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INNOVATION JOURNAL



**TO FIND AND NOT BE FOUND,
THAT IS THE QUESTION!**
Surveillance and low observability

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 **LEONARDO**

TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

Surveillance and low observability

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TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

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editorial

In a world dominated by digital technologies and Artificial Intelligence (AI) there will always be space for electromagnetism when the application needs some sensing or transmitting function to interact with the real world. Sensors like radars cannot exist if there is no device that transduces data (and signals) into an Electro-Magnetic (EM) wave, as well as a device that, viceversa, senses the radiated EM wave and transduces it into data.

Such data are then processed with the most advanced algorithms, even with AI, but transducers data-to-EM and viceversa are mandatory. No bit can be radiated if not converted into an EM wave.

Within this scenario, the knowledge behind transducers data-to-EM (antennas), EM wave propagation and EM wave interaction with real objects is, besides AI and big-data, also strategic and core for a company like Leonardo.

At Leonardo we are aware that all disciplines related to EM (antennas, microwave, etc.) are core to design state of the art radars and sensors. Furthermore, nowadays the need is growing to enhance also the knowledge regarding the interaction of EM waves with real objects, which is mainly aimed at finding strategies aimed to minimise the interaction of such waves with real world objects, thus reducing the probability of detecting such an object by a radar. The first aim of this edition of Polaris is to exploit the state of the art within electromagnetism in Leonardo, especially focusing on those technologies and strategies to “find” and “not be found”.

A further aim of this issue of the Journal is to provide a general overview on measures to keep undetected. These include a number of solutions ranging from the electronic ones for electromagnetic sensors, to those suitable to Optical and InfraRed systems that detect targets basing on the IR and visible radiation they emit, up to camouflage and other physical solutions based on materials and varnishing.

Within that range, active detection capabilities of sensors and their resilience from cyber and electro-magnetic attacks must be traded-off and ensured, while minimizing their detectability.

Some articles in this issue also deal with aspects of the integration of radiating sensors and optical systems into their relevant platforms, which challenges designers and manufacturers as well as requires seamless interaction and effective feedback among providers, customers and end-users on the field.

As we harness the power of controlled electromagnetic manipulation, we are not just advancing technology; we are even rewriting the rules of what's possible in sensing, communication, and data transmission.

The future of applied electromagnetism is here. Its waves are already shaping the world of tomorrow. Let's surf them!

Head of AESA Antenna Engineering
Stefano Mosca

TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

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Invisible Eyes for Information Superiority

Mario Adimari

Leonardo-Corporate

Seeing while not being seen has always been an objective of military and security operations. The growth of technological capabilities, combined with unpredicted geopolitical changes and the complexity of the related challenges they pose, push the need for quickly acquiring greater and better information, from diverse sources. Information superiority increasingly serves the effective use of the military instrument, pushing towards a real multi-domain vision that enables decision makers to make optimum use of digitized information. Such challenges are widely reflected in the market sales volumes, as well as in NATO requirements. Leonardo possesses the multi domain expertise and skills needed to develop solutions suitable to those challenges, thanks to its continuous and significant investments in research and development including disruptive technologies, and thanks to its broad product portfolio that has few equals in the world.

INTRODUCTION

Seeing while not being seen, or communicating while not being heard. These needs have always been the basis of military operations or, more generally, of all operations aimed at security and protection of populations, infrastructure, material and intellectual assets in those situations – if not of real conflict – at least of competition or crisis amongst states and non-state entities. In the world of defence and security, this theme has been addressed as information superiority for some time now (Figure 1).

In the past, scouts, optical and acoustic signals and messengers were already used to acquire and distribute information throughout the command chain. There is no doubt, however, that the dramatic explosion of technological capabilities, the sudden and somehow unexpected geopolitical changes we are witnessing, and the complexity of their related challenges, constantly push to acquire more and more information, from diverse and dispersed sources at an ever greater pace. The so called “speed of relevance” has to be re-considered as well; new threats like hypersonic vehicles, swarming drones, non-kinetic effectors and hybrid warfare events impose faster and more accurate data collection and

processing. In this respect, Artificial Intelligence, High Power Computing and next generation Modelling and Simulation tools will play a key role in achieving real Assisted Decision Making (ADM) command and control systems.

Such challenges are widely reflected in the market volumes. Just to provide an example, NATO is pushing on digital transformation and a significant increase in budget for the acquisition of modern C4ISTAR systems is planned. In 2023, the expenditure on common funds increased by 26.6% on a Year-over-Year basis compared to 2022. Within the NATO 2030 program, the growth rate is expected to gain momentum and the NATO Security Investment Programme (NSIP) funding is expected to reach close to 5.7 BEuros in 2030 [see “[The 2024-2028 Common Funding Resource Plan](#)”].



1– The Information Superiority Challenge

INFORMATION SUPERIORITY: WHY DO WE NEED IT

Information superiority has been declining over the years in various forms yet it is increasingly functional to the effective and efficient use of the military instrument as a whole.

The “network-centric” approach has made communications networks a real force multiplier, thanks to the ability of federating platforms and systems. In recent times, we are moving from that approach towards a true multi-domain operations (MDO) vision that extends the concept to previously unaddressed domains (underwater, space, cyberspace) and focuses on “how” digitalized information can be used more effectively by the decision makers in a coordinated manner. (Figure 2).

It is therefore essential to protect such sources of information from detection and thus from deception or destruction by opposing forces, by making them intrinsically capable of self-protecting and becoming “invisible” by design. Ultimately, acquiring coherent, reliable, complete, timely information, in a safe, undetectable and protected way is now more vital than ever.

The complexity of ISTAR systems and their coexistence on often “constrained” platforms, such as manned and unmanned aircraft, require that the specific design and engineering criteria be defined for both the sensors and the platforms hosting them. (Figure 3). Last, but not least, the optimal integration and mutual compatibility between co-located assets is essential to safeguard the required performances across the manufacturing and deployment processes.



2 – Multi Domain Operations aim at setting timely and effective synchronised effects across the different domains



3 - JSF-F35 A Multirole Stealth Fighter

SENSORS



4 – Integrated Multifunctional Radar and Optical Sensors on an extendible mast for Surveillance and Counter-Drone Missions

Modern sensors operate mainly in the radio frequency ranges, from radio/micro waves (radars, telecommunication and navigation systems) to infrared and visible light (thermal and daylight TV cameras). Radar systems are (with some exceptions) active systems that detect targets based on the echoes they generate when hit by the electromagnetic pulses. Optical and IR cameras are passive systems, as they detect targets based on the IR and visible radiation they emit.

In both cases, the challenges are multiple: we need to increase sensor detection capabilities and resilience against cyber and electro-magnetic attacks, while minimizing visibility and detectability of the platforms that host them (Figure 4).

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Other non-functional requirements, such as low Size Weight and Power (SWaP) factors, integration of radiating and optical systems into the platform structures, and increased local data processing capabilities, complete the framework within which the new generation ISTAR systems must be designed and delivered (Figure 5).

As a matter of fact, low detectability features as part of the so-called “digital continuum”, that is to say, the ability to fuse a variety of sensors into a unique, coherent, available data oriented digital architecture.



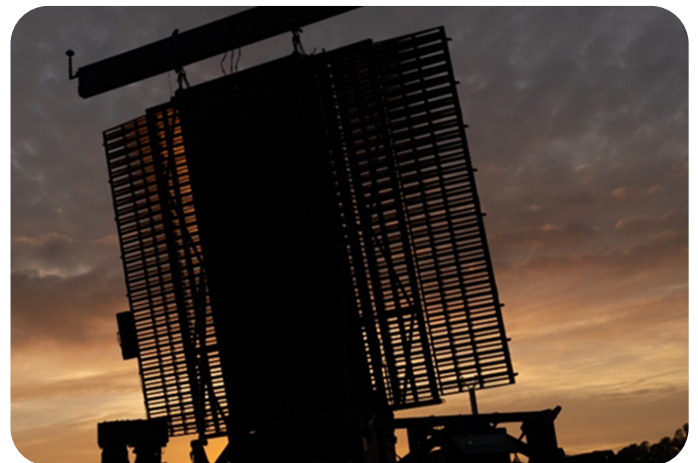
5-Shipborne Conformal Radar Antennas (Square Faces for Primary Radars and toroidal Secondary Radar antenna on top)

LEONARDO'S STANDING

Leonardo is a world leader in the Aerospace, Defence and Security sectors. It is among the few companies that can boast the multi domain expertise and skills needed to develop these solutions, thanks to its continuous and significant investments in research and development and to the ongoing exploitation of so-called disruptive technologies.

Within a broad product portfolio that has few equals in the world, one cannot fail to mention the wide range of surveillance and multifunctional radars based on Active Electronically Scanned Array (AESA) antennas for all types of land, naval and airborne applications. In a necessarily “networked” context, Leonardo's secure communications solutions for mobile and deployable systems are the necessary complement to the creation of ubiquitous, reliable, available ISTAR systems.

Leonardo surveillance radars (known as FADR - Figure 6 and DADR in the NATO taxonomy) are the result of continuous technological evolution and are in fact a standard for air defence systems. The multifunctional radars of the Kronos family for land and military applications can simultaneously perform different functions, from surveillance to target tracking and guidance of interception systems. Electronic counter-counter measures (ECCM) are also implemented.



6-The Fixed Air Defence Radar (FADR) RAT-31/DL

In all cases, Leonardo's radars are able to drive multiple scanning beams in a very precise and directive manner, in order to minimize their own electromagnetic signature (what makes the radars detectable) and optimize their overall energy balance. Active ECCM capabilities against intentional and unintentional interfering signals, as well as the steerable nulling of reception sectors, make the systems suitable for operating in contested and hostile RF environments.

INTEGRATION

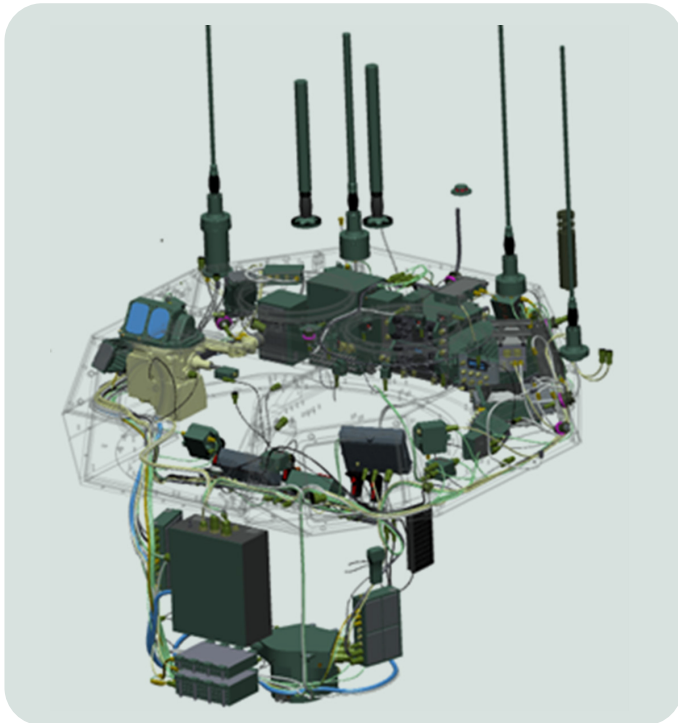
As mentioned above, realizing active but hardly detectable systems such as radars, communications and navigation systems unavoidably pass through the search for the best integration with the hosting platforms. Improper integration invariably leads to degraded performances as compared to the “open field” case, due to electrically susceptible materials and complex shaping. Many combined factors may cause phenomena of

mutual interference between different systems and generate refractive phenomena and spurious emissions that finally increase the RF signature of the digitized platforms (Figure 7).

These matters can only be addressed through: a deep knowledge of electromagnetic physics; the availability of calculation models suitable for the detailed design of radiating systems; the availability of in-house “core” technologies (including in our case the silicon foundry for radar components); and finally the study of electromagnetically passive materials or, on the

contrary, capable of active “thermal” camouflage to hide from IR detection systems.

In this respect, the Global Combat Air Programme (GCAP), which is not just a Sixth-Generation aircraft but a real Multi-Domain System-of-Systems, is among the most notable examples of Leonardo’s endeavours towards the realization of innovative disruptive solutions (Figure 8). At the same time, this programme can be taken as an example of how important it is to set up multinational cooperation programs in any defence domain, thus achieving investments and long term sustainability, interoperability and the highest effectiveness.



7 – Co-located Communications and Self Protection Systems Antennas in an Infantry Armoured Fighting Vehicle Turret



8 – The GCAP A/C will deliver multispectral unconventional Data Fusion thanks to AI

CONCLUSIONS

Can we face an enemy if we don’t see him? And what if we don’t even know what he is capable of? And what’s in his mind, what plans does he want to pursue? All this knowledge is essential for us to protect ourselves and must be acquired silently, discretely, constantly. That’s the challenge. That’s part of the Leonardo’s mission. The articles in this Polaris issue address in detail many of the topics briefly touched on herein. Deep dives will be made on radar antenna technologies, integration issues for collocated radio sources and on new generation camouflage materials.

Mario Adimari: mario.adimari@leonardo.com

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An Overview of AESA Antennas in Leonardo Italy

Matteo Ciattaglia¹, Giuseppe Colangelo¹, Alfonso Farina², Leopoldo Infante¹, Giancarlo Prisco¹

¹Leonardo – Electronics Division, ²Leonardo Consultant

This paper exploits the developments of AESA antennas in Leonardo Italian sites. The AESA are classified with respect to their beamforming architecture, being it either analog or digital. The main characteristics of each AESA are briefly described.

INTRODUCTION

Active Electronically Scanned Arrays (AESAs) are advanced antennas, mostly used in radar and/or communication systems. They consist of thousands of electronically controlled antenna active elements - typically called Transmit/Receive Modules (TRM) - that steer and shape the antenna beam with no need for physically moving the antenna itself.

The TRMs are integrated just behind the antenna array elements (one per element) and they are connected to suitable beamforming network. In comparison to traditional array antennas, its configuration allows for increased sensitivity (low noise figure) and low transmission losses. Furthermore, each TRM allows precise control over the phase and amplitude of the transmitted and received signals, which results in improved beamforming performance and flexibility.

TRMs leverage the solid-state technology, including gallium arsenide (GaAs) and, recently, gallium nitride (GaN) semiconductor materials, which offer higher power efficiency, improved reliability, and extended operational lifetime compared to vacuum tube-based systems. Furthermore, since the power amplification is distributed in transmission over the TRMs, the AESA benefits of graceful degradation, i.e. the failure of a certain amount of TRMs is not critical for the operation of the AESA, as it just works in degraded mode. Finally, since the power amplification is allocated into the TRMs, the AESA beamforming networks don't manage any high RF power thus they can be miniaturized by using the stripline technology, instead of the bulky waveguides as in Passive Electronically Scanned Arrays (PESA).

In the references section, some of the most relevant papers are reported dealing with the AESA developed by Leonardo during the years [1]-[6].

Several systems have been developed during the years in Leonardo that satisfy the radar system capabilities in accordance with the emerging technologies. Ranging from the narrow band waveform monopulse tracking beam to the wideband waveform simultaneous beams, a table can be organized as done in [7]. In Table 1 a ranking of phased array architectures is reported by mapping the systems performances and the number of digital channels considering a test case array made by 1000 radiating elements.

The analogue beamforming networks are used to create antenna beams that can reach a maximum of 32 considering the monopulse tracking beams with the aid of auxiliary channels used for side lobe blanking and anti-jamming adaptive processing.

When the number of digital channels starts to increase up to the 25% of aperture size, the concept of one single beamforming is changed into a two-stage beamforming realized by subarray AESA architectures. Thanks to improvements in key digital technologies, such as the analog-to-digital converters (ADCs), digital-to-analogue converters (DACs), the field-programmable gate arrays (FPGAs), the direct synthesis chips, the digitization of the AESA has become a reality that enables the Digital-Beamforming (DBF) as a key asset.

The digital transmit/receive channels can be configured into subarrays (Figure 1) and the digitization occurs at subarray level. The beamforming is made partially in analogue domain within the subarrays and partially in digital domain among subarrays. Several types of subarrays architectures can be considered: contiguous,

randomized and overlapping, just to mention the most known ones. The choice of the subarray shape is selected in accordance with the radar capabilities expressed in terms of wideband performances and/or the The subarray architecture collapses when the digitization occurs at element-level, thus performing the complete digitization of the array (fully digital AESA) - i.e., each subarray contains a single antenna element. This latter AESA architecture is the best in terms of radar capabilities since there are no constraint among the radiating element and the beamforming that is made into the digital domain only without any mixing with analogue beamforming.

The advantages of digital AESA architectures can be seen also in terms of efficiency, reduced number of critical items involved into the beamforming and the calibration procedure. The calibration tasks are key features for all AESA architectures and become more and more useful when they can be updated as frequently as possible once the systems are deployed in the operative scenario.

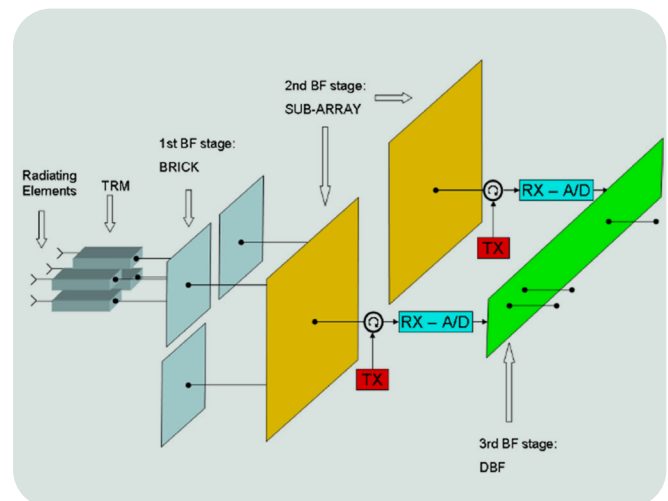
Since the late 90s, Leonardo has been involved, as other radar industries, into the evolution of AESA antennas architectures for radar application by following the steps previously discussed.

By moving from the fully analogic to the digital one, passing throughout the subarray architectures, a survey of the main AESA, developed by Leonardo during the years, is reported and grouped in the following sections:

- in the Analog Beamforming AESA section, we present Leonardo AESAs for naval and land applications, starting from the RAT-31, a 1D AESA, up to the 2D AESA with MFRA and DBR-C/X. This section also presents an airborne radar called EuroMale that has been recently developed including lightweight and compact design
- the Subarray and DBF AESA section includes two examples (MEADS, TMMR) in which the DBF is made at subarray level
- the last section is dedicated to the AESA that is based on complete digitization at element level.

AESA Architecture	n° of digital channels	Digital Processing Level	Radar Capabilities
RF Analogue BFN	1	Total Antenna Pattern	Detect Target
RF Analogue BFN	2-3	Total Antenna Pattern	Above & Angle Estimation
RF Analogue BFN	3-4	Total Antenna Pattern	All of above & Side Lobe Blanking
RF Analogue BFN	5-32	Total Antenna Pattern	All of above & Jammer Nulling
RF Analogue Subarray BFN + Digital Array of Subarray	32-64	Subarray Pattern	All of above & improved dynamic range Wide Instantaneous Bandwidth
RF Analogue Subarray BFN + Digital Array of Subarray	64-256	Subarray Pattern	All of above & multiple beam forming
Element digitized (Fully Digital Array) DBF	1000	Element Pattern	Better everything

Table 1-Relation among AESA architecture and Radar capability for large AESA antennas with 1000 radiating elements



1-The Analog beamforming combines array elements into a number of subarrays that are then digitally beamformed together

ANALOG BEAMFORMING AESA

A family of long range AESA radar has been developed for Fixed (FADR), Deployable (DADR in Figure 2) and Naval (RAN40L) configurations. These L-band AESA are based on a row by column analog beamforming called horizontal beamforming network (H-BFN) and vertical beamforming network (V-BFN). Amplitude and phase, in the air-cooled TRMs of this AESA, are components represented with 6 bits respectively. The antenna front-end is composed of a radiating row in an open array ground plane, to reduce the wind load. Four dedicated independent monopulse VBFNs can be tuned to different elevation beamforming angles and frequencies. The DBR-X and DBR-C systems are based on analogue

AESA architectures and can be considered as the evolution of MFRA (Figure 3). Their building block is a multichannel TRM: 8 TRMs for DBR-X (Figure 4), 4 TRMs for DBR-C, with amplitude and phase control on each channel. The analogue beam forming network provides Sum, Delta Azimuth, Delta Elevation, Four Lobes and Side Lobe Blanking channels.

The EuroMale Radar (Figure 5) is composed by an AESA based on a tile architecture. Each tile includes several TRMs with independent control on amplitude and phase. The selected chipset and the beamforming network design enable the array performances over an extremely wide instantaneous bandwidth, for SAR applications.

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SUBARRAY AND DBF AESA

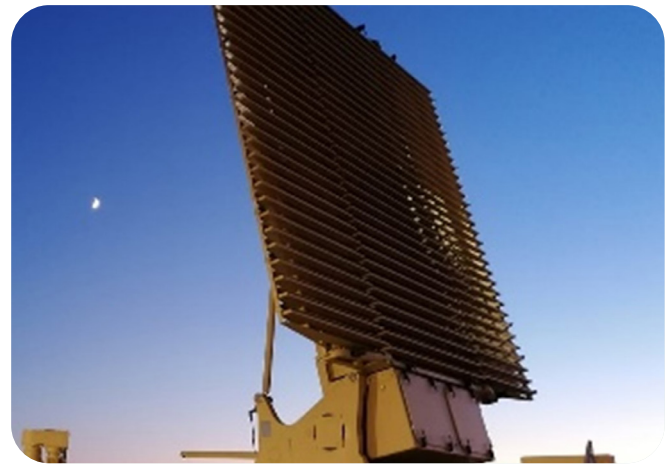
The MEADS system (Figure 6) had been developed jointly with other companies. It is based on a subarray AESA architecture as reported in Figure 1. The subarrays are irregular, i.e. of different sizes, and group together an appropriate number of TRMs, while an RX front end is used to digitise the IF signal forwarded to a DBF to form the receive beams.

The TMMR is a digital radar with direct sampling of RF carrier. Its monopulse antenna belongs to a Digital Beam Forming AESA, whose BFN (Beam Forming Networks) is not realized only with RF analog combiners, as a part of it is also implemented at digital level. The advantage of the DBF AESA, if compared to the analog BFN one, lies in increased reconfigurability, which for a monopulse antenna means increasing the efficiency of the monopulse slope and of the delta beam patterns in general. Obviously, the reconfigurability that can be achieved depends on the number of available digital receiving channels.

The active part of the Front End had been obtained with large reuse of the DBR-C Radar components, in particular the high efficiency GaN Amplifier. It is an antenna with very limited number of radiating elements (<100), which differs from previous products mainly due to its passive thermal dissipation. In particular, by means of a thermal conduction based on "heat pipes", the heat generated internally by the active boards is conveyed to the housing section and is then exchanged by convection with the external air. The reduced number of radiating elements and the possibility of using the radar functions with reduced duty cycle, has enabled to obtain purely passive heat dissipation, thus reducing costs and improving its reliability.

The cost design drivers of the antenna have limited to eight subarrays the number of its receiving channels, in addition to the weight and bulk requirements. These have led to increase the integration between parts of the antenna. To this end, the BFN is assembled directly behind the TRMs with no use of cables. The receiving section of the back-end, which is integrated with the A/D (Analog/Digital) and D/A (Digital/Analog) conversion sections (whose digital board includes the firmware and software radar processing), is (in turn) assembled within the rear housing section and is then directly connected to the Front End. This enables obtaining a very compact radar, whose overall dimensional are 40cmx40cmx30cm, as reported in Figure 8.

A further step on the level of integration is obtained by implementing a RF low-loss solder-less transition into the radiating section, which allows direct connection of the radiating element with the TRM by eliminating connectors and/or welding processes (Figure 9). This provides significant impacts not only on cost and size but also on the RF performance (in particular the Noise Figure and the Radiated Power, Figure 10).



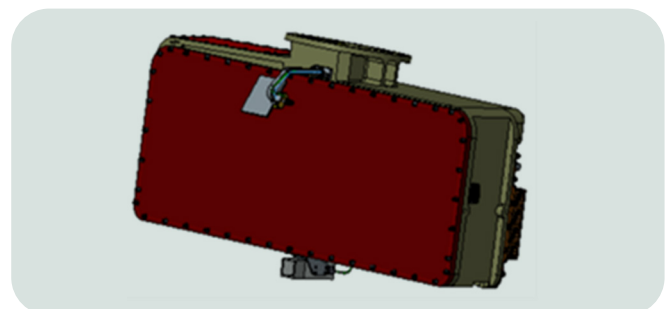
2-Deployable Air Defence Radar



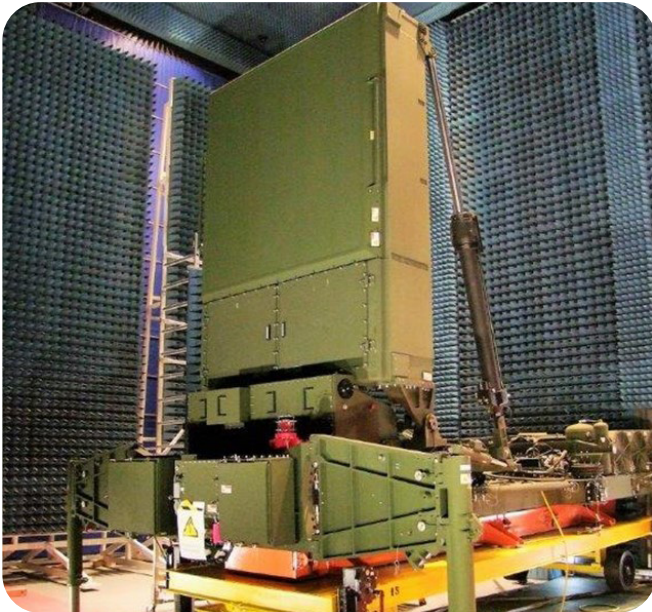
3-MFRA



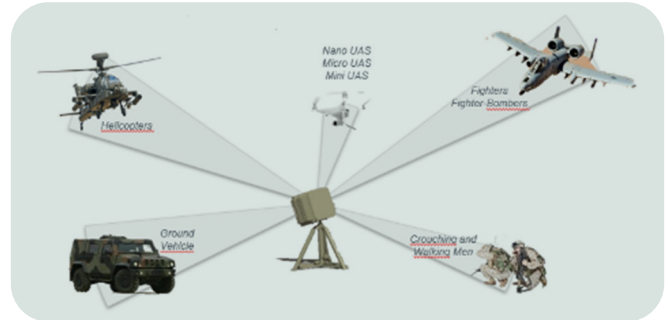
4-DBR-X on ship



5-EuroMale Radar



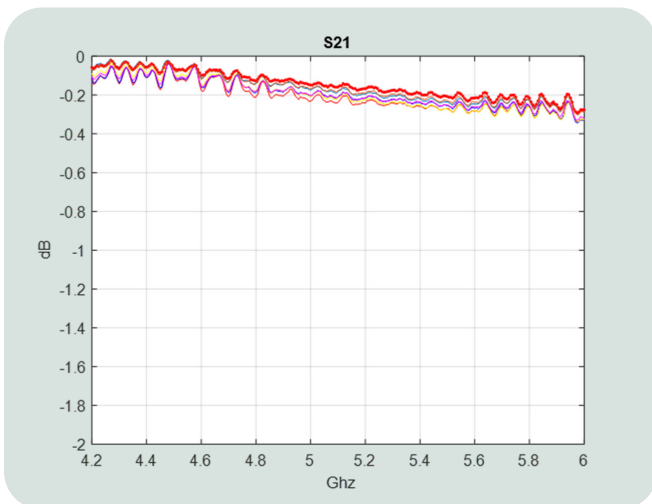
6-MEADS



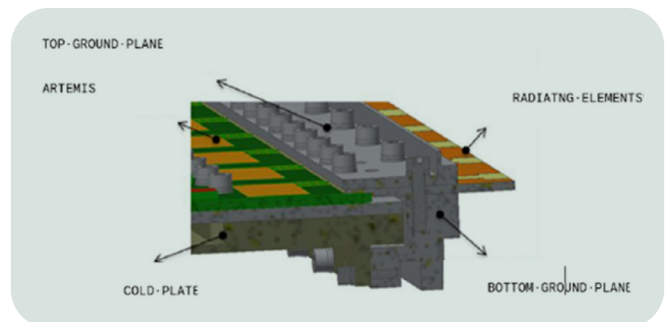
7-Tactical Multi Mission Radar



8-Compact TMMR



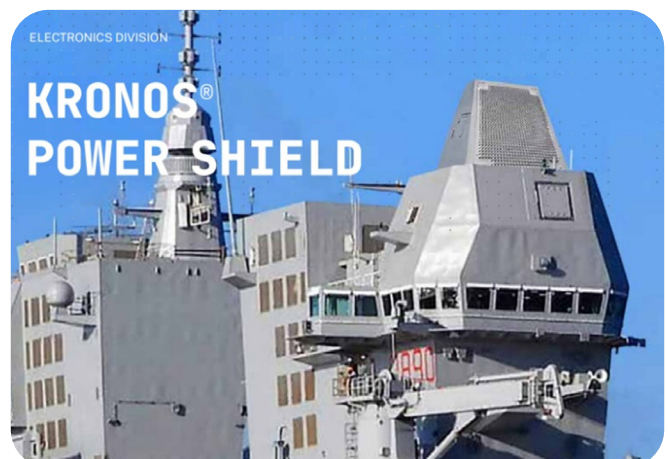
10-RF Transition Radiating Elements-TRMs Losses for eight channels (~0.2 dB)



9-RF Transition Radiating Elements-TRMs

FULLY DIGITAL AESA

The Kronos Powershield multifunctional radar for naval application is depicted in Figure 11 and it is composed by high power TRMs whose digital backend provides direct RF digitization at element level for both the transmitted and the received waveform. The TRM module of this Fully Digital AESA is the Digital Active Tile (DAT) depicted in Figure 12. The DAT is a 4x4 tile located immediately behind the radiating elements, thus performing the digitization of the RF signal at array element level. The received signals are then routed to the DBF located inside the Array Cabinet Unit. The DBF can deploy all its potentiality by creating multiple receiving digital beams and wide band time-controlled beams.



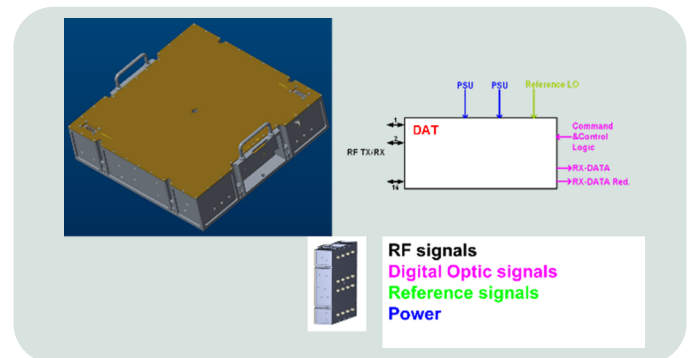
11-Kronos Powershield

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ACRONYMS

ADC	Analog Digital Converter
AESA	Active Electronically Scanned Arrays
BFN	Beamforming Network
C4I	Command Control Communication Computing Intelligence
DADR	Deployable Air Defence Radar
DAT	Digital Active Tile
DBF	Digital-Beamforming
DBR-C	Dual Band Radar – C Band
DBR-X	Dual Band Radar – X Band
EuroMALE	European Medium Altitude Long Endurance
FADR	Fixed Air Defence Radar
FPGA	Field Programmable Gate Array
GaAs	Gallium Arsenide
GaN	Gallium Nitride
H-BFN	Horizontal - Beamforming Network
MEADS	Medium Extended Air Defence System
MFRA	Multifunctional Radar Active
PESA	Passive Electronically Scanned Arrays
RF	Radio Frequency
SAR	Synthetic Aperture Radar
TMMR	Tactical Multi-Mission Radar
TRM	Transmit Receive Module
V-BFN	Vertical - Beamforming Network



12-Digital Active Tile

CONCLUSIONS

In this paper we have reported the progresses in development of the AESA Antennas designed in Leonardo-Italy, also classifying them by analog or digital beamforming. We have reported on our first AESA which is the FADR (Fixed Air Defence Radar), going through the DBR-C and DBR-X, which are the most recent and advanced AESA antennas with analog beamforming. Then we have reported two examples of digital beamforming AESA, i.e. the MEADS and the TMMR. Finally, we have described the KRONOS Powershield which is a fully digital AESA.

Leopoldo Infante: leopoldo.infante@leonardo.com

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AESA Test Revolution

Matteo Ciattaglia¹, Giuseppe Colangelo¹, Alfonso Farina², Leopoldo Infante¹, Giancarlo Prisco¹

¹Leonardo – Electronics Division, ²Leonardo Consultant

In this paper we describe a set of advanced AESA antenna test methods that are used to fully characterize and assess their performances. In particular, we report a method and test bench to characterize full digital AESA, a robotic arm to perform arbitrary near field measurements, a method to accurately evaluate AESA beam pointing error in many directions and, finally, a method and algorithm to obtain a plane wave generator with minimum number of elements.

INTRODUCTION

Precision and efficiency are paramount in the technology domain, in which the Active Electronically Scanned Array (AESA) antennas have emerged as the game-changers. These sophisticated antennas are revolutionizing radar systems across military, aerospace, and civilian sectors, thanks to their ability to electronically steer beams and to adapt to dynamic environments. To exploit the full potential of AESA antennas, is required a deep understanding of their performance characteristics. That is where AESA radar antenna measurements come into play, as they provide the critical tool for assessing, optimizing, and advancing the capabilities of AESA antennas. Conventional near field measurements are usually utilized to characterize the far field (FF) radiation pattern of such antennas, but often other more sophisticated methods and/or hybrid measurement methods are needed. In particular, every time we need to investigate antenna performances that are not directly obtainable from near field methods, or to generate a Plane Wave on the AESA, or to measure an AESA with digital input/output.

In this article, we delve into the realm of advancements in AESA radar antenna measurements, by exploring specific methods that have been developed. The first section of this paper describes a method and test setup for measuring full digital AESA antennas. Next, we describe an innovative near field setup that uses a

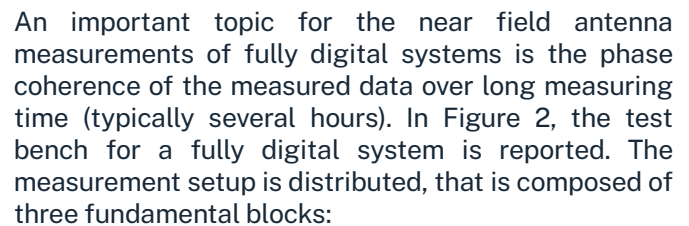
robotic arm. Then, we describe a method for rapidly characterizing the AESA pointing error in many hundreds of different pointing directions, and finally a method to generate a Plane Wave onto the AESA radiating face is reported.

NEXT GENERATION AESA TEST BENCH

Digitization is exponentially growing, which pushes several companies of the Defence sector to develop fully digital AESA for radar systems that go ranging from L-band to X-band operative frequency, exploiting the recent development in the state of the art for Digital-to-Analogue (DAC) or Analogue-to-Digital (ADC) converters. The fully digital AESA provides significant advantages, such as direct sampling of the radiofrequency signal, which implies loss reduction with signal quality improvement, and the Digital Beam-Forming (DBF). This evolution of radar systems has been ongoing at LED (Leonardo Electronic Division) with the design, development and deployment of the Kronos Powershield, a fully digital L-Band radar with more than one thousand radiating elements. These form a completely distributed architecture that is controlled at single element level thanks to a DBF architecture. In Figure 1, a Kronos Powershield deployed inside the antenna test facility is reported.

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MEASUREMENT METHODS



- Even if we'd suppose that all the measurement errors occurring during the near field setup were kept under control, for the fully digital AESA systems the phase coherence among the measuring equipment and AUT is however a key issue.

All these blocks must be locked in phase and properly managed in time, in order to coherently measure the transmitted and received signals from the AUT. Since several Phase-Locked Loop (PLL) components are present in the test bench, a phase drift becomes immediately evident due to frequency locking error, even in the short term. Moreover, other error contributions are dominant in the long term, which are mainly due to temperature variation of all the active devices. In order to mitigate these unbalances, a compensation mechanism must be considered that is based on generating an extra signal used for the phase recovery. Two schemes are presented in the following.

With this methodology, the phase drift compensation for the transmitted-received signals from the AUT and the received-transmitted from the VSA and VSG is resolved in terms of transfer function. By following the test bench block diagram reported in Figure 2. The expression to compute the transmitted signal from the antenna can be written as follows:

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Where $TX'_{AUT}(f)$ is the on-air transmitted signal from the AUT operating in TX-mode, and $RF_{REF}(f)$ is an extra signal generated outside the AUT, but derived from the AUT system clock. For the received signal the computation can be performed through the following equation:

$$RX_{AUT}(f) = \frac{RX'_{AUT}(f)}{VSG_{VSA}(f)} * RF_{REF}(f) \quad (2)$$

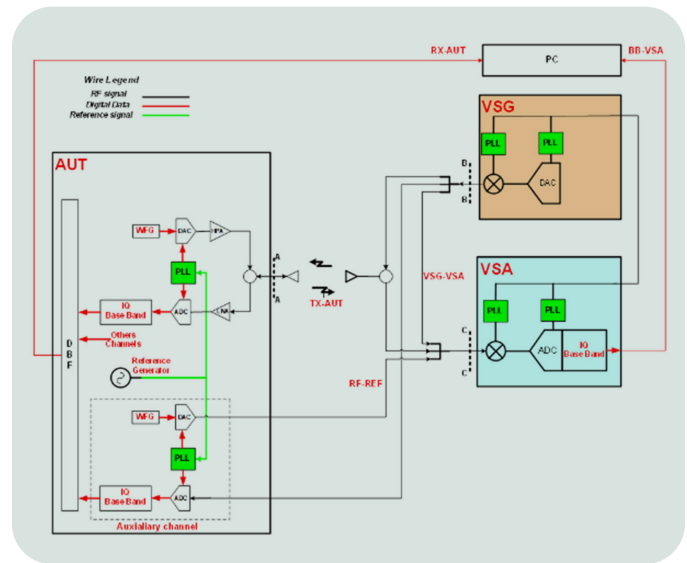
Where $RX'_{AUT}(f)$ are the digital received data for the radiofrequency signal from the AUT, and $VSG_{VSA}(f)$ is the on-air test signal generated by the VSG towards the AUT when in RX-mode.

The Auxiliary channel method

This method is based on a dedicated digital back-end embedded on the AESA systems, as depicted in Figure 3. This channel aim is also for calibration and diagnostic. Using a dedicated firmware task, the auxiliary channel transmits the normalization tone to the VSA and receives the normalization tone from the VSG. This happens by means of a dedicated cabled RF network to avoid any interference with the AUT over the air signal. With this methodology, the computation for the compensated transmitted signal is the same as in (1), but the $RF_{REF}(f)$ is now derived from the auxiliary channel of the system. In the receiving mode the normalization is is:

$$RX_{AUT}(f) = \frac{RX'_{AUT}(f)}{RF_{REF}(f)} \quad (3)$$

Where $RX'_{AUT}(f)$ are the digital received data for the radiofrequency signal from the AUT.



3-Fully digital AESA test bench based on auxiliary channel

ROBOTIC NEAR FIELD

Near field scanning techniques have become very common since the '80s for testing phased arrays. Usually, planar arrays are tested with a Planar Near Field (PNF) setup, while phased arrays featuring more complex shapes could require Cylindrical Near Field (CNF) or Spherical Near Field (SNF). Each one of such three types of near field setups requires its own dedicated mechanical positioner that is strictly related to the shape of the scanning surface (planar, cylindrical, or spherical). Thus, with traditional scanners it would not be possible to execute measurements with only one setup on all of the three different types of scan surface. With the advent of sophisticated anthropomorphic robotic arms, the latter paradigm has changed, as with a single anthropomorphic arm we can move the probe on a surface with arbitrary shapes. For small antennas on which the arm geometry allows to move the probe in the far field of the antenna, this setup allows to perform also a far field measurement. Finally, the ability to position a measurement probe in any position and with

different orientation with respect to the Antenna Under Test (AUT), paves the way to new measurement capabilities. The measurement setup for antennas based on an anthropomorphic robotic arm is installed in the Leonardo AESA Test Lab in Rome plant. A video demonstration can be seen in [\[5\]](#).

The following section reports some measurement results that have been obtained during the experimental activities.



4-Near field setup with anthropomorphic robotic arm

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The Measurement Setup

The anthropomorphic arm selected for this setup must feature at least six or more Degrees Of Freedom (DOF). The arm allows to move the near field probe on scanning surfaces with arbitrary geometries with the only constraint of respecting the ergonomics of movements of the robotic arm.

The complete near field measurement setup depicted in Figure 4 is composed by:

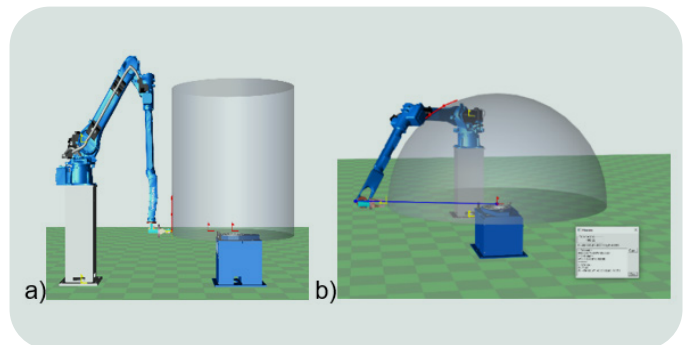
- Anthropomorphic six-axes Robotic Arm model Motoman GP120RL made by Yaskawa. It is a shelf-mounted robot with extraordinary reach of over 4 m, which is capable to handle up to 120 kg. The work envelope of the robot is expanded both downwards and upwards with the aid of a pedestal that is fixed to the ground. The wrist joint is designed for high movements of inertia with a Yaskawa's patented vibration control system enabling fast and precise positioning. The repositioning error is ± 0.05 mm (ISO9283), the overall weight of the robot in itself is 1830 kg.
- Turning Table model VMS-1000 made by Yaskawa, featuring maximum payload of 1000 kg. The sleeve bearings make enable endless rotation. RF Cables and signals devoted to the AUT are fed into a fixture from the hole in the center of the fixture disc. A suitable cable winding recovery system avoids the use of rotary joint. The repositioning angular error is $\pm 0.005^\circ$.
- Vector Network Analyzer (VNA) to perform the RF measurement.
- Test antenna probe mounted on the arm.
- Robot Controller YRC1000 with a programming pendant made by Yaskawa enables to command and control the six axes of the robot, together with the azimuth angle of the turning table. To reach a measuring point in front of the AUT, an inverse kinematics analysis trajectory plan is made by the controller.

The built-in software runs on a dedicated PC. It calculates the rotation matrices for the six axes by satisfying joint configurations. Three canonical configurations have been tested:

- Planar Scanning: obtained by fixing the azimuth turning table and performing a raster scan with the six axes robot achieving a rectangular measuring plane of 4100 mm along the horizontal plane and 1800 mm along the vertical plane;
- Cylindrical Scanning (Figure 6a): the robot moves over a central column in front of the AUT achieving 1800 mm of vertical scanning and 360° scanning angle along the azimuth;
- Spherical Scanning (Figure 6b): the robot moves over a circle arc spanning an elevation angular sector greater than $\pm 90^\circ$ thus achieving at least a half



5 - Alignment setup with camera



6-a) Cylindrical scanning b) Spherical scanning

spherical cap over the AUT. In this latter configuration, the AUT boresight is pointing over the top of the anechoic chamber.

- In order for the scan surface to be aligned to the AUT, an alignment procedure is required. The alignment setup is depicted in Figure 5 and it is composed by the following equipment:
- High precision laser range finder model LK-G407 made by Keyence;
- High resolution VGA CMOS camera 640x480 Pixelmodel CA-035M made by Keyence;
- Reference points over the AUT frame.

The alignment procedure starts by locating the reference point over the AUT by using the high resolution camera. Once the pointing laser is on the target, the distance is acquired by using the high precision laser range finder. By repeating the procedure for each reference point, we can find the best frame in the Robot coordinates approximating the required scan plane. To summarize, the anthropomorphic robotic arm for near field scanning offers several advantages, including:

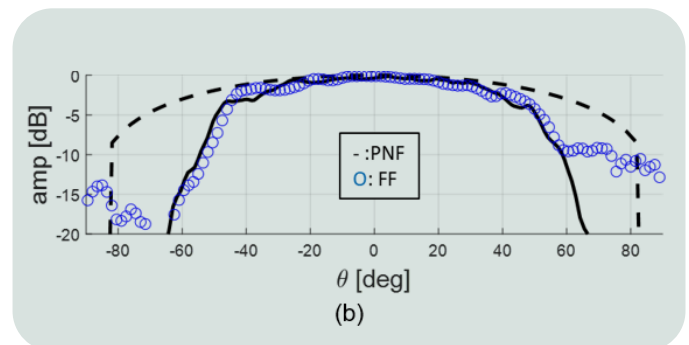
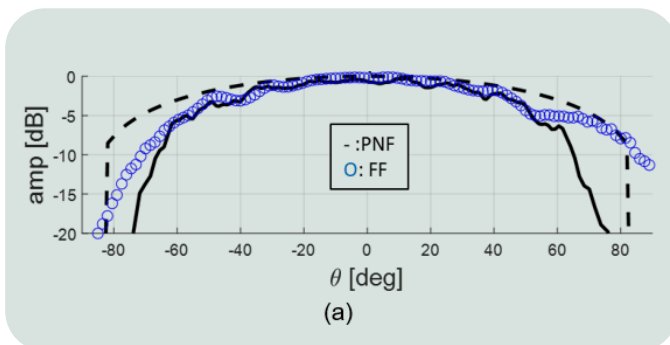
- Flexibility: The robotic arm can be programmed to follow complex scanning trajectories, allowing for

versatile measurement configurations.

- Precision: Modern robotic arms offer high precision and repeatability, ensuring accurate positioning of the antenna under test.
- Automation: Once programmed, the robotic arm can perform scanning tasks autonomously, reducing the need for manual intervention.
- Adaptability: The same robotic arm can be used for scanning different types of antennas, making it a versatile tool for antenna testing and characterization.
- Cost: For PNF, the cost of the robotic arm, with respect to a scanner needed to cover an equal surface is at least 50% lower.
- Probe payload: the payload of the selected arm is 120 kg, this allows to install a variety of probes.

Furthermore, if the AUT size (D) is such that its far field distance ($2D^2/\lambda$) is shorter than maximum probe to AUT distance, the setup can be used also as a traditional far field range. Of course, if the AUT far field distance exceeds the maximum probe to the AUT distance, the far field measurement is not anymore feasible. This problem can be overcome by taking advantage of the high payload of the arm, which allows to install on the arm not only a simple and light measurement probe, but also a phased array which can be configured to act as a Plane Wave Generator (PWG).

Positioning the PWG at the proper distance from the AUT allows to measure its far field pattern, thus acting as a Compact Antenna Test Range (CATR). Finally, if the AUT is for radar application, by connecting the radar processing hardware to the AUT, the PWG can be used as a target echo simulator for Hardware In the Loop measurements. Many tests have been done in order to characterize the test setup depicted in the previous section. The test reported in this section is based on the radiation pattern measurement of a small antenna by using the setup in two different configurations, one for PNF and the other far field measurement. Figure 7(a) and Figure 7(b) show measurement results with both those measurement configurations. As it can be seen, there is good agreement between measurements within an angle of about 120° around boresight. Outside this angular range, the PNF measurement goes down for scan plane truncation issues, while the far field measurement continues to measure correctly the AUT radiation pattern. Of course, with respect to the far field measurement, the PNF provides the AUT radiation pattern in a wide solid angle centred on the AUT boresight (the limits of such solid angle are related to the PNF scan plane size), not only in the plane depicted in Figure 7. Nevertheless, the PNF has a longer measurement time with respect to the far field measurement.



7-(a) Comparison between PNF and FF measurements at frequency F1;
(b) Comparison between PNF and FF measurements at frequency F2

ROBOTIC NEAR FIELD

Some types of mono-pulse AESA antennas require an accurate beam pointing capability, thus they must be characterized in terms of pointing error over the entire scan coverage and with a reduced angular pitch in order to allow the correct processing of quantities also linked to error variations.

The pointing error of a monopulse AESA is normally obtained by processing the near field measurements passing through the computation of the far fields of the sum and difference patterns. This technique, applied to a large number of antenna configurations (phase states for each element of the array), would require the execution of a very large number of acquisitions involving Near Field measurements over a large area, in order to allow correct calculation of the sum/ difference in the far field zone, in each of the pointing directions.

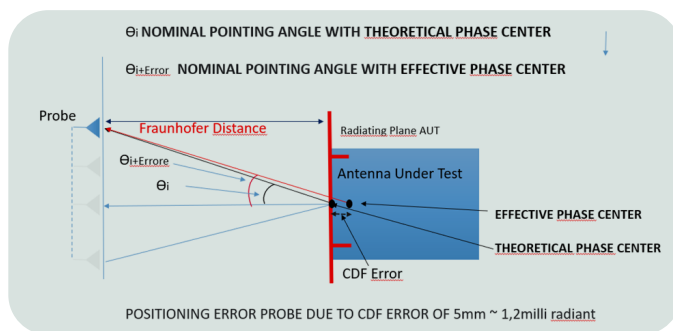
However, for antennas whose far field distance is relatively short, i.e. easily reproducible in an indoor environment, it is possible to adopt alternative techniques that allow significantly reducing the test execution times. The proposed technique consists in measuring the field patterns radiated or received by the AUT, with a probe moved by a scanner which is positioned in the far field zone with respect to the AUT.

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By using the same facility as canonical near field measurements, the probe is positioned from time to time (through a scanner) in order to carry out acquisitions corresponding to each of the nominal pointing directions in which it has been decided to electronically set up the theoretical nulls of the AESA difference patterns. To obtain complete characterization of the error throughout the scanning area, it is necessary to repeat the procedure for each pointing direction, by synchronizing the electronic pointing of the beams with the mechanical positioning of the scanner probe. To make the procedure described effective for the purpose of determining the error pointing parameters, it is necessary to:

- Determine the correct Phase Center (PC) of the AUT in the preliminary phase;
- Verify the correctness of the approach described (analytically/ experimentally).



8-Position error probe due to CDF error

Phase centre evaluation

In the antenna design theory, PC is the point from which the electromagnetic radiation spreads spherically outward, whose phase of the signal is equal at any point on the sphere. The PC position does not always coincide with the geometric centre of the antenna, due to physical reasons. When the probe used for the far field measurement is located near the far field zone of the AUT, any relatively small error in the evaluation of the PC leads to an alignment error comparable with the pointing error that one wants to calculate.

In far field measurements carried out at a distance of several kilometres (and much longer than the Fraunhofer zone) the indeterminacy between the geometric centre and the actual phase centre does not cause any detectable misalignments. The Figure 8 shows an example of how an error between theoretical and actual PC of a few millimeters over far field distances of a few meters causes a misalignment of the order of a milli-radian. The estimate of the phase centre is obtained through post-processing of a canonical NF measurement for a fixed pointing.

The estimated of the antenna's PC is verified experimentally by moving the beam in the specular directions, once only in two directions $\pm AZ$ and then only in $\pm EL$ and by verifying the symmetry of the respective pointing obtained.

Verification approach

The procedure may not prove to be strictly valid, as for the evaluation of an antenna parameter in a specific direction (u_0, v_0) (pointing error), it is required varying the antenna configuration itself. Actually, the rigorous FF measurement procedure would require for each nominal pointing to move the probe around that direction, to identify the nulls of the difference patterns without altering the antenna configuration. Obviously, this procedure presents unavoidable increases in time, as well as considerable implementation difficulties. To validate the approach, a complete model of the entire AESA antenna has been built, capable of faithfully reproducing the Full-Wave Electromagnetic behaviour of the radiating section, by integrating it with statistical parts determined by the active sections (phase and amplitude errors due to amplifiers, phase shifter, temperature gradients inside the antenna, couplings, error due to plank positioning, measurement uncertainties). Based on the antenna model, it can be verified that the technique adopted provides an extremely small measurement error that is kept within the tolerance range.

Causes of error included in the model:

- EF Element factor array from FULL Wave Analysis;
- Quantization due to the finite number of bits of the phase shifter;
- Phase actuation error, due:
 - to the different phase shifters components
 - to the different phase states;
- positioning radiating elements;
- Gradients of temperatures inside of antennas;
- Calibration of TRM;
- Calibration of Subarrays;
- Coupling inside TRM and Subarrays.

Comparison of evaluation methods

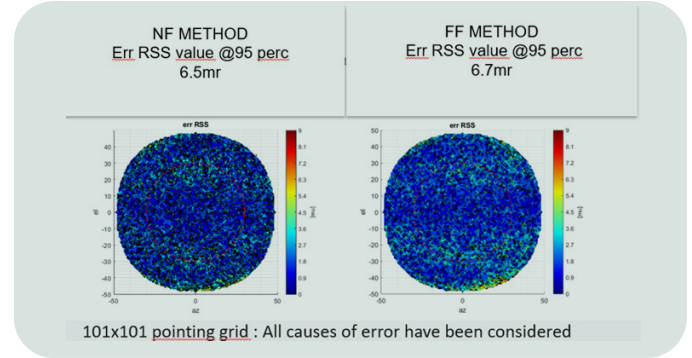
NF method: once the nominal pointing is fixed in the direction (u0,v0), we search for the null of the delta h/v pattern
 FF method implemented: once the observation point is fixed at (u0,v0), we search for the null of the h/v cut obtained by varying the beam steering around (u0,v0):

- $err_{RSS} = \sqrt{(err_h)^2 + (err_v)^2}$
- $err_h = achievedh - desiredh$
- $err_v = achieved - demanded$

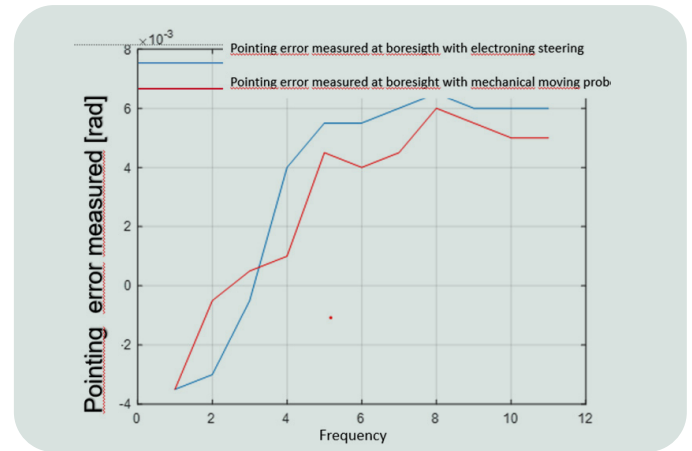
The implemented approach has been also validated experimentally by comparing the pointing errors obtained from the measurement with electronic scanning and those obtained through a mechanical movement of the probe for some pointing directions. The direct FF technique with electronic scanning of the beam on an extended grid of 101x101 points in azimuth and elevation for a set of 10 frequencies allows the measurement of pointing errors to be obtained in approximately 20 hours. With the same grid and set of frequencies, the measurement time required with traditional NF-FF techniques or with direct FF measurements involving probe movements would be longer than 250h.

PLANE WAVE GENERATOR

In order to focus a plane wave across the Antenna Under Test (AUT) aperture by using arrays [1]-[4] for generating an incoming signal from a target in far field zone, different solution strategies have been proposed in the literature. The solution methods can be subdivided in two main class. The first one contains those solution methods that adopt global optimization algorithms to determine the optimal array element locations and weighting coefficients, such that the synthesized architecture is able to radiate plane waves in the near zone. In this case, a Genetic Algorithm based procedure is adopted [1]. To the second class belong all those solution methods that feature no need for global optimization schemes (convex programming). In [4] a new flexible synthesis method has been introduced. Once the AUT dimensions have been fixed, the synthesis method in [4] allows to determine the minimum PWG dimension, the minimum number of radiating elements in the probe antenna, their locations and weights, such to achieve the desired plane wave distribution in the near zone. In order to assess capabilities of the synthesis procedure, its robustness with respect to the mutual coupling effects has been investigated by means of a full-wave analysis.



9-Err RSS point error with NF Method and FF Method



10-Bore-sight pointing error with direct measurement of FF: comparison of electronic scanning of the beam and movement of the probe

The case of one dimensional AUT

In this section, we first briefly recall the synthesis strategy recently proposed for 1D AUT. The overall method is divided into three main steps: in the first one the problem of determining the minimum PWG dimension such to achieve the plane wave distribution, is discussed. Then, in the second step, the procedure for determining the optimal element locations is discussed. The last step deals with determination of the optimal weighting coefficients.

Estimating the minimum PWG dimension

The minimum PWG dimension enabling to achieve the plane wave distribution in the near field region is determined by introducing two quality factors that allow to quantify how the PWG radiated field across the AUT, approximates the plane wave distribution.

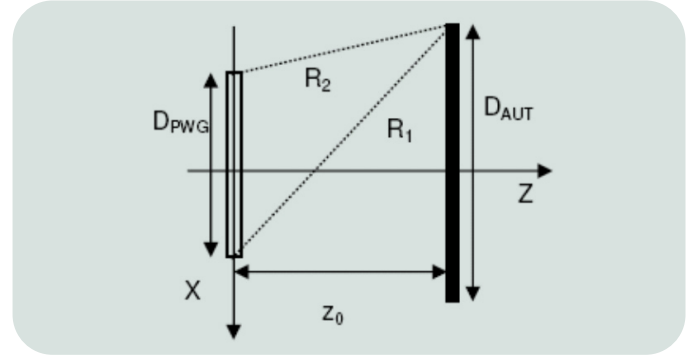
The quality factors are [1]-[4]:

$$\varepsilon_p = \frac{2\pi - |\delta_{MAX} - \delta_{MIN}|}{2\pi}, \quad \varepsilon_a = \frac{a_{MAX}}{a_{MIN}} \quad (4)$$

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where δ_{MAX} and δ_{MIN} are the maximum and minimum phase values of the synthesized field on the AUT, respectively, a_{MAX} and a_{MIN} the maximum and minimum field amplitude values across the AUT aperture, respectively. Once the quality factors are introduced, it is possible to estimate DPWG. In particular, as shown in [4], a PWG dimension equal to the AUT allows to achieve a plane wave distribution across the AUT with both quality factors larger than 0.8, which is generally enough in actual applications [1][2].



11-Reference 1D geometry. Linear AUT

Optimal PWG synthesis via sparse arrays

Once the PWG dimension has been estimated, we have to determine the minimum number of radiating elements and their locations able to fulfill the design specifications. According to [3]-[4], this number is only determined by the size and shape of the PWG, and is given by:

$$N \cong \text{int} \left[\frac{2}{\lambda} (R_1 - R_2) \right] + 1 \quad (5)$$

wherein R_1 and R_2 are reported in Figure 11, while int denotes the integer part of its argument [4]. The locations of the PWG sources are then determined as follows [4]:

$$\begin{aligned} \varphi_{\text{MAX}} &= \tan^{-1} \left[\frac{D_{\text{AUT}}}{2z_0} \right] \\ \varphi_n &= -\varphi_{\text{MAX}} + (n-1) \frac{2\varphi_{\text{MAX}}}{N-1} \quad n = 1, \dots, N \\ x_n &= z_0 \tan(\varphi_n), z_n = z_0 \end{aligned} \quad (6)$$

wherein N is chosen according to (5). The z -coordinate is now equal to $z_0=0$. Equations (6) also allows to fix the positions of the sampling points on the AUT, i.e. the location and number of sampling points wherein the synthesis constraints will be considered. Also note the sampling points in (6), uniformly spaced in the φ -coordinate, are not-uniformly spaced both on the AUT aperture (non-uniform constraints locations) and on the PWG. This means that the corresponding PWG is a sparse array, with reduced number of radiated elements. In such arrays, the minimum distance between sources is greater than λ , with a reduced interaction between sources.

Optimal synthesis of the complex weights

The synthesis of the optimal weighting coefficients for the PWG array has been solved by exploiting a Singular Values Decomposition procedure, as detailed in [4]. In particular, due to linearity of the constraints with respect to the unknowns, the synthesis problem is recast to the solution of a linear system with $2N$ (independent) equations and $2N$ (real) unknowns, ([4] for more details). Notably, its solution, unlike existing procedures, does not require the adoption of any global optimization schemes, thus saving time and computational requirements.

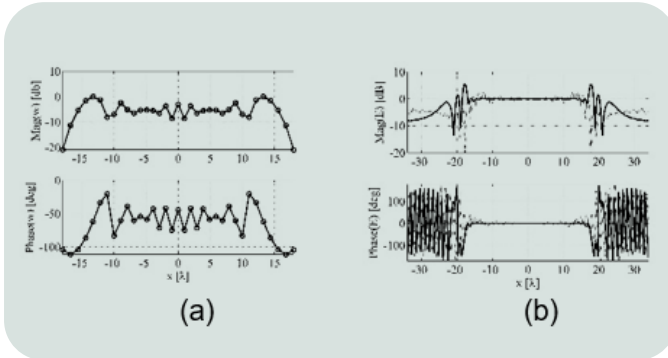
As an example, let $D_{\text{AUT}} = 34\lambda$. Then, by requiring $\epsilon_a = \epsilon_p = 0.88$, and by assuming $z_0 = 34\lambda$, we get $D_{\text{PWG}} = 34\lambda$.

Now, by using (5), an estimation of the minimum number of required PWG sources is achieved, while their locations are achieved with (6). In this example, $N = 33$, whereas, if a $\lambda/2$ uniformly spaced array is adopted, the number of elements should be 69. Now, 0.9λ is the minimum distance between the sources.

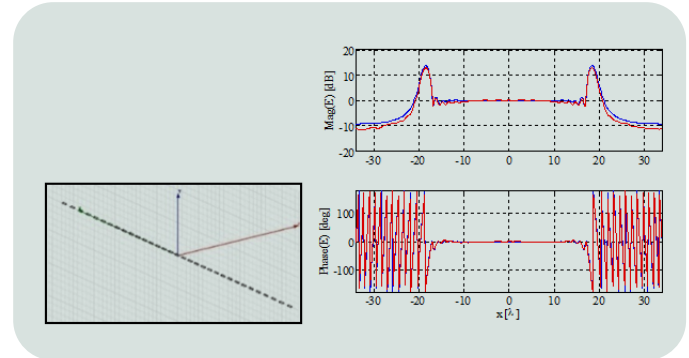
Notably, the above method allows to design non-redundant PWG array with a reduced number of sources with respect to other techniques, thus saving costs and computational time.

The array weights, after modelling the PWG sources as horn antennas, are achieved by exploiting the procedure of Sect. II.C.

The results are shown in Figure 12, wherein the PWG near zone pattern one achieves by adopting a number of sources lower than (5) is also reported. As it can be seen, the plane wave across the AUT is well approximated by the PWG near field zone, thus emphasizing capabilities and usefulness of the developed tools.



12-(a) Synthesized optimal weights. (b) Amplitude and phase of the PWG field radiated across the AUT, when 33 (solid line) and 31 (dotted line) sources are considered



13-PWG array geometry (left) and the actual field distribution radiated by the PWG across the AUT (right)

Mutual coupling effects

In this Section, the performances one achieves in presence of real sources acting as radiating elements of the synthesized PWG array are discussed. In particular, a real element factor has been simulated by means of a full-wave analysis performed with a commercial electromagnetic simulator (Ansys HFSS). A PWG consisting of $N=33$ elements, with a rectangular aperture acting as the radiating element, has been simulated.

The PWG array geometry and the field distribution radiated by the PWG across the AUT region are reported in Figure 13. The full-wave results well fit the analytical ones, thus confirming robustness and accuracy of the proposed technique. Moreover, the results also confirm that, due to the sparse nature of the synthesized PWG, the performances actually achieved are not affected from mutual coupling, because the average distance between the two adjacent sources is greater than λ .

CONCLUSIONS

In this paper we have reported several advanced test techniques to characterize and assess the performances of advanced AESA antennas. For full digital AESA we have reported a test-bench that is able to measure the S_{ij} (between the AESA port and the near field probe) where the input is RF and the output is digital data, or vice versa. Furthermore, we reported an extremely flexible near field set-up that uses a robotic arm to move the near field probe, thus allowing potentially to have a scanning surface of any arbitrary shape.

For AESA, whose beam pointing error is very demanding, we reported a method to perform quickly an extensive characterization of the beam pointing errors. Finally, we have described a method to develop an array to realize a plane wave generator with the minimum number of array elements. The PWG is useful when we need to emulate a far field signal on an AESA under test, in order to test it in a real-world scenario.

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Co-siting Analyses on Large Structures in Leonardo

Gabriella Bernardi, Giancarlo Prisco, Maurizio Felaco, Elio Petrillo, Stefano Mosca
Leonardo - Electronics Division

It is well known that the most of the real-world electromagnetic problems cannot be solved via analytical approaches, due to the complexity of geometries and materials of the actual devices. The Computational electromagnetic modelling process uses computers to model interactions among electromagnetic fields, physical objects and the environment. This approach enables to overcome the inability to derive closed form solutions of Maxwell's equations under various constitutive relations of media, and boundary conditions. This makes the Computational Electromagnetics (CEM) crucial for Leonardo that is at the forefront of complex system design, as systems it delivers are often installed on electrically large structures such as military vehicles, aircraft and ships, on which radar, satellite and other communication systems have to be located in the same places. In such cases, the CEM is the only way to perform EMI/EMC analyses that can be challenging, due to a number of reasons, first of all the difficulty of finding the best trade-off between the computational costs of accurate models and the reliability of simulated results. This paper provides an informative overview of the main CEM techniques and their applications in the Electronics Division of Leonardo company.

INTRODUCTION: the problem at hand

Upon the most recent decades, technological development has led to the creation of complex systems that are characterized by the integration of an ever-increasing number of subsystems. The unavoidable consequence of this trend is the need for a large number of electromagnetic sensors that must coexist in limited spaces, i.e. the need to avoid strong electromagnetic interferences that could affect the performance of the whole system. This requires predicting the electromagnetic behaviour of all the involved sensors and their electromagnetic interactions, since the design stage. Military vehicles, aircraft or naval ships are examples of such complex systems of systems that employ armament and electromagnetic (EM) sensors, such as radars and other systems, many of which use antennas. Thus, such EM environments are dense, as they contain a great number of EM sensors used for communication, navigation, detection, electronic warfare, homing, direction finding, and tracking purposes [\[1\]\[2\]](#). Each of those sensors is required to offer

certain capabilities specific to the contexts and missions. Therefore, the increasingly challenging goal is to not degrade their performances. The occurrence of electromagnetic interference (EMI) depends on the unwanted amount of power of the transmitted signal that goes through a receiving system. If the magnitude of the signal exceeds the compression point for those devices, a high-power signal may saturate one or more receiving system devices, such as mixers or amplifiers [\[3\]\[4\]](#). This could prevent from receiving the desired signals through reduction in the signal-to-noise ratio. In case this affects one or more systems, they could be made unavailable to operations. The consequence of not addressing EMI issues timely can be potentially catastrophic, depending on the given mission and context. Thus, it is needed to allocate sensors in such a way that the various subsystems coexist and perform their desired operations without causing mutual interference or adversely affecting each other's operations.

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This is among the main challenges the EM engineers must face since the concept phase of system design. In this paper, emphasis is placed upon the importance of discovering and alleviating EMI problems in the concept phase of system design, when configuration choices are still very fluid. Thus, should electromagnetic compatibility (EMC) problems be discovered, it is necessary for the designers to modify the design until acceptable level of EMC is achieved. It becomes clear the importance, for a company like Leonardo, of having an internal area that deals with these problems using state-of-the-art knowledge and tools that are as innovative as possible.

The EMI-EMC problem features very complex and nonlinear nature. This makes it difficult for system designers to accurately predict magnitude of the interference problems among antenna systems. Past EMI data from similar systems or comparable EM sensors on different platforms may be useful, but it is unlikely to be sufficient. This is due to the inevitable differences between the configurations of the different designs, as the parameters of the antenna systems involved could also differ for similar systems and could therefore significantly change the EM environment.

This problem can be addressed by using a suitable Computational Electromagnetic (CEM) tool to model the platform and EM environment and then to simulate the EM interactions of the antennas against each other on the concept-level definition in order to quantify the interference between them. Leonardo adopts some commercial software to run such numerical simulative study, like HFSS (Ansys) [5], CST Microwave Studio (Dassault Systems) [6] or Feko (Altair) [7].

After achieving the required representations in the simulation environment for each antenna, the antenna models are simulated on the platform model to determine the coupling between them, using the most appropriate approach. For example, if the predicted interference between two antennas is above a given tolerable level, the antennas are relocated on the platform, consistently with other design constraints, and the simulation is then launched again. This process is repeated until a configuration is achieved, whose severity of interference is reduced to a level that is acceptable to the designers.

In the present paper, we show how Leonardo deals with evaluating the performance and the coupling that exists among disparate antennas that are installed close to each other (co-sited) on-board any platform.

Quantitative expressions are quoted when needed, but the mere mathematical aspects of derivations are not provided, as these are discussed in general terms in the scientific literature. In particular, Section 2 (*Antenna Coupling On The Platform*) describes the inter-antenna coupling in the presence of a platform and recalls some mathematical expressions useful for determining coupling parameters from field solutions, whether the latter are exact or approximate. Section 3 (*CEM Techniques in Leonardo*) provides a reminder of just enough of complexity of the three fundamental full-wave CEM methods [8], to understand why there is a limit to the electrical size of the platform and of the antenna structures. Section 3 also summarizes how the range of application of the CEM modelling has been extended beyond what could be achieved via the full-wave route through the use of hybrid CEM asymptotic methods. All these concepts are shown by means of the results of some examples implemented in Leonardo. Section 4 (*Conclusions*) concludes the paper.

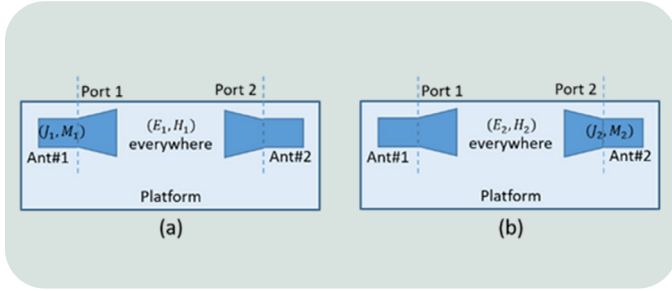
ANTENNA COUPLING ON THE PLATFORM

In this Section we consider the simple case of two antennas (antennas #1 and #2) in the presence of a platform, that form the “original environment.” Two situations are considered [21].

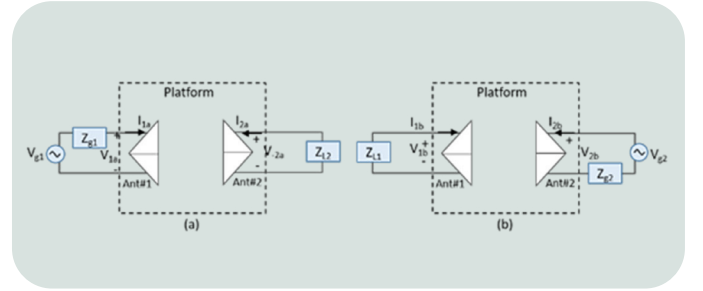
In situation (a), shown in Figure 1(a), the impressed source ($J_r M_1$) is turned on (it generates incident fields (E_{inc1}, H_{inc1}) everywhere), while ($J_r M_2$) is removed. We can also say that in this situation, antenna#1 is transmitting and antenna#2 is receiving, i.e. some of the power arriving at its port can be extracted to some load. We can say that port#1 is excited and has a load connected to port#2. The total fields everywhere are (E_r, H_r), due to the impressed source in the presence of the physical structures of both antennas, as well as of the platform. Voltages and currents can be defined at both ports, in

terms of the fields there. These will be denoted by V_{1a} and I_{1a} at port#1, and by V_{2a} and I_{2a} at port#2. Those are the circuit quantities determined from fields (E_r, H_r). We can associate (V_{1a}, I_{1a}) and (E_r, H_r).

In situation (b), shown in Figure 1(b), we keep unchanged the structure of the two antennas and the platform. Nothing is removed. The environment remains the same. However, the impressed source ($J_r M_2$) is turned on, it generates incident fields (E_{inc2}, H_{inc2}) everywhere and ($J_r M_1$) is removed. Antenna#2 is transmitting, and antenna#1 is receiving. The total fields everywhere are now (E_r, H_r). The circuit quantities at the two ports, once again defined in terms of the fields there, will be called V_{1b} and I_{1b} at port#1, while at port#2, in this situation #b, the corresponding quantities are V_{2b} and I_{2b} .



1-General sketch of two antennas in presence of a platform



2-Transmitters and receivers (generators and loads) attached to the ports in the circuit model in terms of their Thevenin equivalent circuits

These are the circuit quantities determined from fields (E_2, H_2) that are generated when the impressed source (J_2, M_2) is turned on and (J_1, M_1) off. We can associate (V_{2a}, I_{2a}) and (E_2, H_2) . We imagine initially that a full-wave CEM solution of the complete electromagnetic formulation described above can be found. The two situations are schematically depicted in Figure 2, where we view all transmitters and receivers (generators and loads) attached to the ports in the circuit model, in terms of their Thevenin equivalent circuits. It is always assumed that $Z_{L1} = Z_{g1}$ and $Z_{L2} = Z_{g2}$. We note that $V_{2a} = -I_{2a} Z_{L2}$ and $V_{1b} = -I_{1b} Z_{L1}$ are in each of the two situations, while the minus signs are arising due to the reference direction assumed for the currents. From the circuit theory, we know the power supplied to port #1:

$$P_{t1} = \frac{1}{2} * \text{Re}\{V_{1a} * I_{1a}\}. \quad (1)$$

$$Z_{in1} = \frac{V_{1a}}{I_{1a}}, \quad Z_{in2} = \frac{V_{2b}}{I_{2b}}. \quad (2)$$

$$P_{r2} = \frac{1}{2} * \text{Re}\{V_{2a} * (-I_{2a})\} = \frac{1}{2} * \frac{|V_{2a}^{OC}|^2}{|Z_{in2} + Z_{L2}|^2} * \text{Re}\{Z_{L2}\}. \quad (3)$$

P_{r2} reaches its maximum under conjugate match condition $Z_{L2} = (Z_{in2})^*$, and so under such matched conditions:

$$P_{r2} = \frac{|V_{2a}^{OC}|^2}{8 * \text{Re}\{Z_{L2}\}} \quad (4)$$

The power coupling between the antennas is P_{r2} / P_{t1} . At this point, we recognize that the input impedances are dependent on everything in their environment, namely the platform and antennas structures. For example, in many co-siting situations, Z_{in1} is not significantly influenced by the presence of the platform or by the structure of antenna#2. Thus, in practice, this assumption is often made for Z_{in1} and Z_{in2} . Clearly, such an assumption would prove itself inaccurate if the antennas are extremely close, as it is the case of the individual elements in an array antenna.

In the on-platform coupling, the situation is similar, even though the two antennas might not be very distant from each other. We can call this weak coupling. Weak coupling does not imply insignificant coupling.

Under the weak coupling assumption (which is valid in many co-sited antenna coupling computations), and the assumption of matched conditions at each antenna, in the sense that $Z_{g1} = Z_{L1} = Z_{o1}$ "and" $Z_{g2} = Z_{L2} = Z_{o2}$, P_{r2} becomes:

$$P_{r2} = \frac{|V_{2a}^{OC}|^2}{8 * Z_{o2}} \quad (5)$$

$$P_{t1} = \frac{1}{2} * |I_{1a}|^2 * Z_{o1}, \quad P_{t2} = \frac{1}{2} * |I_{2b}|^2 * Z_{o2} \quad (6)$$

By performing some simple computations we obtain:

$$\frac{P_{r2}}{P_{t1}} = \frac{\left| \oint_{S2} (E_2 \times H_1 - E_1 \times H_2) * \hat{n} dS \right|^2}{16 P_{t1} P_{t2}} \quad (7)$$

Surface S2 must enclose only the impressed source, and what we consider the structure of antenna#2. If both the antennas are lossless, Eq. (7) can be written as:

$$\frac{P_{r2}}{P_{t1}} = \frac{\frac{1}{4} \left| \oint_{S2} (E_2 \times H_1 - E_1 \times H_2) * \hat{n} dS \right|^2}{\left\{ \oint_{S1} \text{Re}\{E_1 \times H_1^*\} \right\} \left\{ \oint_{S2} \text{Re}\{E_2 \times H_2^*\} \right\}} \quad (8)$$

Let's observe that the power coupling between the two antennas can be found solely in terms of the transmitted fields of each antenna, albeit in the presence of each other and of the platform. No port voltages or currents are required. Some other useful forms of the power coupling theorem are provided in [9].

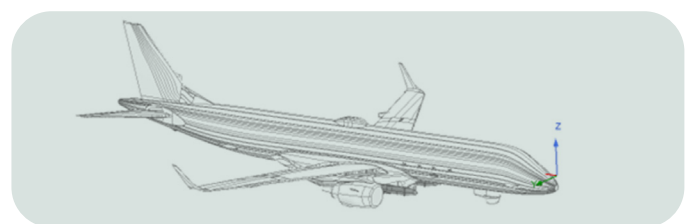
CEM TECHNIQUES IN LEONARDO

The previous section has shown the full-wave CEM viewpoint for modelling a coupling environment consisting of two antenna structures in the presence of a platform. If some full-wave CEM model can supply the fields ($E_r H_r$) and ($E_z H_z$) at the antenna ports and over the surfaces defined in Section 2 in relation to the reaction integrals, for this original environment we have everything we need to accurately evaluate their coupling. The three-fundamental full-wave CEM techniques are: the method of Moments (MoM), the Finite Element Method (FEM), and the Finite-Difference Time-Domain (FDTD) method with its variants such as the Finite-Integration-Technique (FIT). Since such kind of calculation is very time-consuming and often unaffordable, the CEM developers have supplemented the fundamental full-wave CEM techniques' capabilities by incorporating improved algorithms, or through their hybridization. A hybrid method is a linkage of two different computational methods. They are indispensable in the on-platform coupling models. If two different full-wave schemes are combined in some way, we could call it a full-wave/full-wave hybrid method (e.g. MM/FEM). If a full-wave method is combined with some asymptotic method (as defined below), we can call it a full-wave/asymptotic hybrid method and explicitly name the specific methods used in doing so (e.g. MM/PO). On such terminology, there is not complete unanimity in the CEM literature.

Method Of Moments (MoM)

Let's first consider the use of the MoM [10] for the full-wave CEM model described in the previous Section. The MoM formulation is based on some integral equation model of the electromagnetics problem. The unknowns of this equation are a set of equivalent currents (in general, both electric and magnetic) on all the physical structures present. This is enabled by the application of the surface equivalence principle. In particular, let us consider a Perfectly Conducting (PEC) object and an impressed source. Let's suppose that the intent is to use the MoM to find the scattered fields everywhere, and hence the total one. The surface equivalence principle allows us rigorously to replace this original problem by the equivalent one, where equivalent currents exist in the free space. By invoking them, we remove the physical object without approximation. An integral equation obtained by imposing the physical condition of zero tangential electric field on surface S_{pec} has the equivalent surface current density J_s as the unknown quantity to be solved for. The MoM procedure discretizes the integral equation into a matrix equation of the form $[A]*[X] = [B]$. The unknown column vector $[X]$ contains the coefficients of a set of expansion functions in terms of which the unknown currents are represented. In practice, the surfaces of the material structures are meshed into geometrical elements that define the geometry, and in terms of which expansion functions can be defined. The larger the surface area (in square wavelengths) of the physical structures being meshed, the more mesh elements, and hence unknowns, are required. For situation (a) in the previous Section, the column vector $[B]$ is known from (E_{inc}, H_{inc}). The equivalent currents generate the scattered fields (E_{scat}, H_{scat}) everywhere. The sum of these fields gives the total fields ($E_r H_r$). Same comments apply to situation (b) through a simple change in subscripts.

Once we know J_s , finding which is the difficult part of the task, these scattered fields can be easily found by employing a radiation integral, over S_{pec} , of the known J . This would be a full-wave CEM approach. We would find that J_s can be nonzero everywhere over S_{pec} , and not only over that portion in line of sight from the location of the impressed source. An example of such a technique application is shown in Figure 3 and Figure 4: the Electromagnetic Interference between all the emitters on board the EMBRAER ER-J190 has been computed by recurring at different methods depending on the analysed frequency bands (Ku, X, C, L, UHF, VHF, HF). The model of the aircraft is visible in Figure 3.

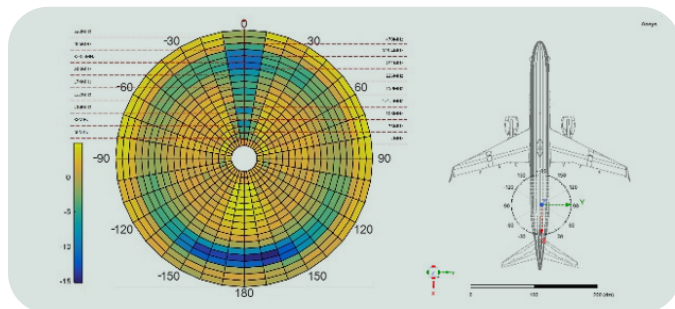


3-EMBRAER ER-J190

The size of the aircraft is such that coupling between emitters at communication frequencies (UHF, VHF, HF) has been numerically computed via the method of moments, i.e. a rigorous full-wave approach. A single frequency analysis has been performed, which means that an antenna able to transmit in more than one band was analysed separately at lower and upper frequencies of each band. At each frequency, every transmitting antenna present in the scene is fed alone while the others are closed on matched load. In this way the effects of the platform and of the antennas on the transmitter radiation pattern are evaluated as the worst-case inter-antenna coupling condition (condition of antenna impedance matching for maximum power transfer). Coverage analysis has also been performed.

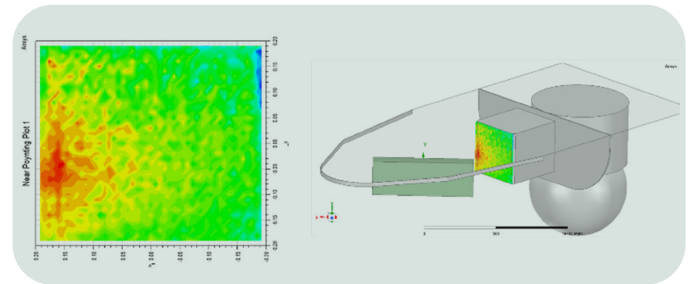
An example is reported in Figure 4, for a communication antenna in VHF/UHF band, installed in front of the rear wing of the aircraft. The graph shows that its radiating performance is limited over an entire angular sector. Such graphs show the antenna azimuthal gain/directivity levels for each angular measurement, shown on the polar axis. In this case only far fields had been computed. As such technique can provide fields everywhere, another example is reported.

By applying the same approach, a near field has been



4-Example of computed coverage for one V/UHF antenna

computed on a surface, in order to evaluate the risk of damage to electronics component exposed to the field radiated by a search radar installed on the drone of the EUROMALE program, as it is shown in Figure 5. For security reasons the field values are not reported and the antenna details are not in the picture. The surface where the field is computed is not included in the modelling of the problem, since the incident near field was the goal of the analysis. It is added only for graphical needs.



5-Example of incident near field on a surface, computed by the method of moments

Finite Element Method (FEM), Finite Difference Time Domain (FDTD) and Finite Integration Technique (FIT)

The FEM [11], the FDTD method [12], and the FIT [13] discretize differential equation models of the electromagnetics problem, the unknowns being the fields (E , H) directly in some region of the space surrounding the material structures. This complete region must be meshed. All these methods end up in principle having to solve a matrix equation, even if this is not done explicitly. In the full-wave CEM methods, “solving the problem” refers to the solution of this matrix equation. Numerical accuracy, convergence rates, scalability, parallelization ability, computation time, memory consumption, are the parameters that make different methods suitable to different applications. Large matrix equations require the use of iterative solvers, and it is the rate at which this converges to an accurate solution that is of concern. If the number of unknowns, and hence $[X]$, is extremely large, the matrix $[A]$ becomes too large, as the solution of the matrix equation is not as reliable as required, or takes too much computation time, or requires excessive memory. There is little sense in thinking that one must simply obtain increased computational resources. Electromagnetic engineering, of which the on-platform coupling is but one aspect, is getting more and more demanding. Ever since the fundamental full-wave CEM methods became established as a technology, it is not surprising that many ingenious ways have been devised to enhance the performance of this technology.

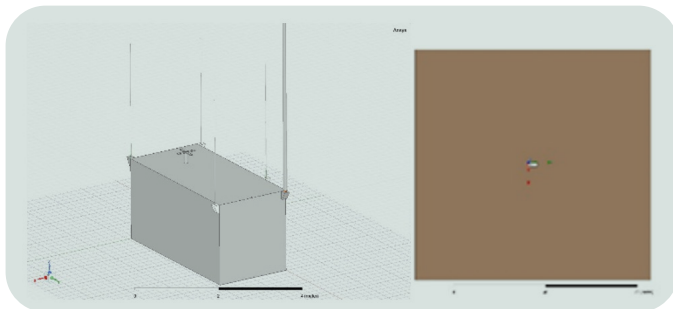
The incorporation of multilevel fast multipole method algorithms (MLFMM) [14] into MoM implementations over the last 20 years, greatly increased the electrical

size of a structure that can be handled by the MoM as a single problem. The MLFMA is referred to as an accelerator of the MM, dealing with the way the matrix $[A]$ is determined and used, and is already incorporated in most commercial CEM codes offering a MoM capability. It is also referred to as the multilevel fast multipole method (MLFMM). The FEM method can only be applied over a finite region of the space and so requires some way to truncate this region. A rigorous way to do this is to adopt a hybrid FEM/MoM approach, in which the FEM is applied within some closed surface, and the MoM is used to deal with the free space region outside the surface, by placing equivalent electric and magnetic current densities on the truncation boundary surface. The internal fields and the surface current densities are found in an electromagnetically consistent manner without any approximation. There can be more than one explicit FEM region, as long as each region is truncated, as we indicated above. This method is most often called the finite element – boundary integral (FE-BI) method. It facilitates the accurate full-wave solution of problems involving detailed geometrical and material make-up, which are also open to free space. The domain decomposition methods (DDMs) are a more general class of approaches useful to extend the size and complexity of problems that the full-wave CEM methods are able to handle. The DDM divides the original problem (the domain) into several separate smaller subdomains and it is often referred to as a divide and conquer approach by CEM developers.

TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

Surveillance and low observability

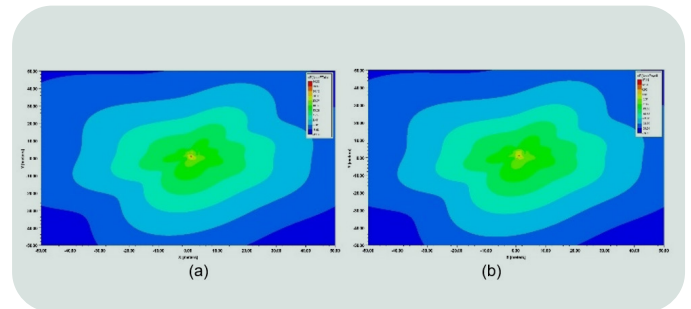
Although the subdomains are rigorously linked, the problem solvers on the subdomains can operate in parallel. A different fundamental full-wave CEM method can be used in each of the subdomains and then it would be classified as a full-wave/full-wave hybrid method [15]. Below an example of FEM application is shown. Under the JDIFSS program, maps of the intensity of the average electromagnetic fields around the shelter have been calculated, in order to allow the assessment of safety of the personnel exposed to electromagnetic radiation, according to Italian laws.



6-(a) CAD model of the whole system, with the antennas placed on the shelter; (b) area of field computation, top-view

The geometrical model of the system, with communication antennas placed on the platform, is reported in Figure 6.

Communication antennas cover bands from HF to UHF. The total average fields have been computed over the area sketched in Figure 6, at the heights of 1.7m and 2.5m above the terrain. An example of such a computation is reported in Figure 7(a-b), where maps of the average Erms field and Pointing vector intensities respectively are shown (at h=2.5m).



7-(a) Erms field [dBV/m] and (b) Poynting vector [dBW/m2] intensity maps

ASYMPTOTIC METHODS

Let's now turn our attention to asymptotic methods. Here and in the rest of this Section, we will think in terms of PEC objects, although penetrable objects can be similarly dealt with. Typically, for a given scattering object size, an asymptotic solution to an electromagnetics problem becomes increasingly accurate as the frequency becomes higher. In other words, the asymptotic modelling approach becomes more accurate as the scattering structures becomes electrically larger, which are also the conditions under which the full-wave CEM methods become computationally more burdensome. The hybrid full-wave/asymptotic CEM methods attempt to combine the strong points of each of them.

Geometric Optics

The fields of any finite-sized source (electric and magnetic fields perpendicular to each other and straightforwardly related by the intrinsic impedance of the medium) become almost locally plane at observation points that are distant enough from the source. The smaller the electrical extent of the source, the shorter the distance at which this occurs. We can divide any source distribution into smaller parts, and each part's fields looks locally plane when it is relatively close to the source. A full electromagnetic solution may be found through superposition. Mathematical expressions for a locally plane field take the form of a reference value, multiplied by a so-called "spreading factor" that

accounts for the amplitude decrease, and a factor that accounts for the phase change, as the field propagates. An additional term deals with polarization. We can also impose a direction-dependent pattern factor. Such fields can be interpreted [16] in a geometrical optics (GO) sense. The field propagation in all directions can be viewed in terms of rays, which fits well with our intuition, with rays undergoing reflection from objects. When using such ray-optic methods to analyse the electromagnetic scattering, some errors can be made, because shadowing of rays occurs that would be neglected by a strict GO approach application.

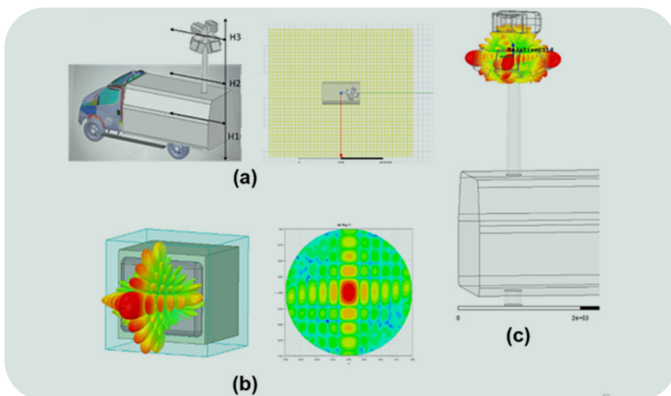
Physical Optics

The physical optics (PO) [17] procedure makes the assumption that, at each point P on the finite-sized PEC structure, the current density is approximated by the expression $J_S^{PO}(P) = 2\hat{n}(P) \times H_{inc}(P)$ over the illuminated portion of the object, and it is assumed zero over the shadowed portion. This current is then used in a radiation integral to find the fields it generates. One finds that the electrically larger the scatterer is, the better the PO current approximates the exact J_s .

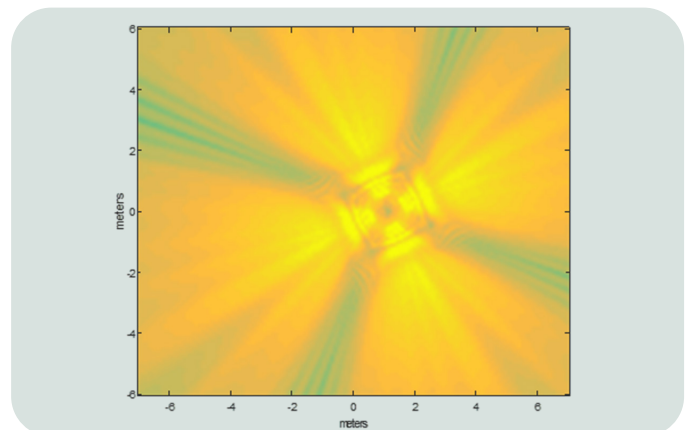
In other words, for a scatterer of a given physical size, this approximation becomes increasingly accurate as the frequency increases. It is an asymptotic method. Let's point out that there is no matrix equation to solve in the approximation of $J_s^{PO}(P)$. Even for an electrically large object, the PO current near edges of the object (where the object's surface is not smooth) differs appreciably from the exact J_s . Accuracy of the approximate current can be improved by adding additional terms to $J_s^{PO}(P)$, using an asymptotic procedure known as the physical theory of diffraction (PTD). The correction terms are inferred from analytically obtained asymptotic solutions of certain geometrically simple problems, albeit being not mathematically simple [18]. Below, an example performed in Leonardo within the MSS program, of near field computation for personnel safety evaluation, where PO approximation has been adopted, due to the complexity of the problem and to the short time available to solve it. In this case the output of the study is the computation of 2D contour maps of the average

rms E-field (as stated by personnel safety current regulations) at different heights above the ground, in a squared area around the vehicle, as reported in Figure 8(a), where the field source is visible as it is composed by four radar mounted on a telescopic pole. In Figure 8(b) the single radar far-field is shown while in Figure 8(c) the total radiated field is shown.

Since the structure was very electrically large, and the radar were designed to electrically scan over a large number of pointings, the required computation was very time-consuming; therefore, a hybrid approach was necessary between the full-wave field solution of the radar and the PO asymptotic approximation over the other regions around the sources. In Figure 9 one of the output of this study is reported, without intensity levels of the fields. Here the average electric rms field has been computed on a squared area around the vehicle at one of the required heights, $H_3=4.3m$ above the ground (Figure 8(a)). In this computation the actual pointings scheduling of the radar has been accounted for.



8-Example of hybrid approach for near field computation and personnel safety evaluation: (a) geometry of the problem; (b) field source pattern; (c) total radiated pattern



9-Average electric rms field, computed on a squared area around the vehicle

SBR

The basic Shooting and Bouncing Rays (SBRs) asymptotic method takes a slightly different way. Many rays are launched from the source (or each part into which a source has been divided, followed by the use of superposition) and are used to "paint" an illuminated surface to find the PO currents. The fields due to these PO currents are then in essence converted into ray-optic format, and themselves, as reflected or diffracted rays, allow to "paint" those parts of the surface lit by each of the many rays. Thus, the SBR extends the standard PO approach by finding an improved approximation J_s^{SBR} to the exact J_s , no longer with physically false "dark" shadowed portions. This J_s^{SBR} is used in radiation integrals to find the fields of the antennas in the presence of the platform, in order to perform coupling computations.

While the final step in the computation might be rapid, tracking the various rays could be very computational expensive. There are variations in the manner the SBR is implemented; a review of many hybrid method ideas is available in [20]. The total field is made by contributions that emanate from specific locations on the scatterer, classified depending on whether they emanate from a point on a smooth part of the surface or from an edge, or they arrive after having travelled ("crept") around the surface and being then launched toward the observation point. The total field at an observation point could be written as the sum of these direct, reflected, edge-diffracted, or smooth-surface diffraction terms. By including these extra rays, we would be able to eliminate the obvious error of the classical GO (namely the prediction of identically zero fields in the shadow region).

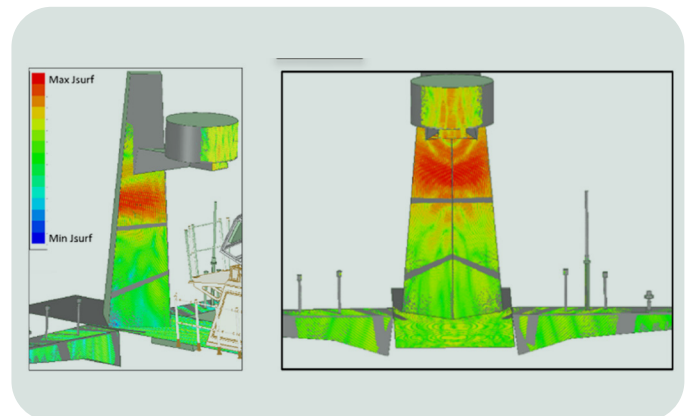
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This whole procedure is called the geometrical theory of diffraction (GTD), and its refinement is called the Uniform GTD (UTD). We would find that the results so obtained improve as the electrical size of the object gets larger or, in other words, the frequency for a given sized object gets higher. It is an asymptotic approach. The actual field value (amplitude and phase) and polarization of a reflected ray at some observation point is written in terms of the field impinging on the object at the point of reflection, the reflection coefficient, and the amplitude and phase terms mentioned earlier, dependent on the distance from the reflection point. The diffraction coefficients replace the reflection coefficients at the observation points, and so on. Such computations are extremely fast. To go searching for the location of relevant reflection points and diffraction points, for given source location and observation point, uses most of the computing time when UTD methods are implemented in codes that allow analysis of objects of general shape.

As it happened with the full-wave CEM methods, much work has been done on algorithms that perform such searching, and many commercial codes include a UTD capability. In the on-platform coupling context, hybrid full-wave/asymptotic methods use some full-wave methods over a part of the platform, which is usually the closest to the antenna(s), while the asymptotic

methods are used for determining the scattering from, and hence the influence of, the remaining portions [19]. The following Figure 10 shows the application of the SBR method in the computation of the induced surface currents on a naval platform, where a sensor is pointing toward the mast. No details are provided about the sensor, the frequencies and the intensity levels of the currents. Figure 10 has the only aim to show the potentiality of the SBR method in solving complex e.m. problem in electrically large environments like a naval system.



10-Side and front view of the induced surface currents on a naval platform, when a sensor is pointing toward the mast

CONCLUSIONS

This paper shows how Leonardo deals with e.m. analyses of electrically large structures, by making use of the state of the art of Computational Electromagnetic Modelling. The presentation of theoretical instruments is completed with examples of computation applied in the context of several Leonardo programs.

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TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

Surveillance and low observability



Microwave Filters in Leonardo

Maurizio Felaco, Giancarlo Prisco, Gabriella Bernardi, Elio Petrillo, Stefano Mosca

Leonardo – Electronics Division

Nowadays, in the geopolitical scenario, the Defence domain acquires increasing strategic importance, due to ongoing delicate conflicts and emerging tensions on multiple international fronts. One of the relevant themes is the technological autonomy of the national defence industry. In this perspective, it is desirable to pursue the internalization of every knowhow, from the design phase to the realization of key devices, which should go ranging from the overall system to every critical subsystem and components. The RF filters are components typical to every microwave assembly; consequently, their importance crosses over all the business areas of Leonardo. Until recently, Leonardo sourced such critical filters from leader companies in that specialized sector. Given the growing need for keeping such autonomous design capacity within Leonardo, ten years ago a twofold activity was launched, consisting first of all in the Fit Form and Function (FFF) redesign of filters whose supply had become critical, and then in the custom design for new systems. This article shows some relevant results that demonstrate our capability to design filters operating in a wide range of frequencies, exploiting the main development technologies.

INTRODUCTION

Nowadays a relevant theme is the technological autonomy of the national defence industry. In such a context, particular attention has been paid to the RF filters that are key components of every microwave assembly. In recent years, Leonardo Electronics Division has launched a program aimed to increase and keep its capabilities required for the design and development of microwave filters. As a result, Leonardo has identified a detailed effective process. The design workflow involves the following phases:

- Identification of the minimum order ideal filter able to satisfy the electrical requirements. The filter order represents the number of its resonators. Obviously, the higher the order of the filter, the greater its dimensions.
- Choice of the most suitable technology based on constraints, e.g. the maximum loss level, the available space, the signal power, etc.
- Identification of the filtering structure and implementation of an electromagnetic simulation model suitable to estimate its performance.

- Design optimization, in order to fulfil the required electrical performance, respecting the overall dimensions and construction constraints (e.g. the minimum feasible distance between 2 strips for a microstrip filter).

Leonardo with its suppliers supervises all the design and realization process phases, in order to obtain the best achievable performance. In particular, the final aim is to reach a project definition that is both technologically feasible and robust with respect to uncertainties of the manufacturing process.

In recent years, filters designed in Leonardo had been developed in the most widespread technologies: microstrip, stripline, waveguides, cavities, ranging in the main operative bands of the Leonardo systems. In the following, an overview of the filters developed in Leonardo is given. The shown projects are representative of different filter types, manufacturing technologies and operative bands. The filters here shown had been designed within the main development programs of the different Leonardo business areas, e.g. the land, naval and airborne domain. The paper is organized as follows. Section CAVITY FILTER shows a project of a bank of cavity filters, section PRINTED CAVITY presents an innovative

design that realizes a cavity filter by exploiting the printing technology, section TUNABLE COAXIAL FILTER shows an example of a tunable filter in the FM operative band. Finally, CONCLUSIONS section follows.

CAVITY FILTER

A cavity filter is typically composed of one or more hollow metal cavities containing conductor resonant structures [1].

Such a cavities may feature different shapes and sizes (cylindrical, square, etc). The design of cavities requires realization resolution of dimensions and shapes that ensure accuracy and stability of the resonance conditions. The higher the operative band frequency, the higher the required manufacturer precision. Typically, the metal employed for cavity filter is high-quality conductive metal. Cavity filters are the best solution for applications that require low loss and high power handling.

In Figure 1 a switched bank of cavity filter is showed. The bank has been designed for a X band radar of the Leonardo's AESA radar family. Thanks to miniaturization of the board-to-board connectors of the new generation, the height of filters mounted on the board features increase of just few millimeters (i.e. height of the connectorized interface is 3-4mm). This innovative connection guarantees the following advantages:

- Easier mounting process;
- No damage to rework;
- Multiple test point on the board.

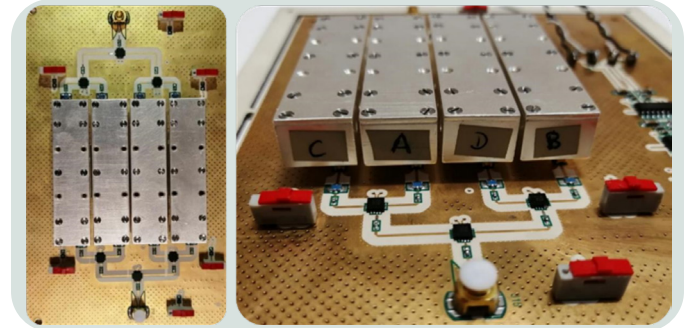
In Figure 2 the detail of the connection filter-to-board is depicted with the CAD model of the used connector.

On the PCB the transition has been appropriately optimized. The resulting layout is shown in Figure 3.

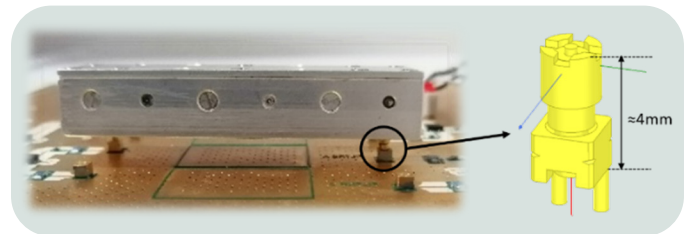
The cavity filters type of the bank is bandpass, whose fractional bandwidth is approximately 3%, centred at different frequencies in X band. Figure 4 depicts the performance of one of the filters. Performances of the measured filter are in excellent compliance with those predicted.

PRINTED CAVITY

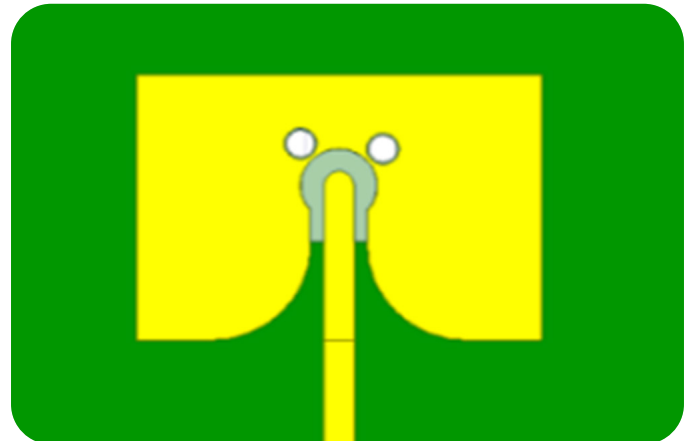
Cavity filters typically require larger physical footprint, which prevents from its usage in applications with limited available space. Furthermore, the high precision required in their manufacturing process brings high production costs. When an application does not require very high power handling or low losses, typically less expensive technologies are used. The printed cavity is an innovative typology of filtering structure. Printed cavities are inspired by geometries of the metal cavities but are manufactured by using the printing technology,



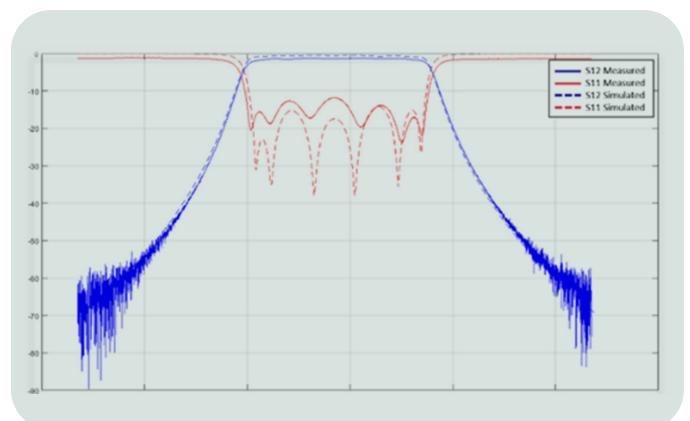
1-Switched X band cavity filter bank



2-Innovative filter mounting using board-to-board connectors



3-Optimized transition layout for the pcb-to-pcb connector



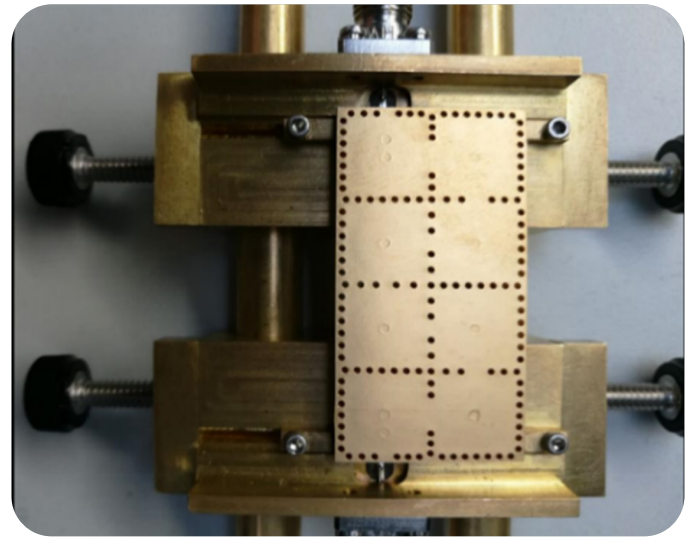
4-Xband cavity filter measure vs simulation

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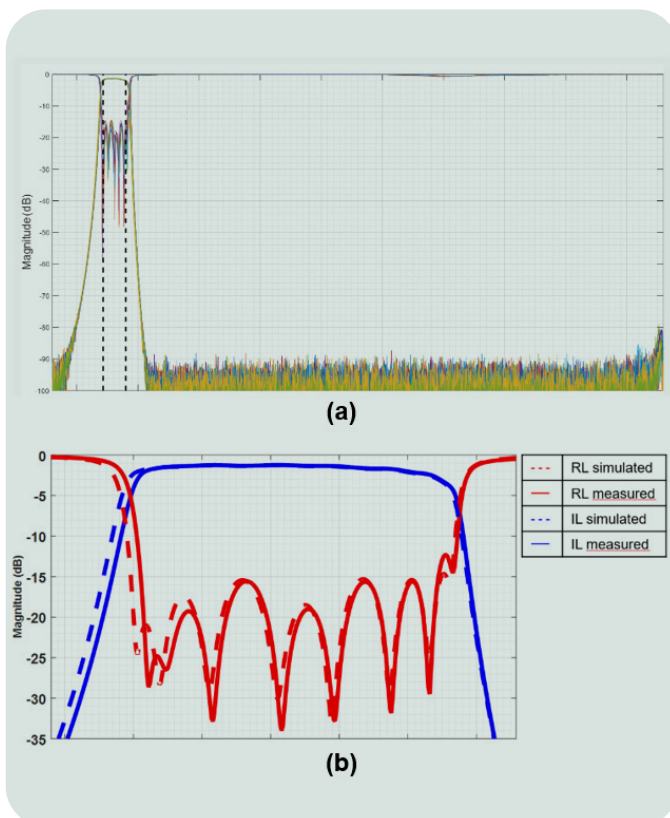
Surveillance and low observability

thus exploiting advantages of the high precision and lower cost that are typical to such manufacturing process. Typically, the resonators are made with vias that are drilled into the dielectric. The dielectric allows to achieve more compact structures but it implies greater losses. In typical printed filters, the resonators geometry is 2D (i.e. in a plane) [2]. In the printed cavity filters, the resonators geometry is 3D. By exploiting the third dimension, printed cavity filters guarantee reduced surface footprint, while requiring higher profile. Since the area available on the PCB surface is usually a critical resource, due to the increasing density of mounted components, printed cavity filters are an attractive choice. From the technical point of view, the printed cavity filters are manufactured as multilayer circuits.

The most widespread assembling technique for multilayer PCB require the introduction of intermediate bonding layers (prepreg). Typically, the tolerance declared on the values of the prepreg electromagnetic characteristics is higher than that of the dielectric layers.



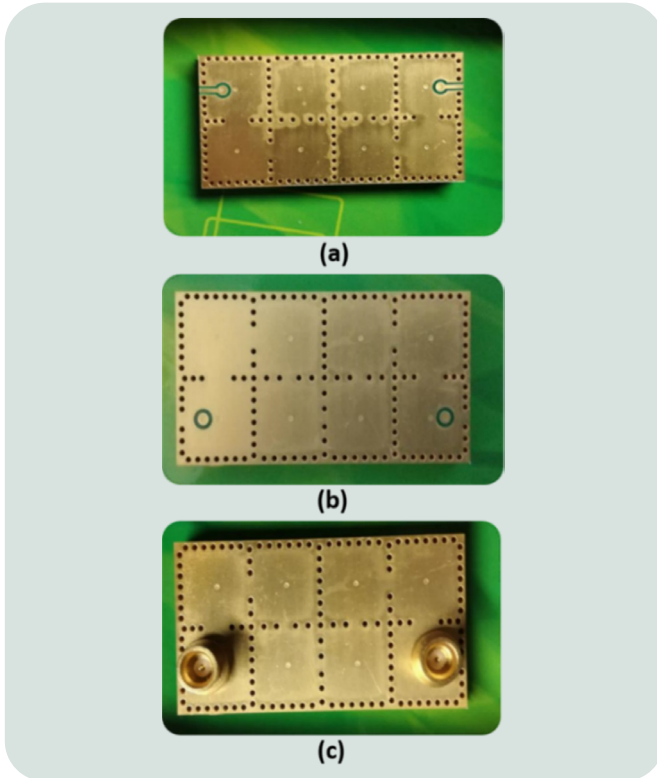
5-L-band passband filter with harmonics rejection



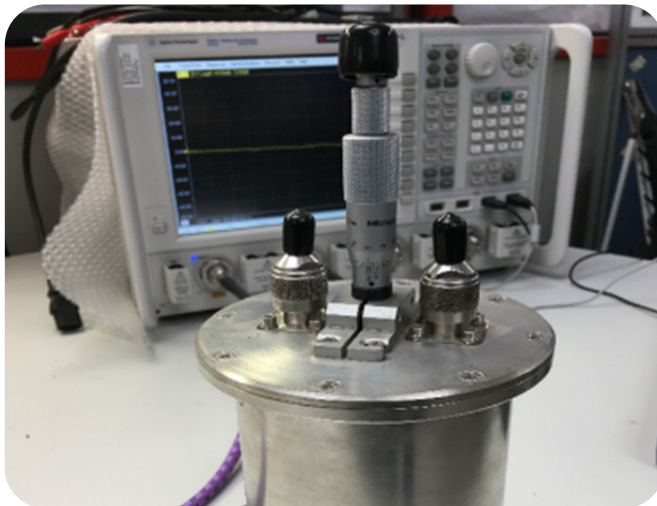
6-L-band printed cavity filter: (a) wide band measure of 50 samples; (b) In band performance, simulation vs measures of a sample

Furthermore, it is very difficult to estimate the final thickness of the bonding layers. Indeed, these uncertainties in printed cavities can affect the resonant frequencies of the resonators and their mutual couplings, thus producing undesired variations of the filter performance. Thanks to an emerging bonding technique, the Fusion Bonding technology, those uncertainties are avoided because no bonding layer is required in the stacking process. The realized laminates are pressed at high temperature, about the melting point of the teflonate dielectrics, which makes the contact surfaces to fuse, thus creating a single solid layer. The Fusion-Bonding technique, is more expensive, but it provides excellent repeatability of the component. The filter on its test jig shown in Figure 5 is an example of a printed cavity filter. It has been designed for a full digital system of Leonardo. It is a bandpass filter with significant rejection immediately out of the pass band. Exclusively filters of high order are able to ensure such a selectivity, i.e. filters with high number of resonators. By exploiting the advantages of the printed cavities, its surface dimensions are very compact, despite the high order of the filter. Figure 6 (a) depicts the wide band performance of 50 samples of the filter. The homogeneity of the measurements demonstrates the high degree of repeatability of the production process. The filter guarantees a rejection of about 80 dBc up to the eighth harmonic. This remarkable result is obtained thanks to a design technique that makes combined use of both the distributed parameter structures and the lumped parameter structures (i.e. quasi lumped structure).

Figure 6 (b), shows an insight of the filter performance in the pass band: it can be seen that the measured performance matches the expected one. Due to its high selectivity, wide band rejection and compact dimensions, this project has also been used for other programs of the Company. By design, a number of versions are available for this filter, thus allowing its use in different board-filter interfaces. Figure 7 shows three variants of the filter: (a) microstrip interface, (b) stripline interface and (c) connector interface.



7-Example of different transition Filter-PCB: (a) SMT microstrip mounting, (b) SMT stripline mounting, (c) connector mounting

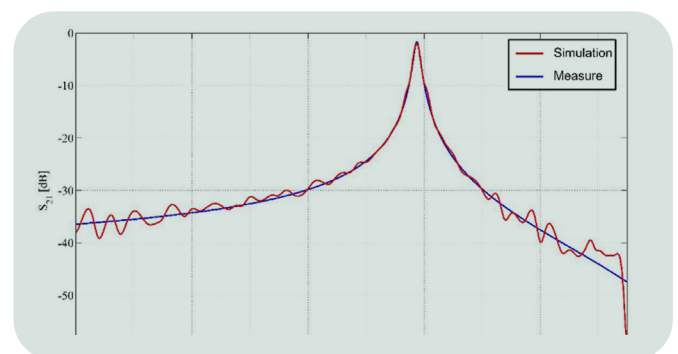


9-Tunable FM filter measure vs simulation

TUNABLE COAXIAL FILTER

The filter in Figure 8 is designed for a passive Passive Coherent Location radar (PCL) exploiting commercial FM broadcast transmitters as the illuminating sources for detecting targets.

Since passive radars cannot control the power levels emitted by a source of opportunity, one of the critical issues of these systems is the low power level of the received signal, which determines challenging specifications in terms of losses for the filters along the receiving chain. The Waveguide technology is the best choice in this case. Another critical issue faced in this project is the space requirements. The filter has to be as compact as possible, because the radar is a portable system. This requirement is particularly challenging, due to the system's bandwidth. The solution applied is a design that matches a distribute structure, in particular a coaxial resonator, to a concentrated high capacitance [3]. The high capacity placed at one end of the resonator allows to reduce its length, thus containing the overall size of the filter. The concentrated capacity is made with two metal discs whose radius is slightly smaller than the radius of the internal part of the cavity. The capacity value can be tuned mechanically thanks to a mechanism that allows the two disks to move apart/closer together. The variation of the capacity induces translation of the resonance frequency of the filter that can thus be tuned to the desired FM frequency. The performance of the realized filter is in excellent compliance with the simulated one, as it is shown in Figure 9.



9-Tunable FM filter measure vs simulation

ON BOARD FILTER

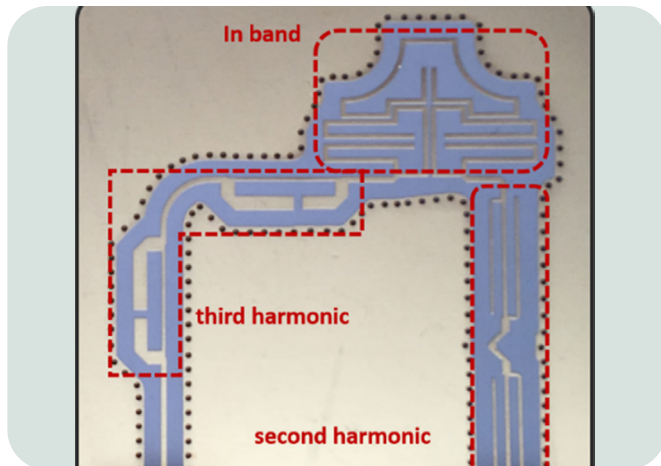
By “on-board” filters we mean filters designed in planar technology directly on the PCB boards. Where applicable, the on-board filter certainly represents the most economical solution. Due to the ever-increasing PCB density, the room available for filtering is not only reduced, but it is also not suitable for hosting the most common layout of planar filters [4] (end coupled, edge coupled, stepped impedance, combline, interdigitated

etc). In these cases, creative solutions are necessary. For example, Figure 10 shows how the geometry of the resonators is shaped to fit the space available on the PCB. The microstrip filter is developed on the upper layer of a multilayer PCB and is made up of three sections. The top section implements an L-band pass band filter, while the other two sections host notches for rejection of the second and third harmonics, respectively.

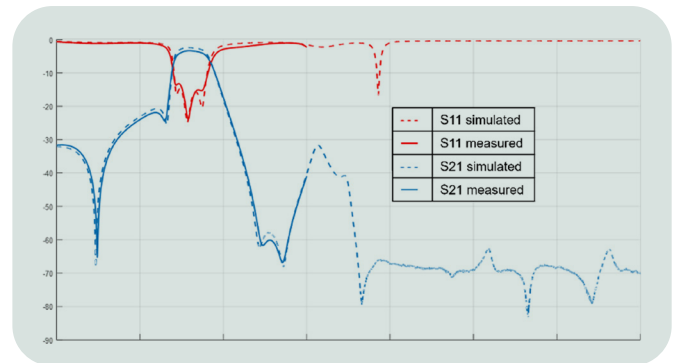
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Figure 11 shows the performance of the simulated filter and of a realized sample. The large overlap between the measure and the simulation highlights the quality of the electromagnetic modelling tools used in the design process. As it was expected, the filter features band-pass behaviour in the L band and ensures a rejection of the second and third harmonic that is greater than 60dBc.



10-On board L-band filter



11-On board L-band filter measure vs simulation

CONCLUSIONS

In the recent years, Leonardo Electronic Division has launched a program aimed at acquiring the capabilities necessary to design and develop microwave filters. As evidence of the achieved results, some successful projects are reported in the previous sections. Thanks to the acquired skills, Leonardo has the capability to develop microwave filters in the most common technologies (e.g. microstrip, stripline, waveguides, cavities)

and to realize typical designs as well as more advanced ones. Furthermore, the examples demonstrate the ability to design filters in different operating frequency bands.

In the future, our objective is to go exploring further technologies, such as dielectric and quartz resonators, as well as expanding the frequency range of the Leonardo filters.

Maurizio Felaco: maurizio.felaco@leonardo.com

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Exploration of Innovative Technologies for Aircraft Stealth Applications

Francesco Acerra, Marco Boccaccio
Leonardo - Aircraft Division

Aircraft survivability in modern warfare crucially depends on achieving low observability that encompasses stealth and advanced electronic warfare capabilities to evade detection across the electromagnetic spectrum, which ranges from radio frequencies to the visible range and acoustic surveillance. Extensive research has been dedicated to developing radar absorbing material solutions, aiming to address future challenges such as achieving wideband absorption and utilizing lightweight materials. This work aims at providing a comprehensive review of the State-Of-The-Art Radar Cross Section reduction technology, with a specific focus on aerospace applications. Our exploration covers innovative technology developments, various electromagnetic scattering mechanisms, treatment methods, and design techniques utilized for evaluating effectiveness of the RCS reduction solutions. Through this examination, this paper aims to shed a light on current advancements and future directions in enhancing aircraft survivability through stealth technologies.

INTRODUCTION

In modern warfare environments, low observability plays a paramount role in enhancing the survivability of military systems. This capability is crucial for evading detection across the electromagnetic spectrum, which goes ranging from radio frequencies to the visible range, as well as acoustic detection [1][2]. Efforts to diminish the detectability of military systems have yielded mature technologies over the years, with radar detection reduction standing as the prominent example. Since World War II, all branches of the military have been actively engaged in development of the Radar Cross Section (RCS) reduction technology [3]. Notably, the strong emphasis of the Air Forces on this technology has laid the foundation for a robust aerospace RCS reduction technology base. Furthermore, the frequency of the detecting radar poses a significant challenge in Radar Absorbing Material (RAM) design [4]. Generally, lower frequencies necessitate thicker and heavier RAMs to meet the electromagnetic absorption requirements. These parameters must be carefully balanced, considering potential conflicts with aircraft aerodynamics and payload capacity.

Radar systems operate across a broad range of frequencies, with higher frequencies offer improved accuracy but feature greater susceptibility to weather interference such as rain or clouds. At lower frequencies, radars can detect targets at long distances but struggle to discern target details, such as aircraft typology and potential threat level. Conversely, higher frequencies provide more detailed information but are effective only over shorter distances. Nowadays, conventional stealth technologies feature limited effectiveness at extremely low frequencies (below 300 MHz), with most applications falling within the 2 GHz to 18 GHz range [1]. The goal of the stealth technology is to obtain wideband radar invisibility, whereas the low observability technology is primarily focused on reducing the radar detection at short distances, particularly within the 8–12 GHz frequency range. This approach allows targets to be detected at longer distances but with poorer details. It is worth noting that especially at lower frequencies, thicknesses of RAM can become excessive [5]. Hence, the ongoing research efforts elaborated in the following paragraphs aim to develop thinner designs relative to the wavelength, to address this challenge effectively.

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Objective of this work is to provide a comprehensive overview of State-Of-The-Art RCS reduction technology, particularly of its evolution within aerospace applications. This overview encompasses the innovative technologies under development by *Leonardo Aircraft Division Strategy & Innovation/Leonardo Labs* (LAD S&I/LLabs) in collaboration with *Leonardo Aircraft Division Engineering* (LAD Eng) e.g. various categories of electromagnetic scattering mechanisms, treatment methods related to these mechanisms, design techniques and novel materials for RCS reduction solutions.

RADAR SIGNATURE

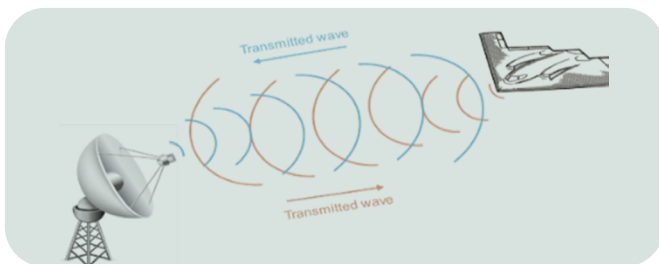
The very basic operational concept of RADAR (Radio Detection and Ranging) consists of emitting electromagnetic energy pulses towards a target, which propagate through space at nearly the speed of light, until they interact with that target [4][6]. Radar gauges the elapsed time for a specific energy pulse to reach the target and return back to the radar station (Figure 1). By utilizing Equations (1) and (2), the distance R (Figure 2) between the target and the radar apparatus can be calculated:

$$R = \frac{c_0 t}{2} \quad (1)$$

Where c_0 and t represent the light speed and the duration required for the energy pulse to return back to the radar system after striking the target, respectively. The basic radar equation is given by [4][7]:

$$P_r = \frac{P_t G_t}{4\pi R^2} \frac{\sigma}{4\pi R^2} A_e \quad (2)$$

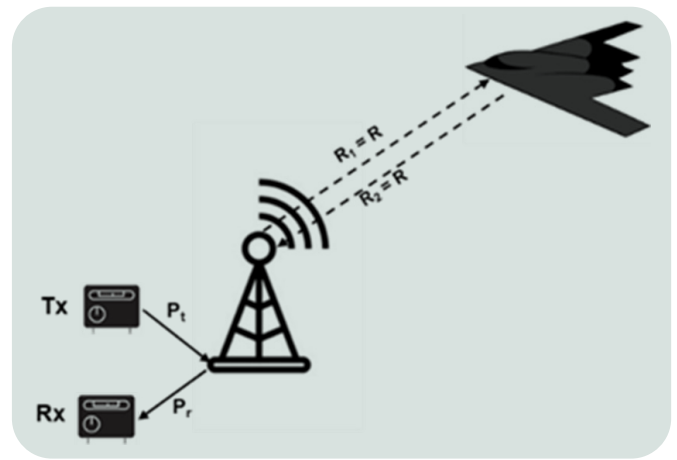
with P_r being the reflected power, P_t is the transmitted power, G_t represents the gain of the transmitter antenna, σ denotes the Radar Cross Section (RCS) of the target and A_e represents the aperture of the receiver antenna.



1-Schematic principle of RADAR

When energy pulses interact with an object, a portion of the energy is absorbed, some passes through the object, and the rest is reflected. The quantity of incident energy that is reflected depends on the radar signal's wavelength and frequency, as well as on the shape, dimensions, and material composition of the object, such as an aircraft. Objects of interest, such as aircraft, missiles, and satellites, are termed "Targets," while false echo returns caused by other objects like buildings, terrain, or bodies of water are termed "Clutter."

Efforts have been dedicated by radar researchers and designers to distinguish clutter returns from actual targets.



2-Target detection principle

The radar beam typically assumes a conical shape. When the target is distant from the source, a greater surface area is illuminated by the radar, which results in a smaller portion of the electromagnetic energy being reflected back to the radar system by the target. Thus, for long-range targets, only a fraction of the original outgoing pulse energy is detected, due to the design of antennas. Increasing the radar power could extend range, but it would be not advantageous as it would lead to increased energy wastage in free space and to heightened system noise. Antennas with larger apertures can mitigate this issue by generating narrower, more intense outgoing energy beams that are capable of intercepting more reflected energy.

Radar detects an object only when the radar antenna receives back a sufficient amount of energy from the object, surpassing the signal-to-noise ratio or the electronic noise threshold required for detection. In the transmission-scattering-reflection system, a number of variables influence the maximum detection range, such as the transmitter effective outgoing energy, the object radar cross-section (RCS), the beam width, the total energy returned after interaction with the object, the antenna aperture size, and the receiver processing capability. Among these variables, the RCS holds particular significance in target detection, as it is depicted in the following paragraph.

RADAR CROSS SECTION

The Radar Cross Section (RCS) of an object characterises its reflective intensity towards the radar system. The RCS is defined as the ratio of the power scattered back per solid angle from the target in the radar's direction to the total intercepted power by the target. The RCS area does not correspond to the geometrical cross-section of the target. RCS of a sphere, remains frequency-independent when operating at very low wavelengths (i.e., $\lambda \ll \text{range}$ and $\lambda \ll \text{radius}$), where the frequency is exceptionally high. Conversely, RCS of a flat plate, is frequency-dependent and is calculated as $4\pi A^2/\lambda^2$, where A represents the area of the plate. This relationship can be expressed proportionally in Equation (3):

$$\sigma = \lim_{R \rightarrow \infty} 4\pi R^2 \frac{|E_s|^2}{|E_i|^2} \quad (3)$$

Where E_s and E_i represent the scattered field strength at radar and the incident field strength at target, respectively. RCS can alternatively be expressed in terms of decibel square meter (dBsm) using the following expression:

$$\sigma_{\text{dBsm}} = 10 \log_{10} \frac{\sigma}{\sigma_{\text{ref}}} \quad (4)$$

Where σ_{ref} denotes the RCS of a reference object with area of 1 m^2 . According to Equation (4), a RCS value of 10 dBsm corresponds to 10 m^2 , -10 dBsm is 0.1 m^2 . Detectability of a target is inversely proportional to its RCS; thus, lower RCS leads to reduced detectability and vice versa.

RCS of a target is generally influenced by several factors:

- Target size: Generally, larger aircraft tend to have larger RCS.
- Target shape: The shape of the aircraft significantly impacts its RCS. Parabolic-shaped radomes can redirect waves without scattering, while stealthy designs with flat and sharp edges, scatter waves in different directions, leading to RCS reduction. Modern stealth aircraft typically feature flat designs with sharp edges and corners.
- Target material: The material used in aircraft construction greatly influences their RCS. Metallic structures reflect waves fully, resulting in high RCS. Substituting metallic structures with Radar Absorbing Structures (RAS) can enhance stealth capabilities.
- Angles: The angle of incidence and reflectance depends on the target's shape and orientation relative to the radar source. However, controlling these angles during real combat missions is often challenging, due to aircraft dynamics.
- Radar frequency: Stealth aircraft should be designed to be stealthy across the frequencies at which radars operate. Broadband RAS can attenuate a wide range of frequencies.
- Orientation: The polarization of outgoing and incoming radiation concerning the target's orientation plays a role in RCS.

Among these factors, selecting low-reflecting materials in the form of RAS offers a viable approach for RCS reduction, without compromising aerodynamic performance or aircraft design required for airworthiness.

STATE-OF-THE-ART RADAR ABSORBING MATERIALS

Radar Absorbing Materials (RAMs) are complex composites consisting of an insulating matrix and a magneto-dielectric reinforcement [7]. Their primary purpose is to diminish the reflected electromagnetic radiation by absorbing a significant portion of the incident electromagnetic energy. This absorption is achieved through various means; the first method allows signals to penetrate and subsequently reduces their intensity. Materials compliant with this approach, suitable for a broad range of radar frequencies, are known as broadband RAMs. Alternatively, RAMs can produce internal reflections of the received signals that impede the reflection of signals from the outer surface. This method, termed resonant RAMs, is advantageous when dealing with specific frequencies. In the context of material responses to electromagnetic waves, three interrelated parameters play crucial roles: permittivity, permeability, and electrical conduction.

Conversely, the microwave absorption of dielectric materials is directly contingent upon the electric permittivity ϵ , which is given by [6]-[8]:

$$\epsilon = \epsilon' - j\epsilon'' \quad (5)$$

where ϵ' and ϵ'' denote the real part and imaginary part of the material permittivity. When microwaves interact with dielectric materials, internal fields are generated that induce displacement of dipoles and free bound charges such as electrons and ions. These movements encounter resistance from inherent frictional and elastic forces within the materials, resulting in losses. These losses are collectively represented by the imaginary permittivity or dielectric loss parameter ϵ'' .

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The loss tangent $\tan \delta$ is defined as the ratio of ϵ'' (imaginary part of permittivity) to ϵ' (real part of permittivity), as expressed by:

$$\tan \delta = \frac{\epsilon''}{\epsilon'} = \frac{\gamma}{2\pi f \epsilon_0 \epsilon'} \quad (6)$$

where γ represents the effective conductivity generated by the electronic conduction and dipole displacement current, and f is frequency. Analogously, the magnetic permeability can be evaluated as follows:

$$\mu = \mu' - j\mu'' \quad (7)$$

with μ' and μ'' being the real part and imaginary part of the magnetic permeability. Hence, the magnetic loss tangent $\tan \delta$ can be defined as:

$$\tan \delta_\mu = \frac{\mu''}{\mu'} = \frac{\gamma}{2\pi f \mu_0 \mu'} \quad (8)$$

The crucial parameter governing the propagation of electromagnetic waves through any medium is its impedance (Z_M). The relationship between material impedance, permeability, and permittivity is expressed by:

$$Z_M = \sqrt{\frac{\mu}{\epsilon}} \quad (9)$$

In the quest to minimize the reflection of incident rays back to the radar system, Equations (10) describing the reflection coefficient r of a surface is given below:

$$r = \frac{Z_M - Z_0}{Z_M + Z_0} \quad (10)$$

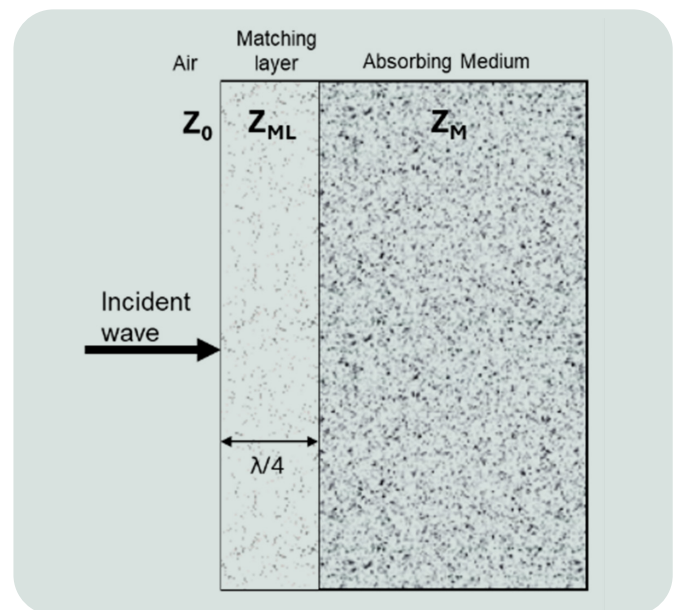
Consequently, in accordance with the principle of impedance matching, the reflection coefficient of a substrate reaches zero when the characteristic impedance of the medium equals that of the surrounding medium (i.e. $Z_M = Z_0$). On this matter, the intrinsic coefficient of free space Z_{fs} is expressed as:

$$Z_{fs} = \frac{E}{H} = \sqrt{\frac{\mu_0}{\epsilon_0}} = 377 \, \Omega \quad (11)$$

the magnetic field factor, μ_0 and ϵ_0 are the magnetic permeability and the dielectric constant of the free space. Hence, the reflection from a substrate will approach zero if its intrinsic impedance is approximately $377 \, \Omega$ [9]. Conversely, the impedance value of metallic aircraft is close to zero. Consequently, when signals encounter a metallic surface, a significant portion of them is reflected. Besides aligning with the impedance value of air, an effective RAM material must rapidly absorb electromagnetic radiation energy as it passes through. To fulfil this requirement, the surface of an aircraft must be coated with RAMs. However, as described in the following sections, the energy absorption that can be obtained by integrating RAMs highly depends on their thickness. The most commonly used RAMs and relative drawbacks are illustrated in the following subparagraphs.

Impedance Matching layer

Matching layer RAMs serve as intermediary layers positioned between the incident surface and the absorbing media to reduce the thickness required for gradual transition materials [4][10]. With reference to Figure 3, thickness and impedance Z_{ML} of the transition layer lie between those of the incident medium and the RAM. The purpose of these RAMs is to equalize the impedance of the incident medium with the combined impedance of the first and second layers. This impedance matching is achieved when thickness of the matching layer is one fourth of the wavelength of electromagnetic radiation within the layer. However, impedance matching is only attained at frequencies equal to the optical thickness, which limits the utility of matching layer materials to narrow-band RAMs. To absorb electromagnetic radiation at microwave frequencies, these RAMs are fabricated with an intermediate impedance transition layer alongside controlled quarter-wavelength thickness of incident waves.



3-Impedance matching layer

Tapered Loading Absorbers

Tapered loading absorbers comprise a blend of a lossy material and a low-loss material, typically arranged in a slab-shaped structure [3][11].



4-Tapered loading absorber

The lossy material is uniformly dispersed parallel to the incident surface, while the impedance gradient increases perpendicularly into the material. A 3D plastic net or open-end foam is fabricated by applying the lossy material through methods such as dipping, spraying, or drainage and drying. However, achieving reproducible fabrication of impedance gradients in the same manner is challenging with these 3D plastic net RAMs. Alternatively, another variant involves stacking homogeneous layers of material with increasing loading of the lossy component in the direction of the electromagnetic wave propagation, thus generating an impedance gradient as a step function (Figure 4). Although these RAMs offer advantages over pyramid RAMs in terms of thickness, their performance tends to be weakened. Additionally, adjusting the impedance gradient over one or more wavelengths can be beneficial.

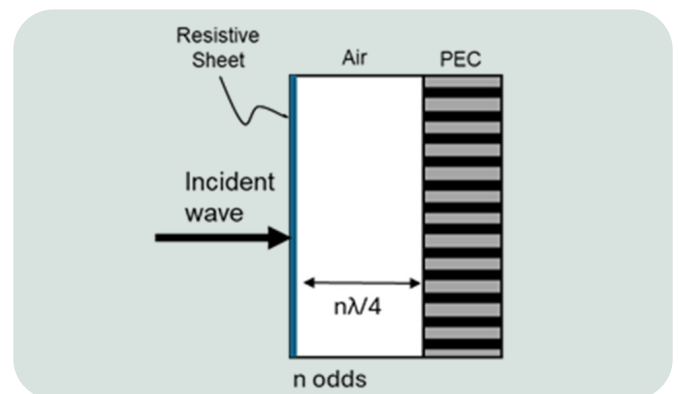
Salisbury screen

The Salisbury screen differs from the quarter wavelength impedance matching layer, since it does not depend directly on the permeability and permittivity values of the bulk layer [4][12]. The screen consists of a resistive sheet positioned at odd multiples of one-fourth of the wavelength in front of a conducting (metal) back, typically kept separated by an air gap (Figure 5). Alternatively, the air gap can be substituted by a material possessing high permittivity, albeit at the expense of bandwidth due to reduced gap thickness. Drawing from the transmission line theory, a quarter-wavelength transmission line transforms short circuits on the metal into open circuits on the resistive sheet. The effective impedance of the structure hinges on the gap between the metal backing and the resistive sheet; halving the gap to the wavelength size reverts short circuits, leading to perfect reflection of electromagnetic waves. Conversely, if the resistance of the sheet equals the impedance of air, an impedance matching occurs that results in zero reflection.

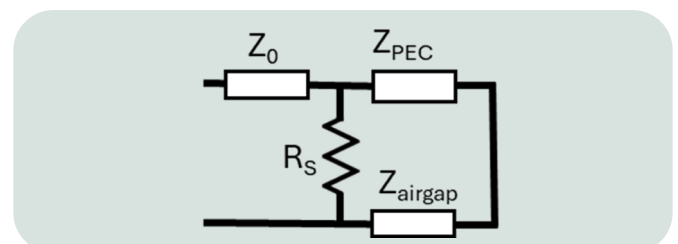
Assuming an electrically thin layer (i.e. $1 \gg k_0 d \sqrt{|\epsilon_l|}$) where $k_l = 2\pi/\lambda_0$ denotes the angular frequency of the resistive layer with thickness d and electric permittivity ϵ_l , whose internal loss is due to its conductivity (i.e. $\epsilon'' \gg \epsilon'$), the optimal thickness of the screen can be determined when the sheet resistance matches the impedance value of the free space, utilizing the following formula [4]:

$$d = \frac{1}{Z_{fs}\gamma} \quad (12)$$

The mentioned approximation may not accurately represent practical devices, as the observed results indicate a shift in resonant frequency towards smaller values with increased thickness of the resistive sheet or permittivity (ϵ_0). This phenomenon has been demonstrated by using a transmission line model with a RLC circuit representing the Salisbury screen (Figure 6) [13].



5-Salisbury Screen



6-Salisbury Screen: RLC equivalent circuit

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INNOVATIVE SOLUTIONS

Advancements in radar-absorbing materials (RAMs) have witnessed a notable shift towards innovative solutions, departing from conventional and State-of-the-Art (SoA) materials. Emerging developments performed by *LAD S&I/LLabs* in synergy with *LAD Eng* focus on leveraging novel materials and metamaterials to achieve unprecedented levels of radar stealth. The identification of novel material solutions allows for improvements in low observability performances, especially in highly restricted and challenging zones of the aircraft, where no further shaping adjustment can be done, due to the aerodynamic factor. These advancements encompass both the passive and the active approaches, including metasurfaces and active metamaterials, as well as novel nanotechnology-based solutions. In this section, the cutting-edge research and development efforts of *LAD S&I/LLabs* are described, which are aimed at achieving high radar absorption performance on a wide and challenging frequency band with thin (subwavelength) solutions, also emphasizing the unique properties of these innovative RAMs in redefining the landscape of radar stealth technology.

Novel nanotechnology-based RAMs

The optimal Radar Absorbing Material (RAM) should exhibit robust attenuation capabilities across a broad frequency spectrum, and feature low density, excellent thermal stability, cost-effectiveness, and minimal thickness. The approach pursued by the department *LAD S&I/LLabs* on this matter relies on the development of hybrid nanofillers to integrate in aerospace materials for improving stealth performance, whose filler concentrations can be tuneable according with the operating frequency range. Therefore, these materials boast low densities, high and adjustable conductivities, expansive surface areas, resistance to corrosion, reduced weight, and remarkable electromagnetic (EM) wave absorption capabilities.

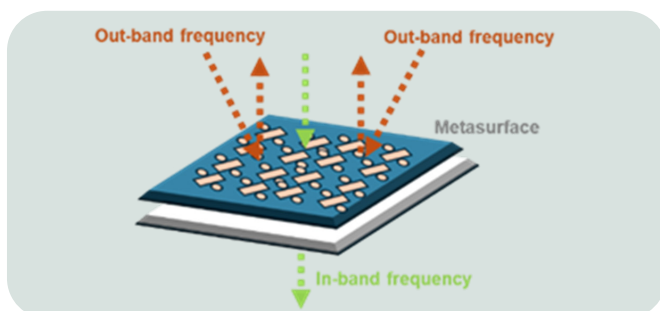
Wideband metamaterials design

Wideband metasurfaces represent the cutting-edge approach for advancing low observability applications, particularly in stealth technology. Leveraging intricate geometrical and material configurations at subwavelength scales, the *LAD S&I/LLabs* research efforts are focused on design of ultrathin (subwavelength) metamaterials to achieve precise control over electromagnetic wave interactions, underscoring their significance in the pursuit of advanced stealth technologies. By modulating reflection, absorption, and transmission characteristics with high precision, passive metasurfaces enable the engineering of the electromagnetic responses tailored to specific stealth requirements. Their ability to manipulate electromagnetic waves at nanoscale dimensions empowers the creation of structures with unprecedented stealth capabilities, effectively mitigating the radar detection across the required frequency bands.

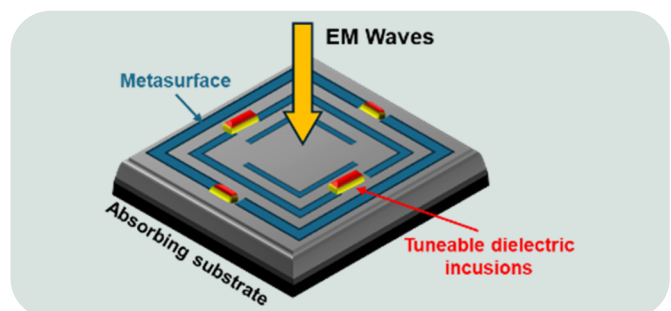
Active tuneable metasurfaces

Active metasurfaces represent a groundbreaking approach for achieving tunable radar absorption capabilities in low observability applications. By integrating active elements such as diodes, transistors, or varactors into metasurface structures, the systems under development by the *LAD S&I/LLabs* can dynamically adjust their electromagnetic properties in response to external stimuli.

This enables real-time control over reflection, absorption, and transmission characteristics, offering unparalleled adaptability and flexibility for radar stealth. Through precise modulation of the active elements' conductivity or impedance, active metasurfaces can actively manipulate electromagnetic waves, effectively camouflaging or concealing objects from radar detection, across a wide range of frequencies. Their ability to dynamically tune radar absorption properties makes them invaluable for next-generation stealth technologies, paving the way for highly adaptive and responsive low observability solutions.



7-Wideband metamaterials



8-Active tuneable metasurface

CONCLUSIONS

The comprehensive overview of the SoA in RCS reduction technology presented in this work highlights the critical role of low observability in enhancing the survivability of military systems. By exploring innovative technologies and treatment

methods, as well as by evaluating design techniques, future advancements in aerospace applications are poised to further refine stealth capabilities and to address emerging challenges in radar detection.

Francesco Acerra: francesco.acerra@leonardo.com

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Antenna Siting Design Approach for Optimization of Antennas Positioning on Fixed Wing Aircraft

Giuseppe Nicolò, Francesco Marano
Leonardo – Aircraft Division

This article introduces a simulation-based approach aimed at optimizing the placement of antennas on aircraft platforms by estimating their radiative performances and assessing the Radio Frequency Interoperability between transmitter and receiver systems. This analysis enables the prediction of potential issues during the design phase, allowing for adjustments to antenna locations on the aircraft or implementation of mitigation actions. Key aspects of this approach include the generation of electromagnetic models for both the aircraft platform and antennas, followed by their validation using commercial computational tools that leverage various solver techniques. Moreover, the article outlines a specific troubleshooting case conducted on a trainer platform: investigating anomalies in Radio Altimeter indications using a numerical approach to replicate failures in a simulation environment. This process facilitated the identification of a solution without encountering critical issues.

INTRODUCTION

Modern air vehicle platforms rely on a wide range of avionics systems that operate across diverse radio frequency spectrums, encompassing communication, safe navigation and self-protection functionalities. These systems have become standard equipment, highlighting the necessity of designing and evaluating their functional and operational performance.

Each system requires one or more dedicated antennas and/or sensors installed on the aircraft's external surfaces. However, this installation may introduce unwanted mutual interaction or distortion effects due to the fuselage. Such effects can significantly degrade antenna radiation patterns and alter antenna electric characteristics, potentially impacting the performance of related radio frequency equipment. Hence, during the design phase, it's crucial to accurately predict and assess these phenomena to develop avionics systems effectively [1].

Aircraft upgrades often involve integrating new systems and antennas, while ensuring the continued smooth operation of existing equipment. The installation of antennas on fixed-wing aircraft must adhere to aerodynamic and mechanical constraints, while maintaining their electromagnetic performance.

In this context, antenna siting design becomes pivotal in verifying electromagnetic performance and in suggesting relocation or mitigation actions to maintain the desired electromagnetic environment. Thus, the antenna location represents a delicate balance between aerodynamic/mechanical constraints and electromagnetic performance [1].

The results of antenna siting analysis serve as inputs for defining the electromagnetic compatibility (EMC) test requirements that are essential for the EMC certification process. Additionally, they aid system engineers in analysing system performances and defining the Blanking Matrix.

Other critical aspects of the antenna siting approach include estimation of the electric fields radiated by transmitter systems, especially in the Near Field, to evaluate the Radiation Hazards requirements concerning personnel, fuel vapour, external stores (such as missiles, bombs, guns), or obstructions (like Electro-Optic Turrets, PODs, etc.).

Moreover, the verification of specific radio frequency certification requirements applicable to particular systems or other regulations is supported by numerical analysis of the radiation pattern degradation.

ANTENNA SITING APPROACH

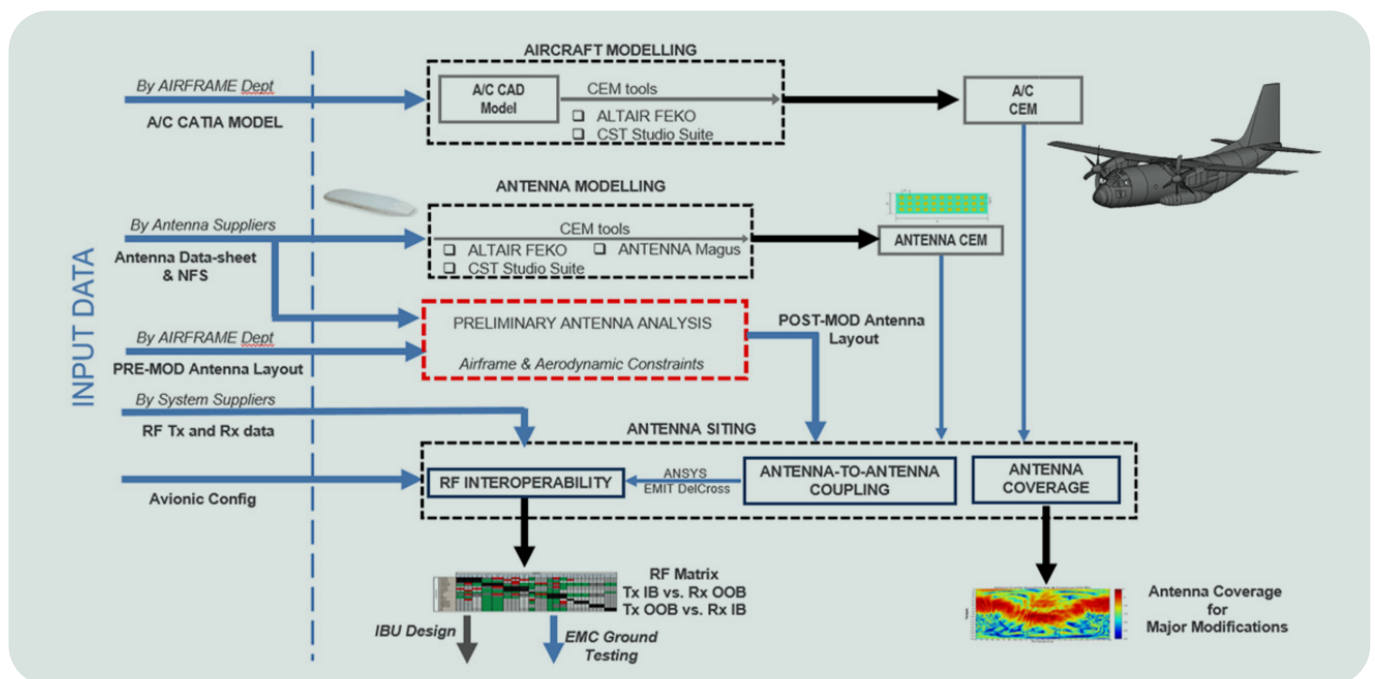
The precision of electromagnetic (EM) numerical analysis and predictive accuracy relies mainly on establishing precise and reliable geometric models for both the aircraft (A/C) and its antennas from an EM perspective. Initial inputs include CATIA A/C geometries and data sheets provided by antenna suppliers. The selection of the most suitable numerical method is guided by factors such as the antenna's operating frequency band and its electrical dimensions.

The utilization of commercially available tools equipped with comprehensive EM numerical techniques enables effective management and control of all potential EM interactions between installed antennas and the entire aircraft structure. Parameters such as Antenna-to-Antenna Coupling values, Antenna Radiation Patterns, and Antenna Near Field Iso-Surface values can be computed numerically, visually represented, and exported for further post-processing analysis.

Results derived from Antenna Radiation Patterns offer system engineers insights into potential losses in the system's link budget and radio frequency (RF) blocking caused by the aircraft during specific flight profiles. Antenna-to-Antenna coupling values serve as inputs for investigating potential interferences between transmitter and receiver pairs, thereby informing EMC ground testing.

The Near Field Iso Surfaces predict the effects of transmitter equipment on external obstructions or personnel and fuel vents. Throughout the design phase and beyond [2], the computational results can streamline the testing process by utilizing numerical predictions as the basis for informed and efficient measurement preparation and planning.

Figure 1 illustrates the typical approach to antenna siting on a fixed-wing aircraft, which is detailed in the following paragraphs.



1-Antenna Siting Approach

MODELLING

The Computational Electromagnetic (CEM) techniques are based on the numerical solution of the Maxwell's laws that describe and predict the complete electromagnetic behaviour in any structure or region, given an initial state and a boundary value. For the numerical analysis, it is necessary to create a virtual geometrical representation of the structure of interest and divide this model into a finite number of special blocks, or elements. The last modelling step is the creation of the EM model of the antenna, which represents the initial condition of the simulation. Two principal kinds of EM model are possible: a full wave and a synthetic model.

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Aircraft Modelling

The virtual geometric model for aircraft begins with the CATIA file, which encompasses intricate structural components such as the platform's framework, avionic bays, bulkheads, windows, and ribs. The level of detail within this file varies, primarily contingent upon the frequencies of interest. Detailing can be intensified in specific areas of concern. Consequently, to address the electromagnetic (EM) challenges at hand, the CATIA model undergoes the refinement process that eliminates structural elements with minimal EM impact.

In the analysis of antenna placement, the final EM model emphasizes external surfaces, excluding minute details whose dimensions are negligible with respect to the investigated wavelength. This model is then segmented into elements – surface meshes or cuboidal volumes – whose sizes are dictated by the frequency of the problem at hand. The choice of the mesh type is contingent upon the computational electromagnetics (CEM) techniques employed.

The precision of the meshing process hinges on fidelity of the CATIA-derived model. Selecting which details to omit is a nuanced task, as there is no universal guideline. The level of the model precision is determined by the specific EM problem and by the frequency range under scrutiny.

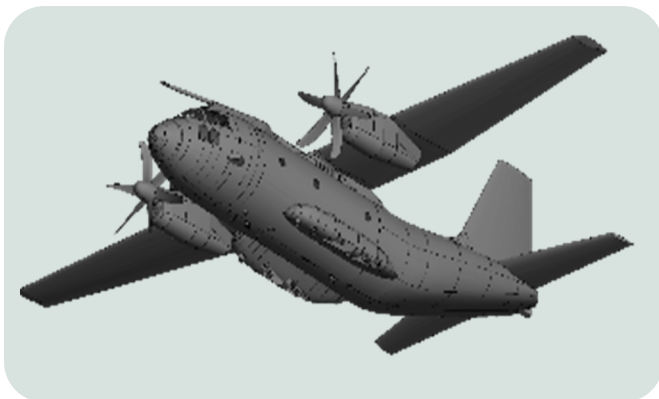
While minor structural elements may be disregarded for far-field antenna analysis, they remain pertinent for scenarios involving near-field effects.

Conversely, at higher frequencies, where detail sizes align with the wavelength under test, the proliferation of mesh elements may prolong the computational runtime and necessitate ample RAM resources.

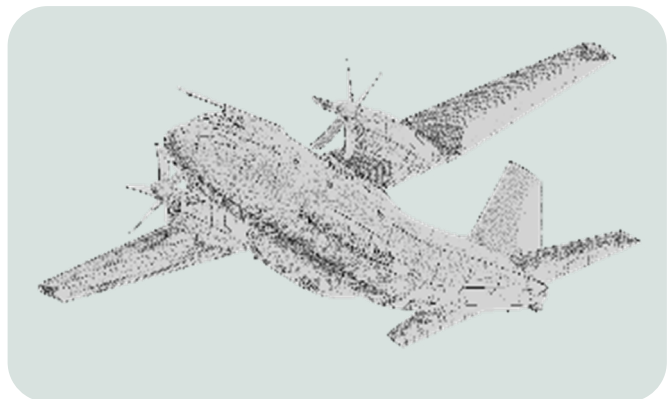
Thus, the EM model is tailored to accommodate the operational frequency band, the nature of the EM problem, and the available computational resources.

Figure 2 and Figure 3 juxtapose the aircraft EM model, streamlined to remove negligible electromagnetic details, with the discretized model that comprises smaller elements slated for analysis.

In preliminary assessments, the aircraft platform is typically modelled by using Perfect Electric material [1]. This is a conservative approach that minimizes reflection effects, and thereby fosters confidence in its results. However, the increasing prevalence of composite materials, aimed at reducing aircraft weight and costs, necessitates the integration of their electromagnetic properties. Key considerations include the impact on antenna radiative performance and the interplay between antennas.



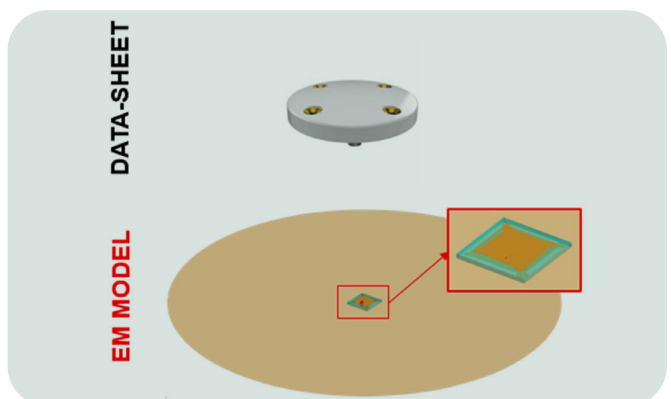
2-CEM A/C model: geometry



3-CEM A/C model: surface meshes

Antenna Modelling

The creation of an EM antenna model relies mainly on both the antenna's complexity and the data provided by the supplier. For standard antennas like blade, patch, monopole, and towel bar antennas used in communication, navigation, GPS, IFF/TCAS, and HF communication systems, respectively, an accurate EM model can be developed by using the fundamental principles of the electromagnetic theory and the electrical specifications outlined in the supplier's data sheet. Validation entails comparing simulation results with crucial parameters such as antenna gain, impedance, VSWR, and return loss. Figure 4 illustrates this process that outlines the proceeding from the data sheet to the final EM model.

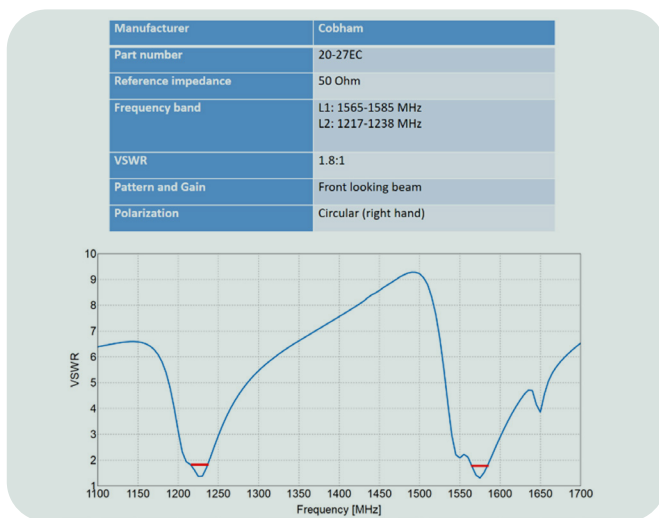


4-Antenna: Geometrical and EM model

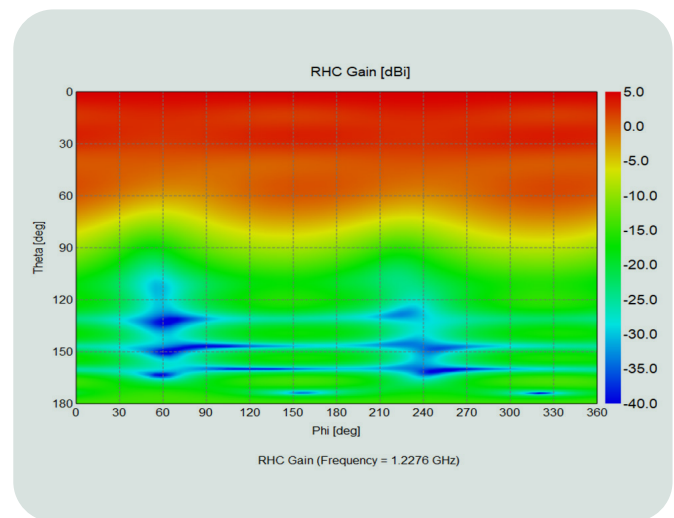
Figure 5 then juxtaposes the supplier's declared electrical characteristics with the EM model's results, focusing on its Voltage Standing Wave Ratio (VSWR) and pattern. Additionally, Figure 6 delves into the Right-Hand Circular Polarization (RHCP) Antenna Gain component, showcasing the gain distribution in 3D space based on the solved EM antenna model, as represented in Mercator's Projection.

However, as depicted in Figure 7, complex antennas such as phased arrays or electrical beam steering antennas, present a different challenge. Here, engineering expertise and antenna design theory are indispensable,

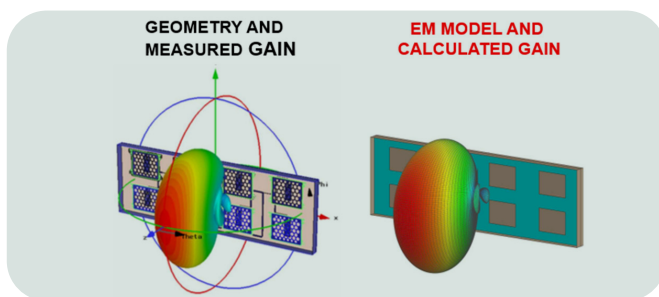
particularly because suppliers often withhold information due to their intellectual property rights. The design process involves selecting parameters such as dielectric material, profile, and size of individual radiating elements, while considering the wavelength under test. Radiation patterns and array dimensions determine the number of radiating elements. To streamline computational complexity, planar multiplayer substrates are utilized. Adjustments to feed amplitude and phase ensure alignment with the desired beam characteristics. Subsequently, the feed network is integrated into the overall antenna model for EM analysis.



5-Antenna: Data-sheet and Simulated performances



6-Antenna: 3D Simulated RHCP [dBi] Antenna Gain



7-Patch Array antenna model

In cases involving highly intricate antennas, suppliers may offer synthetic sources that include calculated or measured electric and magnetic fields [3] around the antenna. These sources serve as representations of the antenna model and enhance accuracy within specified operating bands. Alternatively, encrypted sources, resembling black boxes that contain the physical antenna without divulging proprietary details, may also be provided. While synthetic sources assure accuracy within operating bands, encrypted sources offer confidence across the entire spectrum.

ANTENNA COVERAGE ANALYSIS

The installation of antennas on an aircraft fuselage can significantly alter their radiation patterns due to the presence of metallic structures, curved surfaces, and external obstructions like fairings, turrets, and PODs. The antenna coverage involves a theoretical study aimed at predicting distortions in antenna gain, potential RF blocking, and depolarization phenomena.

To establish a baseline, engineers refer to the concept of the antenna In Isolation, which represents the antenna installed on a ground plane according to specifications provided by the supplier. For specific antennas like IFF interrogators and RADAR antennas, the In Isolation performance is evaluated with the antenna in free space. This evaluation entails the comparison of the radiation patterns calculated for the antenna In Isolation with those estimated for the antenna installed on the aircraft. Figure 8 illustrates such comparison, depicting a blade antenna installed on a ground plane (In Isolation) alongside the same antenna installed on an aircraft.

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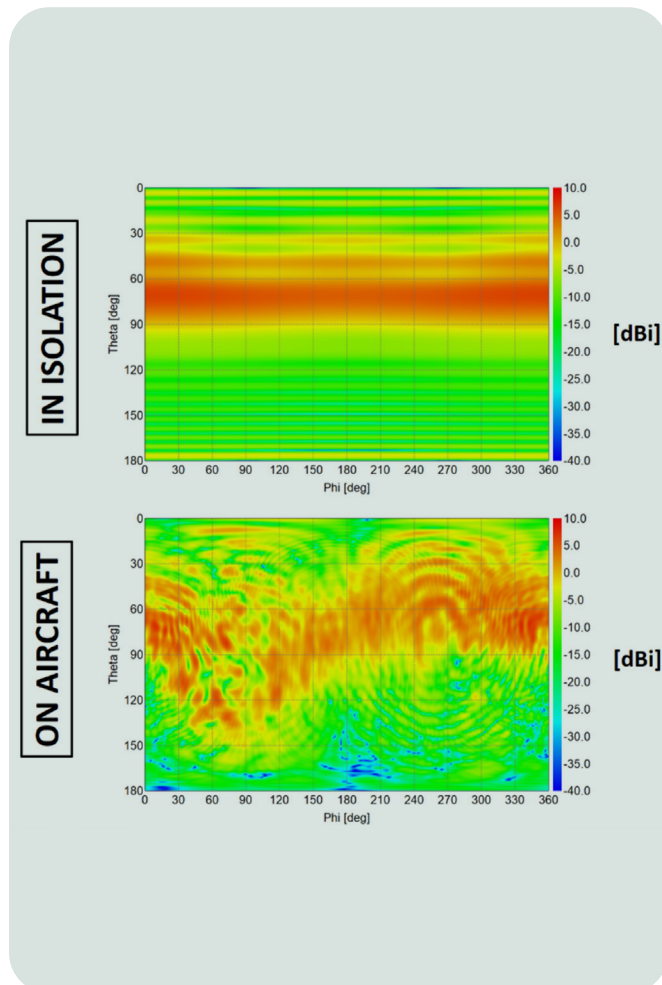
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These results, presented as the distribution of the radiated field in 3D space, furnish vital insights to system engineers. This data aids in defining system performances, such as the system range and the shielding effects at various flight attitudes. Additionally, Figure 9 illustrates an example of RF blocking caused by the presence of a refueling probe on the radiation

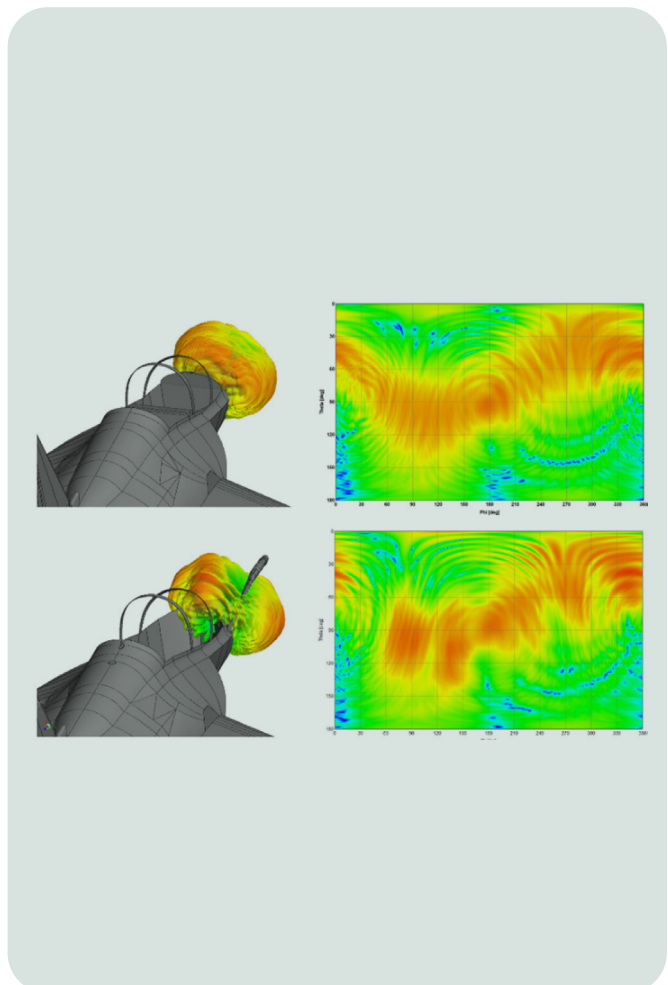
pattern of an antenna.

Simulations conducted on aircraft without the refueling probe enable validating the initial hypothesis of RF degradation.

The external obstacle induces RF blocking, resulting in gain holes and a redistribution of the radiation patterns, and ultimately amplifying antenna gain ripples.



8–Mercator's projection representation



9–Trainer Platform: example of geometry obstruction on antenna pattern

ANTENNA COUPLING ANALYSIS

The theoretical antenna coupling study is carried out to estimate the isolation between transmitter and receiver antenna pairs. It represents the portion of the transmitted power (useful signal or harmonics and spurious) captured by the receiver antenna, the *unwanted signal*. The Antenna Coupling Factor values are the inputs for the Radio Frequency (RF) Interoperability Analysis, for estimation of the Electromagnetic Margin of Interference.

RF INTEROPERABILITY ASSESSMENT

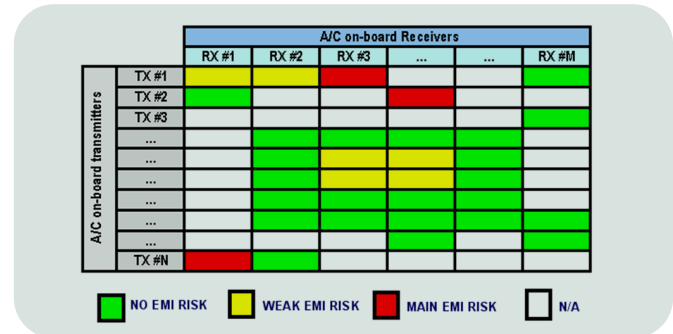
Using the results of the antenna coupling analysis and taking into account the RF transmitter and receiver characteristics (transmitted power, receiver sensitivity, EMI certification levels), an RF interoperability assessment is produced, to predict potential interferences between on board transmitter and receiver.

The analysis is performed in different scenarios, which are:

- **Transmitter In Band vs. Receiver Out Of Band:** when the transmitter operates in its band, fundamentals could cause a spurious response of the receiver;
- **Transmitter Out Of Band vs. Receiver In Band:** transmitter spurious/harmonics could fall inside the receiver operating band;
- **Transmitter In Band vs Receiver In Band:** when there is an overlap, partial or global, between the functional bands of the transmitter and the receiver systems.

For each scenario, a threat/victim matrix is generated. Finally, one threat/victim matrix can be built by summarizing the degree of interference risk between transmitter/receiver pairs. Figure 10 shows an example with the typical colours code.

The threat/victim matrix allows identifying possible mitigation actions, like the integration of filters along TX and/or RX lines, to reduce or avoid the interference [4]. In addition, the threat/victim matrix is an input to the EMC department to address the ground test campaign.



10 – Threat/Victim Matrix

SPECIFIC EXAMPLE: RADIO ALTIMETER TROUBLESHOOTING ACTIVITY

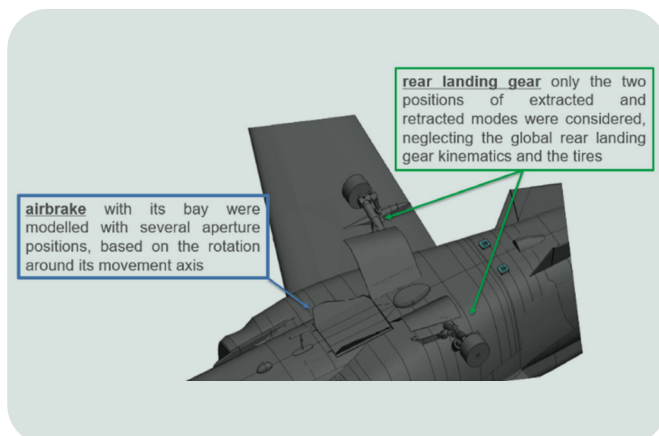
During flight operations on a trainer platform, anomalies were observed in the Radio Altimeter's altitude indication from the ground. Analysis of the flight telemetry data revealed that such irregular readings were triggered by the deployment of the airbrake and/or the extension of the main landing gear. Specifically, when the aircraft exceeded the Radio Altimeter's maximum altitude threshold of 5000 ft, even a slight opening of the airbrake or landing gear caused the false indication of a zero altitude, effectively resetting the maximum altitude reading.

To address this issue, a simulation approach was employed to identify a solution that would prevent these anomalous indications, thus reducing the need for additional investigation through flight tests [5]. The complexity of the simulation approach stemmed from:

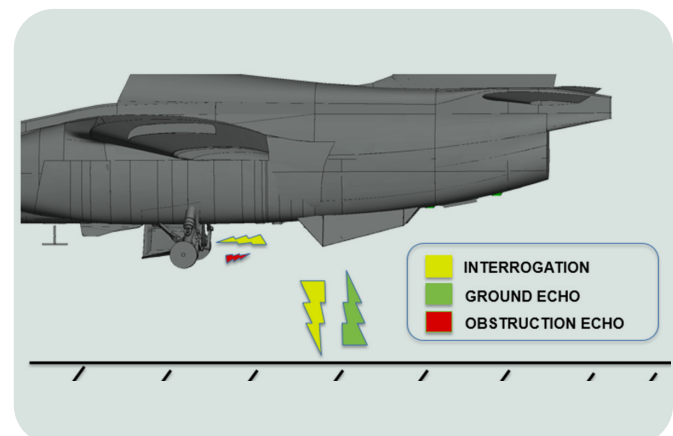
- Incorporating movable geometries into the aircraft model, to accurately capture all the contributing factors affecting the Radio Altimeter receiver antenna (refer to Figure 11);
- Ensuring that the electromagnetic (EM) problem could be solved efficiently within shorter run-times, particularly considering the higher operating frequency of the Radio Altimeter.

Depending on system functionality, incorrect altitude measurements within the operational range could stem from spurious echoes produced by the deployment of the airbrake and/or by the extension of the landing gears. These echoes might surpass in magnitude the signal reflected by the terrain.

Interestingly, the radio altimeter filter appears ineffective in eliminating these spurious echoes when the maximum measured altitude of 5300 ft is exceeded.



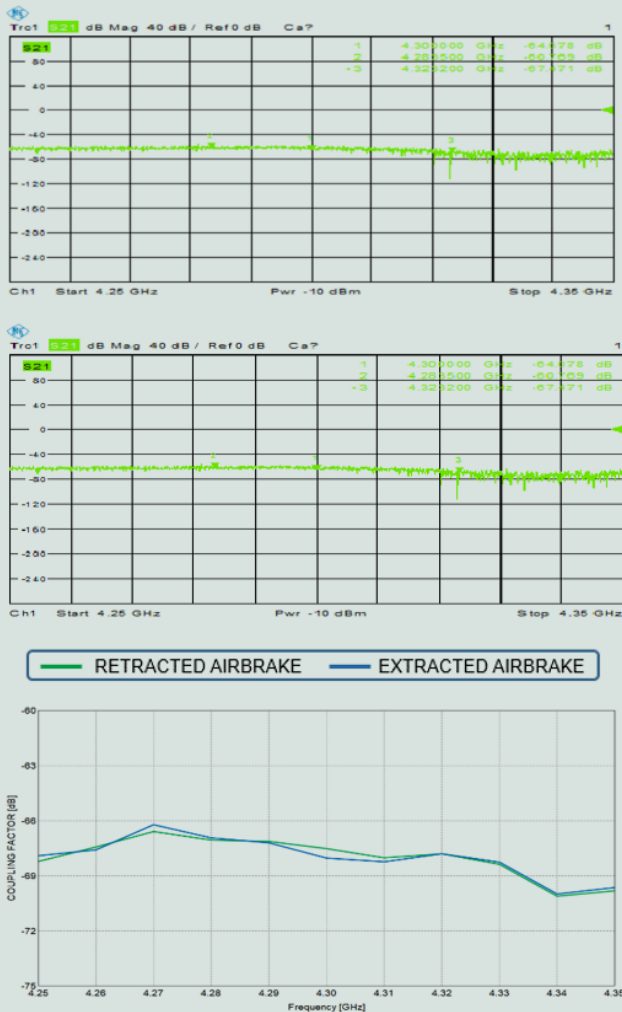
11 – Trainer platform: movable geometries model



12 – Terrain and aircraft echoes at Radio Altimeter Antenna

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13-Validation of EM Model

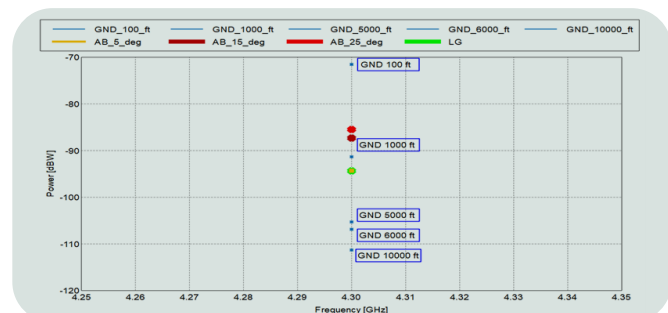
When the aircraft flew below that maximum limit, the equipment filtered the spurious echo. However, when the aircraft flew over 5000 feet the equipment continued its functionality, but its data was considered invalid and the first helpful signal was detected (Figure 14).

Based on the results of this analysis, the Radio Altimeter receiver antenna was relocated, by increasing its distances from landing gears and airbrake. Such new antenna location allows avoiding the Radio Altimeter anomaly in indication of the quote.

These considerations address the simulations in the estimation of the received signal at the radar altimeter antenna, caused by the single contribution of:

- *The terrain*, varying the altitude from 100ft to 10000ft. The terrain is modeled like an infinite PEC ground plane and the radar altimeter transmits the signal with the known power. The signal reflected is used to estimate the power received by the radar altimeter antenna;
- *The airbrake*, extracted at different inclination that varies from 5° to 25°. By increasing inclination, the reflected area grows up. The simulations were performed in free space to avoid any interference from terrain. The signal transmitted by the Tx radar altimeter antenna causes a reflection on the airbrake. This reflected signal is applied to estimate the power received by the radar altimeter antenna;
- *The landing gears*, completely extracted. The simulations were performed in free space to avoid the terrain contribution. The signal transmitted by the Tx radar altimeter antenna causes a reflection of the landing gears. This reflected signal is applied to estimate the power received at the radar altimeter antenna.

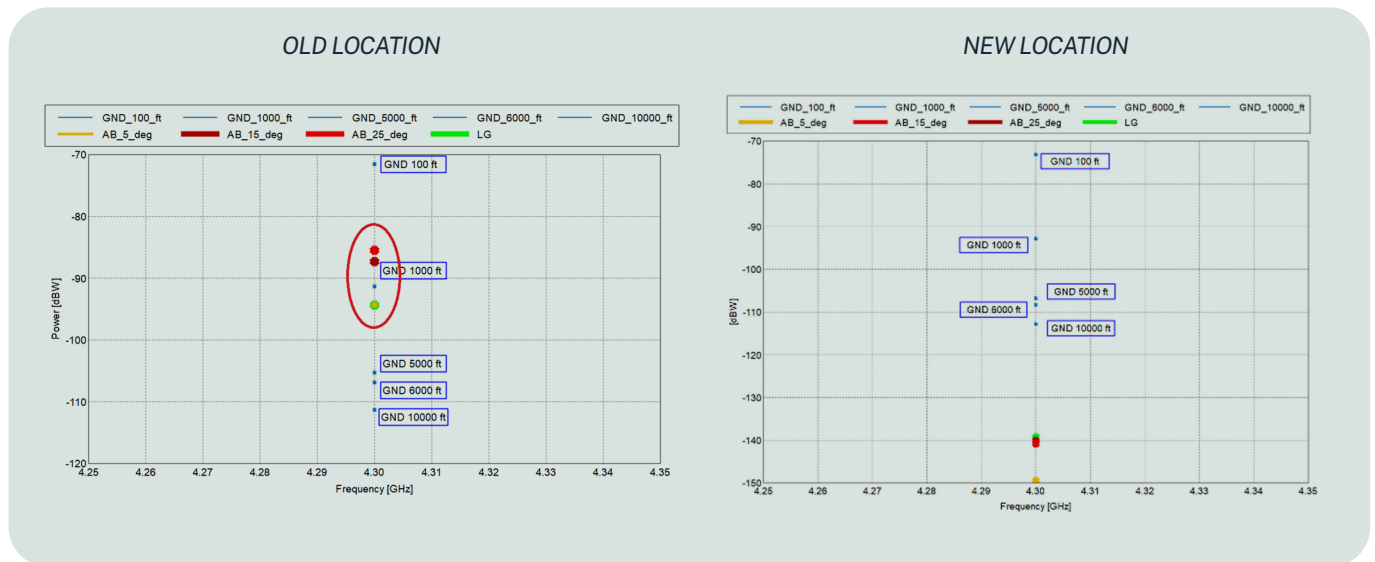
In order to compare the received power levels caused by the terrain, the airbrake and the landing gears, the reflected signal was calculated over an area of the same dimensions, at the same distance between the receiver radar altimeter antenna and the obstacle (terrain/airbrake/landing gear). The ground was modeled as an infinite Perfect Electric Conductor and its contribution at the receiver Radio Altimeter antenna was estimated by varying the aircraft quote, over 5000ft. These values represent the benchmark to identify potential sources of spurious echoes (Figure 12). Clearly, the simulation campaign was carried out by validating the EM models of aircraft and antennas. In particular, a measurement of the coupling between transmitter and receiver Radio Altimeter antennas was compared with the calculated one, by applying a numerical simulation (Figure 13). The simulation campaign shows that the airbrake or landing gear echoes detected by the Radio Altimeter receiver antenna was greater than the signal reflected by the terrain over 5000 feet.



14-Calculated Received Power at RA RX antenna

Figure 15 shows that in the former antenna location the spurious echoes (red circle) are higher than the ground echo, in particular at 5000ft, while in the new antenna location the level of spurious echoes keeps lower.

This trouble-shooting analysis addresses the relocation of the Radar Altimeter antenna on the aircraft, while the flight tests confirm its good results in avoiding any anomalous indication of the quote.



15–Calculated Received Power at RA RX antenna: OLD vs. NEW LOCATION

CONCLUSIONS

This paper presents an approach to antenna siting used to study the antenna placement on fixed-wing platforms from the electromagnetic (EM) perspective. This approach involves modelling both the platform and the antennas, by using commercial tools. Through simulation campaigns, the antenna radiation performance and the radio frequency interoperability between on-board transmitters and receivers can be explored. This helps to mitigate design risks and to identify critical issues beforehand.

Moreover, this type of analysis enables investigating broader electromagnetic problems.

By utilizing EM models of aircraft and antennas, various aspects such as support for specific radiofrequency regulation certifications, Radiation Hazard analysis, and evaluation of electric fields around external obstructions can be analysed. Consequently, simulations have the potential to pre-empt and, in many cases, to substitute testing campaigns, thereby reducing efforts in terms of time and costs.

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Giuseppe Nicolò: giuseppe.nicolo@leonardo.com
 Francesco Marano: francesco.marano@leonardo.com

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Variable Emissivity Coatings Based on Emerging Nanostructured Materials for Adaptive Thermal Camouflage

Bruno Paulillo¹, Alessio Bassano², Marco Morini², Paolo Arrighi², Abhishek Kumar¹

¹Leonardo – Innovation Labs, ²Leonardo – Electronics Division

Adaptive thermal camouflaging requires materials capable to control efficiently and in real time the thermal radiation emitted from a surface. A promising strategy relies on coatings whose tuneable infrared emissivity can dynamically reshape the thermal signature of the objects they cover. Recently, several active surfaces have been reported that are based on emerging nanostructured materials showing large modulation of infrared optical properties upon electrical actuation. In this paper, the working principle, performances and potential application of the most some promising platforms are described, some of which are currently investigated in Leonardo.

INTRODUCTION

“Pretending not to be there or, at least, to be something different”: this is the common backbone of camouflage strategies evolved in nature. It is as well the driver of the search for new “invisibility” technologies for military applications.

To devise a camouflaging system, it is necessary to remember first what ‘seeing’ means. Our perception of objects depends on how they interact with visible light: the surface of an object partially reflects, transmits, diffuses and/or absorbs the ambient light. The part of visible light that reaches our eye “encodes” information about the surface optical properties, that renders its visual appearance. The possibility of ‘being seen’ is closely related to the concept of optical contrast. Our eye is a sophisticated sensor capable of detecting the presence of something if there is even a minute contrast of optical properties with respect to the background. In turn, all camouflaging strategies found in nature or applied to military operations are based on creating illusions, which means reducing optical contrasts by playing with colours and shapes. *Static camouflage* patterns are conceived to blend objects into a certain landscape, whereas *adaptive camouflage* solutions allow to disappear into a dynamically changing background.

Masters of this art are cephalopods, which feature sophisticated, neurally controlled adaptive skin that contains microstructures capable of generating colour patterns (chromatophores) and iridescence (iridophores) to mimic the external environment [1].

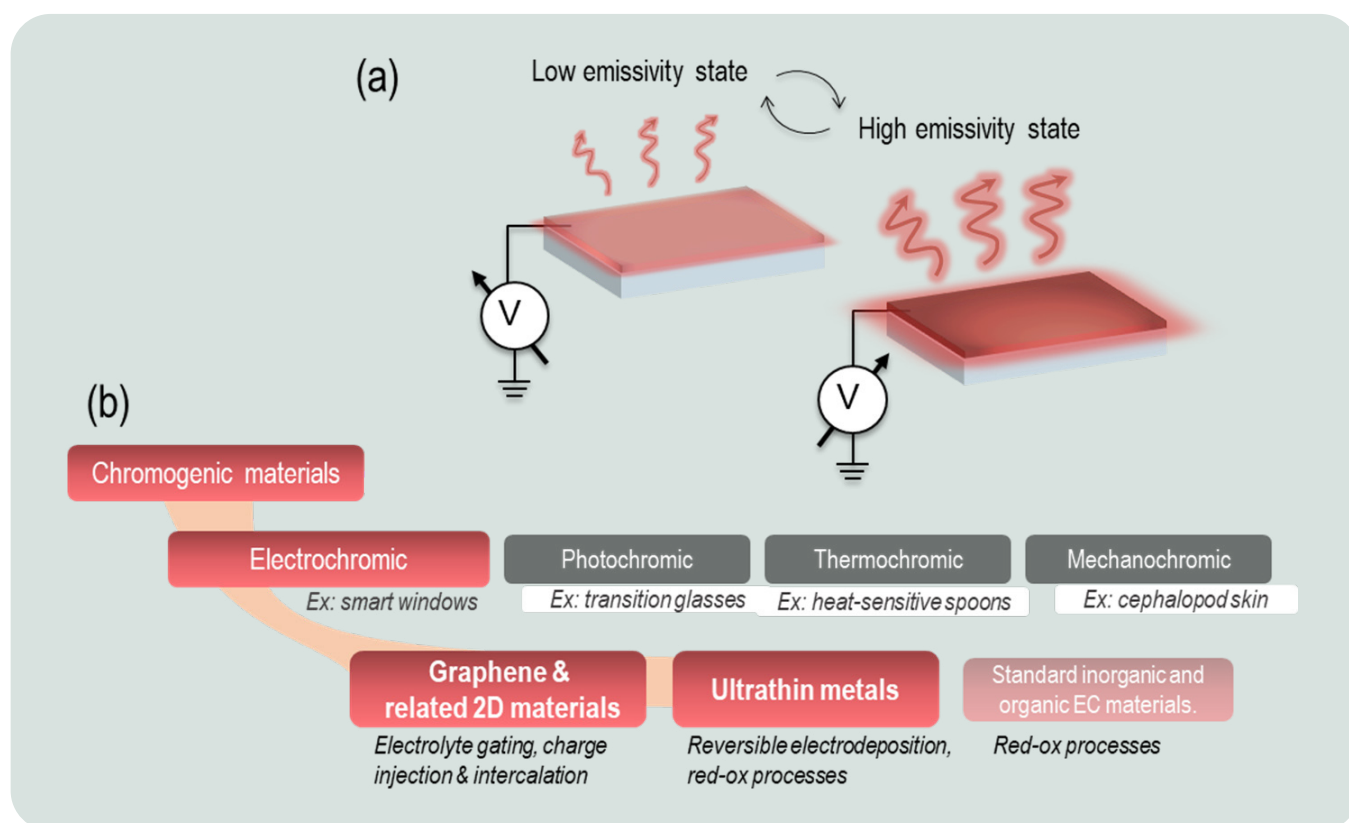
In particular, an artificial skin featuring the above adaptive camouflaging functionality and operating at infrared (IR) wavelengths, would represent a disruptive countermeasure against IR imaging systems operating in the medium wave (MW) and long wave (LW) IR bands. The latter are today widespread in all operational domains to passively detect and identify targets owing to the fact that all bodies near room temperature emit radiation in the so-called thermal IR range (whose wavelengths are approximately between 0.75 and 15 μm). The emitted thermal radiation depends on both the object temperature (T) and its thermal emissivity (ϵ) according to the Stefan-Boltzmann law [2]. It's worth noting that ϵ is related to the IR absorbance (A) of the emitting surface according to the Kirchhoff law $\epsilon(\lambda)=A(\lambda)$ and that reflectance, transmittance and absorbance sum to one ($R+T+A=1$) owing to energy conservation.

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Hence, manipulating the surface IR optical properties of an object provides a powerful way to modulate the thermal radiation it emits and change its perceived 'IR signature', i.e. the contrast of apparent temperature between the object and the background measured by an IR camera [3]. Fabricating an active coating able to dynamically switch between high and low emissivity states upon an external control, as it is sketched in Figure 1(a), would then allow to reconfigure the IR signature of an object placed underneath. This approach is gaining momentum over more conventional strategies such as directly altering the surface temperature or projecting IR images on a low emissivity screen [4].

In this paper, we review some promising technologies recently employed to demonstrate variable emissivity coatings for active thermal radiation management applications, focusing in particular on electrically-tunable surfaces based on emerging nanostructured materials.



1-(a) Conceptual view of a tuneable emissivity surface;
(b) Chromogenic materials classes with some of their application (first row) and EC materials platforms with their underlying working principles (second row)

ELECTROCHROMIC SURFACES FOR IR RADIATION MANAGEMENT

Chromogenic materials

Implementing the above concept requires materials that can change their optical properties (e.g. spectral absorbance and reflectance) in the thermal IR as a function of an external stimulus, also known as chromogenic materials [5]. For example, the cephalopod skin can be considered a mechanochromic system because it is actuated by mechanical (neuro-muscular) stimulation. On the other hand, materials that change colour as a function of temperature (thermochromic) are employed in thermal sensing systems (e.g. heat-sensitive baby spoons) whereas photochromic glass

allows to regulate the transparency of transition lenses under external illumination.

However, mechanical, temperature and light stimuli are often impractical for reconfiguring the response of a device. Electrochromic (EC) materials allow instead to manipulate IR and/or visible optical properties by using an inexpensive electrical control (e.g. a low voltage) and are in general preferred for applications. Figure 1(b) summarizes the classes of available chromogenic materials and, in particular, the emerging EC materials platforms, this paper focuses on.

Electrochromic device architecture

EC materials explored in literature are typically sandwiched into a layered EC device (ECD) structure that allows to electrically reconfigure their optical properties [6]. It consists in:

- a gate electrode to apply the control voltage;
- an underlying electrolyte medium for electrically actuating the tuneable material;
- an “active” EC layer made of a material whose opto-electronic properties (e.g. the electrical conductivity σ for a conductive film) can be tuned by exploiting a certain physicochemical process.

Several tuneable materials have been explored in literature. In most cases, the change in IR optical properties stems from a change in the electrical conductivity of the active layer. In fact, the electrical conductivity σ and the IR reflectance R are related by

the Hagen-Rubens relation $R=1-4\sqrt{(f\pi\epsilon_0/\sigma)}$, meaning that for a conductive layer modulating σ will result in a change of its IR reflectivity [7]. This, in turns, directly affects the IR absorptivity since, for an opaque surface ($T=0$) $A=1-R$.

Let us stress here an important point: being optical properties so tightly related, a variable emissivity surface is *at the same time* a variable reflectivity surface. As a consequence, by dynamically tuning the emissivity of a surface, we are able to balance the quantity of thermal radiation emitted by the object placed behind, with respect to the quantity of thermal radiation reflected from the background scene.

In the following we discuss some promising materials platforms emerged in the latest years for this kind of applications.

Conventional EC materials

Standard EC materials include relatively thick inorganic and organic films. The prototype inorganic materials for ECDs are metal oxides. In particular, WO_3 films have been widely used e.g. for EC smart windows, thanks to their excellent tuneable coloration properties upon electrochemical reduction or oxidation. Furthermore, they have been employed in EC variable emissivity devices capable of modulating the thermal emissivity in the medium and long wave IR bands [8].

As per the organic EC layers, viologens and conductive polymers such as polyaniline (PANI) and PEDOT:PSS are

the most used films [9].

Such materials are more and more found in widespread applications. For example, the Boeing 787 Dreamliner is equipped with electronically dimmable EC windows that replace the traditional blinds. Thanks to advances in nanotechnology such “standard” materials are today available in nanostructured form (e.g. nanowires, nanoparticles) and/or combined in innovative composites (e.g. metal–organic frameworks) which further expands their applications.

EC surfaces based on graphene and related nanocomposites

Two-dimensional (2D) materials, such as graphene, hexagonal boron nitride (hBN), transition metal dichalcogenides (TMDs) and semiconductor quantum wells (QWs) offer extraordinary and unconventional opto-electronic properties that arise from nanoscale quantum confinement effects.

In particular, graphene is the atomically-thick allotrope of carbon and shows unique features such as broadband optical transparency, mechanical flexibility, ambipolar transport and high carrier mobility [11]. Furthermore, the band structure of graphene allows to control its doping level by electrostatic or electrochemical gating, yielding widely tuneable electrical conductivity and optical properties at IR wavelengths [3]. For this, it has become widespread for applications in transparent and flexible opto-electronic devices (e.g. sensors, modulators) and coatings. However, producing defect-free monolayer graphene is challenging and its large scale growth via chemical vapour deposition (CVD) is still highly expensive.

Furthermore, the base infrared absorptivity of single-layer graphene is insufficient to achieve the large emissivity modulation required for the IR camouflage application. Multilayer graphene (MLG) that comprises tens or hundreds of layers and is synthesized by CVD has instead demonstrated to be a promising platform for variable emissivity IR surfaces, when embedded into an ECD architecture and electrically actuated via electrolyte gating.

Salihoglu et al. [12] fabricated, for the first time, a thin and flexible ECD consisting of a three-layer stack including (i) a gold electrode, (ii) an IR transparent polymeric membrane soaked with an ionic liquid and (iii) a MLG active layer with optimized thickness. When a voltage is applied across the device, the ions contained in the electrolyte medium migrates towards the active layer and move in between the graphene planes. Upon this so-called intercalation process, charges are transferred to the MLG electrode, thus modifying its electrical conductivity

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and, in turn, its IR emissivity. With a bias of ≈ 3 V the surface emissivity is reduced from an initial value of ≈ 0.8 to a value of ≈ 0.3 , yielding a lower measured apparent temperature of the surface when observed through an IR camera. Authors show that by electrically actuating the surface one can reach a point where a “hot” object located behind the EC surface becomes “thermally invisible”. Such active surfaces allow to modulate the emissivity relatively fast (reported response time < 1 s) and can be adapted to operate in other spectral bands (visible, microwaves) [13]. However, the MLG used as

active layer remains so far a niche material produced on small scales and at a high cost. More recently, other graphene-based nanocomposites produced with low cost, scalable and industrially compatible techniques, have been developed to serve as active layers in IR tuneable emissivity surfaces and have shown the same functionality and similar tuning ranges of the MLG layers prepared by CVD [14][15]. A detailed review on graphene-based surfaces for adaptive thermal camouflage was recently published by IPCB-CNR, ENEA and Leonardo researchers [10].

EC surfaces based on ultrathin metal layers

Noble metal (e.g. gold, copper, silver) thin films (typically tens or hundreds of nm thick) made with large-scale physical vapour deposition (PVD) methods are widely employed in optoelectronic components such as electrodes, mirrors, filters, plasmonic waveguides and sensors. However, a new realm of possibilities popped up thanks to emerging PVD strategies that allow to deposit over large areas so-called ultrathin metal films that are just a few nm thick. At such small thicknesses, close to the percolation threshold of the material, these layers become semi-transparent, behaving in part as a metal and in part as a dielectric. In particular, one can play with the thickness, surface morphology and micro/nano structuring to obtain the desired trade-off between electrical conductivity and optical transmittance/reflectance, making it possible to devise innovative transparent conductors and optical surfaces. Recently, it was shown that the electronic and IR optical properties of ultrathin metals ($t < 10$ nm) can be tuned by electrolyte gating [16] approaching the graphene paradigm. Although the electrolyte gating scheme employed in these experiments is similar to the one used for MLG, its underlying mechanism was associated to reversible nanoscale reduction/oxidation processes rather than to a doping process as in the graphene case. Such experiments are at a very early stage and the modulation ranges demonstrated so far are quite limited, currently hindering applications.

Conversely, a different and potentially disruptive approach

for EC surfaces based on ultrathin metals is quickly emerging and relies in the so-called reversible metal electrodeposition (RMED). It consists in the reversible deposition and dissolution of a very thin metal film on a (semi)transparent electrode that allows switching from a transparent to an opaque state [17]. This concept was developed for display and smart windows applications at visible frequencies and mostly employed common metals such as Ag and Cu that can be electro-deposited in a highly reversible manner.

Liu et al. proposed for the first time an Ag-based RMED structure with excellent IR-EC properties and great potential for adaptive IR camouflage devices [18]. The EC structure behaves as a tuneable metallic mirror in which the Ag particles dissolved in the electrolyte medium can be deposited to / released from a semi-transparent metallic electrode upon the application of a voltage. A re-edition of such technology based on Cu RMED has recently shown the best IR emissivity modulation reported so far (between 0.07 and 0.92) [19]. Moreover, such active surfaces based on RMED can be modified to show tuneable colors in the visible spectrum, by embedding thin dielectric films into the stack and playing with optical interference effects. However, these surfaces feature response time larger than 1s that may be not compatible with IR camouflaging applications requiring fast adaptation to the background. Moreover, their long-term durability has to be assessed more carefully.

CURRENT PERFORMANCE AND OPEN CHALLENGES

Emerging nanomaterials such as ultrathin metals, graphene and related nanostructures are paving new “avenues” to the development of innovative EC active surfaces for adaptive IR camouflage and thermal radiation management. However, despite the increasing amount of literature data available on these technologies, several points to clarify remain about their actual performance, durability, scalability and production cost. Furthermore, it is difficult to compare reported results for laboratory experiments performed with different device architectures and in different measurement conditions.

As a matter of fact, the real challenge in order to move these technologies out of the laboratories is fabricating large-area and reliable devices, as well as developing comprehensive and standardized tests to benchmark performance indicators.

ONGOING RESEARCH IN LEONARDO

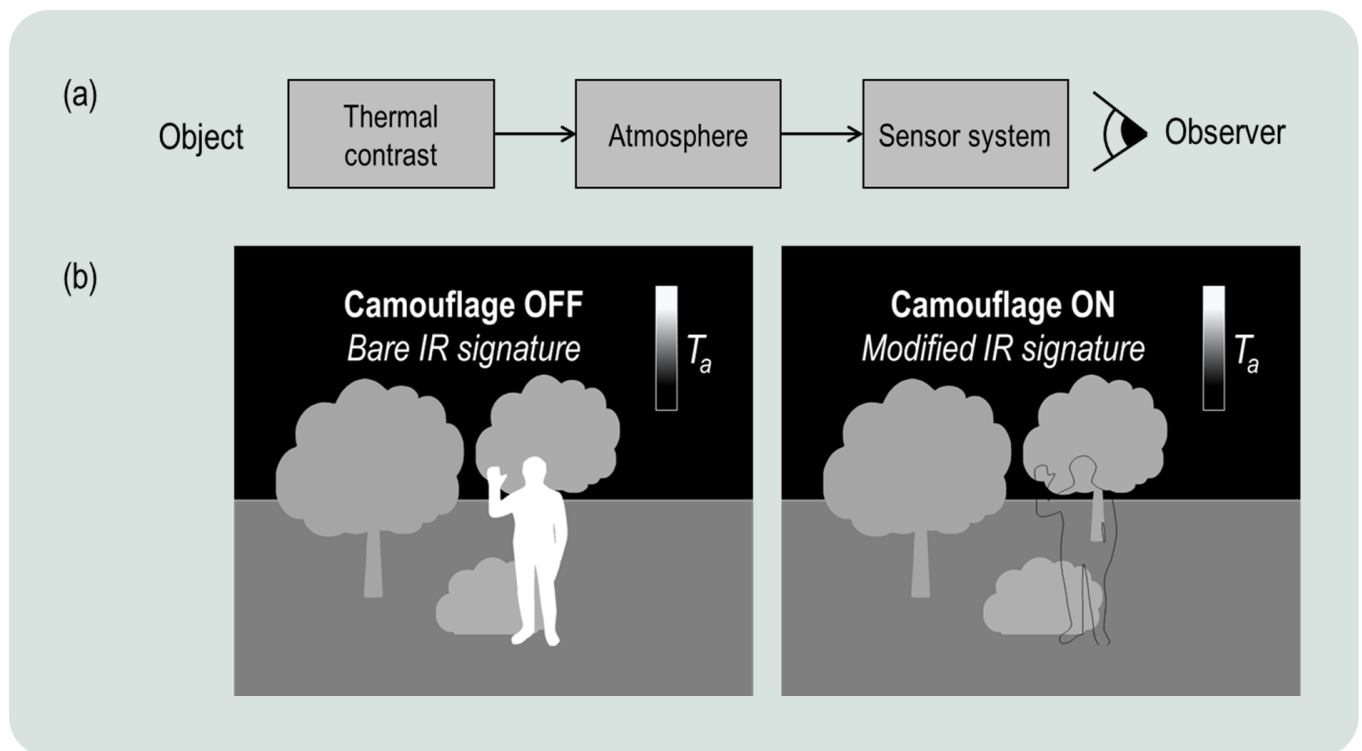
We have seen that engineering surface optical properties by using nanotechnology allows to devise optical coatings that can act as an “artificial skin” that dynamically controls the thermal radiation emitted from an object by tuning its emissivity. Such possibility is appealing for several potentially disruptive technologies such as active thermal management, radiative cooling and thermal camouflage.

For instance, an EC surface could be implemented to develop an adaptive thermal camouflage system capable to dynamically reshape the thermal signature of an object to mimic the background or another object’s thermal appearance, as conceptually shown in Figure 2.

Some proposed solutions for this task are based on direct

modulation of surface temperature or projection of thermal images from a IR-emitting screen that feature high energy consumption and/or lack of practicality. Thermal camouflaging based on EC active surfaces is a promising approach with low power consumption and adaptability to a variety of platforms.

For this reason, some of the above discussed IR EC surfaces are currently being studied by Leonardo Innovation Labs, Leonardo Electronics and external partners at the IPCB-CNR and ENEA research center in Portici (Naples). Furthermore, an Advanced Materials Lab has been set up in the Campi Bisenzio site and is equipped with several state-of-art tools to perform a variety of electrical and optical characterizations on nanostructured materials and surfaces.



2-Adaptive IR camouflage: conceptual view

CONCLUSIONS

In this paper the concept of adaptive thermal camouflage and possible technological solutions based on engineering of infrared optical properties were introduced. In particular, the most promising classes of nanostructured electrochromic surfaces providing the largest variation of thermal emissivity upon electrical actuation were described.

Finally, the open challenges of these technologies and the current research effort carried out in the company were outlined.

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Bruno Paulillo: bruno.paulillo@leonardo.com

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Telemetry Infrastructure for Real-Time Flight Monitoring of Experimental Helicopters

Attilio Colombo, Davide Porrini, Antonio Natola, Moreno Corio
Leonardo - Helicopters Division

The development and certification of helicopters is a complex process that requires a significant amount of flight testing activities with instrumented prototypes. These experimental flights are often monitored in real time in order to provide the highest level of safety and to increase the efficiency of flight operations. Due to the large amount of sensors and data streams that are usually collected on experimental aircraft, dedicated ground infrastructures are deployed and operated to receive and extract all data from the aircraft telemetry downlink. These facilities must ensure the proper level of performances and capabilities to the team of specialists involved in the flight test program. This paper presents an outline of the current telemetry hardware and software infrastructure within Leonardo Helicopter Division, which has been developed along the time to fulfil such needs.

INTRODUCTION

Flight testing has experienced an increase in the level of complexity during time due to both technical and program requirements, with the related schedule constraints that are always more tight, and with the complexity and enhanced level of safety of the products that has to be demonstrated to the Authorities during the certification process. In this context, telemetry is of paramount importance to speed up the activities and to guarantee high level of safety of the flight testing. Thus, it has become the standard tool for large part of the experimental flights.

To execute a good telemetry, it is fundamental to have a complex system consisting of different components: the instrumentation of the aircraft, the system to decode and display the received flight data, the engineers from the technical areas who have to interpret the data in real time and communicate to the flight crew every important information, which in some cases may be safety critical. All these goals are made achievable thanks to the core part of the system: the antennas and the engineers who connect and maintain the connection between the flying aircraft and the telemetry room.

SCOPE OF THE TELEMETRY

The telemetry can have different scopes, depending on the type of flight. According to a very rough classification, such scopes can be divided in two main groups: those devoted to provide efficiency and those aimed to ensure safety.

The telemetry for safety reasons is requested by the formal airworthiness document. It is aimed to reduce as much as possible the risk of compromising integrity and safety of the test aircraft and of the crew. Examples of telemetry for safety reasons include:

- the first ground run and the first flight of a new aircraft model;
- the flight envelope expansions;
- operations in particular environments (icing);
- execution of development and certification flights classified as high risk by the hazard assessment (e.g. stall tests, autorotation tests close to the boundaries of the certification envelope).

On the other side, the telemetry for efficiency has the scope to reduce the effort and number of flights that are needed to achieve a given target. This is for example the case of the system identification tests, consisting of flights containing a number of conditions that are needed to tune the analytical model of the aircraft.

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Such activity is growing in importance, because the next future “certification by simulation” approach needs a fully reliable analytical model to become a viable path. During these flights the conditions requested have to be flown in a very precise and repeatable way. For this reason, the engineers in the telemetry room perform a quick near-real time analysis, to confirm the validity of the test points or to request for their repetition, also giving indications to the flight crew on how to improve the test point. This can save several flights that should be needed to repeat conditions that could be not perfect or not suitable for the intended purpose, and it also helps the flight crew in reducing the time to obtain the required result.

Another example is the automatic flight control system testing. During it, different gain sets are loaded directly on-board by the flight crew and are tested, with rapid

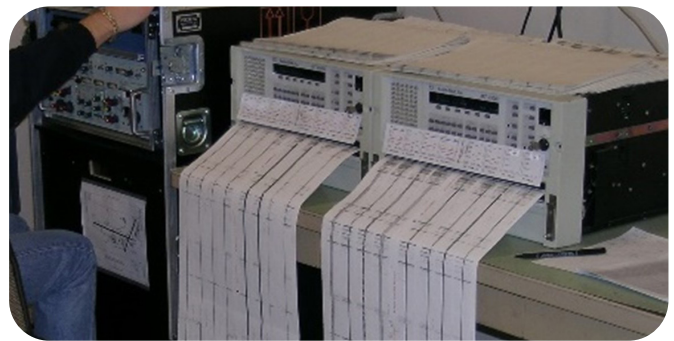
analysis of the results and their discussion in real time by the crew and the engineers in the telemetry room, to decide the next test point or gain set on the basis of the outcomes just obtained.

All these flights are performed with heavily instrumented helicopters and are requested for the development and certification of the aircraft. For the experimental prototype of a modern helicopter, “heavily instrumented” could mean up to 100,000 recorded parameters, a good percentage of which, up to 10-20 thousands, to be real-time transmitted in telemetry, including data from instrumented structural parts and components (e.g. rotors, transmissions, fuselage, engines and flight controls), air data unit parameters, data flowing on avionics and flight computer digital buses, navigation data, videos, comms, etc.

HISTORICAL BACKGROUND

The concept of real-time flight monitoring by means of telemetry data link has been used since long time in the aviation industry and has always maintained its fundamental scope and principle of operation. However, the developments in technology and the changes of requirements have led to a significant transformation of the involved equipment, with enormous increase in the system capabilities.

This transformation has involved both the on-board segment of the flight test instrumentation system and the corresponding telemetry ground infrastructure. On one side, the number of sensors and the variety and size of the different data sources has been increasing from a few tenths of purely analogic signals up to several thousands of parameters of various nature. On the other side, the ground based infrastructure has evolved in terms of complexity and technology, to cope with the increased data rate to be processed and to provide more sophisticated and more powerful tools that are needed not only to improve the pure flight monitoring but also to increase the real time analysis features in support of the entire flight test team, both on board and on ground. Figure 1 shows a typical configuration of a legacy mobile telemetry system. In it, the data of interest are plotted on strip charts for a limited number of parameters under monitoring (less than 20). The ground segment of the telemetry system has successively been improved by adding some screens lively presenting the main flight parameters of the aircraft in order to provide the ground test crew with better situation awareness. In addition, also the number of monitored parameters has been increased, but still plotting on paper during the flight the corresponding time histories, as shown in Figure 2 and Figure 3.



1–Old strip-chart plotters



2–Legacy telemetry console



3–Legacy telemetry control room



4–One of the tower-mounted telemetry antennas at Cascina Costa airfield, in the background of an instrumented tail rotor test rig



5–Portable telemetry antenna during cold trials

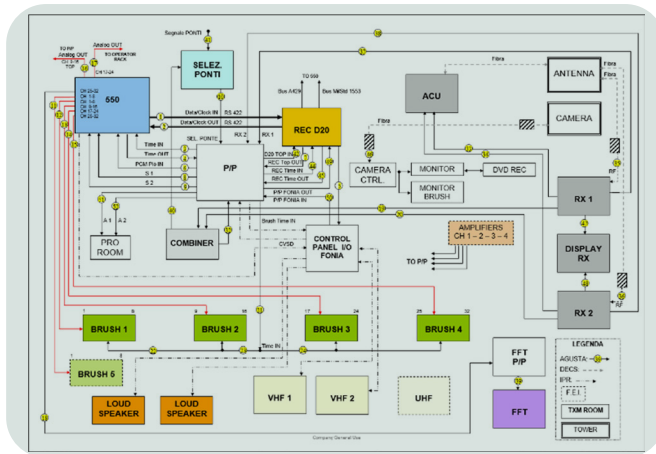
That step improved the way the telemetry was performed, by making the flight conduction parameters available to engineers in the telemetry room, to enable them to better evaluate the flight condition being executed and to correlate data from the relevant parameters with the flight phase conduction. A major step in the evolution of Leonardo Helicopters Division (LHD) telemetry infrastructure has derived from the swap of the radiofrequency band from the L-band to the C-band. This change has been driven by the continuous increase in the mobile phones traffic, with its related increase in data usage, which has requested the availability of new radiofrequency bands. As a consequence, the L-band has been primarily assigned to the mobile phones communications, whilst the C-band has been assigned to the aviation telemetry. The associated change of the carrier frequency adopted for telemetry increased from about 1500MHz to about 5100MHz. This has required a substantial refurbishment of the ground infrastructure that has been realized throughout different phases.

In the first phase, the transmitting and receiving antennas of the telemetry link, as well as all related equipment inclusive of downconverters, receivers, radios, RF couplers, fiber optics, etc. and the main processing hardware, have been upgraded. The main part of the ground section of the new data transmission system now includes two 6-foot diameter directional antennas, tower-mounted at Cascina Costa airfield, as shown in Figure 4. During telemetry, the azimuth and elevation of these antennas are continuously regulated by the corresponding antenna control unit in order to keep track of the aircraft flight path. Different control modes are available to the operator: manual tracking, auto-tracking, and slave mode, in which the orientation of the antenna is governed by the GPS signal coming from the aircraft. With these antennas a range of up to 300/400 km can be covered, with data-rates up to 15/18 Mbps depending on the applicable instrumentation requirements and the related configuration of the data acquisition system installed on the prototype.

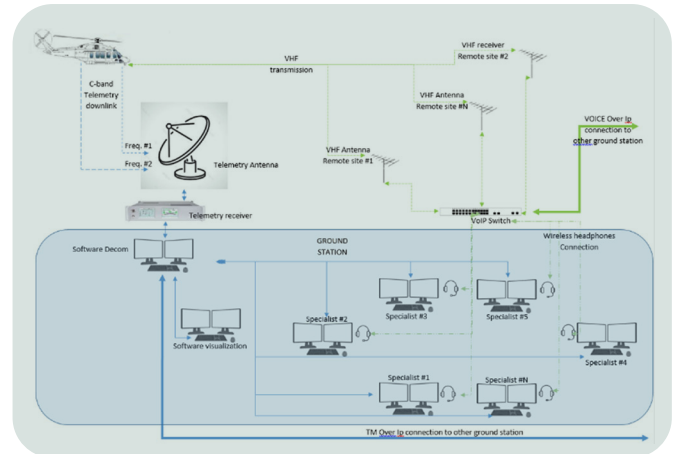
For off-site operations requiring telemetry, smaller tracking antennas are available. These items are utilized also in severe climatic environments, for example during cold and icing flight test campaigns, as shown in Figure 5. In the second phase of the C-Band migration project, also the telemetry rooms have been substantially modernized, by realizing more functional and interconnected control rooms that are multiuser and reconfigurable. In this refurbishment also the environmental impact of the control room operation has been improved, thanks to the replacement of the old and highly energy consuming air conditioning units, with more efficient systems, free of ozone-depleting gases. In terms of architecture, the new control room solution is based on the substitution of the previous hardware-based decommutation of the input signal shown in Figure 6.

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6-Block diagram of the legacy control room architecture



7-New control room architecture

These have been replaced by a software-based decomm application that has been developed internally by LHD and runs on standard workstations fully compatible with all Network and IT requirements. This has enabled the Flight Ops team to incorporate into such software package also the ability to convert additional kind of data streams generated by new components of the on-board data acquisition system (IENA data packets, wireless slings, Ethernet based videos, etc.), thus making the corresponding new set of data available to the team of engineers in the control room. For audio communications between the flight crew and the ground team in the control room, a new VOIP based intercom has been also developed. This system, fully

configurable via software, allows audio communications among the test conductor, the various engineers involved in the flight test activity and the flight crew, by means of dedicated VHF links. The block diagram of such new solution is shown in Figure 7. Furthermore, the new control room solution has enabled integrating the data coming from the test aircraft with additional services already available, like weather stations, air traffic information services or images from airfield cameras. In addition, also the voice data coming from the on-board inter-communication system can be received in the telemetry room and used to increase the awareness of the involved test team. The final layout of the new control room is shown in Figure 8 and Figure 9.



8-New Telemetry room layout



9-Configurable multifunction displays

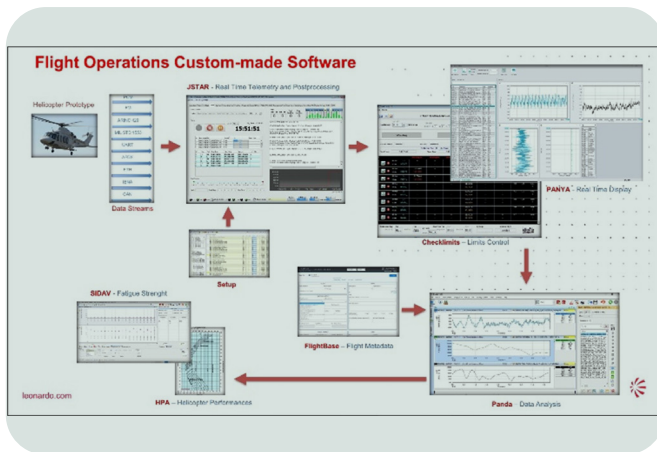
DATA PROCESSING SOFTWARE

The complexity of the data acquisition process, both in real time during telemetry and in the post-processing phase, after completion of the flight, is managed by LHD through a suite of various software applications that allow to manage the different phases of the acquisition process:

- Setup of the Acquisition system;
- Control of the acquisition system;
- Real Time Acquisition (Telemetry);
- Real Time data display;
- Data Post-processing;
- Post-processing analysis;
- Adding metadata;
- Archiving of acquired data.

For each flight test, thousands of parameters are acquired. Therefore, the speed of execution of the software is crucial. Especially for flight safety, the response time of the system must be combined with robustness of the process, as the software is executed in real-time during telemetry.

All the LHD-proprietary applications have been created by the Flight Operations team with the aim of maximizing the integration with the physical machine, the aircraft, the Experimental Flight Ops environment and the needs of users of the data. In Figure 10 a schematic of the data flow is shown, with the different data streams that are acquired on board the aircraft and the various software items that are involved in the process.



10-Flight Data flow within LHD software

The raw data coming from various connections such as PCM, Analog, Arinc 429, Mil-Std-1553, Ethernet AFDX and IENA, involving the use of different protocols is decommuted and converted into Engineering Units by means of the JStar application. The same executable application is used to process the entire data flow both during the Real Time telemetry sessions and during the Post Flight data processing, starting from the raw data saved onboard the aircraft on suitable storage media (Compact Flash cards or Solid State removable Disks).

CONCLUSIONS

For sure, the central role of Telemetry in the Experimental Flight Testing context will be maintained also in the future. The present paper has illustrated the main characteristics of the current implementation of the LHD Telemetry infrastructure. It has been put in place in order to modernize the legacy solution, with the intent to provide LHD with future-proof facilities that are able to incorporate new technologies and new solutions to further improve quality and effectiveness of its Flight Test Operations.

Thanks to its integration and modularity, the **JStar** application allows to make the data available to commercial and proprietary applications for Real Time data visualization and analysis. Furthermore, automatic data checks are performed by means of another dedicated application running in parallel to verify that the acquired parameters remain within the design limits, and to highlight possible exceedances, in order to promptly identify potential anomalies.

Once converted, the data are then tagged and stored on common data servers, independently of the device or stream they come from. They generate standard archives (Panda archive) that can be analyzed and exported through the **JPanda** application, available on the Leonardo Software Center and are used by numerous users, mainly of the Engineering organization, from various geographies and sites. Flight data of experimental aircraft is archived using binary files with a proprietary format, while the associated metadata are stored in relational databases.

Once the data has been stored, the metadata describing the conditions in which the test are performed must be added to the archive. For this purpose, the **FlightBase** Web App has been designed, which allows the FTE to enter these information both from a PC or from a mobile device on the runway. At the end of the process, ad-hoc applications are used to obtain aggregate information from the processed and stored flight data: **HPA**, to calculate the helicopter performance that are used to generate the flight manual diagrams, and **SIFAT** to calculate the fatigue characteristics of mechanical parts.

In addition to the above standard process that is applied to all experimental flights, more recently also a dedicated pipeline to export specific subsets of the flight data archive to the DaVinci-1 supercomputer has been realized. This with the intent to boost the development of AI-based applications and advanced flight data analytics to be exploited both for the development and certification of the products and for the delivery of new services and new capabilities to the operators of the fleet of Leonardo's helicopters.

TO FIND AND NOT BE FOUND, THAT IS THE QUESTION!

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These may include various possibilities, some of which are already under development, like:

- virtual sensors to complement the suite of physical sensors of the test aircraft;
- AI-based applications to create virtual assistants that help the team of engineers in monitoring the test aircraft behaviour in real-time;

- virtual cockpit that provide a ground-based pilot with an environment and a set of information as close as possible to what is normally available in the actual aircraft cockpit.

All these characteristics can be seen as contributors to allow LHD to fulfill its mission and goals as a leader in the helicopter and vertical take-off market.

Attilio Colombo: attilio.colombo@leonardo.com

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Editor in Chief

Vincenzo Sabbatino

Editorial office

Emidio Di Pietro
Giovanni Cocca
Marco Morini
Patrizia Pozzoni

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