1.552 m × 2.538 m. The array is illuminated by a plane wave propagating in the *z* direction with the electric field polarized in the *x* direction. The frequency is set to 2.0 GHz, making the array larger than $10\lambda \times 17\lambda$, where λ is the operating wavelength. It is desired to find the scattered fields in the far zone.

3. Solution to Problem SOLBOX-09

3.1 Solution Summary

Solver type (e.g., noncommercial, commercial): Noncommercial research-based code developed at CEMMETU, Ankara, Turkey

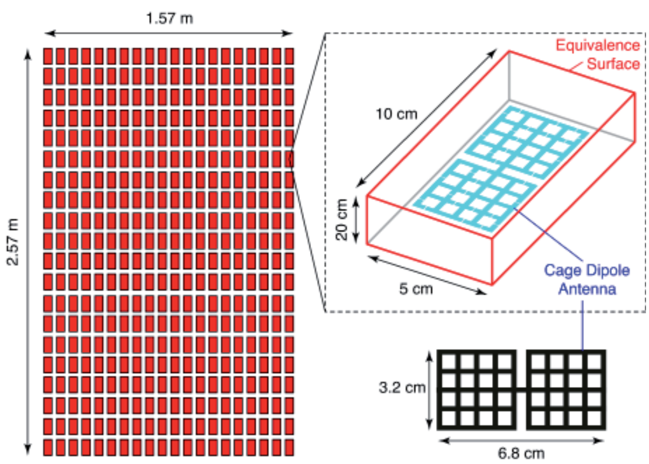

Figure 2. An array of 20 × 20 cage-dipole antennas. The antenna geometry, which was designed for RFID applications, was taken from [13]. Equivalence boxes that were used in the NEPA-MLFMA solution are also shown.

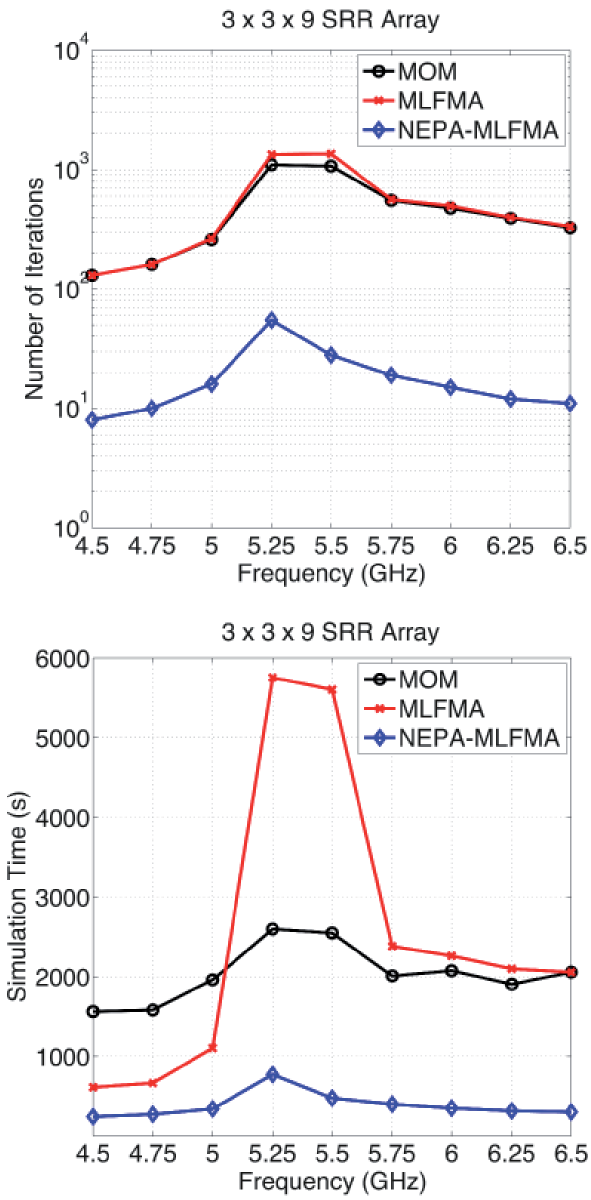

Figure 3. Three different solutions of SOLBOX-09 by using NEPA-MLFMA, in addition to conventional implementations of MoM and MLFMA. The number of GMRES iterations and the solution time (in seconds) are plotted with respect to frequency.

Solution core algorithm or method: Frequency-domain Equivalence-Principle Algorithm (EPA) accelerated with the Multilevel Fast Multipole Algorithm (MLFMA)

Programming language or environment (if applicable): *MATLAB + MEX*

Computer properties and used resources: 2.5 GHz Intel Xeon E5-2680v3 processors (using single core)

Total time required to produce the results shown (categories: < 1 sec, < 10 sec, < 1 min, < 10 min, < 1 hour, < 10 hours, < 1 day, < 10 days, > 10 days):
< 1 hour for 9 frequencies

3.2 Short Description of the Numerical Solutions

The problem SOLBOX-09, involving a 3 × 9 × 9 array of SRRs, was solved by using a *MATLAB* implementation that was based on the hybridization of EPA [8] and MLFMA [14] in the frequency domain. Solutions were performed at nine discrete frequencies, i.e., with 250 MHz intervals from 4.5 GHz to 6.5 GHz. The EPA implementation used the formulation in [8] with N-type projections of fields via the Gram matrices, while the surface-to-surface translations were accelerated via the MLFMA. The overall implementation was hence called NEPA-MLFMA. The electric-field integral equation

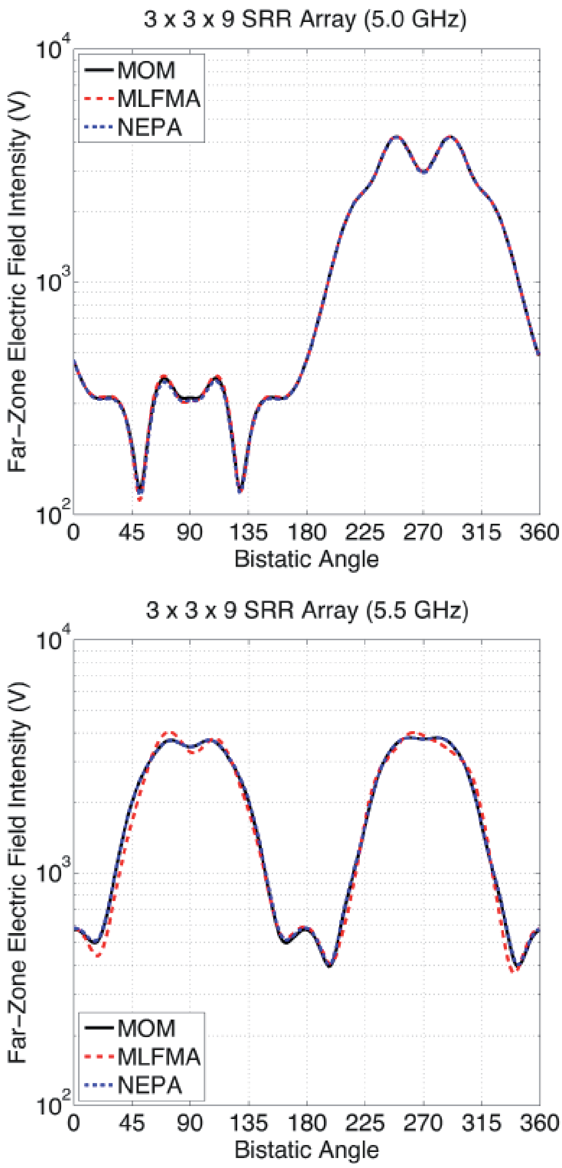

Figure 4. Three different solutions of SOLBOX-09 by using NEPA-MLFMA, in addition to conventional implementations of MoM and MLFMA. The far-zone electric field intensity is plotted with respect to the bistatic angle on the *z*-*x* plane at 5.0 GHz and 5.5 GHz.

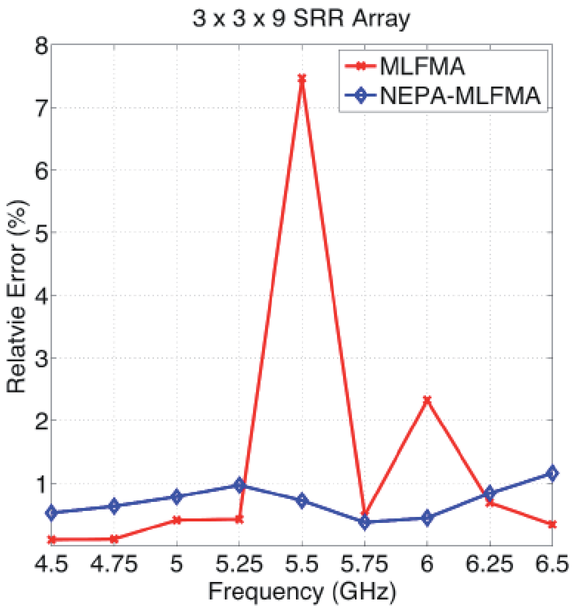

Figure 5. The relative error in two different solutions of SOLBOX-09 using a conventional MLFMA and NEPA-MLFMA. The relative error was plotted with respect to frequency from 4.5 GHz to 6.5 GHz. The error was calculated by considering scattered electric field intensity on the *z*-*x* plane and using MoM results as references.

(EFIE) was used to formulate the scattering problems. Each SRR was discretized by using 1 mm triangles, on which Rao-Wilton-Glisson (RWG) functions [15] were defined to expand the induced current density. Employing 73 unknowns per SRR, a total of 17,739 RWG functions were used for the overall array. For the solutions with NEPA-MLFMA, nine equivalence surfaces (rectangular boxes) were defined, each enclosing a layer of 3 × 9 SRRs, as also depicted in Figure 1. The size of each equivalence surface was 40 mm × 105 mm × 12 mm, i.e., the distance between the surfaces of the neighboring equivalence boxes was 18 mm – 12 mm = 6 mm. The equivalence surfaces were also discretized by using the RWG functions. Using 1356 RWG functions per equivalence surface, 2 × 12,204 unknowns were used to expand equivalent electric and magnetic currents. For comparisons, solutions were also performed by using a conventional MoM and a standard MLFMA [16]. All iterative solutions were performed by using the generalized minimal residual (GMRES) algorithm without preconditioning. The threshold of an iterative convergence was set to 0.0001. Finally, all electromagnetic interactions in MoM, MLFMA, and NEPA-MLFMA were computed with a maximum 1% relative error.

3.3 Results

Figure 3 presents the number of GMRES iterations and the solution time at nine different frequencies from 5.5 GHz to 6.5 GHz. It could be observed in the top plot that the iteration counts increased drastically at 5.25 GHz and 5.5 GHz for all solutions, i.e., using MoM, MLFMA, and NEPA-MLFMA. At the same time, NEPA-MLFMA dramatically reduced the number of iterations in comparison to MoM and MLFMA at all frequencies. It was remarkable that the discrepancy between MoM and MLFMA (specifically at 5.25 GHz and 5.5 GHz) was due to the ill-conditioning, which made the iterative solution extremely sensitive to errors in matrix-vector multiplications. At these frequencies, MLFMA was even slower than MoM due to larger numbers of iterations. As depicted in the bottom plot of Figure 3, NEPA-MLFMA significantly reduced the solution time, especially at 5.25 GHz and 5.5 GHz, in comparison to MoM and MLFMA. At 5.25 GHz, the speedup provided by NEPA-MLFMA was 3.36, and 7.45 with respect to MoM and MLFMA.

Figure 4 presents the scattered electric field intensity in the far zone at 5.0 GHz and 5.5 GHz. The amplitude of the field was plotted with respect to bistatic angle from 0° to 360° on the *z*-*x* plane. We observed that at both 5.0 GHz and 5.5 GHz, the MoM and NEPA-MLFMA solutions were consistent with each other. As shown in the top plot, MLFMA also provided consistent results at 5.0 GHz. However, at 5.5 GHz (bottom plot of Figure 4), the scattering values obtained by using MLFMA visibly deviated from those obtained with MoM and NEPA-MLFMA. This was again caused by ill-conditioning and accumulation of errors during iterative solutions when MLFMA was used for matrix-vector multiplications. Finally, Figure 5 provides an overall plot for the relative errors in MLFMA and NEPA-MLFMA solutions with respect to MoM solutions. It could be observed that the error was mostly below 1% for NEPA-MLFMA, while it reached more than 7% for MLFMA. It appeared that MLFMA solutions must be accelerated via strong preconditioners as practiced in [17], not only to improve the efficiency but also to avoid excessive errors at critical frequencies.

4. Solution to Problem SOLBOX-10

4.1 Solution Summary

Solver type (e.g., noncommercial, commercial): Noncommercial research-based code developed at CEMMETU, Ankara, Turkey

Solution core algorithm or method: Frequency-domain Equivalence-Principle Algorithm (EPA) accelerated with the Multilevel Fast Multipole Algorithm (MLFMA)

Programming language or environment (if applicable): *MATLAB + MEX*

Computer properties and used resources: 2.5 GHz Intel Xeon E5-2680v3 processors (using single core)

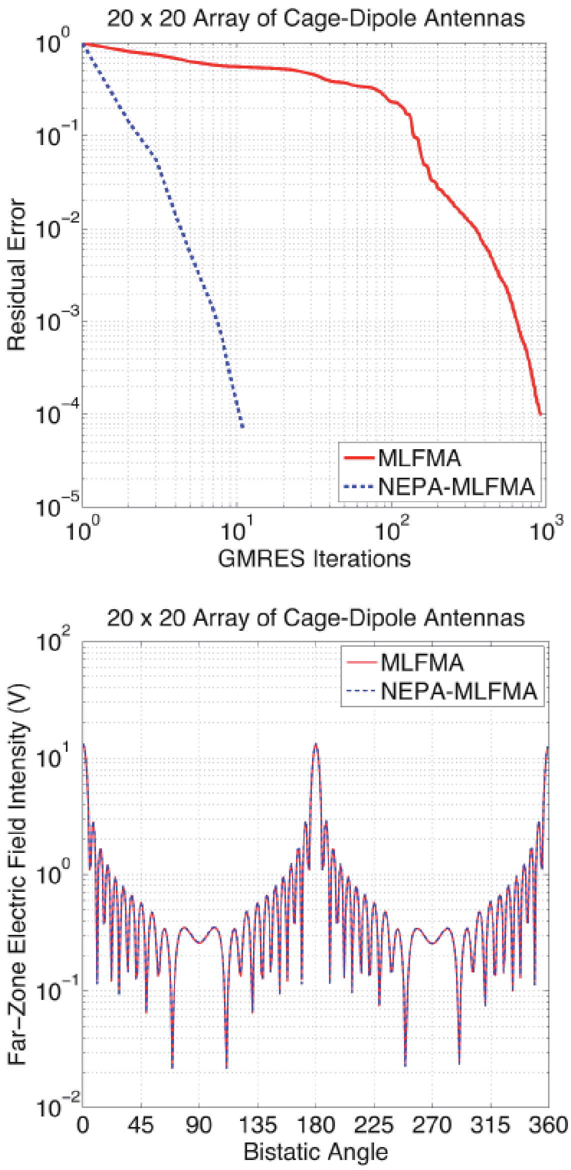

Figure 6. Two solutions of SOLBOX-10 using a conventional MLFMA and NEPA-MLFMA. In addition to convergence histories of iterative solutions, the electric field intensity in the far-zone is shown with respect to the bistatic angle on the *z*-*y* plane.

Total time required to produce the results shown (categories: < 1 sec, < 10 sec, < 1 min, < 10 min, < 1 hour, < 10 hours, < 1 day, < 10 days, > 10 days): < 1 hour

4.2 Short Description of the Numerical Solutions

In order to solve the problem SOLBOX-10, the NEPA-MLFMA implementation described in Section 3.2 was used. For the solution at the given frequency of 2.0 GHz, each cage-dipole antenna was discretized with 521 RWG functions defined on triangulated surfaces. This made a total of 400 × 521 = 208,400 unknowns for expanding the electric current density on the antenna surfaces. For employing EPA, each antenna was symmetrically located in a 5 cm × 10 cm × 2 cm equivalence surface, as also depicted in Figure 2. Considering the distances between the antennas (8 cm and 13 cm in the *x* and *y* directions), the surface-to-surface distance between the equivalence surfaces was 3 cm in both *x* and *y* directions. The equivalence surfaces were also discretized with the RWG functions. In comparison to the discretization of the antennas, a coarser mesh was used, leading to 186 RWG functions per equivalence surface and a total of 2 × 74,400 unknowns to expand the equivalent electric and magnetic currents. For comparisons, a standard MLFMA implementation was also used, where the discretized antenna array was directly solved without employing EPA. For both NEPA-MLFMA and MLFMA solutions, the antennas were formulated with EFIE. All electromagnetic interactions were computed with a maximum 1% error. Iterative solutions were performed by using the GMRES algorithm without preconditioning.

4.3 Results

Figure 6 presents the results of simulations performed by using MLFMA and NEPA-MLFMA. In the top plot, the residual error was plotted with respect to GMRES iterations. The advantage of NEPA-MLFMA was clearly visible as improved convergence rate and reduced number of iterations from 917 to only 11. In the bottom plot of Figure 6, the scattered electric field intensity in the far-zone was plotted with respect to bistatic angle on the *z*-*x* plane. The NEPA-MLFMA and MLFMA solutions were very consistent, verifying the accuracy of solutions.

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The Moscow Embassy Microwave Signal

Lately, there has been a great deal of scientific, public, and media attention about the health and safety of prolonged human exposure to radio-frequency (RF) radiation, especially with regard to cancer. The situation has been acerbated, in part, by dissemination of a report from a US government-sponsored experimental study [1, 2]. Indeed, release of the laboratory-rat cancer result was deemed as something of a public-health obligation by the reporting government entity: the National Institute of Environmental Health Sciences/National Toxicology Program [3].

The recent interest in this topic prompted me to recall an event that occurred in the mid 1970s, when the biological and health effects of RF and microwave radiation were brought to the fore by revelation of two US ambassadors to Moscow dying of cancer, and the then-ambassador Walter Stoessel threatening to resign in 1976, because of widespread staff concerns regarding potential health implications of an unusual microwave signal the Soviets were directing at the US Embassy in Moscow [4]. (Stoessel died of leukemia in 1986. He was 66 years old [5].)

Available reports suggest that since the 1960s, it had been known that the Soviets were targeting the upper floors of the central wing of the embassy in Moscow with low-level microwave radiation from nearby buildings, first from one, and then two sites [4, 6]. Indeed, it actually may have begun as soon as embassy staff had moved into the complex on Tchaikovsky Boulevard back in 1953, and persisted through at least 1988.

The levels of microwave exposure went up and down over the years. The direction and intensity of the 0.5 GHz to 10 GHz microwave signal changed in 1975 from 0.5 mW/m^2 for nine hours per day to 0.1 to 1.5 mW/m^2 for 18 hours per day, but it was always directed toward or aimed at the upper floors of the embassy. Only background levels were detected elsewhere in the embassy complex.

It is noteworthy that these levels were actually below what was considered as safe for human exposure in the US at the time. As a matter of fact, between 1966 and 1982, the American National Standards Institute (ANSI) guideline for safe human exposure to microwave radiation was 100 W/m^2 . Essentially, the same standard was adopted by the US Occupational Safety and Health Administration (OSHA). In contrast, the Soviets' standards were 1 mW/m^2 and 10 mW/m^2 for general public and occupational exposures, respectively [7]. The Soviet standards were derived from observations made in laboratory experiments using small animals, and surveys of people occupationally exposed to microwave radiation.

The existence of the microwave signal had been kept secret for years. This was because at the time, no one knew exactly why the Soviets were doing it, or that there might be any health consequences (a point I will get back to later).

Conjectures on intent and purpose abounded; the reasons why the Soviets did it remained a mystery. Speculations ranged from the Soviets' attempt to disable or interrupt communication signals from American electronic listening devices, to energizing electronic snooping gadgets that may have been implanted in the building prior to American occupancy, to influencing the psychological states or brainwashing embassy staff.

At the time, Soviet publications and scientific understanding concerning microwave interaction with animal behavior and neurophysiology, including the central nervous system, were replete with accounts of direct effects of low-level microwave exposure on animals and humans [8, 9]. There was also broad interest in clinical and hygienic or public-health findings, especially under occupational conditions. Most of the reported hygienic effects were of the nature of subjective complaints, commencing two to five years after the start of work involving microwave radiation. Specific categories included "asthenic syndrome," characterized by depression, fatigue, headache, irritability, loss of appetite, and memory. Another category, "autonomic syndrome," featured fainting spells, heart enlargement, and pulse and pressure liability. The third group, "diencephalic syndrome," was represented by digestive abnormality, insomnia, and sexual dysfunction.

This episode harkens back to 1952, when American diplomatic security personnel discovered a tiny electronic eavesdropping device in the American Ambassador's Moscow office [10]. It was concealed inside a carved Great Seal of the US given by the Soviets to the US Ambassador to Moscow in 1945. The Ambassador proudly hung it in the embassy. It was a passive RF transponder-type sensor: a predecessor of current-day RFID sensor technology, activated by RF energy launched from an external source, which allowed Soviet agents to eavesdrop on secret conversations for seven years before detection. From that point on, the embassy was under periodic surveillance for electronic signals. There thus was a distinct reason for the discovery of the microwave signal at the embassy complex on Tchaikovsky Boulevard back in 1953.

In any case, according to reports, when Ambassador Stoessel learned about the microwave signal, he threatened to resign unless the embassy community was informed. As a result, existence of the microwave signal was finally made public in a press conference called by the Ambassador in 1976. Still, the embassy community felt betrayed about being kept in the dark for so long [6].

It is conceivable that more of the embassy community were anxious about the effect the microwave signal might be having on their health, especially since the US Congress enacted the Radiation Control for Health and Safety Act in 1968 [11]. The deliberations had highlighted a general lack of current scientific knowledge on biological effects and health implications of both ionizing and non-ionizing radiation exposure, and unveiled the considerable amount of unnecessary radiation that people were exposed to each year, including microwaves.

The US Congress had declared then that the public's health and safety must be protected from the dangers of radiation from electronic products. *The act* authorized the federal government to set *radiation* standards, monitor compliance, and undertake research. It directed the US Department of Health, Education and Welfare to establish and carry out an electronic product radiation control program designed to protect the public's health and safety from radiation emitted by electronic products, including microwaves.

An immediate benefit of Stoessel's press conference was the installation of metallic screens on the embassy's outer windows to provide a substantial degree of shielding against microwave penetration through windows into the building [5]. The nominal screening efficiencies were about 1,000 to 100,000, depending on such factors as mesh size and type of materials.

Furthermore, while apparently unknown to the embassy staff, other activities were already underway. The US government had initiated a research program, code named "Operation Pandora," at the Walter Reed Army Institute of Research. Herb Pollack, a physician, Joseph Sharp, a behavioral psychologist, and Mark Grove, a microwave electronics engineer, were principals [10]. The program subjected trained Rhesus monkeys to microwave exposures, mimicking characteristics gathered by monitors at the Moscow embassy. Microwave-induced interruption in the monkey's performance for food was studied in a classic operant-conditioning experimental protocol. The study was terminated in 1969. Reported findings included some microwave irradiation associated "aberrant behavior." However, agreement on an unambiguous consensus for behavioral psychological changes was not reached among project personnel.

Apparently, soon after the highly publicized press conference where existence of the Moscow Embassy microwave signal was finally made public in 1976, a two-year epidemiological research was initiated by the US Department of State at Johns Hopkins University, near Washington DC [12-14]. The study involved 1,827 employees and 3,000 dependents at the Moscow Embassy, and 2,561 employees and 5,000 dependents at comparable other US Foreign Service posts in the Soviet bloc as a control population, namely, Belgrade, Budapest, Leningrad, Prague, Sophia, Warsaw, and Zagreb, during the period from 1953 to 1975. The control or comparison group was chosen to match the study group in selection criteria and environmental factors such as climate, diet, disease, and general social milieu, except that the postings were not subjected to microwave irradiation.

The purpose was to assess any differences in morbidity and mortality between the Moscow and comparison groups. Extensive efforts were devoted to identify and trace the populations. Information on illness, conditions, or symptoms was gathered and validated. Death certificates were used to ascertain mortality. Standardized mortality ratios and morbidity indices for various groups were developed for the study. At completion of the study in

1978, the investigators' conclusion was that the Moscow and comparison groups did not significantly differ in overall and specific mortality, and no compelling evidence was observed to implicate the Moscow microwave signal in any adverse health effect. Nevertheless, investigators had noted that the study population was relatively young, and it might have been too early to detect long-term health and mortality outcomes.

The most perplexing question remains: what was the Soviets' purpose in microwaving the Moscow embassy?

At a later point, there was an indication that the Soviets had suggested it was a jamming signal calculated to thwart US electronic spying devices. This is a plausible but simplistic rationalization; however, the signal strengths were rather weak for jamming maneuvers.

One possibility is that the Soviets were using microwaves to activate numerous snooping devices they had implanted in the building prior to American occupancy. This obviously was an expected scenario, given the success of the bug concealed in the carved Great Seal, especially for the reason that electronic fabrication capability had advanced by leaps and bounds since 1945. However, ever since discovery of the Great Seal bug, surveillance for electronic signals had become routine at the embassy complex.

Another intriguing proposition was that the Soviets were bouncing the microwave signal off the embassy's window glass in an attempt to eavesdrop on conversations taking place inside the office. The theory was as follows: conversation-generated sound waves would set the glass window pane into tiny vibrations, which does happen, in principle. Reflection of the microwave signal impinging on the window pane would be modulated by the vibrations in amplitude and phase, which might then be electronically demodulated to reproduce conversations taking place inside the office. While feasible, there are technological challenges in converting the faint electronic signals to voice, although it is not impossible. At best, the reflected microwave intensity would be down about 90% while also being in a potentially noisy electronic operating environment. Nevertheless, given the demonstrated sophistication of the Great Seal bug from an earlier period, it could be well within the Soviets' capability.

Finally, it might have been a designed exploitation of the Soviet understanding of how prolonged exposure to low-level microwaves affects the mental state of exposed subjects. If that was the intent, the Moscow microwave signal indeed may have accomplished its intended purpose, in part. The embassy staff clearly showed anxiety after learning about the existence of the microwave signal. They were concerned about potential health effects the microwave signal might be having on them or their children. It became somewhat of a morale problem for a time.

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From Science to Management

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My career path in radio science seemed to start mostly by accident. However, I was probably programmed from birth to where I eventually found myself. I had always intended to be an engineer, and was brought up with electronics as the natural field to be in. When it came time to select a university, I chose a field that combined science and engineering, ending up at Rhodes University in Grahamstown, South Africa, doing an undergraduate degree in Physics and Electronics.

The Physics Department at Rhodes had a strong radio science interest, with an active ionospheric physics group. By the time I was in my honors year, I had already started to engage in those activities. I received my PhD in Space Physics from Rhodes University in 2003 for a thesis entitled, "The Development of a Neural Network Based Ionospheric Model for the Bottomside Electron Density Profile Over Grahamstown, South Africa." The thesis work cemented the role that I would play in the radio-science arena over the next decade, in the areas of ionospheric physics and space weather.

After my PhD, I completed a postdoctoral year as a research fellow in the Graz University of Technology in Austria. This was an amazing experience, and made me a strong believer in the value of postdoctoral fellowships. My postdoctoral project was on the use of neural networks for modeling the lower ionosphere (D layer), utilizing incoherent-scatter-radar data and rocket-based measurements. One of the most amazing experiences for me was assisting on a Norwegian project and visiting the EISCAT radar. It was my first and only experience of the Northern Lights, and the power of science in general and space weather in particular has been with me ever since.

After returning to South Africa, I was offered a joint appointment as a Researcher with Rhodes University and the Hermanus Magnetic Observatory (HMO). The HMO was a national facility of the National Research Foundation (NRF). It was a dream job, as I had the best of both worlds: academic life focusing on research that I loved and aspired to do, and student development. I am extremely passionate about student development, and had an active group of students and postdoc fellows from all over Africa working with me on various aspects of ionospheric modeling. I also became integrated into the URSI community, having been the recipient of an URSI Young Scientist Award twice (1996 and 1999), and regularly attending the URSI General Assemblies. I have attended every URSI General Assembly between 1996 and 2017 (a total of eight in a row). I am also very involved in the National URSI Committee of South Africa, first as the representative for Commission G, and in recent years as the Chair of the Committee.

Through my research interests, I also became a contributor and member of the International Reference Ionosphere (IRI) Working group, which is a joint working group of URSI and COSPAR. Several model developments that my group worked on where incorporated into the IRI, and I was Chair of the group for four years, from 2010 to 2014.

During the period between 2004 and 2010, I sat on a number of international committees and published extensively. I received a B3 rating from the South African NRF, which indicated international recognition for my research. Over the years, I like to believe that I contributed to the space-research focus in South Africa, bringing the concept of a space-weather capability to the country as well as developing capability in this scarce skills area.

My research focus was also in assisting with the development of capability within Africa through active collaborations with Nigeria, Zambia, and Kenya. I also supervised a number of students from different countries within Africa. A major project for me was in June 2009, when I co-chaired the Local Organizing Committee for the International Heliophysical Year (IHY)-Africa 2009 Workshop, held in Livingstone, Zambia. This workshop aimed to promote an awareness of Space Science in Africa, and assist African space scientists to develop international contacts in the field.

In 2009, I was fortunate to receive three awards, including the African Union Regional Women Scientist Award for Basic Science, Technology and Innovation in the Southern African Region, and a Department of Trade and Industry (DTI) 2009 Technology award in the Advanced High Technology Category for the industry linked project "Ionospheric Model: Phase 5."

In 2007, I started to move towards management of science. In 2010, I was asked to lead the HMO through a migration process to the newly established South African National Space Agency (SANSA). SANSA became operational on April 1, 2011, and I accepted the position of Managing Director for the Space Science program (essentially the new version of the HMO). Rhodes University offered me a Visiting Research Professorship that allowed me to continue supervising students, even though I had moved permanently to SANSA.

Eventually, after struggling to balance both research and management, I realized that I needed to be either a good researcher or a good director, but that there was no way I could be good at both. A key aspect was the opportunity to make a difference. So, in 2013, I registered for an MBA with the Business School Netherlands. This was a wonderful learning experience for me, coming from a science background. I graduated with my MBA in September 2015. During the past decade I have led a number of successful research, science-advancement and infrastructure-funding proposals, and I have tried to mentor and nurture talent within the fascinating world of radio science and space. I miss the pure-research side of things. However, I take strength in the fact that I am able to influence focus and policy that will allow many young researchers in South Africa to grow the country's knowledge economy.

Currently, I am the Managing Director for Space Science within the South African National Space Agency (SANSA). The position involves providing both strategic direction and operational oversight to the space science program and facility located in Hermanus, South Africa. In South Africa, space science is defined as the area between the sun and the Earth and everything in between, referred to as the near-Earth space environment, as well as the impact of solar activity on Earth systems. The space-science program is responsible for research, infrastructure, and products related to the near-Earth space environment, and so, although I am not undertaking research myself these days, I am still able to influence and direct focus areas in space science for the organization and nationally. One of the leading focus areas for the next few years will be the "operationalizing" of space-weather models to provide robust reliable prediction and forecast products to various sectors. SANSA has embraced this move, and is currently putting things in place to ensure an operationalized environment for its space-weather center. This will involve moving our focus to a more regional focus, and steering our research towards the eventual operational requirements. Both the area of space weather as well as the directing of research for the African continent have always been passions of mine, and I look forward to playing a significant role in these areas in the future.

INTERNATIONAL SYMPOSIUM ON ELECTROMAGNETIC THEORY (EMTS 2019)

May 27-31, 2019, San Diego, CA, USA

First Call For Papers

The International Symposium on Electromagnetic Theory (EMTS 2019) will be held May 27-31, 2019, in San Diego, CA, USA. It is organized by Commission B (Fields and Waves) of the International Union of Radio Science (URSI), and is financially cosponsored by the United States National Committee for URSI (USNC-URSI) and the IEEE Antennas and Propagation Society (IEEE AP-S). EMTS 2019 is the 23rd event in the triennial series of international EMT symposia, which has a long history since 1953. Its scope covers all areas of electromagnetic theory and its applications. It is the major scientific event of Commission B, along with the URSI General Assembly and Scientific Symposium, Atlantic Radio Science Conference (AT-RASC), and Asia Pacific Radio Science Conference (AP-RASC). The venue is the hotel Westin San Diego, which is minutes from downtown activities including the San Diego Zoo, Balboa Park and its numerous museums, and the Gaslamp district for dining and nightlife. San Diego is the eighth largest city in USA, and is often referred to as "America's Finest City." Known for its great hotels, beautiful weather, pristine beaches, friendly people, and a plethora of entertainment, San Diego is a favorite destination for visitors across the globe. The San Diego airport is conveniently close to our symposium venue, so transportation to the conference will be quick and easy.

Welcome to San Diego in May 2019! The conference will offer plenary talks by distinguished speakers, regular oral and poster sessions, and a one-day school for young scientists (May 27), focusing on a topic in electromagnetics. A number of Young Scientist Awards will be offered, covering the registration fee and accommodation during the conference. In addition, business meetings, receptions, and a conference banquet will be organized. EMTS 2019 will focus on electromagnetic fields and their applications. Contributions on any aspect of the scope of Commission B are solicited. Some suggested topics are listed below. Special-session topics will be listed later on the Web site. All submissions (two to four pages in IEEE two-column format) will be reviewed by the Commission B Technical Advisory Board. Accepted and presented papers may be submitted to IEEE Xplore.

Important dates

- Paper submission site opens: July 15, 2018
- Deadline for paper submission: **October 22, 2018**
- Notification of acceptance: January 10, 2019
- Early-bird and author registration ends: March 30, 2019

Contact: Technical Program: Kazuya Kobayashi <kazuya@tamacc.chuo-u.ac.jp>
Local Organizing Committee: Sembiam Rengarajan <srengarajan@csun.edu>

Suggested Topics

1. Electromagnetic theory
- Analytical and semi-analytical methods
- Mathematical modeling
- Canonical problems
- Scattering and diffraction
- Inverse scattering and imaging

2. Computational methods
- Integral equation methods
- Partial differential equation methods
- High-frequency and hybrid methods
- Fast solvers and high-order methods
- Time-domain techniques
- Computational algorithms

3. Materials and wave-material interaction
- Metamaterials and metasurfaces
- Plasmonics and nanoelectromagnetics
- Electromagnetic bandgaps and other periodic structures
- Optical devices
- EMC and EMI
- Bioelectromagnetics

4. Antennas and propagation
- Antenna theory
- Antenna measurements
- Multi-band and wideband antennas
- Antenna arrays and MIMOs
- Wireless communication systems
- Guided waves and structures
- Random media and rough surfaces
- Millimeter wave/5G propagation
- Millimeter-wave antennas
- MIMO for 5G communication

5. Other topics
- History of electromagnetics
- Education in electromagnetics

www.emts2019.org

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Associate Editor's Introduction

At the triennial occasion of the General Assembly and Scientific Symposium (GASS), URSI acknowledges scientific achievements in the field of radio science by awarding outstanding individuals with a number of URSI awards. Two of those awards are "early career" awards, meaning that they are awarded to a young scientist not more than 35 years old on September 30 of the year preceding the URSI GASS. These early career awards are the Issac Koga Gold Medal and the Santimay Basu Prize.

The Santimay Basu Prize is awarded to a young scientist who has made an outstanding contribution to research that furthers the understanding of radiowave propagation in random media and its application for the benefit of society. The award takes into account the excellence of the research, the merit of the candidate in achieving his or her results, and the efforts required to accomplish the research. At the 2017 URSI GASS in Montreal, the Santimay Basu Prize was presented to Dr. Jamesina Simpson with the citation, "For advancing three-dimensional finite-difference time-domain (FDTD) solutions of electromagnetic wave propagation within the global Earth-ionosphere waveguide applied to space weather, remote-sensing, and very-low-frequency propagation."

The Issac Koga Gold Medal is awarded to a young scientist who has made an outstanding contribution to any of the branches of science covered by the Commissions of URSI. The award is for career achievements of the candidate, with evidence of significant contributions within the most recent six-year period. At the 2017 URSI GASS in Montreal, the Issac Koga Gold Medal was presented (Figure 1) to Dr. Yue Li with the citation, "For contributions to the development of electromagnetic metamaterial circuits and antenna designs in mobile communication systems." It is a great pleasure to put Dr. Li in the spotlight in an interview in this column.

Interview with Yue Li: Winner of the 2017 URSI Issac Koga Gold Medal

What was your educational background?

Although I did not get a good grade in the college entrance examination, I was quite fortunate to be the last student (with the lowest grade) enrolled by Zhejiang University, a famous Chinese University in the field of engineering. To prove I was not the worst student, I worked hard in the university and won the qualification of Tsinghua University as a graduate student in 2007. Tsinghua University is the best university in China, and it is a big challenge to graduate. I joined the microwave group of Prof. Zhenghe Feng, who became my PhD supervisor. There, my major was microwave antenna design. After I finished the PhD degree in 2012, I stayed in the same group and started the new topic of metamaterials and circuit design. In December 2013, I joined the group of Prof. Nader Engheta at the University of Pennsylvania, in pursuit of new techniques of fields and waves with a wide range of optics such as metamaterials and nanophotonics. In January 2016, I returned to Tsinghua University as an Assistant Professor.

How would you describe your career so far?

With challenges, but interesting and meaningful. The experiences at the University of Pennsylvania and Tsinghua University had taught me how to think about science and technology, how to develop new concepts, and how to apply them in practical engineering. The curiosity of science always pushes me to learn new topics. There are challenges, but your passions will help you to overcome these hurdles. Being an Assistant Professor in Tsinghua University is an interesting career, and you will not easily predict what you will do in the next five years. You are always learning new things and creating new things.

How did you become interested in your current research field?

My current research field is technologies in electromagnetics, including antennas, circuits, and metamaterials. The amazing physics behind this applied science made me become interested in this field. You should imagine that by engineering different metallic or dielectric structures, we can control and tame the electromagnetic fields and waves. For example, you can design an antenna to totally radiate in one direction, rather than omni-directionally, or you can tune the scattering light at will just using a layer of metal. Such counterintuitive phenomena fascinate me in pursuit of the technologies in electromagnetics.

What motivates you in your field of study?

The difference between pursuing novelties in theory and engineering. For an engineering department, we may focus more on how to put a new technique into practical usage, such as solving new problems, application to new environments, or designing new structures. But for theoretical innovation, a new concept or a new phenomenon is more important, with strict theoretical analysis, numerical simulations, and ideal experiments. The motivation for my research is to overcome this difference, to develop new theory and apply it in practical systems or devices. Challenges are always there, but this is my goal, creating new concepts from theory to engineering.

Which challenges have you faced during your career?

In the field of electromagnetics, the topics are quite wide and you need to learn a lot of new concepts. For instance, you need to learn techniques from dc to RF, and from microwaves to optics. You also need to learn fabrication processes and measurements. Learning new things always comes with challenges, but if you are passionate about it, it is quite interesting.

What is the current "hot field to study" in your area, and what do you think it will be in the future?

In my opinion, the circuit concepts inspired by metamaterials form a promising topic in the field of electronics and optics. As we know, integrated-circuit techniques change our daily life as we use such lumped circuits in electronic devices, such as computers, mobile phones, and so on. However, existing circuit techniques only operate up to a few gigahertz, i.e., microwave range, which limits the signal-processing rate and wireless-data-transfer speed. If we can achieve the integrated-circuit idea in the optical domain, we may easily design a supercomputer with the dimensions of a cell phone, and we may transfer an HD movie in just one second. However, the materials hindered the development of such great circuits in the optical domain, as we no longer have metal or less-dispersive dielectrics in the optical domain. Even though nature did not provide the materials we need, we can artificially design materials with such properties, and this is the motivation for the topic of metamaterials. Based on the concept and advanced design of metamaterials, we may translate the techniques from the microwave range to the optical range, leading to the development of new technologies and theories. Among them, the lumped-circuit design is the most significant in the optical domain, not only to design components and devices at sub-wavelength scale, but to increase the rates of signal processing and data transfer. Wireless communication will come to a new generation.

What fascinates you most in your current position?

The new ideas in science and technology, the smart students, and the creative colleagues. The most fascinating moment is the feasibility of a new idea, which is successfully published in a paper or adopted in engineering projects.

Which achievements earned you the Issac Koga Gold Medal?

I think there were three contributions for the Koga Gold Medal.

First, there was the idea of lumped circuitry inside a waveguide, also named "waveguide metatronics." Based on this idea, we can design lumped circuits using waveguide metamaterials in different ranges, such as the microwave domain, the terahertz domain, and the optical domain, breaking the limitations of material selections.

Second, the development of photonic doping, which transplants the doping idea from the microscopic scale to the macroscopic scale. We can easily control the magnetic properties of an epsilon-near-zero medium by tuning the parameters of an inserted dielectric rod, such as the diameter and permittivity. This is also a new paradigm of metamaterials without using periodic structures.

Third, for the contribution to mobile antenna design. In this topic, advanced techniques of multiple-domain cooperation (such as the spatial, time, and frequency domains) are used to design multiple antennas within a small volume, with the merits of high isolation, small dimensions, and omnidirectional radiation. Such ideas can be adopted in the design of mobile-phone antennas, base-station antennas, and access-point antennas.

Which opportunities/persons have been instrumental for your career so far?

There are three persons with a significant impact on my career. The first two I met in my graduate-student pursuits in the microwave group of the Department of Electronic Engineering at Tsinghua University. I met my supervisor, Prof. Zhenghe Feng, and my co-supervisor, Prof. Zhijun Zhang. They brought me into the antenna world. They were experienced professors in microwave techniques, and taught me how to find and solve problems in practical projects. By accumulating experience in engineering, I can find new problems from the engineering point of view. The third person I met during my days in Prof. Nader Engheta's group in the University of Pennsylvania. Prof. Engheta is a scientific giant in the field of metamaterials and nanophotonics. He taught me how to theoretically solve a problem and helped me to think theoretically. These three great professors taught me not only in science and technologies, but also how to be a good person. They started my career of academic life.

What do you hope to achieve in the coming years?

First, as a professor, I hope to successfully teach. In my class of antennas and metamaterials, I hope more students will become interested in such areas, and I can help them to do research on these topics.

Second, I hope my research will have good progress. We have lots of interesting ideas, but need to prove them theoretically and experimentally. I hope my group members will have more great achievements in the fields of antennas and metamaterial-inspired circuits.

What advice would you give to students or early-career researchers?

I am also an early-career researcher. I am glad to share some experience with other early-career researchers, and also with students. At the beginning, everything is tough, and we need to be optimistic, even though the reality is different. We may have to face more failures than successes. We need to maintain our passion. The joy of discovery in science and technology comes from passion. Even though you choose a "cold" area, passion will lead you to excellent research and great achievements. Please believe your passion, which will be leading you to get your goals.

A Short Biography of Yue Li

Yue Li received his BS in Telecommunication Engineering from Zhejiang University, Zhejiang, China, in 2007, and his PhD in Electronic Engineering from Tsinghua University, Beijing, China, in 2012. He is currently an Associate Professor in the Department of Electronic Engineering at Tsinghua University. In June 2012, he was a Postdoctoral Fellow in the Department of Electronic Engineering, Tsinghua University. In December 2013, he was a research scholar in the Department of Electrical and Systems Engineering, University of Pennsylvania. He was also a visiting scholar at the Institute for Infocomm Research (I2R), A*STAR, Singapore, in 2010, and at the Hawaii Center of Advanced Communication (HCAC), University of Hawaii at Manoa, Honolulu, HI, USA, in 2012. Since January 2016, he has been with Tsinghua University.

He has authored and coauthored over 80 journal papers and 30 international conference papers, and holds 15 granted Chinese patents. His current research interests include metamaterials, plasmonics, electromagnetics, nanocircuits, mobile and handset antennas, MIMO and diversity antennas, and millimeter-wave antennas and arrays. He was the recipient of the Issac Koga Gold Medal at the URSI General Assembly and Scientific Symposium in 2017; the Second Prize of Science and Technology Award of China Institute of Communications in 2017; the Best Student Paper Award (Third Prize) from APCAP 2017; the Young Scientist Award from URSI AP-RASC 2016; the Young Scientist Award from EMTS 2016; the Best Student Paper Award form ICMMT 2016; the Best Paper Award form ISAPE 2016; the Young Scientist Award from URSI GASS in 2014; the Outstanding Doctoral Dissertation of Beijing Municipality in 2013; and the Principal Scholarship of Tsinghua University in 2011. He serves as an Associate Editor of the *IEEE Transactions on Antennas and Propagation* and the *IEEE Antennas and Wireless Propagation Letters*.

Figure 1. The Issac Koga Gold Medal being presented to Dr. Yue Li (r) by Prof. Kazuya Kobayashi during the opening ceremony of the 2017 URSI GASS in Montreal, Canada.

Report on IEEE RADIO 2017

Prof. Weng Cho Chew, IEEE Antennas and Propagation Society (AP-S) President Elect 2018, officially opened the fifth edition of the 2017 IEEE Radio and Antenna Days of the Indian Ocean (IEEE RADIO 2017) international conference (Figure 1). This was held at the Protea Hotel by Marriott Cape Sea Point, Cape Town, South Africa, from September 25-28, 2017. The conference was organized by the Radio Society (Mauritius), and financially sponsored by IEEE AP-S. IEEE RADIO 2017 was technically cosponsored by the Union Radio Scientifique Internationale (URSI) and IEEE Region 8.

The local organizing committee consisted of researchers from CentraleSupélec, France; Université de La Réunion; University of Mauritius; and Square Kilometre Array (SKA) Africa. We were grateful to SKA Africa, Newton Fund, Development in Africa for Radio Astronomy (DARA), Mobile & Wireless Forum (MWF), and Computer Simulation Technology (CST) for providing financial sponsorship for the conference.

IEEE RADIO 2017 gathered participants from over twenty different countries, coming from all five continents. The conference featured twelve regular sessions, covering diverse topics such as radio astronomy, antenna design, wave propagation and scattering, couplers, waveguides and resonators, wireless communication systems, devices and circuits, computational electromagnetics, and biological effects and medical applications of electromagnetic fields. A first workshop was dedicated to the important SKA project, and a second workshop addressed the technical and financial implications of counterfeit devices. Several internationally recognized scientists delivered invited talks during the conference.

In line with the previous editions, the oral presentations of young scientists and students were evaluated by a jury. A cash prize of 300 Euros and a shield were awarded to the Best Young Scientist less than 36 years old. Moreover, three cash prizes of 300 Euros, 200 Euros, and 100 Euros as well as shields were awarded to the first, second, and third best student papers, respectively.

The next edition of the RADIO international conference will be held on the beautiful island of Mauritius in 2018.

Vikass Monebhurrun
General Chair, IEEE RADIO 2017
E-mail: vikass.monebhurrun@centralesupelec.fr

Figure 1. The opening ceremony of IEEE RADIO 2017 on Monday, September 25, 2017.

USNC-URSI Student Paper Winners

The US National Committee of URSI held its annual National Radio Science Meeting January 3-6, 2018, in Boulder, Colorado. The winners of the Student Paper Contest at the meeting were as follows:

First Prize: **Yahya Mohtashami** (Figure 1)
"A Balun-Free Hybrid Helix/Monopole Antenna for Microwave Ablation"
University of Wisconsin, Madison
Prize amount: $1000

Figure 1. Yahya Mohtashami (c), winner of the First Prize in the USNC-URSI National Radio Science Meeting Student Paper Contest is shown with USNC-URSI Chair Sembiam Rengarajan (l) and Erdem Topsakal (r), Chair of the Student Paper Contest.

Second Prize: **Daisong Zhang** (Figure 2)
"A Novel Flexible Electro-Textile 3T MRI RF Coil Array for Stroke Prevention: Design, Characterization and Prototyping"
University of California, Los Angeles
Prize amount: $750

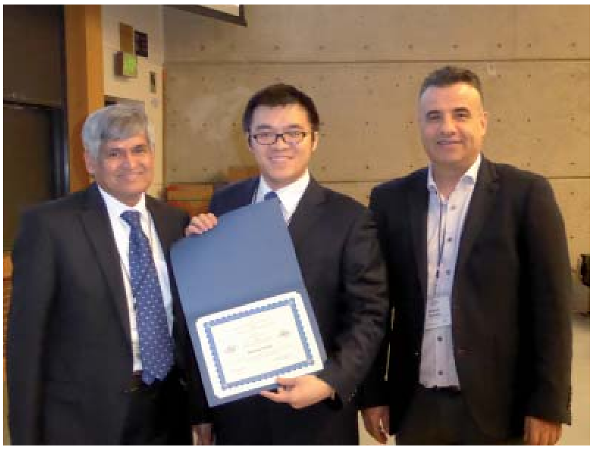

Figure 2. Daisong Zhang (c) won the Second Prize in the USNC-URSI Student Paper Contest.

Third Prize: **Wenyang Li** (Figure 3)
"Comparison Redundant and Sky Model Based Interferometric Calibration: A First Look with Phase II of the MWA"
Brown University
Prize amount: $500

Figure 3. Wenyang Li (c) won the Third Prize in the USNC-URSI Student Paper Contest.

ISAPE 2018

The 12th International Symposium on Antennas, Propagation and EM Theory

December 3 - 6, 2018, Hangzhou, China

www.meeting.org/isape2018

ISAPE, a serial symposium on antennas, propagation, and EM theory, always offers an active forum for exchanging creative ideas and experiences on the latest developments and designs in the areas of antennas, propagation, and electromagnetic theory for professors, researchers, engineers, and excellent students all over the world. The 12th International Symposium on Antennas, Propagation and EM Theory, ISAPE 2018, will be held in Hangzhou, China. All prospective papers in the areas of antennas, propagation, electromagnetic theory, computational electromagnetics, and EMC are welcome. All accepted papers will be indexed by the INSPEC database and EI Compendex. **Paper submission deadline: September 1, 2018**

SUGGESTED TOPICS

A. Antennas & Related Topics
A1. Microstrip & Printed Antennas
A2. Active & Integrated Antennas
A3. Array Antennas, Phased Arrays and Feeding Circuits
A4. Small Antennas
A5. Adaptive & Smart Antennas
A6. Multi-Band/Wideband Antennas
A7. Wire & Slot Antennas
A8. Aperture Antennas & Feeds
A9. Millimeter Wave & Sub-Millimeter Wave Antennas
A10. Optical Technology in Antennas
A11. Antennas in Mobile Communication
A12. Antenna Measurements
A13. FSS, Polarizers & Radomes
A14. Reconfigurable Antennas & Arrays
A15. Reflector/Lens Antennas & Feeds
A16. Others

B. Propagation & Related Topics
B1. Mobile & Indoor Propagations
B2. Mobile Channel Characterization & Modeling
B3. Millimeter & Optical Wave Propagation
B4. Earth-Space & Terrestrial Propagation
B5. Radio Meteorology
B6. Remote Sensing
B7. SAR Polarimetry & Interferometry
B8. Tunnel Propagation
B9. Propagation in Ionized and Non-Ionized Media
B10. Radio Astronomy
B11. Ionospheric Modification
B12. Tropospheric, Stratospheric and Ionospheric Sounding
B13. Incoherent Scatter Radar & Observations
B14. Others

C. EM Theory & Related Topics
C1. Bioelectromagetics
C2. EM Fields in Complex Media
C3. Geo-Electromagnetics
C4. Theoretical Electromagnetics & Analytical Methods
C5. Transient EM fields
C6. High-Frequency Techniques
C7. Nonlinear electromagnetics
C8. Random Media & Rough Surfaces
C9. Waveguiding Structures
C10. Time-Domain Techniques
C11. Inverse Problems & Imaging
C12. Scattering, Diffraction, & RCS
C13. Metamaterials & Electromagnetic Bandgap Structures
C14. Measurement Techniques
C15. Nano-Electromagnetics
C16. Seismo-Electromagnetics
C17. Others

D. Computational Electromagnetics
D1. Integral Equation Methods
D2. Differential Equation Methods
D3. Hybrid Techniques
D4. Optimization Techniques for CEM
D5. Asymptotic & High-Frequency Techniques
D6. Low-Frequency Electromagnetics
D7. Computational Bioelectromagnetics
D8. Pre- & Post-Processing
D9. Nondestructive Techniques
D10. NEC Modeling & Analysis
D11. FEKO Modeling & Analysis
D12. CST Modeling & Analysis
D13. MEFiSTo Modeling & Analysis
D14. Object-Oriented Computational Electromagnetics
D15. Transmission-Line Theory
D16. Others

E. Electromagnetic Compatibility & **Related Topics**
E1. Probe & Sensor
E2. Absorbing Materials
E3. Test Chambers
E4. EMC Test & Measurement
E5. Coupling & Crosstalk
E6. EMC Standards
E7. EM Environment
E8. Automotive EMC
E9. EM Bioeffects
E10. EMC in Communications
E11. EMC in Power Engineering
E12. Lightning, ESD & EMP
E13. EMC in Computer & PCBs
E14. Shielding, Filtering & Grounding
E15. EMC in Microelectronics
E16. Immunity & Susceptibility
E17. Spectrum Management
E18. EMI Prediction Analysis & Reduction Technique
E19. EMC Education
E20. Others

F. Others
F1. High-Power Microwave Applications
F2. UWB & Impulse Applications
F3. Ubiquitous Network Systems
F4. Satellite Communication Systems
F5. Radio Technologies for Intelligent Transport Systems
F6. Subsurface Sensing
F7. MEMS-NEMS & MMIC
F8. Passive & Active Circuits
F9. Power Amplifiers, Linearization, & Active Components
F10. Millimeter Wave & Sub-Millimeter Wave Components, Circuits & Systems
F11. Signal Processing for Communications
F12. Advanced Process, Packaging & Integration Technologies
F13. 3D RF Technology
F14. Electromagnetic Materials
F15. Electromagnetic Environment Effects (E3)
F16. Earthquake Precursors & Monitoring
F17. THz Technology

Sponsor: Chinese Institute of Electronics (CIE). **Cosponsors:** CIE Radio Propagation Society, CIE Antenna Society
Local Organizer: Hangzhou Dianzi University. **Working Language:** English.
Paper Submission: See the conference Web site:

www. meetingcn.org/isape2018

URSI Conference Calendar

March 2018

Gi4DM 2018
Kyrenia, Turkish Republic of Northern Cyprus, 14-18 March 2018
Contact: K2 Conference and Event Management Kosuyolu Mh. Ali Nazime Sk. No: 45 Kosuyolu 34718 Kadikoy / Istanbul Phone: +90 (216) 428 95 51 - Fax: +90 (216) 428 95 91 E-mail: gi4dm@k2-events.com, http://www.gi4dm2018.org

8th VERSIM Workshop
Apatity, Russia, 19-23 March 2018
Contact: Dr. Andrei Demekhov andrei@appl.sci-nnov.ru http://www.iugg.org/IAGA/iaga_ursi/versim/

Geolocation and navigation in space and time
The URSI-France 2018 Workshop
Paris, France, 28-29 March 2018
Contact: http://ursi-france.org/

May 2018

Baltic URSI Symposium (part of Microwave Week 2018)
Poznan, Poland, 14-16 May 2018
Contact: Prof. Dr. Andrzej Napieralski, Baltic URSI Symposium 2018 Chairman & Dr Przemysław Sękalski, http://mrw2018.org/ursi2018/national-ursi-2018-topics/

AT-RASC 2018
Second URSI Atlantic Radio Science Conference
Gran Canaria, Spain, 28 May – 1 June 2018
Contact: Prof. Peter Van Daele, URSI Secretariat, Ghent University – INTEC, Technologiepark-Zwijnaarde 15, B-9052 Gent, Belgium, Fax: +32 9-264 4288, E-mail address: E-mail: peter.vandaele@intec.ugent.be, http://www.at-rasc.com

July 2018

COSPAR 2018
42nd Scientific Assembly of the Committee on Space Research (COSPAR) and Associated Events
Pasadena, CA, USA, 14 - 22 July 2018
Contact: COSPAR Secretariat (cospar@cosparhq.cnes.fr) http://www.cospar-assembly.org

August 2018

Metamaterials 2018
12th International Congress on Artificial Materials for Novel Wave Phenomena
Espoo, Finland, 27-30 August 2018
Contact: http://congress2018.metamorphose-vi.org/

October 2018

COMPENG
2018 IEEE Workshop on Complexity in Engineering
Florence, Italy, 10-12 October 2018
Contact: compeng2018@ino.cnr.it , http://compeng2018.ieeesezioneitalia.it/

RADIO 2018
IEEE Radio and Antenna Days of the Indian Ocean 2018
Mauritius,15-18 October 2018
Contact: http://www.radiosociety.org/radio2018/

ISAP 2018
2018 International Symposium on Antennas and Propagation
Busan, Korea, 23-26 October 2018
Contact: http://isap2018.org/

November 2018

APMC
Asia-Pacific Microwave Conference 2018
Kyoto, Japan, 6-9 November 2018
Contact: http://www.apmc2018.org/

LAPC 2018
Loughborough Antennas and Propagation Conference
Loughborough, United Kingdom, 12-13 November 2018
Contact: Poppy Seamarks, Tel: +44 (0)1438 767 304, Fax: +44 (0)1438 765 659, Email: lapc@theiet.org
https://events.theiet.org/lapc/

March 2019

C&RS "Smarter World"
18th Research Colloquium on Radio Science and Communications for a Smarter World
Dublin, Ireland, 8-9 March 2019
Contact: Dr. C. Brennan (Organising Cttee Chair) http://www.ursi2016.org/content/meetings/mc/Ireland-2017-CRS Smarter World CFP.pdf

AP-RASC 2019
2019 URSI Asia-Pacific Radio Science Conference
New Delhi, India, 9-15 March 2019
Contact: Prof. Amitava Sen Gupta, E-mail: sengupto53@yahoo.com

May 2019

EMTS 2019
2019 URSI Commission B International Symposium on Electromagnetic Theory
San Diego, CA, USA, 27-31 May 2019
Contact: Prof. Sembiam R. Rengarajan, California State University, Northridge, CA, USA, Fax +1 818 677 7062, E-mail: srengarajan@csun.edu

URSI cannot be held responsible for any errors contained in this list of meetings

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Prof. T. Tanzi (France)
Prof. S. Wijnholds (Netherlands)
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