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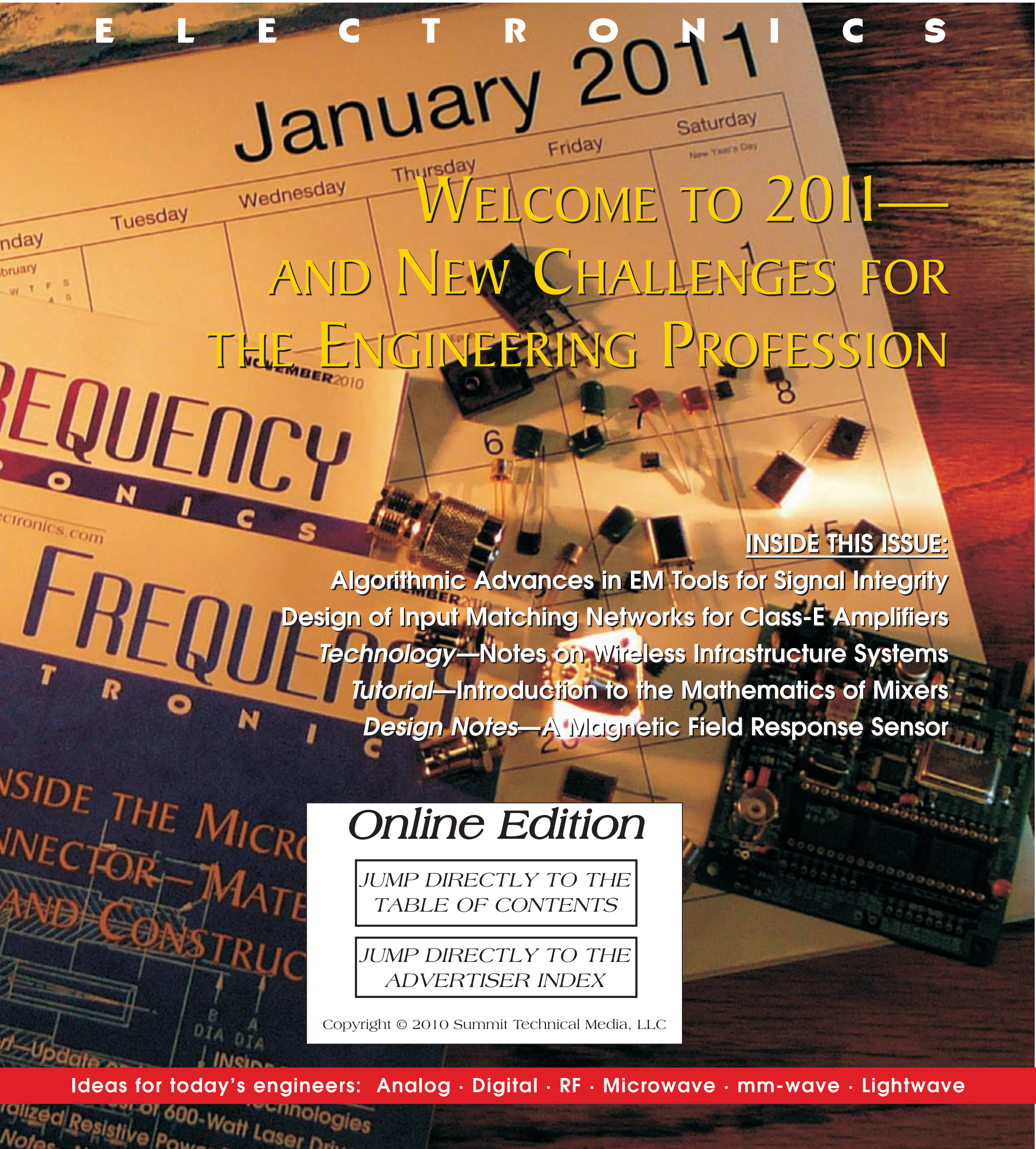
JANUARY 2011

HIGH FREQUENCY ELECTRONICS



Welcome to 2011— and New Challenges for the Engineering Profession

INSIDE THIS ISSUE:

Algorithmic Advances in EM Tools for Signal Integrity
Design of Input Matching Networks for Class-E Amplifiers
Technology—Notes on Wireless Infrastructure Systems
Tutorial—Introduction to the Mathematics of Mixers
Design Notes—A Magnetic Field Response Sensor

Online Edition

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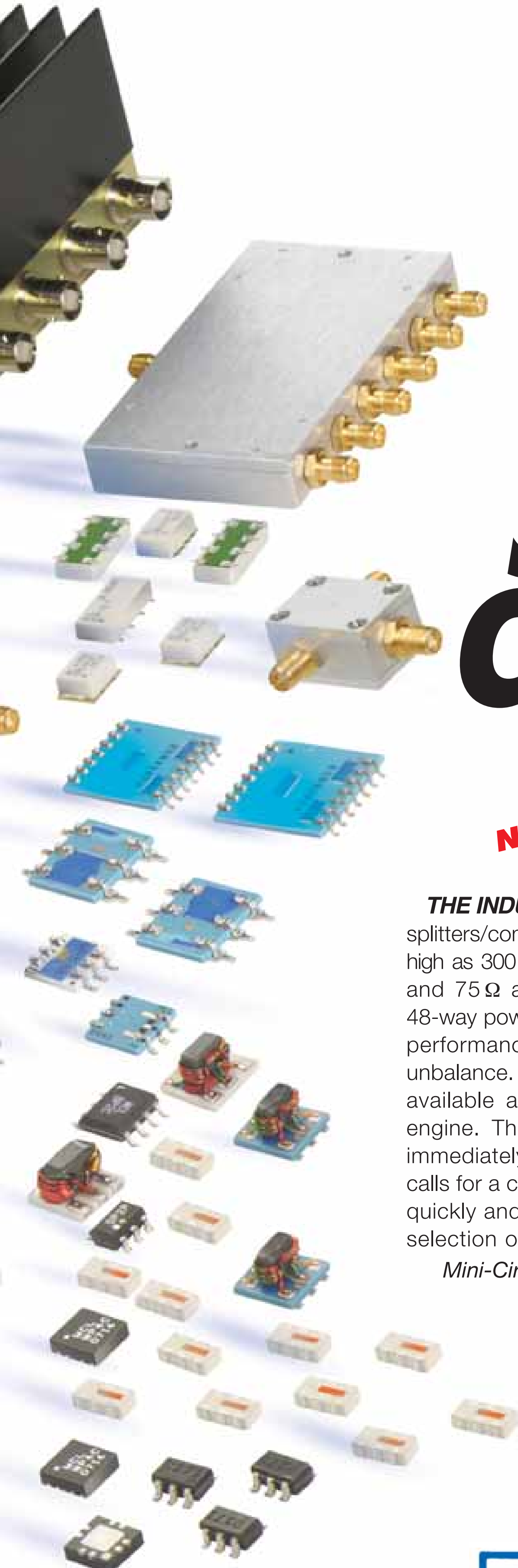

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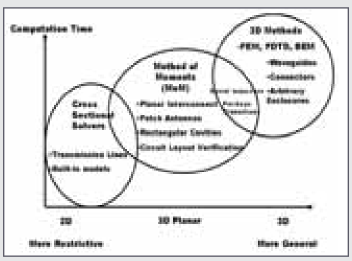
JANUARY 2011
Vol. 10 No. 1

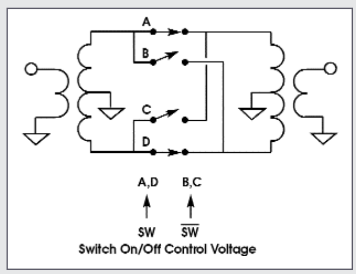
HIGH FREQUENCY ELECTRONICS

You can view this issue page-by-page, or click on any of the articles or columns in the Table of Contents below

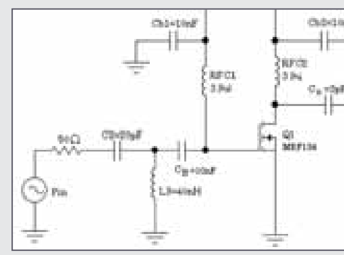
16
signal integrity
High-Frequency Algorithmic Advances in EM Tools for Signal Integrity—Part 1
John Dunn



34
tutorial
The Mathematics of Mixers: Basic Principles
Gary Breed



40
class-E amplifiers
Design of Input Matching Networks for Class-E RF Power Amplifiers
Firas Mohammed Ali Al-Raie



26
product coverage
Featured Products

30
technology report
Notes on Design and Deployment of Wireless Infrastructure Systems

50
product coverage
New Products

62
product coverage
Product Highlights

64
design notes
A Magnetic Field Response Sensor For Measurement of Liquid Levels

Regular Columns

6 Editorial	**12** In the News	**62** Product Highlights
8 Meetings & Events	**50** New Products	**63** Advertiser Index

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A Few Things to Think About for the New Year

Gary Breed
Editorial Director

We have a lot on our minds as 2011 begins. Big picture things include the economy and politics, which can have major effects on our personal lives. Global power shifts and continuing conflicts are things we are concerned about, as well. At the personal level, there many short-term and long-term issues—both promising and worrisome—that affect our lives. Technology is a big part of it, with Internet communities expanding our concept of "local" and wireless capabilities helping eliminate the need for us to have a specific "location".

There are many more things that are getting our personal and professional attention. I've collected a few notes on some of those that are pertinent to the high frequency techniques and technologies we practice.

Resources and Environment

Although the specific definition of "green" keeps changing (exactly how did CO_2 become such a big villain?), the general idea remains the same: natural resources are not unlimited, dangerous things should not be in our environment, and the size and scope of human activity has an effect on the entire planet. Scientists, politicians and activists will keep arguing, but regardless of viewpoint, there are things that engineers are involved in. These include the European Reduction of Hazardous Substances (RoHS) that has changed the design and fabrication of many electronic components and products. And research into the interaction of electromagnetic radiation with the human body continues—as it should.

On a smaller scale, we will be involved in the design of sensors, communications links, and networks that monitor the environment, both for research and for practical operations such as improving the efficiency of homes and commercial buildings. Wireless technology is already a major help in reducing the need for local commutes and long distance travel.

Social Networking and Personal Communication

At first, you might not think of this subject as engineering-related, but Internet-based social networking is already having a major influence on the way we stay in touch with one another, and how we see the world at a personal level. Extending this influence into our professional lives in inevitable, but the right way to use it is still being developed.

Has it moved from personal use to your everyday work? I would love hear your stories of how Facebook, Twitter, LinkedIn and other services are being used for professional development, information sharing and other activities.

Virtual Laboratories

Near-total reliance on computer simulation and analysis has moved from the future to *right now*. Sure, the intuition of personal experience is still essential, but the software tools and powerful hardware at our disposal is replacing more lab bench engineering tasks every day. 2011 may be the last year we consider this an area deserving special attention—and it is this final step into everyday usage that I think is newsworthy.

Education

The traditional approaches for education from early childhood through advanced university degrees are changing. Exponential growth of human knowledge makes it impossible to teach a sample of everything, and changing social attitudes have required schools to take on some responsibilities that parents, families and communities once provided.

Problem solving skills that once were learned through unsupervised neighborhood and school playground activities are being replaced by classroom exercises. Hands-on learning by helping your dad, a neighbor or an uncle work on a car, home repair or other project has largely disappeared.

In this latter case, it means that new engineering students may not be accustomed to working with tools. This means that getting started in the EE school lab takes longer than it used to, and professors need to adjust their teaching methods. It may sound contradictory to my comments on computer simulation, but I think it is completely appropriate to learn things first in a hands-on environment, using real hardware. Simulation is learned as well, to compare the real world with the mathematics. What is different is that the ratio of time spent in the instrument lab and computer lab is changing to ensure that students become familiar with the most important tools.

This month's photo was taken while I was visiting the new engineering building at the University of Wisconsin-Platteville.

Next month, I'll have more notes on our changing times, and how we respond to those changes.

Conferences

February 7-9, 2011
4th Annual Military Radar Summit
Washington, DC
Information: Conference Web site
http://www.MilitaryRadarSummit.com

February 7-9, 2011
IDGA's Software Radio Communications Summit
Vienna, VA
Information: Conference Web site
http://www.SoftwareRadioSummit.com

February 20-24, 2011
IEEE International Solid-State Circuits Conference
San Francisco, CA
Information: Conference Web site
http://www.isscc.org

March 21-23, 2011
7th Military Antennas Summit
San Diego, CA
Information: Conference Web site
http://www.MilitaryAntennasEvent.com

March 22-24, 2011
International CTIA Wireless 2011
Orlando, FL
Information: Conference Web site
http://www.ctiawireless.com

April 3-6, 2011
Int'l Symposium on Power Line Communications
Udine, Italy
Information: Conference Web site
http://www.ieee-isplc.org/2011/

April 9-14, 2011
2011 NAB Show
Las Vegas, NV
Information: Conference Web site
http://www.nabshow.com

April 11-15, 2011
European Conference on Antennas and Propagation
Rome, Italy
Information: Conference Web site
http://www.eucap2011.org

April 18-19, 2011
12th WAMICON 2011—IEEE Wireless and Microwave Technology Conference
Clearwater, FL
Information: Conference Web site
http://wamicon.org

May 2-4, 2011
Sarnoff 2011—34th IEEE Sarnoff Symposium
Princeton, NJ
Information: Conference Web site
http://ewh.ieee.org/conf/sarnoff/2011/

May 16-19, 2011
APEMC 2011—Asia-Pacific EMC Symposium
Jeju Island, Korea
Information: Conference Web site
http://www.apemc2011.org

Short Courses

UCLA Extension
Engineering, Information Systems and Technical Management
10995 Le Conte Ave., Suite 315
Los Angeles, CA 90024-1333
Tel: 310-825-3344
E-mail: shortcourses@uclaextension.edu
http://uclaextension.edu/shortcourses
Transitioning from Technical to Managerial Responsibilities
February 7, 2011
Advances in Satellite Communications: Efficient and Reliable Transmission Systems
February 20, 2011
Software Defined Radio
February 24, 2011

Tektronix/Kimmel Gerke EMC/SI Seminars
Kimmel Gerke Associates, Ltd.
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South St. Paul, Minnesota 55075
Tel: 1-888-EMI-GURU
http://www.emiguru.com
EMC/SI + Optional Troubleshooting Workshop
February 7-9, 2011, Orlando, FL
February 22-24, 2011, San Diego, CA
EMC/SI
March 22-23, 2011, Dallas TX
April 5-6, 2011, Boston, MA
April 11-12, 2011, Baltimore, MD
May 2-3, 2011, Rochester, NY

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1250 Peterson Drive
Wheeling, IL 60090
Tel: 847-537-6400
http://www.dlsemc.com
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April 12-14, 2011, Hilton Hotel, Northbrook, IL

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Or send by mail to the Editorial & Production Office address shown on page 6.

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http://www.besserassociates.com

Phase Noise and Jitter
February 8-11, 2011, Web class

Designing High Efficiency RF Power Amplifiers
February 28-March, 2, 2011, San Diego, CA

Antennas & Propagation for Wireless Communications
February 28-March 2, 2011, San Diego, CA

Applied Design of RF/Wireless Products and Systems
February 28-March 2, 2011, San Diego, CA

Power Conversion & Regulation Circuits for VLSI Systems
February 28-March 2, 2011, San Diego, CA

RF Measurements: Principles & Demonstration
March 14-18, 2011, San Jose, CA

Applied RF Techniques I
March 21-25, 2011, San Jose, CA

Applied RF II: Advanced Wireless and Microwave Techniques
March 21-25, 2011, San Jose, CA

Practical Digital Wireless Signals – Measurements and Characteristics
March 21-25, 2011, San Jose, CA

LTE Mobile Access
March 24-25, 2011, San Jose, CA

European School of Antennas
Prof. Stefano Maci
Dept. of Information Engineering
University of Sienna
53100 Siena ITALY
macis@ing.unisi.it
http://www.esoa-web.org

Antenna Project Management
March 21-25, 2011, EPFL—Lausanne

Propagation for Space Application
March 28-30, 2011, ESTEC—Supaero, Toulouse

Industrial Antenna Design
April 4-8, 2011, IMST—Duesseldorf

Leaky Waves and Periodic Structures for Antenna Applications
April 26-29, 2011, SAPIENZA—Rome

Antenna Measurements at Millimeter and Submillimeter Wavelengths
May 16-20, 2011, AALTO—Helsinki

Propagation and MIMO
May 30-June 3, 2011, UNISI/KIT—Siena

Compact Antennas
June 6-10, 2011, UPC—Barcelona

Terahertz Technology and Applications
June 13-17, 2011, UPC—Barcelona

Advanced Near-Field Antenna Measurement Techniques
June 20-24, 2011, DTU—Copenhagen

Body Area Network
June 27-30, 2011, QMUL—London

Calls for Papers

IEEE Transactions on Terahertz Science and Technology is now accepting paper submissions

IEEE has announced a new publication specifically targeted at the THz field, *Transactions on Terahertz Science and Technology.* The goals of the editors include the wide spread release and distribution of a high quality, high impact-factor publication for THz papers, rapid posting of submitted articles on IEEE Xplore (within 3 months of submission), and a broad scope that cuts across discipline boundaries and covers science, technology and applications. The journal web site is now open, and we are accepting submitted articles through a simple upload process at http://www.thz.ieee.org. All submissions will be managed and peer reviewed through a committed and well recognized group of Topical Editors. The Inaugural Issue is scheduled for print distribution in September 2011 and will contain overview and review articles covering a wide range of THz topics from well established THz researchers. A list of papers and authors for this issue will be released in late January. Subsequent issues are open for general submissions now. Accepted articles will be posted through IEEE Xplore before the release of the inaugural print issue. We hope the new journal will serve as a focal point for the THz community and that you will support it both with your contributions and your readership.

International Journal of RF and Microwave Computer-Aided Engineering—Special Issue

The *International Journal of RF and Microwave Computer-Aided Engineering* announces a special issue on EM-CAD models and their integration into industry-standard CAD tools contributing to engineering design and innovation. The special issue encourages articles that present EM analysis and modeling techniques which clearly demonstrate a transition from theory to practice. One way to showcase this concept is to apply the proposed EM modeling techniques to industry-standard design and optimization examples, emphasizing engineering design innovation (EDI), improvements in product reliability and signal integrity, performance under realistic operating conditions, etc. The special issue encourages EM-CAD techniques capable of modeling industry-standard testing environments.

Information:

This issue will appear in January 2012. Manuscripts should conform to the requirements for regular papers to the journal. Authors wishing to have their articles considered must submit their contributions in PDF format by e-mail before February 28, 2011 to one of the guest editors: Vijay Devabhaktuni, Ph.D., Vijay.Devabhaktuni@utoledo.edu; Chuck Bunting, Ph.D., reverb@okstate.edu; or James Rautio, Ph.D., rautio@sonnetsoftware.com. A complete list of topics may be found in the call for papers, available online at www.sonnetsoftware.com.

Business News

Computer Simulation Technology (CST) announces the signing of a joint marketing agreement with **Cadence Design Systems.** Cadence and CST are collaborating in order to offer an effective workflow for PCB and package layout co-design in high-speed and mixed RF systems. Cadence and CST applications teams will be working together to address customer requirements and offering best in class solutions.

austriamicrosystems and **NXP Semiconductors N.V.** announced that they have jointly developed the first in a range of reference solutions for product authentication in embedded consumer applications. Leveraging the AS399x UHF RFID Reader IC, and the UCODE G2iL series, the reference design delivers the industry's lowest power consumption, lowest solution cost, lowest complexity and highest level of integration.

RF Micro Devices, Inc. announced it has teamed with **Freescale Semiconductor** to deliver ZigBee® solutions for a broad range of smart grid applications. RFMD's newly introduced RF6535 ZigBee front end module (FEM) has been combined with Freescale's MC1321x System-in-Package (SiP) to create the RF6535/MC1321x reference design. The RF6535/MC1321x reference design simplifies RF design requirements, while reducing product cost and complexity. Working together, RFMD® and Freescale address the need for aggressive size reductions in IEEE 802.15.4 designs with a reduced solution footprint and minimized component count.

Peregrine Semiconductor Corporation and **Soitec,** a supplier of silicon-on-insulator (SOI) wafers and advanced solutions for the electronics and energy industries, announced the joint development and ramp in production of a new, bonded silicon-on-sapphire (SOS) substrate which has been qualified for use in manufacturing Peregrine's next-generation STeP5 UltraCMOS™ RF IC semiconductors.

UK-based low power radio solutions provider **Radiometrix** announces its collaboration with distribution firm **APC Technologies.** Located in Gurgaon, close to New Delhi, APC Technologies specializes in the RF and microwave supply chain activities. It serves the industrial, wireless infrastructure, avionics, broadcast, defence and medical sectors, acting as a source for radio modules, passive components, analogue/mixed signal semiconductors, microwave generators, power amplifiers and GPS units.

Laird Technologies announced its acquisition of **Cattron Group International, Inc.** Cattron's sales revenues for the fiscal year ending September 30, 2010 were $53.8 million. The purchase expands the scope of Laird Technologies' wireless machine-to-machine (M2M) product offering and the markets those products serve. Cattron is in the design and supply of high-reliability wireless remote control systems serving the railroad, mining, and industrial markets globally. In addition to providing custom wireless Remote Control (RC) systems that enable external operation of heavy equipment and vehicles as well as ongoing service for these RC systems, Cattron provides fully hosted networked applications software enabling customers to monitor their remote controlled assets in real-time.

Agilent Technologies Inc. announced that it has been awarded a $1.8 million contract by the **U.S. Navy.** Under the terms of the contract, Agilent will supply handheld spectrum analyzers (HSAs) to the Navy for five years. The instruments will help the Navy's technicians install, monitor and maintain RF electronic systems in the field.

TriQuint Semiconductor announced that it has been awarded a Defense Production Act Title III gallium nitride (GaN) manufacturing development contract by the **US Air Force Research Laboratory** (AFRL). The overall goal of the contract is to increase yield, lower costs and improve time-to-market cycles for defense and commercial GaN integrated circuits. The contract was awarded based on TriQuint's success and experience developing new gallium nitride technologies and products.

Giga-tronics Incorporated announced that it has received orders in excess of $4.8M to supply microwave test equipment for the automation of production at contract manufacturers in China. The products supplied are part of Giga-tronics' microwave signal switching family. These orders are expected to ship this Fiscal Year.

Honeywell Electronic Materials announced that it is expanding production capacity for 300mm sputtering targets and related metals as part of a long-term investment plan to support rebounding semiconductor industry growth. Honeywell is increasing targets production capacity, as well as its internal capacity to produce the copper, cobalt, titanium, and tungsten raw materials used to make the targets. The expansion will allow Honeywell to meet growing demand for targets, which are used as the source of metal to make circuits in advanced semiconductors.

APIC Corporation announces that the **Naval Air Systems Command** (NAVAIR) has awarded APIC a SBIR Phase I contract in addition to its current NAVAIR project funding. The goal of this new SBIR Phase I project is design of an integrated optical CDMA (OCDMA)—Code Division Multiple Access network system that addresses multi-level security. APIC will act as a subcontractor for design and analysis of OCDMA network architectures/ components for legacy federated avionics, current Integral Modular Avionics (IMA); and future multi-core processor architectures avionic processing systems in this program.

AWR® Corporation announced that it has added nearly 55 product, application and tutorial videos on Engineering TV, expanding its social media outlets beyond the current AWR.TV and AWR Channel on YouTube. AWR's multime-

dia content covers fundamental engineering topics such as *S*-parameters, filter synthesis and yield analysis to advanced subjects such as WiMAX mobile solutions and AXIEM™ three-dimensional (3D) planar applications. These videos make it easy to learn more about using AWR's high-frequency EDA tools including Microwave Office®, Analog Office®, Visual System Simulator™ (VSS), APLAC™, and AXIEM™ software.

America II Electronics, Inc., one of the world's largest independent distributors of semiconductors and passive components, has finalized a global authorized distribution agreement with **Aviel Electronics,** a division of **RF Industries.**

SouthWest NanoTechnologies, Inc. (SWeNT) has received an **Environmental Protection Agency** (EPA) consent order that will permit SWeNT to manufacture and distribute Single-Wall Carbon Nanotubes (SWCNT) for commercial applications. Previously the Company distributed its SWCNT via a Low Release, Low Exposure (LOREX) PMN exemption from the EPA. SWeNT is currently the only U.S. manufacturer permitted to commercially distribute SWCNTs.

JP Sercel Associates Inc. (JPSA) announced that construction for expanding their laser manufacturing facility is complete. This expansion is an additional 24,000 sq ft, giving JPSA a total of over 58,000 sq ft of work space. The addition provides JPSA the ability to meet customers' increasing demand for their LED, Solar, and excimer laser micromachining systems. The new expansion doubles the size of the existing laser production area used to manufacture JPSA's laser micromachining systems. The expansion will also provide state-of-the-art clean rooms, R&D laboratories to develop cutting-edge micromachining applications and ergonomic office space to accommodate growing customer service and engineering teams.

Sales Appointments

Delta Microwave announces the appointment of **Micro Lambda** as the company's exclusive representative in the Mid-Atlantic territory. Micro Lambda will represent Delta Microwave in Delaware, Maryland, Virginia, West Virginia, the District of Columbia, Southern New Jersey, and Eastern Pennsylvania.

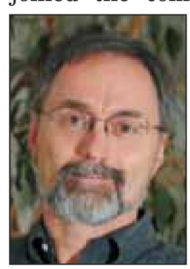
People in the News

Hibernia Atlantic announces the appointment of **Lloyd Jarkow** to the position of Chief Financial Officer. Mr. Jarkow has 16 years of finance experience within international telecommunications. Mr. Jarkow has spent the past 13 years at AboveNet Communications, most recently as Vice President of Corporate Development. Prior to AboveNet, he served as Finance Director for Metromedia International Telecommunications. He holds an undergraduate degree from Cornell University and an MBA from UCLA Anderson.

AR Modular RF has announced that **Jason Kovatch** has joined the company as a Development Engineer. His

responsibilities will include creating and supporting all automated test systems as well as supporting amplifier system product development and testing. Jason began his career as an associate at the Jet Propulsion Laboratory in Pasadena, CA, where he worked on microgravity containerless processing technology. Also while at JPL, working for the NASA Deep Space Network, he was responsible for creation of cryogenic maser amplifiers and instrumentation used for deep space probe communications and research. Jason was the Cognizant Design Engineer for Ground Microwave equipment in the Deep Space Network until 1995, at which time he moved to Hewlett Packard as a custom systems engineer. He later worked in the Signal Analysis division of Agilent Technologies until joining AR Modular RF. In his spare time, Jason is a ham radio operator (his call is NJ7K) and he spends as much time on or near the waters of Puget Sound as he can.

TRM Microwave announces that **Ken Greenwood** has joined the company as Director of Engineering. Ken



brings over 25 years of extensive experience in engineering management, systems engineering, product development and customer relations to TRM and he has co-authored multiple patents as a registered Electrical PE (Professional Engineering) in NH and CA. His background includes the design and development of terrestrial, air-to-ground communications systems and ESM (Electronic Support Measures) systems, with Hughes Aircraft Co., Lockheed, Sanders and various commercial start-ups involved in telecommunications and optical infrastructures. Ken earned an Electrical Engineering BS in Communication Systems from WPI, after 4 years service in the Air Force, and a Master's of Science in (Microwave) Engineering through a Fellowship with Hughes Aircraft.

Georgia Tech Executive Vice President for Research Stephen Cross has named Battelle Memorial Institute's **Robert T. McGrath** the new director of the Georgia Tech Research Institute (GTRI) and Georgia Tech vice president. McGrath, currently a consultant on National Laboratory/University Partnerships, STEM Education, and Race to the Top initiatives for Battelle Memorial, will begin his new responsibilities Feb. 1, 2011. McGrath served as the Senior Vice President for Research at The Ohio State University from 2004 to 2008 and as Associate Vice President for Research and Director of Strategic and Interdisciplinary Initiatives at The Pennsylvania State University from 1998 to 2004. During his tenure, he built extensive partnerships between academic, industry, and government sectors on a variety of applied research projects.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High-Frequency Algorithmic Advances in EM Tools for Signal Integrity—Part 1

By John Dunn
AWR Corporation

This two-part series discusses new advances in the algorithms underlying EM simulation techniques for signal integrity designers. Part 1 examines fast solution techniques for Method of Moments solvers

Only 30 years ago, electromagnetic (EM) simulation software was not a mainstream tool of the practicing electrical engineer, yet today it is essential for successful circuit design. EM simulation is used throughout the design process—creating and verifying models, checking the layout of the circuit for parasitic coupling issues, and verifying the final circuit's performance. The most obvious reason for the growing use of EM simulation is its increasing ability to simulate large circuits of practical interest, made possible by more powerful computers. What is not as well known is that a number of new conceptual breakthroughs in EM theory have also contributed greatly to the increased power of EM simulators.

The purpose of this two-part article is to introduce the reader to two of the more successful of these ideas. It is important that users of EM software understand the basic algorithms being used if they are to use the tools effectively and avoid common pitfalls. This first article focuses on advances in simulation methods for planar solvers, or as they are sometimes called, 3D planar or 2.5D simulators. We will see how the time required to solve a problem can be dramatically sped up by using fast, compressed, iterative solution methods. The methods have been developed over the past 20 years and are now reaching fruition in commercially available software. The second article explains how fast frequency sweeping allows for accurate solution of a problem over a frequency range with fewer frequency points.

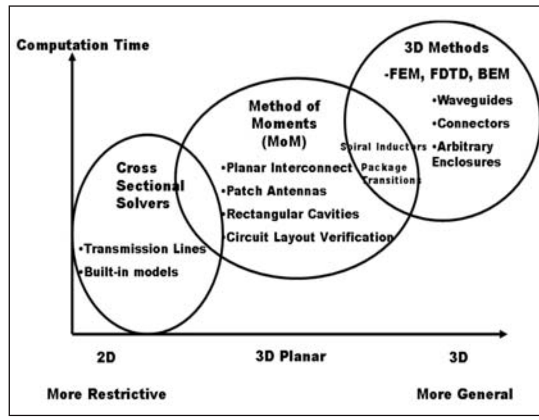

Figure 1 · Comparison of different classes of EM simulators.

The Basic Challenge

EM problems do not scale well with increased problem size. This means that as the problem becomes larger with more unknowns, the computational resources required increase at faster than a linear rate with the problem size. Planar simulators work by meshing the conductors in the circuit up into small cells, usually triangles or rectangles. The most popular EM technologies require the problem to scale by the power of 3, or $O(N^3)$. By this we mean that if the number of cells is doubled, the solution time increases by a factor of eight (2^3). Because of this faster-than-linear scaling, increasing the size of the problem quickly becomes impractical. The fast methods explained in this article scale as $O(N\log N)$, which is much better than $O(N^3)$.

Classes of EM Simulators

Three classes of EM simulation technology are shown in Figure 1—cross sectional

solvers, 3D methods, and the method of moments, which is the focus of this article. The horizontal axis indicates how general a geometry the solver can analyze, and the vertical axis shows relative computation time.

3D methods work by meshing the geometry into small cells, and the electric field is then solved in each cell. The cells are usually 3D triangular shapes known as tetrahedra, but other cell types such as cubes are possible. The finite element method is the most well known, but other good choices exist. The main advantage of these methods is their generality with regard to the geometries they can solve. The main disadvantage is they are computationally very intensive.

Cross sectional solvers are available in many commercial circuit simulators. Their purpose is to characterize transmission lines by taking the cross-section of the line and solving for its electrical parameters per unit length. This is accomplished by solving a simplified set of Maxwell's equations, the quasi-static approximation for the cross-sectional geometry. These solvers are extremely fast and make it easy to get a line's electrical properties. The user normally does not even know that he or she is using electromagnetics when one of these models is used. The disadvantage is that they work only for transmission line models.

Moment methods (MoM), the technique used in the middle bubble of Figure 2, work by solving for the currents on the circuit's conductors. Moment method codes are typically written so that they can solve for currents on horizontal conductors and vertical conducting walls and vias. This is the reason for the terms "2.5D" and "3D" planar solvers. Moment methods require a Green's function in order to calculate the coupling between all the current elements on the conductors. Unfortunately, the Green's function can only be solved for certain geometries. In particular, the conductors must be on infinite, homogeneous, dielectric layers.

It is possible to have an infinite ground plane on the bottom of the dielectric stackup, and a metallic cover if desired. A variation of the method also allows the problem to be bounded by a conducting rectangular box. The requirement of planar, dielectric layers is why the method is not as general as the 3D methods mentioned earlier. However, the planar simulators cover a wide variety of practical technologies—printed circuit boards, modules, and chips, for example. They can usually solve larger problems than full 3D simulators, as they are solving only for currents on the conductors compared to the electric fields throughout the entire space of the problem.

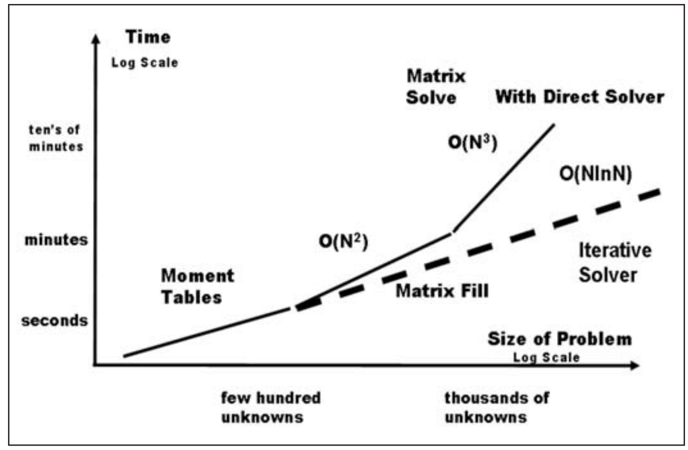

Figure 2 · Simulation times for the MoM.

The Solution Time for Conventional Planar Solvers

The method of moments solves for the unknown currents on the conductors. This is accomplished by first meshing the conductors into a finite number of cells and then solving for the unknown current on each cell. The unknown current is simplified on each cell, and a linear approximation is mostly used. The job of the simulator is to find the unknown linear varying current on each cell. A matrix equation is set up and solved for the unknown currents.

Each matrix entry gives the interaction between the currents and charges of two cells by using the Green's function. This set of interactions can be thought of as mutual capacitance and inductance between the cells. In addition, there are self-capacitance and inductance terms and resistance terms. Radiation and phase delay between the elements is also included. After the interactions between the mesh elements have been calculated, the matrix can be formed. If there are N cells in the problem, the size of the matrix is $N \times N$ and there are N^2 unknowns. The matrix is solved for the unknown currents on the conductors.

How long does it take for MoM codes to simulate a given size structure? The answer is shown in Figure 2. The Green's functions are first calculated for the given geometry, which takes about $O(N)$ time, where N is the number of meshes (Note that $O(N)$ means order N). The time it takes to simulate with N unknowns is $K \times N$ where K is a positive number. Compare this to $O(N^2)$: The time is equal to $M \times N^2$, where M is a positive number. For a large problem, the $O(N^2)$ process will dominate the $O(N)$ process, but the constant K could be larger than M, so the $O(N)$ could dominate the time of simulation for a small problem. The matrix next needs to be filled (i.e., each element in the matrix must be calculated), which takes $O(N^2)$.

The matrix equation must then be solved, and with conventional techniques, and this requires $O(N^3)$. For a large problem, the solution time will dominate. MoM is an $O(N^3)$ process, so if the number of meshes is doubled, the simulation will take eight times longer. The amount of

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SPDT (R)	0.3	44	39.8 (0.8 dB)	0.45	23	40.5 (0.1 dB)	SKY13290-313LF
SP3T (R)	0.3	49	26	0.45	27	29	SKY13317-373LF
SP4T (R)	0.3	49	26	0.6	28	30	SKY13322-375LF
SP4T (R)	0.3	54	41	0.45	24	38 (0.1 dB)	SKY14151-350LF

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consumed memory is also a factor, as the matrix must be stored while the matrix equation is solved. Virtual memory is used when the physical memory is consumed, swapping data in and out of the hard drive. This is a very slow process, taking milliseconds instead of microseconds each time memory is accessed. In practice, the simulation becomes impractical once the computer begins to use virtual memory.

To determine how large a problem can be solved on today's computers, assume each unknown is represented by 16 bytes of data (double-precision complex floating point). 1000 unknowns requires $1000 \times 1000 \times 16 = 16$ Mbytes of RAM, and 10,000 unknowns takes 1.6 GBytes of RAM. A 32-bit computer can access up to 4 GBytes of RAM, with much of that reserved for the operating system and other applications, and a practical limit for RAM available for the simulation is 2 GBytes, which corresponds to about 10,000 cells. This memory limitation is overcome by 64-bit operating systems, and 16 GBytes of RAM is typically installed in these computers.

Fast Iterative Solvers for MoM

The biggest bottleneck in simulating large problems with MoM simulators is solving the matrix equation. A direct solve, which requires the direct solution of the matrix, is $O(N^3)$, so doubling the number of unknowns increases the simulation time by a factor of 8. This produces a near order-of-magnitude increase in solution time and memory requirements and does not lend itself well to larger physical problems. Fortunately, new developments in solver technology have overcome several hurdles and have reduced solution time. Globally, these techniques attempt to bypass a direct solution of the entire matrix and in doing so avoid the $O(N^3)$ computation cost.

The methods work by iteratively solving a smaller, approximate representation of the matrix. These methods rely on a well-conditioned matrix. The condition number of a matrix is a figure of merit that indicates how easily the matrix can be solved with finite precision arithmetic on a computer. A poorly conditioned matrix (large condition number) indicates that the solution will not be accurate, or perhaps not even solvable. EM numerical methods usually result in poorly conditioned matrices because of the nature of the equations. The biggest challenge for developers of fast EM solvers is overcoming the poor conditioning of the matrices that arise for various reasons. To understand these poorly conditioned matrices and how to mitigate them, it helps to focus on one typical cause of poor conditioning, the so-called DC catastrophe.

MoM codes solve for currents and charges on conductors, and there is one equation that involves both current and charge. Current and charge are not independent quantities for frequency domain problems. A current changing in space means that there is a charge buildup.

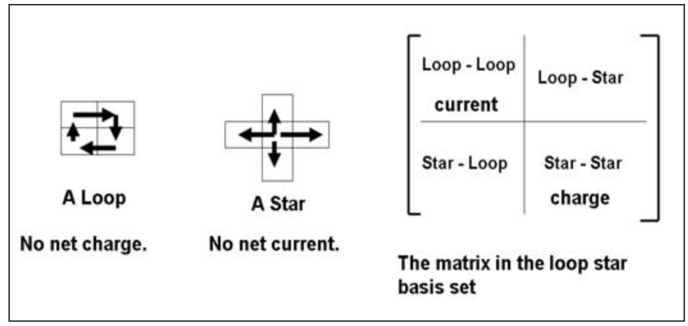

Figure 3 · Loop-star preconditioning of cells.

For example, when current goes around a bend in a line there is charge buildup at the corner of the bend.

At DC, charge and current are independent of each other, with charge the source of electric fields and current the source of magnetic fields. It is only when the frequency is non-zero that electric and magnetic fields produce each other and current is related to charge. If solving one equation for current and charge at DC, this equation has no unique solution. At low frequencies, the solution is problematic and the condition number of the matrix is very large. The ultimate example of this problem is at DC, where there is not even a unique theoretical solution—resulting in the "DC catastrophe." This problem is common to almost all EM simulators and at low frequencies is not practically solvable.

"Preconditioners" that operate on the problem in a way that improves the condition number before trying to solve it can be used to overcome this problem. The idea is to take the meshes (or basis functions) and recombine them into something that allows the charge and current terms to be separated in the equation. This allows the designer to approximately solve for the two separate problems of charge and current, producing the approximate DC electrostatic and magnetostatic solutions. These approximate solutions can then be used as a good way to precondition the original matrix to make its solution practical.

There are many forms of preconditioners. This discussion will focus on the loop-star type, which gets its name because of its regrouping of the cells, as shown in Figure 3. The loop of cells has no net charge, as the current is flowing in a closed loop. The star grouping has no net current, as the current modeled by this configuration of cells flows outward from the center. In the loop-star preconditioner, the problem can be reformulated by using the loop configuration and the star configuration as two new basis functions to describe the problem, This is much like the way vectors (1,0) and (0,1) can be used to describe points in the Cartesian plane. The preconditioner should thus give the matrix a lower condition number because the matrix produced by the new basis should have more zero

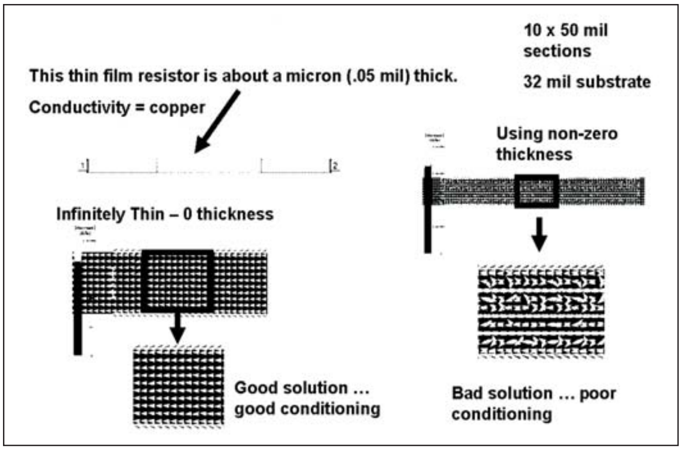

Figure 4 · Thick line that is poorly conditioned.

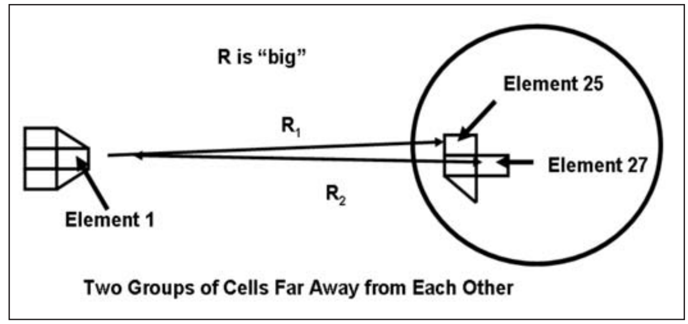

Figure 5 · Two groups of cells in a multipole method.

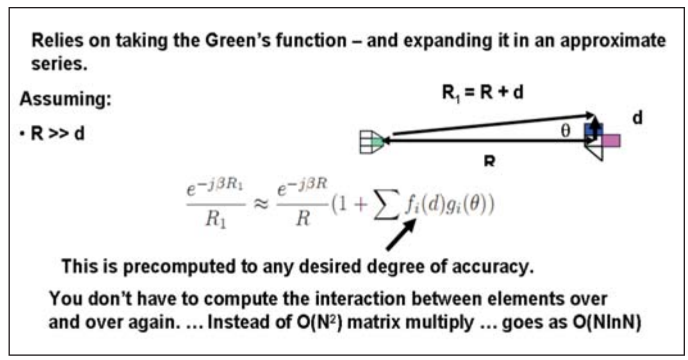

Figure 6 · Expanding out the Green's function in the multipole method.

entries then the initial formulation.

Poorly conditioned matrices can occur in a variety of other ways besides low-frequency issues. For example, there could be a situation in which a transmission line has non-zero thickness. There are currents on both the top and bottom sides of the line, but if the line is not very thick, the currents are almost on top of one another (Figure 4).

If the line becomes infinitely thin, there are two unknown currents to be solved for but only one equation for the total current, so there is no unique solution to the problem. Figure 4 shows how the currents become very poorly behaved as a result of the poor conditioning of the matrix. It is important to avoid thick lines that are too thin, or in the case of boundary element methods, regions that are too thin. Modern planar simulators therefore need to employ a variety of preconditioners to cover the various causes of poorly conditioned matrices.

The Compressed, Fast Solve Algorithm

Once the matrix has been properly conditioned, the next challenge is to solve the better-conditioned problem in faster than $O(N^3)$ time. Methods have been developed that can go as fast as $O(N \ln N)$ time. The two most common methods are fast multipole techniques and compressed matrix techniques. These methods are based on the fact that the interaction between groups of cells that are far apart have two important properties: the interactions between the groups are small, and they are all about the same value. To understand this concept, referring to the two groups of cells separated by a large distance in Figure 5, imagine the size of the interaction between a cell in the left group and a cell in the right group, R. Compare it to the interaction between the same cell in the left group and a neighboring cell in the right group. They (R1 and R2) are probably about the same. This fact is exploited in fast multipole methods (Figure 6) in which the Green's function is expanded out as a series in d, where d is the local distance of the second group from its center of mass.

The fast multipole method has become extremely popular for antenna and radar codes but has three limitations when applied to SI problems. First, the method relies on large distances compared to wavelength, which is normally not the case for an SI problem. Second, the method relies on expanding the Green's function out into a series of functions, but codes for SI engineers have more complicated Green's functions, making this procedure difficult. Third, fast multipole methods can become numerically unstable when the size of the problem is not electrically large.

Fast matrix compression methods are more commonly used in the SI community. These methods are similar to multipole methods because they rely on groups of cells that are far away and have about the same interaction strength as shown in Figure 7. However, they do not rely on analytical manipulations of the Green's function. Rather, they numerically exploit the fact that the matrix entries for a given block of the matrix are about the same because of the previous observation that in this given block the interactions are between two far away cells. For example, Figure 7 shows a block in the matrix with cells 25 and 27 interacting with cell 1. Elements A1,25 and

A1,27 will have about the same values as the interaction is between cells in two different blocks.

Since the elements in the block all have about the same values, it is possible to represent the elements with many fewer members. Using mathematical algorithms very similar to those from compression standards like JPEG or MPEG, the number of members needed to represent the block can be reduced. For the world of images, compression methods permit digital or video cameras to represent pictures using less data than the uncompressed original. The matrix representing the photograph or video is represented by a series of reduced matrices that provide an approximation that is "good enough" in most cases. Of course, what is actually good enough is in the eye of the beholder, but suffice it to say that this technique can be applied also to SI problems to gain measurable improvements in EM performance while achieving acceptable accuracy.

Once the matrix blocks have been reduced in size using these compression algorithms, the matrix is solved using an iterative solver. This is perhaps the key difference from the direct solve technique. The iterative solver never solves the original matrix equation and is therefore not necessarily bound by the $O(N^3)$ scaling but instead multiplies the compressed matrix by a vector. If performed correctly, the entire process can potentially proceed in $O(N \ln N)$ time. It should be noted that to get $O(N \ln N)$ scaling the actual matrix is never calculated, as this takes $O(N^2)$ time. Rather, successive approximations of the matrix are calculated in a consistent manner. Achieving overall improvement with a particular implementation of an iterative solver depends on whether the whole is better than the direct solver.

All the pieces of the iterative solver must when working in concert take less time than the direct solver with its brute force approach. The exact increase in speed achieved by the iterative solver depends on the details of the geometry being studied. The method relies on groups of blocks that are far from one another. Geometries in which there are a large number of nearby cells (overlapping shapes on different layers in a package, for example) will not compress as well as geometries in which most shapes are spread out on one layer.

With all the components of an iterative solver working properly and for problems of a few thousand unknowns, an increase in speed should be achieved when compared to the direct solver shown in Figure 2. However, the direct solver continues to perform well, especially at RF and microwave frequencies with several hundred unknowns. This point should not be overlooked because some modern commercial EM solvers incorporate both direct and iterative solvers with algorithms that analyze a particular problem and then select the fastest or most accurate method.

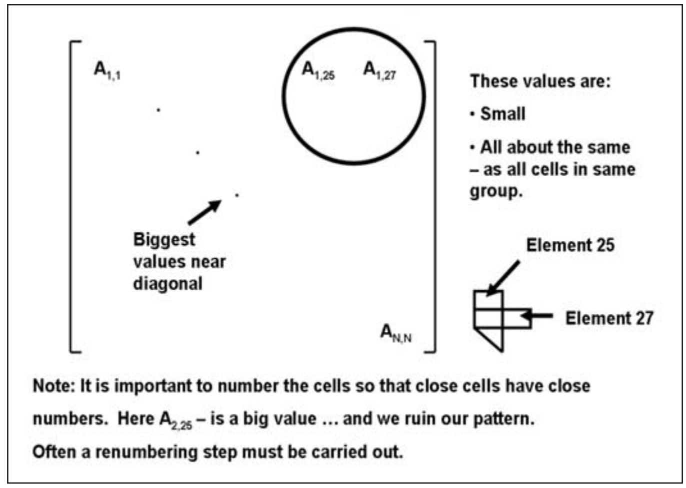

Figure 7 · A matrix block between two groups of cells.

Challenges Remaining

Fast solvers promise to give designers a tool that can solve complex problems that are simply not approachable with traditional methods. However, a number of research challenges remain. As discussed in this article, preconditioners must be used for the matrices to be well conditioned and the methods to work. Right now, there is no one best preconditioner. It is also not always clear which one will work well for a given problem. As the methods enter into more into mainstream usage, this problem will become more relevant to the designer. Therefore, the challenge for EM tool developers is to make the methods more robust and automated.

Next Month

The second article in this two-part series will describe ways to reduce solution times by using fewer frequencies while maintaining the frequency resolution of the resulting dataset. It will address intelligent selection of simulation frequencies such as fast frequency sweeping algorithms that reduce the number of frequencies required to obtain the desired frequency range as well as efficient methods to map EM solutions from the frequency-domain to the time-domain.

Author Information

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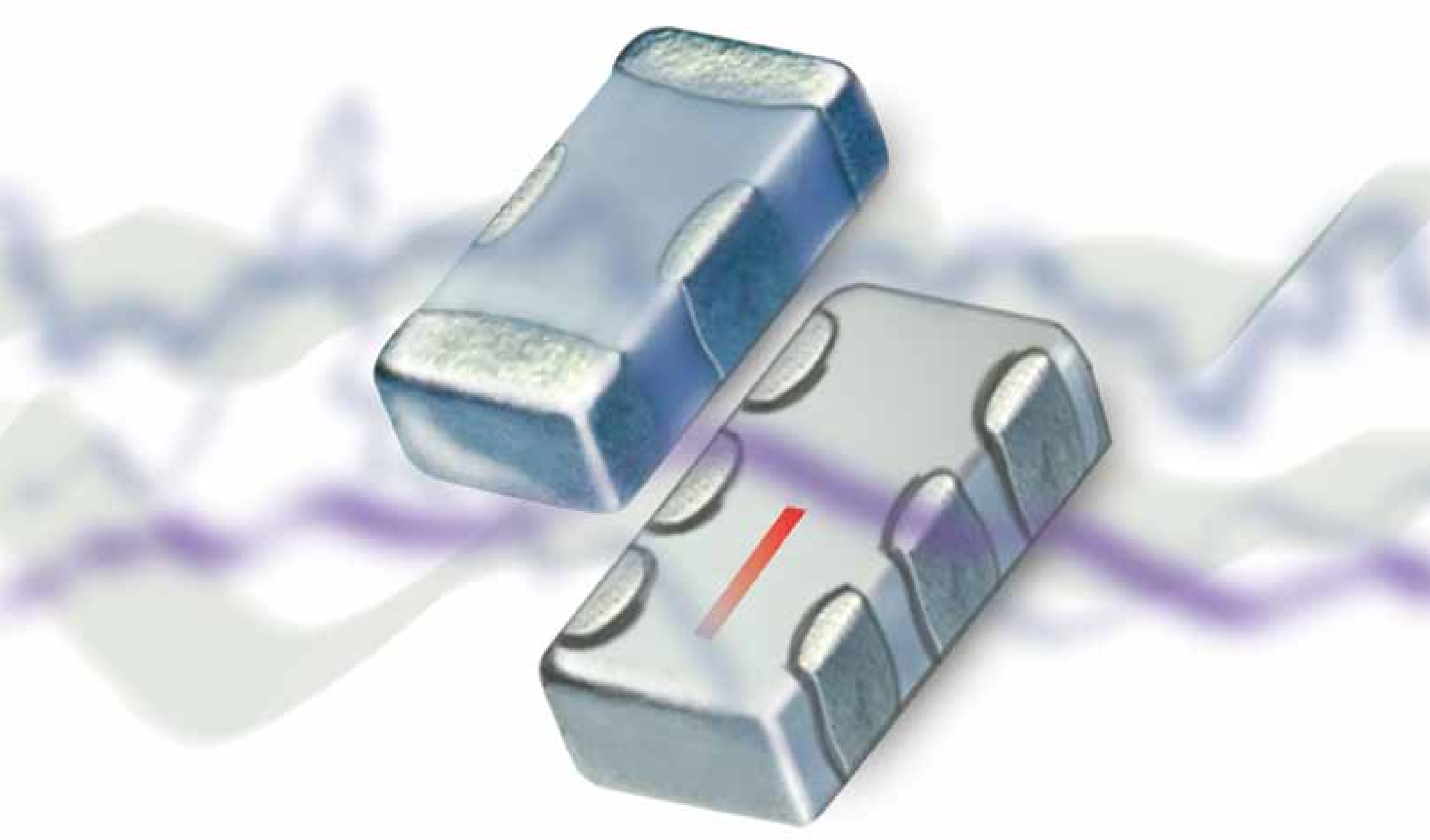

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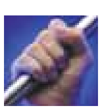

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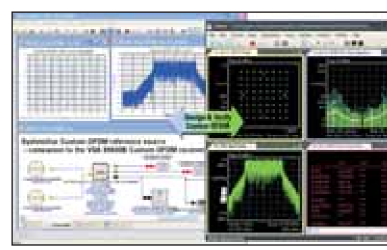

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Notes on Design and Deployment of Wireless Infrastructure Systems

Most recent news regarding wireless infrastructure has been about deployment of 4G (WiMAX and LTE) systems. Although this upgrade to wireless networks is progressing, it is still limited to the largest markets, with many big cities still not yet developed, despite the messages delivered in the advertising of the major wireless providers.

3G systems are more common, with coverage to all populated areas. However, some of these areas are underserved and unless located along interstate highways, many rural areas have service that is best characterized as “voice only” or “slow 3G data”.

Cost of new equipment is certainly a factor in the speed of buildout of 4G wireless systems—for both wireless company investment and consumer purchase of advanced devices and upgraded service plans. However, technical issues are a bog factor, too. The design of base station radios, antenna systems, portable devices, and system routing/switching equipment is very different when an individual connection is expected to pass data at megabit rates instead of kilobits. In addition, higher data rates bring new uses beyond voice, text and the occasional photo. Equipment back end processing must also be upgraded to support these new services and applications.

Higher Data Rate Wireless Requirements

Let’s take a look at some of the technical issues involved with the level of service promised for 4G wireless systems:

System support

Backhaul capacity—connecting cell sites to the operators’ switching centers is undergoing a major upgrade to support the huge increase in data throughput. This work is underway as fast as crews can plow optical fiber into the ground and/or install high capacity mm-wave point-to-point radio systems. Such work does not happen instantaneously, and is subject to slowdown by weather, supply chain and manpower limitations.

System management—4G wireless provider central office equipment more closely resembles an Internet server farm than a telephone network. New management processes to handle the increased traffic are required, including both data handling and cell site management.

Modulation Bandwidth

Adaptive radios—Making best use of the available bandwidth can mean changing the radio channel between low bandwidth uses such as voice and text, medium rate uses such as web browsing and low resolution video phone, plus the highest data rate services like video download or streaming, online gaming, or video conferencing. PC adapters that use the wireless networks for a high speed connection will add another level of usage beyond that of smartphones and their apps.

Regulatory cooperation—Although out of the headlines, regulatory support and technical standards remain a big part of the behind-the-scenes work in wireless technology. Frequency allocation may be more important in international markets where wireless services are still developing, but even in the U.S., 700 MHz and TV “white space” spectrum is being explored for high speed data services. New policies on “net neutrality” promoted by the U.S Federal Communications Commission (FCC) will affect the business plans of wireless providers in the same way it will affect other broadband Internet portal services.

Signal-to-Noise Margin

High radio performance—Communication theory shows us that higher data rates can be achieved through increased bandwidth and/or higher signal-to-noise ratio (SNR). Achieving the highest possible bandwidth in a given channel bandwidth demands the highest possible SNR performance over the path from transmitter input to the data output of the receiver. Complex modulation schemes that support high data rates require a channel that has maximum transmitter effective radiated power (ERP) and receiver dynamic range, usually determined by the difference between minimum discernible signal (MDS) and the level where intermodulation distortion (IMD) prevents reliable demodulation.


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Optimized coverage

The radio channel that carries high data rate signals can be improved in several ways that increase the SNR:

Smart antennas—Steered-beam antennas and multiple input/multiple output (MIMO) technology are part of the solution for achieving high SNR. However, these technologies require additional antenna hardware and control systems to make them work. Steered-beam antennas are implemented only at the cell site, but MIMO requires dual antennas and radio front-ends in portable devices that use the technology. This will affect cost, size and battery lifetime, which is already a design challenge for supporting radio function and data processing required for the new range of high speed applications.

Microcells and picocells—Another means of increasing SNR is a shorter signal path, taking advantage of the $1/r^2$ rolloff of signal level versus distance. Thus, a signal path that is half the distance requires one-fourth the power level to achieve the same SNR. Adding additional small cells to the system accomplishes this improvement, but the tradeoff is significant, since those microcells and picocells require backhaul just like a major cell site. In many cases, the reduced capacity of a small cell may make backhaul easier where a good wired infrastructure exists, but it will be difficult in other areas.

Dynamic system management—Better management of resources will be a major part of all wireless providers' efforts. Some software enhancements will be relatively simple to implement, but it is expected that there will be more reliance on human decision-making by system operating personnel until new traffic and hardware management patterns and control algorithms are developed. Examples of management improvements include better location tracking of users to optimize handoffs between cell sites, and separate operating schemes for inbound and outbound rush hour mobile users.

Increased Reliability

User habits are already changing dramatically where high speed networks are currently deployed, challenging the ability of the network to keep up with traffic in some cases.

Longer connect times—With greater capability, users spend more time using their smartphones, pad computers and other devices over the wireless network. Even at this very early stage of 4G deployment, weaknesses have been revealed that must be addressed before usage increases further.

High-rate apps—Video streaming for television programming, movies and games was not possible before 4G, and this new family of applications represents entirely new challenges for the evolving wireless networks. Combined with longer "on" times, these applications are driving the dramatic order-of-magnitude growth in the amount of data being handled by the networks.

Higher consumer expectations—As implied in the above paragraphs, consumers accustomed to their "always on" home high-speed Internet connections will expect the same level of service when operating wirelessly. This is encouraged by the wireless providers, who are promoting their services as a replacement for the present wired services. In these early months of 4G deployment, service problems have made the news more often than kudos for providing new wireless capabilities.

Summary

Wireless infrastructure to support high speed data services still has technical and operational challenges to meet before achieving the goal of "anywhere, anytime" service. Of course, some of the difficulties are simply a result of trying to get systems operating as quickly as possible—and some lessons must be learned from experience. Despite the significant challenges, there is little debate over the ability of engineers to solve them. The only question is how long it will take.

The Mathematics of Mixers: Basic Principles

By Gary Breed
Editorial Director

This month's tutorial is a first introduction to the mathematical principles that describe the operation of frequency mixers

Mixers are classic RF/microwave circuits that make it possible to translate RF signals from one frequency to another. Ideally, they implement this frequency change with no effect on the amplitude and frequency components of the signal's modulation.

Frequency Translation

Mixers are nonlinear circuits; they rely on near-perfect nonlinearity. This sounds like a contradiction, but it means that perfect switching—discontinuity being the ultimate nonlinearity—will result in ideal mixer behavior. We will describe how this switching takes place in a circuit later on, but first let's review the overall behavior of the mixing process.

Nonlinear response creates new signals where none previously existed. In the case of two unmodulated signals applied to the input of a nonlinear device, there will be a series of output signals that contain multiples of the input signals (harmonics), plus sums and differences of ALL signals, fundamental and harmonic, as described by [1]:

$$f_{out} = |nf_1 \pm mf_2|$$

where f_{out} represents all output signals, f_1 and f_2 are the two input signals, n and m are the order of the harmonics, from zero (fundamental) to infinity.

Mathematically, this is an infinite Fourier type of series, where the amplitude of each discrete output frequency dependent on the order. Higher order results are lower in amplitude, with the actual rate of decrease versus order determined by the quality of the mixing circuit. In all cases, the second order responses will have the highest amplitudes:

$$f_1 + f_2$$
$$f_1 - f_2 \quad \text{(actually: } |f_1 - f_2|\text{)}$$

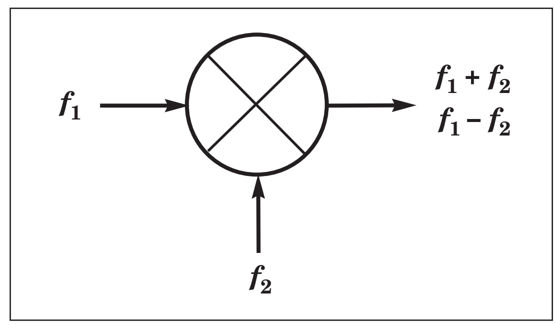

Figure 1 · The frequency translation scheme that is the goal for a frequency mixer.

$2f_1$ and $2f_2$ are also second-order outputs, but nearly all practical mixers use a *balanced* design to suppress these outputs, as well as all other even-order harmonics.

Figure 1 shows the frequency translation scheme we want to obtain from an ideal mixer. If there are no other outputs, if components are ideal (lossless), then the circuit performs the function of multiplication [1], represented as the trigonometric identity:

$$\cos(\omega_1)\cos(\omega_2) = [\cos(\omega_1 + \omega_2)]/2 + [\cos(\omega_1 - \omega_2)]/2$$

where $\cos(\omega_1)$ and $\cos(\omega_2)$ are the time-domain representations of f_1 and f_2. The 1/2 factors simply show that the input amplitude is divid-

ed between the two output terms. In practice, this represents a 6 dB *conversion loss*.

Usually, we want only one of the mixer's outputs, so the unwanted signal must be removed, either by filtering, or by implementing an *image-reject* mixer topology that is actually two mixers with phase shift circuitry that results in a single sum or difference output. Filters have finite stopbands, and image-reject mixers have finite rejection of the unwanted signal. In a sensitive receiver, these imperfect responses may allow strong signals outside the desired passband to be detectable. To minimize this possibility, the relationship of input and output signals must be considered. $f_1 + f_2$ should be chosen so higher-order responses do not fall within the passband of the *intermediate frequency* (IF) filter. Rather that repeat the equations and charts for this type of analysis, References [2, 3] should be consulted.

Real Circuit Performance

An ideal mixer requires perfect switches, as illustrated in Figure 2. In this double-balanced circuit, switches A-D, and B-C are alternately activated at the *local oscillator* frequency, which is the unmodulated signal that determines the amount of frequency difference between input and output signals. In this ideal mixer, the local oscillator signal is not a sine wave, but an ideal square wave with normal and inverted polarity providing the "push-pull" or balanced LO control to the switches.

However, practical circuits do not have zero loss resistance or instantaneous transition times, so an analysis of performance must include these terms. Oxner [4] provides the following description:

An ideal square wave drive will result in switching action according to the Fourier series:

$$F(x) = \frac{1}{2} + \frac{2}{\pi}\sum_{n=1}^{\infty}\frac{\sin[2n-1]\omega t}{[2n-1]}$$

The switching function is derived from this equation as a power function by squaring the first term. Thus the output power deliverable to the output (IF) is:

$$P_{out} = \frac{V_0}{R_L} \qquad \text{or,}$$

$$P_{out} = \frac{V_{in}^2 R_L}{\left[\frac{\pi^2}{4}\left(R_g + R_{SW}\right) + R_L + R_{SW}\right]^2}$$

where R_L is the load impedance, R_g is the internal loss and R_{SW} is device loss (diode junction, or FET R_{DS}).

Conversion efficiency is obtained by the ratio of P_{avg} and P_{out}:

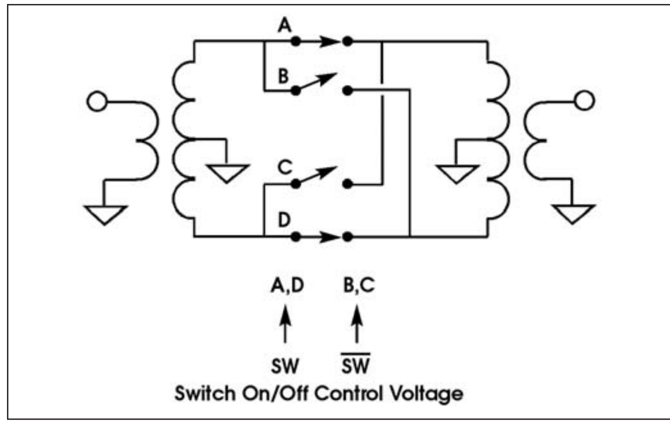

Figure 2 · An ideal mixer has devices (diodes or transistors) that act as perfect switches.

$$L_{conv} = 10\log\frac{\left[\frac{\pi^2}{4}\left(R_g + R_{SW}\right) + R_L + R_{SW}\right]^2}{\pi^2 R_L R_g}$$

Using the above equation, an ideal switching mixer would have a conversion efficiency (in dB) of:

$$L_{conv} = 10\log\frac{4}{\pi^2}$$

which is –3.92 dB. Thus, all mixers will have greater than 3.92 dB conversion loss.

Finally, Oxner provides the following expression that describes the switching function relative to the rise/fall time of the LO switch driver signal (for FET switches):

$$20\log\frac{\left[t_r\omega_{LO}\frac{V_s}{V_c}\right]^2}{8}$$

where V_c is the peak oscillator voltage, V_s is peak signal voltage, and t_r is the rise/fall time of V_c.

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Design of Input Matching Networks for Class-E RF Power Amplifiers

Firas Mohammed Ali Al-Raie
The Polytechnic Higher Institute

Class-E power amplifer design usually emphasizes output network design, but this article examines amplifier performance when an untuned input is replaced with a matching network

Class-E power amplifiers have high efficiency, which makes them attractive in modern wireless mobile communication systems. There are several techniques and approaches developed solely to design the output load networks of such amplifiers to shape the RF power device's output voltage and current for minimum power loss. However, little attention is brought to the design of the input matching network and to the device bias conditions, with their effects on the overall circuit performance. This paper attempts to discuss these topics through a systematic design and simulation approach for a typical 5 watt class-E power amplifier operating at 150 MHz.

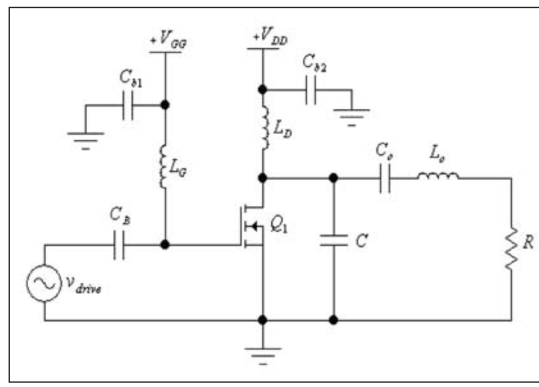

Figure 1 · Typical class-E power amplifier with shunt capacitance configuration.

Introduction

Several methods have been developed for the design of the load network for class-E RF power amplifier. Among those are the shunt capacitance [1], shunt inductance [2], finite DC feed inductance [3], and parallel circuit [4] techniques. The most popular configuration is the shunt capacitance technique due to its simplicity and *designability*, which means that when the amplifier is built as designed, it works as expected [1].

The schematic diagram of the class-E power amplifier with shunt capacitance configuration is presented in Figure 1. In this circuit L_G and L_D represent the gate and drain bias RF chokes respectively, C_B is a DC blocking capacitor, C_{b1} and C_{b2} are bypass capacitors, V_{GG} is the gate bias voltage, V_{DD} is the drain supply voltage, C is the capacitor shunting the active device Q_1, L_o and C_o constitute a series resonant circuit tuned at the operating frequency, and R is the optimum resistance seen by the load network for the required output power. The active device Q_1 (power MOSFET in this case) operates as an ON/OFF switch.

In class-E power amplifier circuit, efficiency is maximized by minimizing power dissipation in the active device, while providing the desired output power. The circuit can be arranged so that high drain voltage and high drain current don't exist at the same time.

For idealized class-E power amplifier operation, it is necessary to provide the following optimum conditions for the drain voltage $v_D(t)$ across the power MOSFET just prior to the start of the device's ON state at the moment $t = T$, where T is the period of the input driving signal [5]:

$$V_D(t)\big|_{t=T} = 0 \qquad (1)$$

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BMZ		1.05 + .01 (f) GHz
BZ		1.02 + .05 (f) GHz
Insertion Loss:		
BMA		.03 x √(f) GHz
BMMA		.04 x √(f) GHz
BMZ		.06 x √(f) GHz
BZ		.15 x √(f) GHz
Float, Inches (mm):		
BMA	Radial: .020 (.51)	Axial: .060 (1.5)
BMMA	Radial: .020 (.51)	Axial: .060 (1.5)
BMZ	Radial: .020 (.51)	Axial: .060 (1.5)
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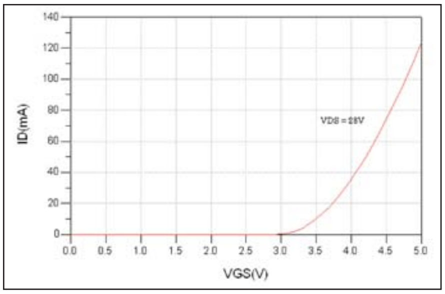

Figure 2 · I_D versus V_{GS} for the MRF134 power MOSFET.

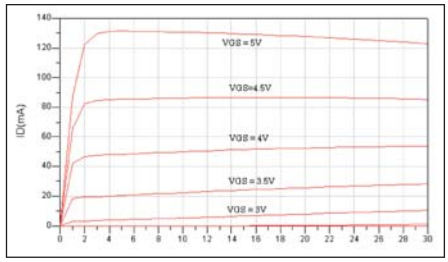

Figure 3 · Simulated drain DC characteristics of the MRF134 power MOSFET.

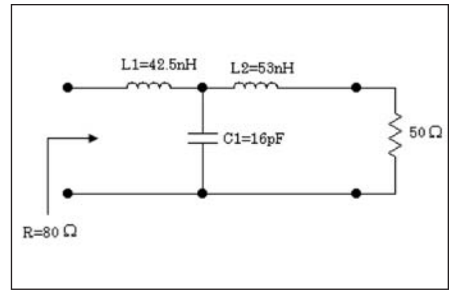

Figure 4 · Configuration of the output matching network.

$$\left.\frac{dv_D(t)}{dt}\right|_{t=T} = 0 \tag{2}$$

Equations (1) and (2) state that the drain voltage should be zero at the turn-on moment, and that the slope of this waveform is zero at the same moment.

Load Network Design Equations

The load network of class-E power amplifier is not intended to provide a conjugate match to the transistor output impedance. Design equations for the load network elements (C, C_0, L_0, and R) can be derived by writing time domain equations for the voltage $v_D(t)$ at the drain of the RF power MOSFET when it is OFF, and the current $i_D(t)$ passing through the RF device when it is ON. A set of simultaneous differential equations can be formed according to the necessary conditions (1) and (2) and solved to determine the network elements [6].

Nathan Sokal, the inventor of this amplifier, has developed explicit form equations to calculate the values of the network elements at any output power and loaded quality factor Q_L.

These equations are formulated as [7]:

$$R = 0.5768\left(\frac{V_{DD}^{\,2}}{P_{out}}\right)\cdot\left(1 - \frac{0.451759}{Q_L} - \frac{0.402444}{Q_L^{\,2}}\right) \tag{3}$$

$$C = \frac{1}{5.44658\omega R}\left(1 + \frac{0.91424}{Q_L} - \frac{1.03175}{Q_L^{\,2}}\right) + \frac{0.6}{\omega^2 . L_D} \tag{4}$$

$$C_o = \frac{1}{\omega R}\left(\frac{1}{Q_L - 0.104823}\right)\left(1 + \frac{1.01468}{Q_L - 1.7879}\right) - \frac{0.2}{\omega^2 L_D} \tag{5}$$

$$L_o = \frac{Q_L . R}{\omega} \tag{6}$$

where P_{out} is the required output power, and ω is the operating frequency.

The value of Q_L can be selected based on a trade-off between operating bandwidth and harmonic distortion of the output signal.

Typical Class-E Power Amplifier Design

For the clarification of the goals of this paper, a power amplifier circuit has been designed and simulated using a commercial microwave CAD program. Design specifications of the amplifier are to achieve an output RF power of 5 W from an input driving level of 0.5 W, and drain efficiency of more than 65% at an operating frequency of 150 MHz.

The following sections describe a step-by-step design procedure with the simulated results obtained from the Agilent's ADS microwave circuit analysis program.

RF Power Device Selection and Characterization

The first step of the amplifier design procedure is the selection of the RF power transistor. For this design, the Motorola's power MOSFET MRF134 has been chosen. This device is capable of delivering 5 W at 400 MHz with a typical power gain of more than 10 dB. It operates from a 28 VDC supply and has a typical drain-to-source breakdown voltage of 65 V. The RF transistor library of the computer program ADS contains a SPICE model for this transistor which simplifies the simulation process. The simulated input DC characteristic (I_D versus V_{GS}) of the power MOSFET is shown in Figure 2 with V_{DS} = 28 V. It can be shown from this curve that the gate threshold voltage $V_{GS(\mathrm{th})}$ = 3 V. On the other hand, Figure 3 presents the simulated output DC characteristic (I_D versus V_{DS}) at several gate voltages. The drain ON resistance $R_{D(\mathrm{on})}$ can be estimated from Figure 3 as 12.5 Ω. This relatively large value of $R_{D(\mathrm{on})}$ will cause a reduction in amplifier's efficiency due to the dissipated power at the drain during the ON period of the power device.

Calculation of the Load Network Elements

The design procedure begins by calculating the component values of the load network using equations (3) to (6). For output power P_{out} = 5 W, operating frequency f =

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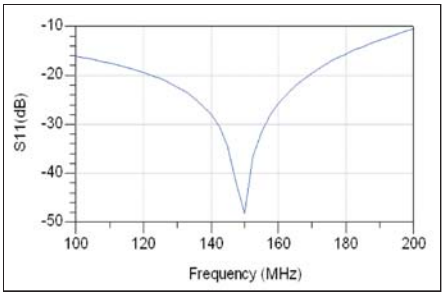

Figure 5 · Input return loss of the output matching network.

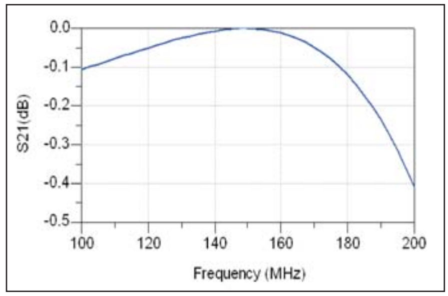

Figure 6 · Insertion loss of the output matching network.

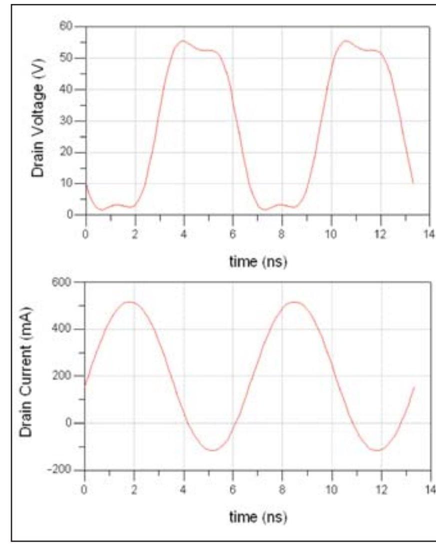

Figure 8 · Simulated drain voltage and current waveforms of the power MOSFET.

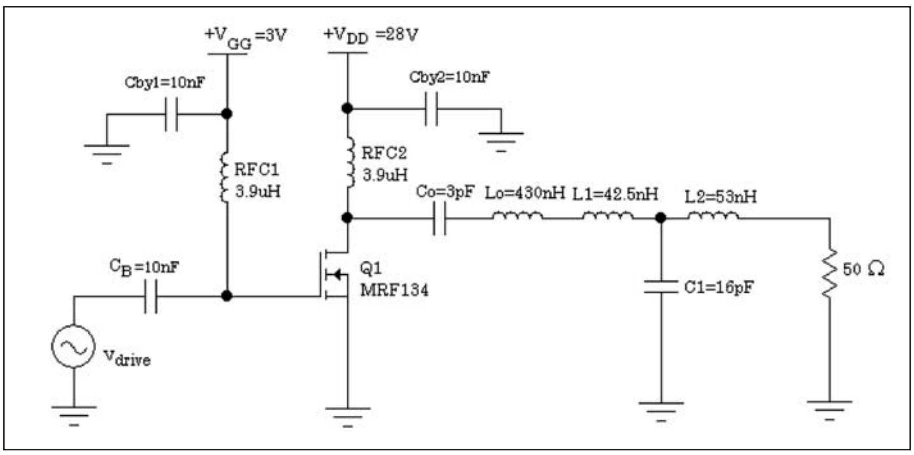

Figure 7 · The designed class-E power amplifier without the input matching network.

150 MHz, loaded quality factor Q_L = 5, and drain power supply V_{DD} = 28 V, the calculated values of the elements of the load network are:

R = 80 Ω, C = 3 pF, C_0 = 3 pF, and L_0 = 430 nH.

The output capacitance of the RF device, C_{out}, is measured as 9.7 pF from the device data sheet. This means that it is greater than the required value of the shunt capacitance. The excess value of 6.3 pF should be tuned out by part of the drain bias RF choke.

Design of the Output Matching Network

An output matching network is needed to transform the 50 Ω amplifier impedance into the required load resistance, which is set to be 80 Ω. This network has been designed with the aid of an immittance Smith chart, and is implemented in a T-section configuration, as shown in Figure 4.

In addition to the transformation function of the output matching network, it also can be used to reduce the harmonic content of the output RF signal. Figure 5 shows the simulated input return loss of this network, while Figure 6 presents its insertion loss versus frequency.

Design of the Biasing Network

The biasing network consists of the drain and gate RF chokes, bypass capacitors, DC blocking capacitors, in addition to the gate and drain bias voltages. For 50% duty cycle operation, the transistor is biased at the threshold point, which means that $V_{GG} = V_{GS(\mathrm{th})}$ = 3 V. This bias point actually corresponds to class-B mode.

Based upon the previous calculations, the schematic diagram of the amplifier circuit is shown in Figure 7.

Amplifier Performance Simulation

The designed amplifier circuit has been simulated using ADS 2006A. With a single tone input signal of 0.5 W power level and an operating frequency of 150 MHz, the RF device's drain voltage and current are sketched as depicted in Figure 8. As shown from this plot, the peak values of drain voltage and current don't exist simultaneously which minimizes the device's power loss. However, at the ON time of the RF signal, the drain voltage is about 3.5V due to the ON resistance at the drain, $R_{D(\mathrm{on})}$. This may degrade the overall efficiency of the circuit. During the OFF interval of the RF signal, a negative current flows through the power MOSFET's output capacitance C_{out}.

In Figure 9 the output signal waveform is plotted, while its spectrum is displayed in Figure 10. It is obvious that harmonics are reduced to acceptable levels due to the

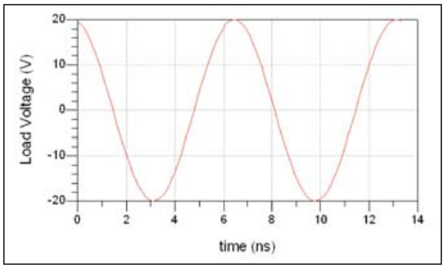

Figure 9 · Simulated waveform of the load voltage.

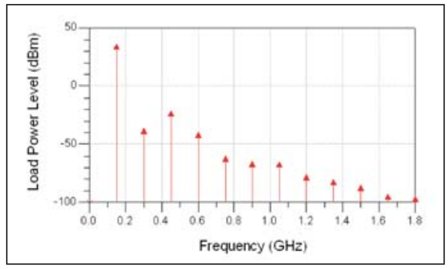

Figure 10 · Power spectrum of the output signal.

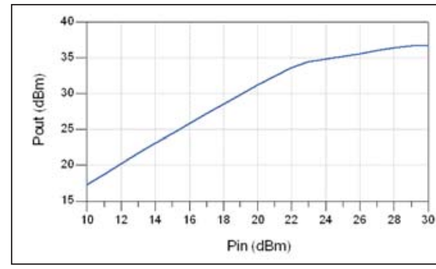

Figure 11 · Amplifier output power versus input power.

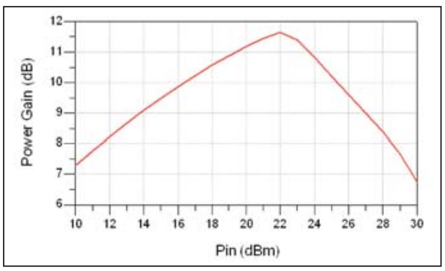

Figure 12 · Power gain versus input power.

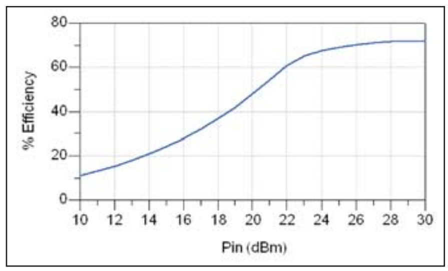

Figure 13 · Simulated efficiency versus input power.

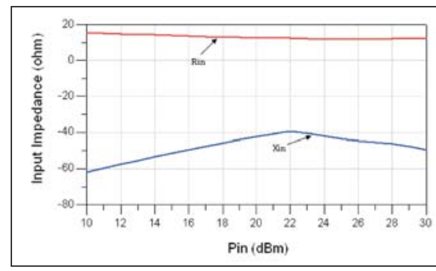

Figure 14 · Variation of the large signal MOSFET input impedance with input signal level.

filtering effect of the load and matching networks.

In order to display the power amplifier's performance characteristics, a sweep of the input power level has been carried out from 10 to 30 dBm at the operating frequency. Figure 11 shows a sketch of the output power versus input power. The output power is obtained from:

$$P_{out} = real\left(0.5 V_L . I_L^*\right) \quad (7)$$

where V_L and I_L are the peak values of the fundamental components of load voltage and current respectively. The output power is about 36 dBm at an input level of 27 dBm.

Figure 12 presents the operating power gain of the amplifier, G_p, versus input power. The power gain is calculated from:

$$G_p(dB) = P_{out}(dBm) - P_{in}(dBm) \quad (8)$$

Notice that the power gain is about 9.0 dB at an input power level of 27 dBm.

Finally, Figure 13 displays a plot of the amplifier's DC to RF efficiency with input power. The amplifier efficiency is about 71.1% at an input power of 27 dBm.

The efficiency of the amplifier circuit has been evaluated from:

$$\eta = \frac{P_{out}}{P_{dc}} \quad (9)$$

where P_{dc} is the DC power consumed by the RF device and is obtained from:

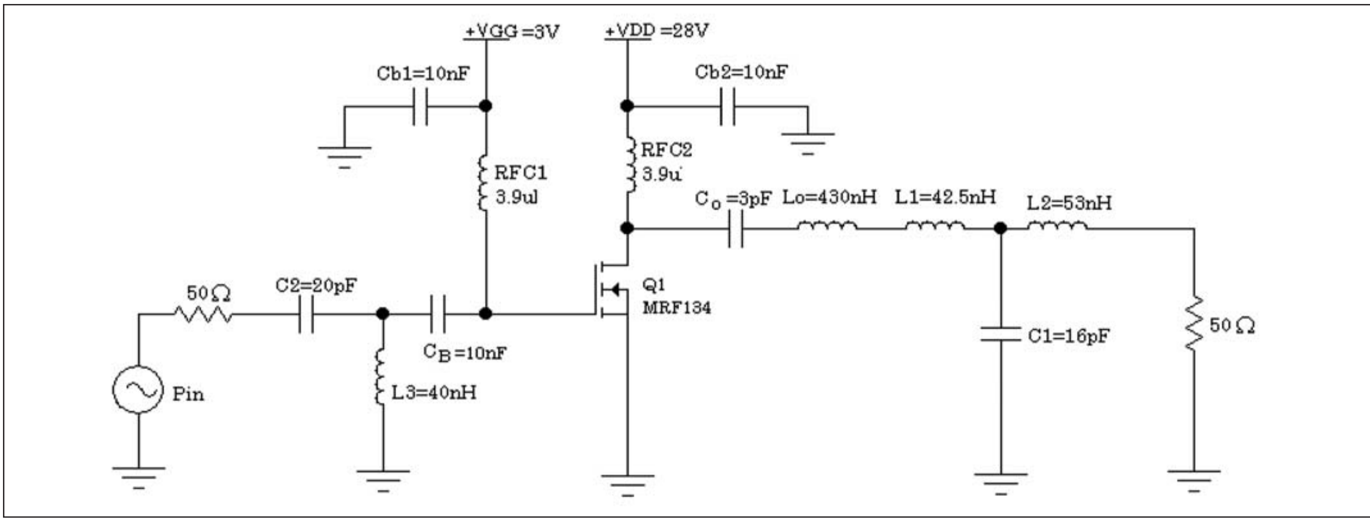

Figure 15 · The designed class E power amplifier after adding the input matching network.

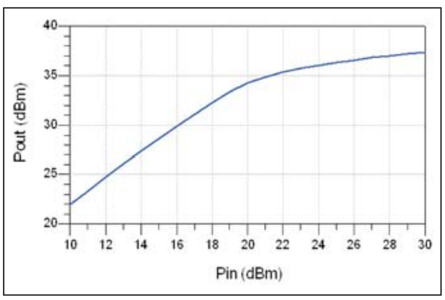

Figure 16 · The simulated output power versus input power for the final amplifier.

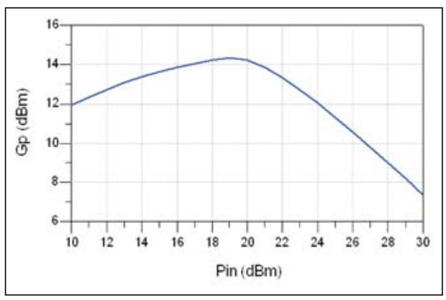

Figure 17 · Operating power gain versus input power.

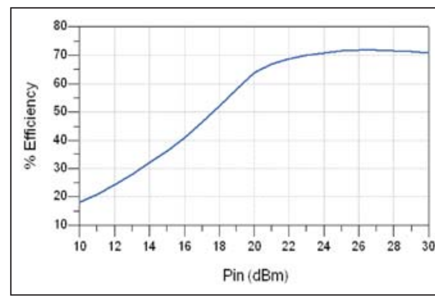

Figure 18 · Amplifier efficiency versus input power.

$$P_{dc} = V_{DD}.I_{DD} \tag{10}$$

where I_{DD} is the DC component of the drain current.

Input Matching Network Design

The input matching network can be designed to match the large signal input impedance of the RF power device with the 50 Ω source impedance. Therefore, the large signal input impedance of the RF transistor should be estimated at the nominal input power, operating frequency, and bias voltages with the existence of the load and output matching networks.

The large signal input impedance of the power transistor consists of two parts, resistance R_{in} and reactance X_{in}:

$$Z_{in} = R_{in} + jX_{in} \tag{11}$$

R_{in} and X_{in} can be estimated from:

$$R_{in} = real(V_{in} / I_{in}) \tag{12}$$

$$X_{in} = imag(V_{in} / I_{in}) \tag{13}$$

where V_{in} is the fundamental component of the input voltage at the gate of the MOSFET, and I_{in} is the fundamental component of the current entering the gate of the transistor. I_{in} can be estimated using a current probe with the aid of ADS simulation capabilities. Figure 14 shows a sketch of the large signal input impedance of the power device versus input power.

As shown from the plot in Figure 14, the input impedance is capacitive. At an input power of 27 dBm (0.5 W), the input impedance is approximately 12 –j45 Ω. The input matching network can thus be designed to match this value with the 50 Ω source impedance. An immittance Smith chart has been used to construct an L-section matching network graphically. Figure 15 presents the final power amplifier circuit after incorporating the input matching network.

There is no doubt that the input matching network improves the net input power delivered to the RF device. The amplifier circuit was simulated again after adding the input matching circuit using ADS. The output power of the circuit is displayed in Figure 16 with a sweep of input power from 10 to 30 dBm. As shown from Figure 16, there is a slight increase in output power being

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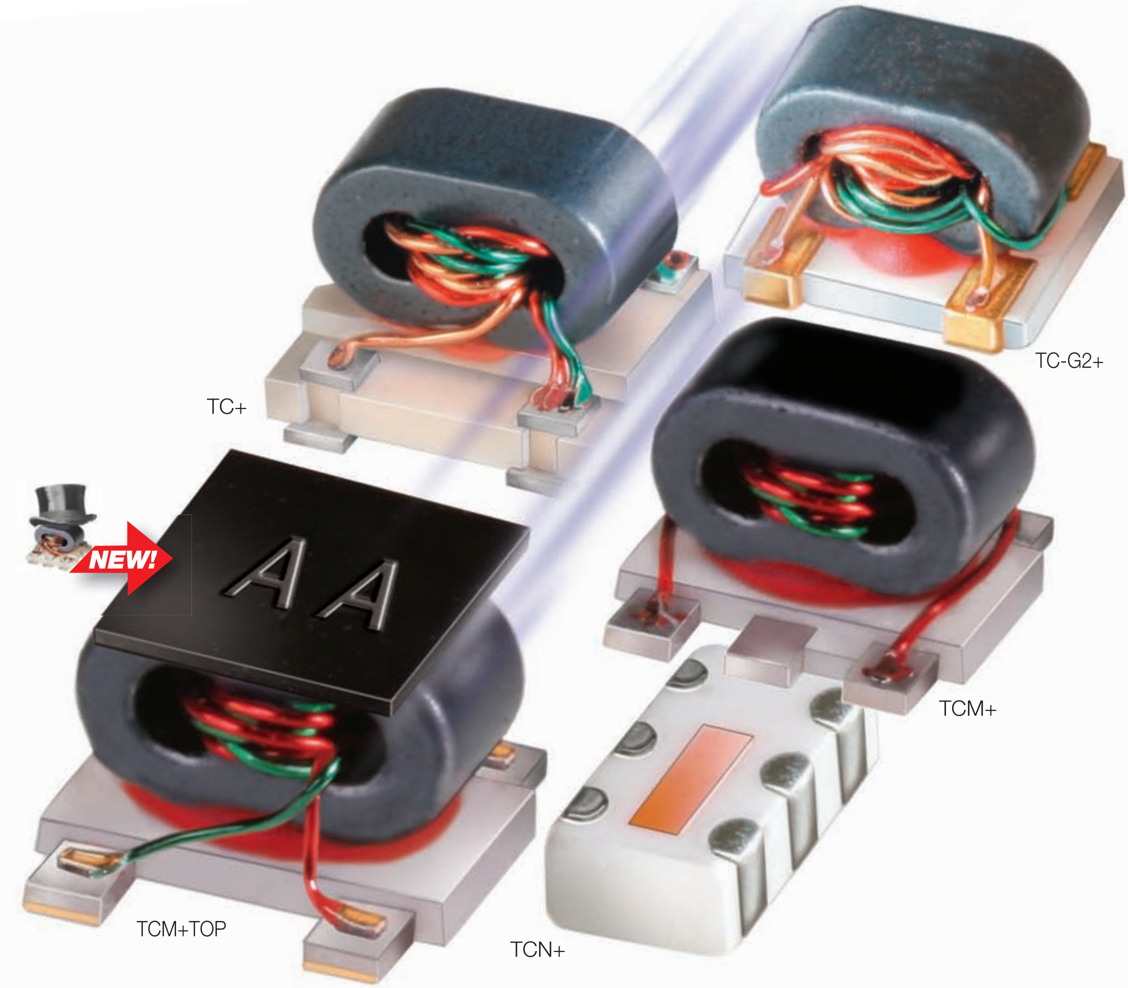

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38.8 dBm for an input power of 27 dBm. The power gain is plotted in Figure 17, and becomes equal to 9.8 dB at the nominal input power. The DC to RF efficiency is sketched in Figure 18 versus input power. The efficiency becomes 71.8% at an input power level of 27 dBm. However, no attempts have been made to optimize the component values of the input matching network for better performance characteristics.

Table 1 summarizes the performance of the amplifier before and after adding the input network.

	Before	After
Output Power	36 dBm	36.8 dBm
Power Gain	9.0 dB	9.8 dB
Efficiency	71.1%	71.8%
P_{in} = 0.5W (27 dBm), f_0 = 150 MHz		

Table 1 · Performance comparison before and after adding the input matching network.

Conclusion

The performance of Class E RF power amplifier with a traditional shunt capacitance load network has been studied thoroughly. It was shown that the high efficiency operation of such amplifiers is determined mainly by the output load network. However, with an accurate and proper design of the input matching network, the performance characteristics of the amplifier can be improved. This article has presented and discussed the main guidelines for synthesizing the input matching circuits for this type of RF amplifier to achieve the improved performance.

Acknowledgement

I would like to express my deep gratitude to Dr. Andrei Grebennikov for his continuous advice and useful suggestions in high efficiency power amplifier analysis, and also for providing me with some technical papers and e-books in the field of RF power amplifiers.

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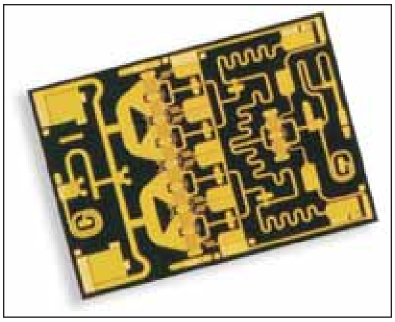

GaN Power Amplifier

TriQuint Semiconductor, Inc. has released a new 30W state-of-the-art gallium nitride (GaN) power amplifier with high power and efficiency for counter-IED (C-IED) and other EW (electronic weapons) systems. The TGA2576 delivers 30W of saturated output power in the 2.5-6 GHz range. The new device is fabricated using TriQuint's production-released GaN on SiC process; it typically offers 30% PAE and 25 dBm of small signal gain. Die-level samples of the TGA2576 are available now; packaged samples are expected in early 2011.

TriQuint Semiconductor, Inc.
www.triquint.com

HFC Broadband Products

RF Micro Devices, Inc. announced the expansion of its broadband product portfolio targeting Hybrid Fiber Coax (HFC) transmission network hardware, cable head end, cable hub equipment, and Multi Dwelling Unit/Multiple Tenant Unit (MDU/MTU) equipment. The new broadband products enhance cable operators' network performance and enable increased bandwidth to support rapidly growing digital-centric services, such as HDTV, 3D HDTV and "mid-split" DOCSIS® networking. The use of mid-split architectures can expand the amount of usable upstream or return path bandwidth by more than 160 MHz. This enables multiple service operators (MSOs) to scale the data services on their existing cable transmission infrastructure, while delivering an enhanced Internet user experience and expanded digital services (e.g. HDTV streaming, HDTV downloads at ten times real-time data rates) to cable subscribers.

RF Micro Devices, Inc.
www.rfmd.com

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Microlab, a wholly owned subsidiary of Wireless Telecom Group, Inc., announces the addition of new 4 × 4 hybrid matrices with exceptional 30 dB isolation and <–160 dBc PIM. This has been achieved with radical new circuitry, eliminating the signal path discontinuities of previous designs while having all inputs conveniently located on the same side. The CM-58D covers the 1710-2700 MHz band, which includes the new LTE and WiMAX band in the 2490-2690 MHz spectrum. The CM-58D is designed for up to 100W average, 3 kW peak power per input, and may be moisture sealed to meet IP67, so is suitable for use in both outdoor and indoor applications such as at radio base stations, in signal combiner boxes and in-building distribution systems. Similar units are available for other wireless bands, and with N connectors.

Microlab/FXR
www.microlab.fxr.com

Battery Powered Wide Band Amplifier

The new BOX-010200 amplifier from AtlanTecRF provides a portable and rugged gain block for situations where cordless operation is either essential or preferred. It is a self-contained battery-powered amplifier providing 25 dB of gain across the whole 1-20 GHz frequency range with typically +13 dBm output power at the 1 dB gain compression point and 3-5 dB noise figure. The MIC design utilises GaAs PHEMT distributed amplifier MMIC technology, together with proprietary matching circuits within a robust die-cast aluminium alloy housing, which also contains a rechargeable NiMH battery. A battery charger is supplied with each unit. The amplifier is fitted with precision type N connectors, male input and female output as standard.

Atlantic Microwave Ltd.
www.atlantecrf.com

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AML Communications, Inc.
www.amlj.com

Grounding and Lightning Protection

Times Microwave Systems has published several technical papers addressing the subject of lightning protection and grounding solutions for wireless networks. These essen-

tial technical resources are available on the Times website and reprints of these technical publications are also available. The December 2011 issue of *Above Ground Level* (AGL) magazine also featured a new article dedicated to the subject of grounding techniques and the relevant effects on the proper operation and performance of in-line lightning protection devices.

Times Microwave Systems
www.timesmicrowave.com/lightning/articles/

Handheld Spectrum Analyzers

The R&S FSH4 and R&S FSH8 handheld spectrum analyzers from Rohde & Schwarz offer many useful new functions for installing and servicing CDMA® and LTE mobile radio networks. In their basic configuration, the analyzers can now perform measurements in the spectral range. They can, for example, detect spurious emissions and display the associated limit values. Thanks to their 20 MHz bandwidth, they are suitable for analyzing LTE signals. Using two new options, R&S FSH-K46 and -K46E, the analyzers are also capable of analyzing CDMA2000® signals. The new R&S FSH-K46 and R&S FSH-K46E software options provide all the functions that service and installation technicians need to check the operationally relevant parameters of CDMA2000 base station signals. The R&S FSH-K46 option measures the power, determines the ratio of peak power to average power, and displays characteristic parameters such as the power in both the pilot channel (F-PICH) and the synchronization channel (F-SYNC), the carrier frequency offset, composite EVM (error vector) and Rho. The R&S FSH-K46E option has been designed for more in-depth analysis and can also perform code domain power measurements.

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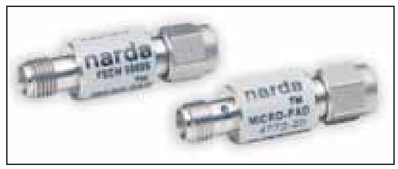

Fixed 3-dB Attenuator

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Crystek Corporation
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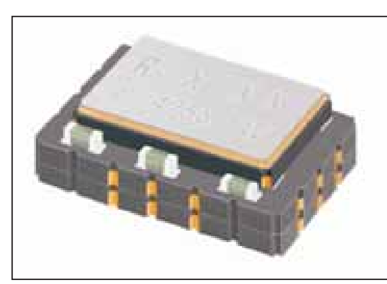
National Instruments
www.ni.com



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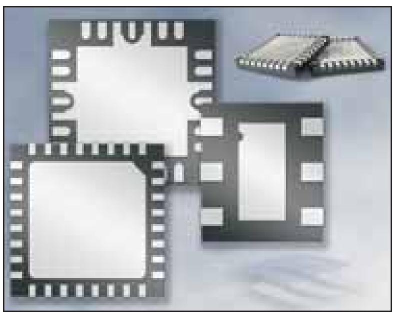

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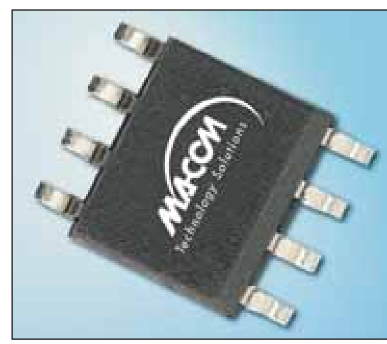
Taiyo Yuden Co., Ltd.
www.t-yuden.com



HBT Driver Amplifier

M/A-COM Technology Solutions Inc. introduced the MAAM-009563, a two-stage HBT driver amplifier for cellular and WiMAX infrastructure applications. This driver amplifier covers a broad frequency range of 250 to 3000 MHz with excellent linearity of 47 dBm output IP3 and typical gain of 19.5 dB. The broad operation is achieved using external matching components, where component values are selected to center the 200 MHz instantaneous bandwidth within the overall frequency range. The lead-free SOIC-8EP surface mount plastic package is RoHS compliant and compatible with solder reflow temperatures up to 260°C. The ESD susceptibility achieves a class 2 ESD rating. Engineering samples, sample boards and production devices are available today from stock.

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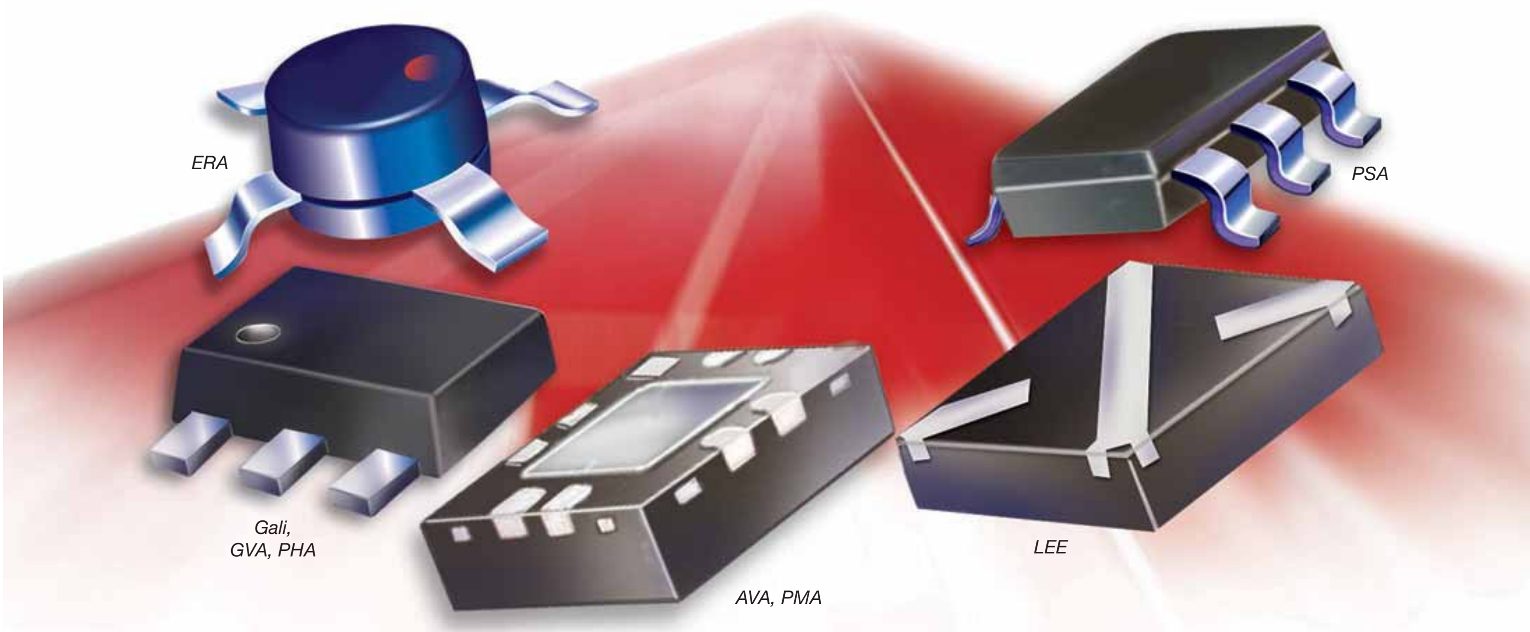

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EMC Technology
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Laird Technologies, Inc. announced the release of its new removable top board-level shield (BLS), ReCovr™. The ReCovr product line incorporates the functionality of a two-piece shield without the need for a separate frame and cover. The shield is specially designed with a locking mechanism that allows for easy removal of the shield cover when access to board-level components within the shield is required. The locking mechanism makes repair of components under the shield quick and easy by eliminating the need for removing the entire shield and reducing board re-work. The removable top shield also integrates Laird Technologies' rigid corner board-level shield technology, which incorporates a new corner design that optimizes component rigidity for increased part and printed circuit board (PCB) firmness.

Laird Technologies, Inc.
www.lairdtech.com

PA Gain Block Family

Avago Technologies announced two new gain block solutions that expand its high-performance power amplifier family targeting cellular infrastructure applications. The new MGA-31589 and MGA-31689 0.5-watt gain blocks feature high linearity, high gain, superior gain flatness and low power dissipation. The MGA-31589 gain block addresses cellular and WiMAX wireless base station and other wireless systems operating between 450 to 1500 MHz, while the MGA-31689 device addresses these applications operating between 1500 to 3000 MHz. The MGA-31x89 power amplifier family is optimized for frequency in order to deliver improved performance across all the major cellular bands—GSM, CDMA, and UMTS—plus next-generation LTE bands. The new gain blocks join with the 0.25-watt MGA-31189 and MGA-31289 devices and the 0.10-watt MGA-31389 and MGA-31489 devices to serve applications from 50 to 3000 MHz.

Avago Technologies
www.avagotech.com

Polarized BNC Cables

The BNC polarized contact panel mount assembly, similar to the twin contact assembly, features a two-stud bayonet locking system to prevent signals from mixing. These assemblies range from 0-200 MHZ and are used with 78-ohm and 95-ohm conductor cables. Unlike the Twin Contact, the BNC Polarized Contact assembly includes a male and female contact to prevent incorrect mating. These assemblies are typically used for balancing low levels/high sensitivity circuits. Applications include computer networking/LAN, process control, military and aerospace. CST Cable manufactures these cable assemblies to any length and quantity with quick turn around times. CST can also manufacture these in box build with harness lay in designs.

CST, Inc.
www.cstcable.com

Broadband Demodulator

Analog Devices, Inc. introduced a new highly integrated demodulator for broadband applications, such as cellular base stations, satellite communications, point-to-point (PtP) radios and defense systems. The ADRF6850 operates over a broad frequency range from 100 to 1000 MHz and supports narrow band and wideband signal modes up to 250 MHz. The device offers best-in-

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VSWR: 1.2:1 maximum
Switching Speed: 2 μsec maximum
Temperature Range: -40 to +75°C
Size: 8.0" x 4.0" x 1.2"

Broadband Airborne IFF SPDT Switch

600 – 1600 MHz

Frequency Range: 600 – 1600 MHz
3.5 kW peak, 35 μsec pulses, 1.6% duty cycle
Loss: 0.33 dB typical, 0.6 dB maximum
Isolation: 40 dB typical, 35 dB minimum
VSWR: 1.35:1 maximum
Switching Speed: 2 μsec typical, 3 μsec maximum
Temperature Range: -55 to +91°C
Altitude: 70,000 feet
Size: 2.3" x 2.0" x 1.2"

Broadband High CW Power

150 – 6000 MHz

Frequency Range: 150 – 6000 MHz
150 W CW up to 1 GHz, 20 W CW 1 – 6 GHz
Loss: 1.6 dB maximum
Isolation: 55 dB minimum
VSWR: 2:1 maximum
Switching Speed: 20 μsec maximum
Temperature Range: 0 to 60°C
Size: 3.5" x 1.25" x 0.75"

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Frequency Range: 9.1 – 9.3 GHz
100 W peak, 1 μsec pulses, .5% duty cycle
Loss: 1.1 dB typical, 1.2 dB maximum
Isolation: 80 dB minimum
VSWR: 1.25:1 maximum
Switching Speed: 70 nsec maximum
Rise/Fall Time: 40 nsec maximum
Temperature Range: -55 to +85°C
Size: 1.5" x 1.5" x 0.4"

Space and ruggedized packaging are also available.

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class RF performance with an input IP3 (third order intercept) of 22.5 dBm, P_{1dB} (input compression point) of 12 dB and a noise figure of 11 dB. The ADRF6850 is in production now and is available for $7.50 per unit in 1,000-unit quantities. It is the companion device to ADI's recently released ADRF6750 modulator.

Analog Devices, Inc.
www.analog.com

PA and DVGA Products

Hittite Microwave Corporation has introduced a new GaAs HBT MMIC power amplifier and a wide dynamic range digital variable gain amplifier (DVGA), which are ideal for high linearity applications in cellular/3G, LTE/4G, WiMAX, military and fixed wireless equipment applications from 400 to 2700 MHz. The HMC921LP4E is a high linearity GaAs HBT MMIC 2 watt PA that is rated from 400 to 2700 MHz. This versatile power amplifier delivers up to 16 dB gain, +33 dBm output P_{1dB}, and +48 dBm output IP3. The HMC926LP5E is a DVGA that operates from 700 to 2700 MHz, and can be programmed to provide from 6.5 dB to 38 dB of gain, in 0.5 dB steps at 900 MHz.

Hittite Microwave Corporation
www.hittite.com

High Power 8-Way Splitter-Combiner

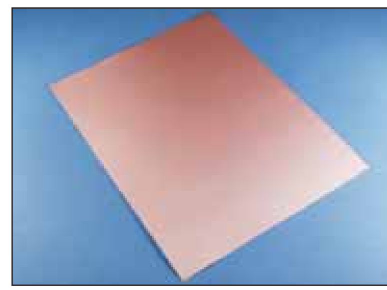
RADITEK's high power 8-way splitter-combiner operates from 1200-1400 MHz, with average input power of 40 watts, 3.5 kW peak, and average output power of 320 watts, 28 kW peak. Isolation is 20 dB min. on adjacent ports and 25 dB min. on non-adjacent ports. Connectors are SMA female (input) and 7/8" EIA (output).

RADITEK
www.Raditek.com

RoHS Compliant VCO

Z-Communications, Inc. announces a new RoHS compliant VCO (voltage-controlled oscillator) model CRO2595A-LF in S-band. The CRO2595A-LF operates at 2570-2620 MHz with a tuning voltage range of 0.5-4.5 Vdc. This VCO features a typical phase noise of –110 dBc/Hz at 10 kHz offset and a typical tuning sensitivity of 21 MHz/V. The CRO2595A-LF is designed to deliver a typical output power of 9 dBm at 5 VDC supply while drawing 27 mA (typical) over the temperature range of –40º to 85º C. This VCO features typical second harmonic suppression of –20 dBc and comes in Z-Comm's standard MINI-16-SM package measuring 0.5 × 0.5 × 0.22 in. It is available in tape and reel packaging for production requirements. The CRO2595A-LF is also ideal for automated surface mount assembly and reflow. CRO2595A-LF is well suited for digital radio and fixed wireless applications that require ultra low phase noise performance.

Z-Communications, Inc.
www.zcomm.com



Halogen-Free Laminates

Rogers Corporation introduced its new XT/duroid™ high performance thermoplastic laminate materials, ideal for high frequency multilayer circuits in the most demanding operating environments. The XT/duroid product line includes XT/duroid 8000 laminates for multilayer designs with as many as five layers and XT/duroid 8100 laminates for constructions with six or more circuit layers. Both laminates feature thin halogen-free dielectrics and are available with low-profile copper foil cladding for use in double-sided and multilayer printed circuit boards (PCBs). Rogers' new XT/duroid 8000 laminates feature a *z*-axis dielectric constant of 3.23 ±0.05 at 10 GHz and a dissipation factor of 0.0035 or less at 10 GHz. They deliver stable electrical performance over wide frequency ranges, with a low thermal coefficient of dielectric constant of +7 ppm/°C from –50 to +150°C. They also exhibit excellent thermal conductivity of 0.35 W/m/°K.

Rogers Corporation
www.rogerscorp.com

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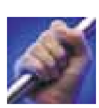
0.5-7200 MHz



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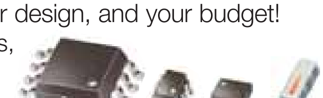

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2011 EDITORIAL CALENDAR

Issue	Products	Technology	Tutorial	Events*
January	Lab Test Equipment Capacitors & Inductors EDA Tools	Wireless Infrastructure	The Mathematics of Mixers	DesignCon
February	Connectors Oscillators Optical Products	Military Communications	Antenna Fundamentals	
March	Microwave Components Resistive Products Switches	Standards & Regulations	Design for EMC Compliance	12th WAMICON
April	RFICs/MMICs Field Test Equipment Couplers & Hybrids	Smart Grid / Smart Home	Designing Below 100 MHz	
May	3G/4G Infrastructure Substrates & Laminates New Literature	Research Update	Wafer Probing Basics	MTT IMS 2011
June	Antennas Cable Assemblies Front-End Components	Wireless Medical Systems	The Optical Interface	AP/URSI Symposium
July	EMC Products Power Devices Sensors	Broadband Everywhere	Power Device Technologies	IEEE EMC Symposium
August	High Speed Digital VCOs & Synthesizers Wireless ICs & Modules	Test Equipment Trends	High Performance Cables	
September	Aerospace & Hi-Rel Signal Analyzers Online Resources	EDA Tool Advances	Custom RFICs & MMICs	European Microwave Week
October	Power Amplifiers Microwave Materials Services	Cable/Connector Update	Connectors on PC Boards	
November	Filters MM-Wave Products Test Accessories	Power Amplifiers	Understanding EDA Models	
December	Discrete Semiconductors Precision Connectors New Literature	ISM Appilcations	MIMO and Smart Antennas	Radio Wireless Week

*At time of publication, bonus distribution is planned

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Meetings & Events · In the News · Design Notes
High Frequency Applications

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Frequency Mixer

The ADE-92+ is passive double balanced mixer featuring a FET quad in a ring configuration providing high dynamic range performance in a small, low cost, RoHS-compliant package. The ADE-92+ provides RF/LO response from 400 to 900 MHz and IF response from DC to 150 MHz. It is especially useful in systems where cost and performance are critical, such as GSM cellular system applications. This mixer is ideal for upconverter and downconverter applications and does not require external matching components or DC power.

www.minicircuits.com

New LTE Videos Web Page

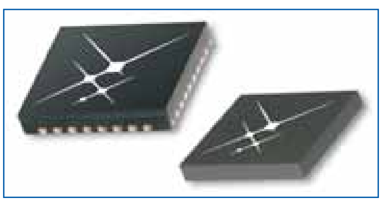
Agilent Technologies, Inc. invites you to gain greater insight into LTE design and test with the new 24/7 Agilent LTE Videos page. The site is designed to give information about LTE test for greater understanding, more rapid development and design into this complex technology. Videos are also available now on YouTube; search for “AgilentLTE.”

www.agilent.com/find/LTEVideos

K-Band Waveguide LNA

MITEQ Inc. introduces a new addition to its family of K-band waveguide LNAs. Model JDM2WK-18002600-25-10P is a very low noise, high dynamic range waveguide front end with WR-28 waveguide input and K(F) connector output. Additional options include pressure windows and three flange configurations, choke, grooved, and cover. This LNA is lightweight and has a small profile and footprint. The aluminum alloy housing is environmentally sealed and also fully EMI shielded. LNA includes reverse voltage protection and full internal regulation.

www.miteq.com



6 and 7 Bit Digital Attenuators

Skyworks is pleased to introduce two digital attenuators with superior attenuation accuracy for base station, cellular-head end, repeater, test equipment and femtocell manufacturers. The 6 bit device has an 0.5 dB least significant bit (LSB) and the 7 bit device has a 0.25 dB LSB. These attenuators provide precise control over multi-standard radio transmitters and receivers, allow the user to configure the device for serial or parallel control, and do not utilize blocking capacitors so that low frequency operation is possible.

www.skyworksinc.com

Ultra-Miniature SMT Synthesizer

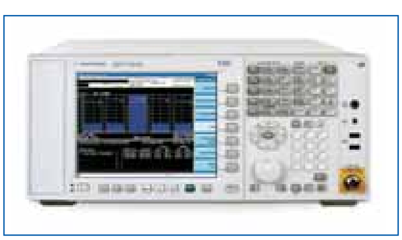
The SLX-1350 from EM Research, which operates over the frequency range 675-1350 MHz in 1 MHz steps, features low spurs (<–60 dBc), low harmonics (<–15 dBc), 0 dBm nominal output power, and low phase noise (<–85 dBc/Hz at 10 kHz). This low-cost unit is ideally suited for use in SWaP-C applications. The SLX Series is available in designs from 50 MHz to 6 GHz, in fixed-frequency or serially-programmable bandwidths to an octave.

www.emresearch.com

RF Detector with Fast Comparator

Linear Technology introduces the LTC5564, a precision RF power detector that operates from 600 MHz to 15 GHz, with exceptionally fast response time of 7 ns from a pulsed RF signal. The LTC5564 can measure RF input signals from –24 to +16 dBm. The LTC5564 can be powered from a single 3.3V or 5V supply. Normal operation draws a nominal current of 44 mA. The device is rated for operation from –40°C to 85°C. The LTC5564 is offered in a small 3 × 3 mm 16-pin QFN package. Pricing starts at $2.90 each in 1,000 piece quantities.

www.linear.com



Express Configurations

Agilent Technologies Inc. introduced express configurations for the popular CXA/EXA signal analyzers and MXG signal generators. Express configuration products provide fast, off-the-shelf delivery of the most popular test and measurement configurations. This service ensures that test equipment is shipped as fast as possible to customers’ research and development and manufacturing lines, ready for immediate use, saving time, effort and expense.

www.agilent.com

Ultra-Broadband LNA

MITEQ Inc. introduces a new addition to its family of wideband high power amplifiers. The AMF-6B-14001535-50-40P is a connectorized Ku-band high power linear amplifier/module, covering 14-15.35 GHz and delivering over 10 W at P_{1dB}. Housing is environmentally sealed, EMI shielded and hermetic sealing option is also available. This module has internal regulation, over voltage, temperature and reverse polarity protection. Single +12 to +15V supply is required and 28V is optional. The power amplifier is fully compliant over a base temperature range of –30 to +60°C.

www.miteq.com

3-500 MHz RF Transformer

This high-performance, low-cost transformer is ideal for use with push-pull amplifiers where balanced-to-unbalanced RF signal transformation is required. It is an ideal match for the inputs of Mini-Circuits dual MMIC amplifiers. When used in this configuration, the high phase and amplitude accuracy provides excellent IP2 and IP3 performance, making it ideal for use in 75Ω CATV return applications or any single-ended 75Ω to balanced 50Ω application.

www.minicircuits.com

Advertiser Index

Company ... Page

ACS ... 28
Aeroflex / Inmet ... 29
Agilent Technologies ... 15
Agilent Technologies ... 53
Anaren ... 21
AWR ... 27
Besser Associates ... 60
Coilcraft ... 11
CST ... 35
C. W. Swift & Associates ... Cover 2
Emerson Network Power ... 4
Empower RF ... 43
EM Research ... 45
J microTechnology ... 54
J microTechnology ... 56
J microTechnology ... 58
Linear Technology ... 13
Micro Lambda Wireless ... 17
Microwave Components ... 32
Mini-Circuits ... 2-3
Mini-Circuits ... 25
Mini-Circuits ... 31
Mini-Circuits ... 37
Mini-Circuits ... 38-39
Mini-Circuits ... 47
Mini-Circuits ... 51
Mini-Circuits ... 55
Mini-Circuits ... 59
MITEQ ... 1
MITEQ ... 57
MITEQ ... Cover 4
Molex ... Cover 3
Renaissance Electronics Corp/HXI. ... 9
RLC Electronics ... 23
SGMC Microwave ... 49
SignalCore ... 33
Skyworks Solutions ... 19
SV Microwave ... 41
Teledyne Cougar ... 7

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Subscriptions are free to qualified technical and management personnel involved in the design, manufacture and distribution of electronic equipment and systems at high frequencies.

A Magnetic Field Response Sensor for Measurement of Liquid Levels

NASA's Langley Research Center (www.larc.nasa.gov) recently reported on the development of a sensor that allows wireless measurement of the level of liquid in closed, electrically-conductive containers such as metal fuel tanks. Magnetic Field Response sensors utilize the absorptive properties of a tuned circuit near the resonant frequency. A simple swept oscillator excites the circuit via magnetic coupling to the inductor, and records the amplitude response versus frequency. The frequency where the response is minimum—due to maximum coupling to, and absorption/radiation by the tuned circuit—corresponds to the value of the tuning capacitance. Thus, a variable capacitance and a fixed inductor are the only components necessary for a passive sensor.

The variable capacitance can be obtained using several structures. Figure 1 shows two of them, an interdigital capacitor and a linear parallel strip capacitor. Immersion in a liquid changes the dielectric constant, which changes the capacitance.

The interdigital structure can have a higher capacitance value than the simpler linear capacitor, resulting in a lower resonant frequency with the same inductor, and generally allowing better measurement resolution. The capacitor can be fabricated on the same substrate as the inductor for a one-piece sensor, or it can be a separate piece, connected by wires to the inductor. This latter arrangement allows installation as shown in Figure 2, placing the capacitor inside the container, while the inductor remains outside where it can be readily accessed by the reader unit. Connection is made by wires that pass through a liquid-tight feedthrough.

The capacitor and inductor must have a spacer between them and the conductive surface of the container to reduce coupling effects. The spacing only needs to be sufficient to permit good coupling to the reader unit. Even if the component values are affected by the conductive container wall, secure mounting will result in stable values and repeatable measurements.

The work is one of several applications developed at NASA Langley using this sensor technology.

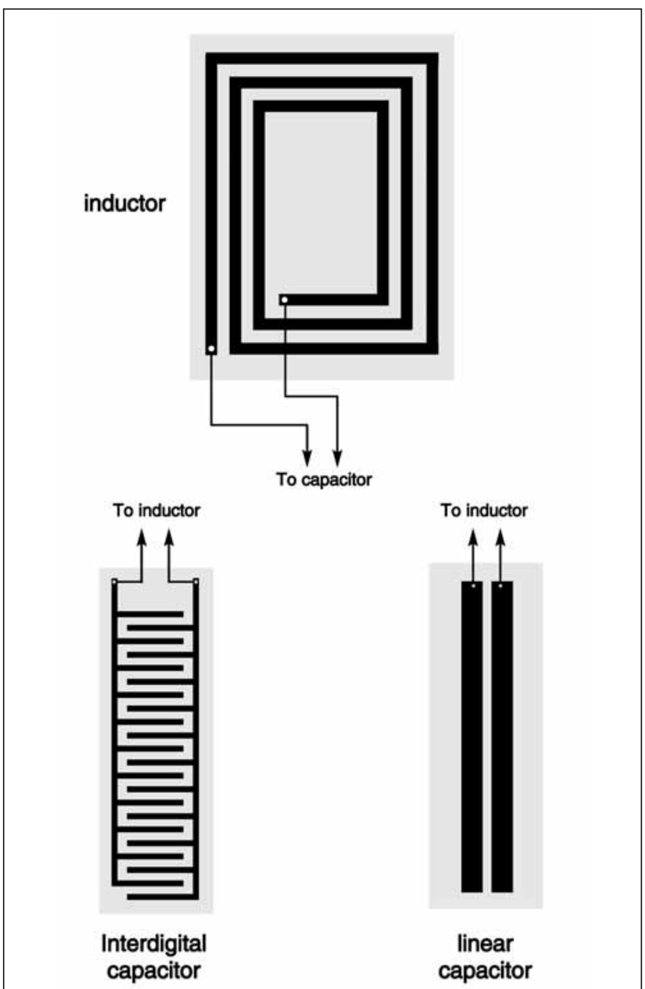


Figure 1 · Printed inductor and capacitor elements for a magnetic field response sensor.

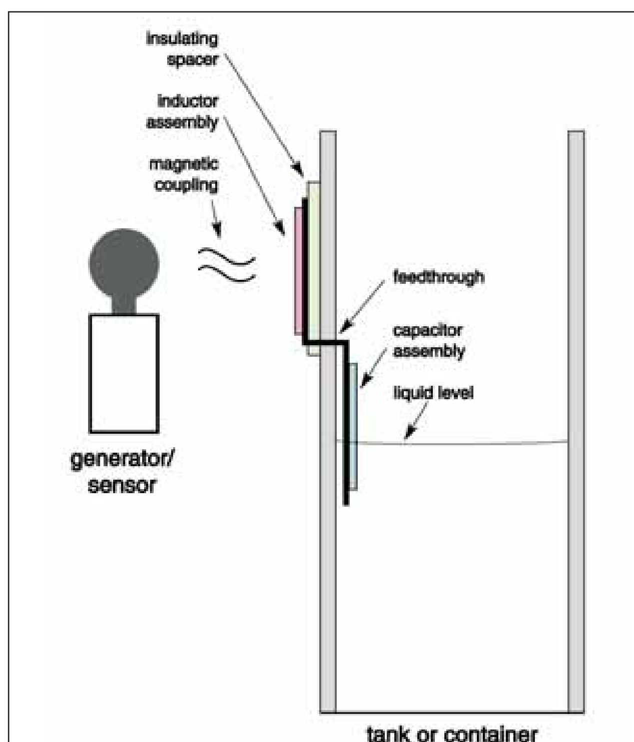
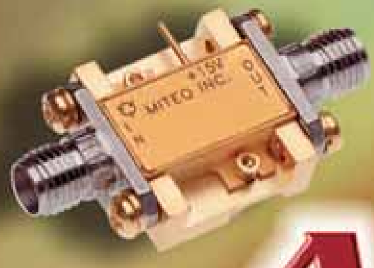


Figure 2 · Mounting method for measurements in a conductive-wall container.



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Model Number	Frequency Range (GHz)	Gain (Min./Max.) (dB)	Gain Flatness (±dB)	Noise Figure (dB, Max.)	VSWR Input (Max.)	VSWR Output (Max.)	Output Power @ 1 dB Comp. (dBm, Min.)	Nom. DC Power (+15 V, mA)
OCTAVE BAND AMPLIFIERS								
AFS3-00120025-09-10P-4	0.12-.25	38	0.50	0.9	2.0:1	2.0:1	+10	125
AFS3-00250050-08-10P-4	0.25-0.5	38	0.50	0.8	2.0:1	2.0:1	+10	125
AFS3-00500100-06-10P-6	0.5-1	38	0.75	0.6	2.0:1	1.5:1	+10	150
AFS3-01000200-05-10P-6	1-2	38	1.00	0.5	2.0:1	2.0:1	+10	150
AFS3-01200240-06-10P-6	1.2-2.4	34	1.00	0.6	2.0:1	2.0:1	+10	150
AFS3-02000400-06-10P-4	2-4	32	1.00	0.6	2.0:1	2.0:1	+10	125
AFS3-02600520-10-10P-4	2.6-5.2	28	1.00	1.0	2.0:1	2.0:1	+10	125
AFS3-04000800-07-10P-4	4-8	32	1.00	0.7	2.0:1	2.0:1	+10	125
AFS3-08001200-09-10P-4	8-12	28	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-08001600-15-8P-4	8-16	28	1.00	1.5	2.0:1	2.0:1	+8	100
AFS4-12001800-18-10P-4	12-18	28	1.50	1.8	2.0:1	2.0:1	+10	125
AFS4-12002400-30-10P-4	12-24	24	2.00	3.0	2.0:1	2.0:1	+10	85
AFS3-18002650-30-8P-4	18-26.5	18	1.75	3.0	2.2:1	2.2:1	+8	125
MULTIOCTAVE BAND AMPLIFIERS								
AFS3-00300140-09-10P-4	0.3-1.4	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS2-00400350-12-10P-4	0.4-3.5	22	1.50	1.2	2.0:1	2.0:1	+10	80
AFS3-00500200-08-15P-4	0.5-2	38	1.00	0.8	2.0:1	2.0:1	+15	125
AFS3-01000400-10-10P-4	1-4	30	1.50	1.0	2.0:1	2.0:1	+10	125
AFS3-02000800-09-10P-4	2-8	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS4-02001800-24-10P-4	2-18	35	2.00	2.4	2.5:1	2.5:1	+10	175
AFS4-06001800-22-10P-4	6-18	25	2.00	2.2	2.0:1	2.0:1	+10	125
AFS4-08001800-22-10P-4	8-18	28	2.00	2.2	2.0:1	2.0:1	+10	125
ULTRA WIDEBAND AMPLIFIERS								
AFS3-00100100-09-10P-4	0.1-1	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-00100200-10-15P-4	0.1-2	38	1.00	1.0	2.0:1	2.0:1	+15	150
AFS1-00040200-12-10P-4	0.04-2	15	1.50	1.2	2.0:1	2.0:1	+10	50
AFS3-00100300-12-10P-4	0.1-3	32	1.00	1.2	2.0:1	2.0:1	+10	125
AFS3-00100400-13-10P-4	0.1-4	30	1.00	1.3	2.0:1	2.0:1	+10	125
AFS3-00100600-13-10P-4	0.1-6	30	1.25	1.3	2.0:1	2.0:1	+10	125
AFS3-00100800-14-10P-4	0.1-8	28	1.50	1.4	2.0:1	2.0:1	+10	125
AFS4-00101200-22-10P-4	0.1-12	34	1.50	2.2	2.0:1	2.0:1	+10	150
AFS4-00101400-23-10P-4	0.1-14	24	2.00	2.3	2.5:1	2.5:1	+10	200
AFS4-00101800-25-S-4	0.1-18	25	2.00	2.5	2.5:1	2.5:1	+10	175
AFS4-00102000-30-10P-4	0.1-20	20	2.50	3.0	2.5:1	2.5:1	+10	125
AFS4-00102650-42-8P-4	0.1-26.5	24	2.50	4.2	2.5:1	2.5:1	+8	135

Note: Noise figure increases below 500 MHz.

3 YEAR WARRANTY

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RoHS compliant Available

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