February 2001

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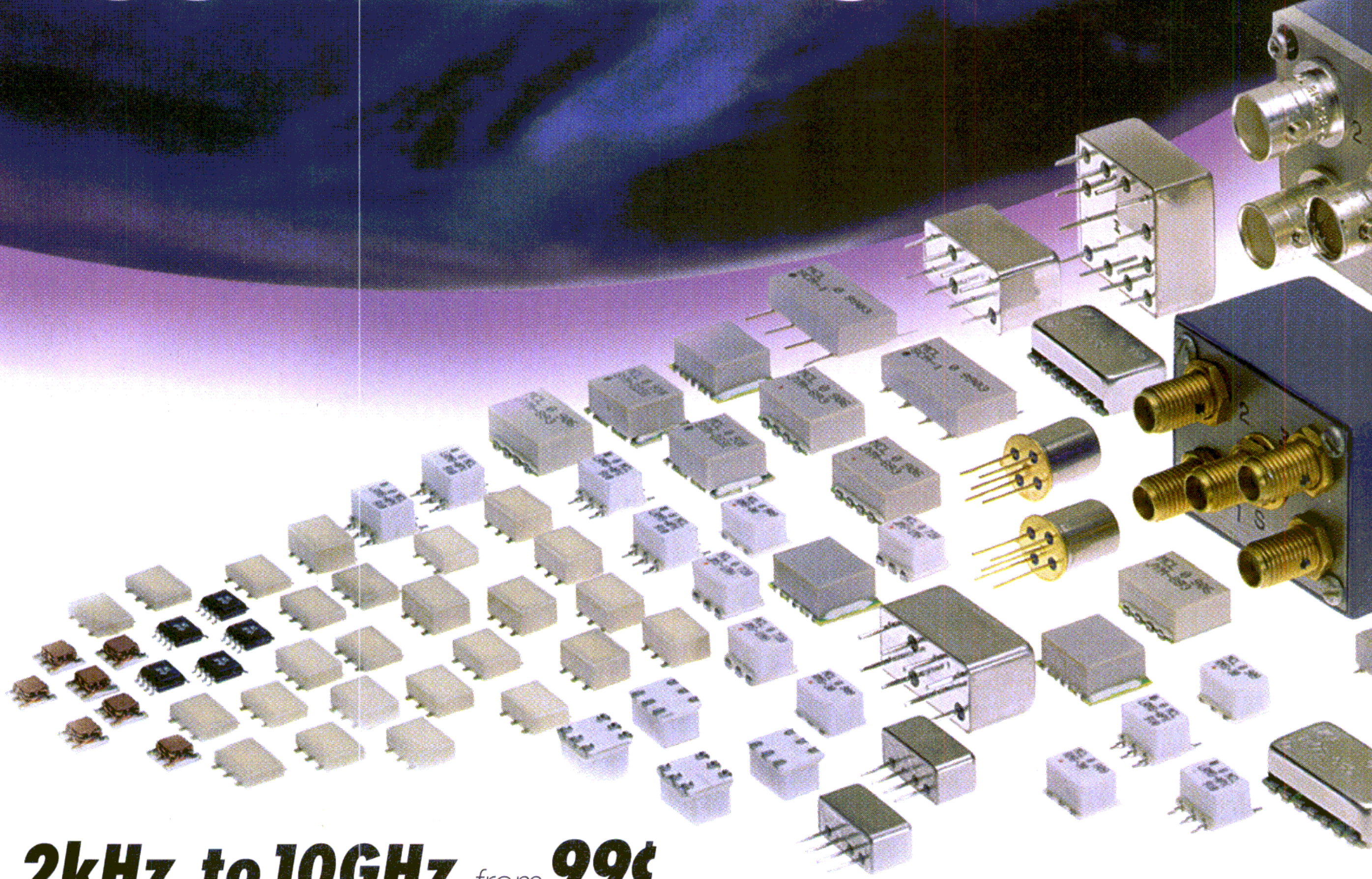

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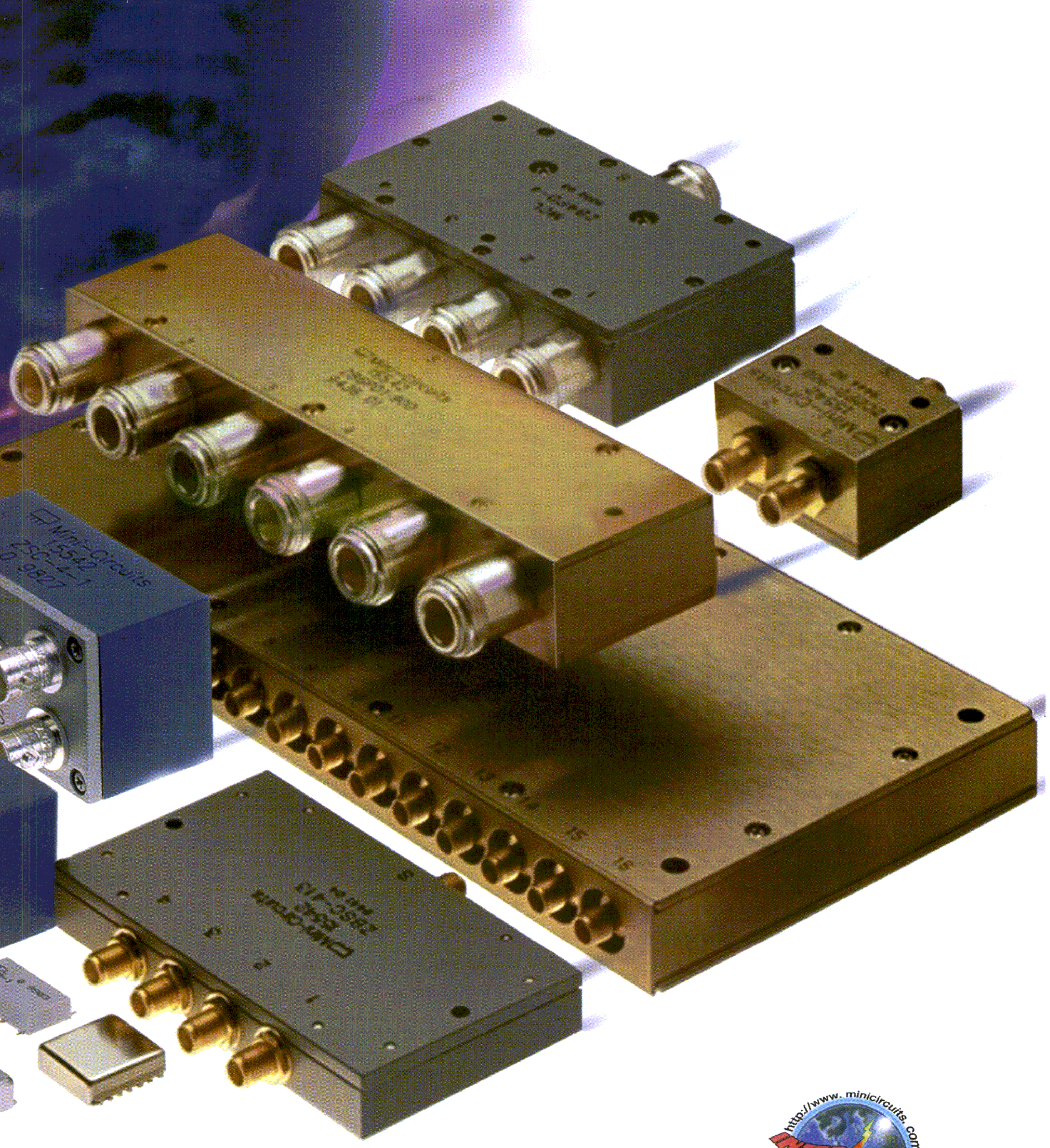
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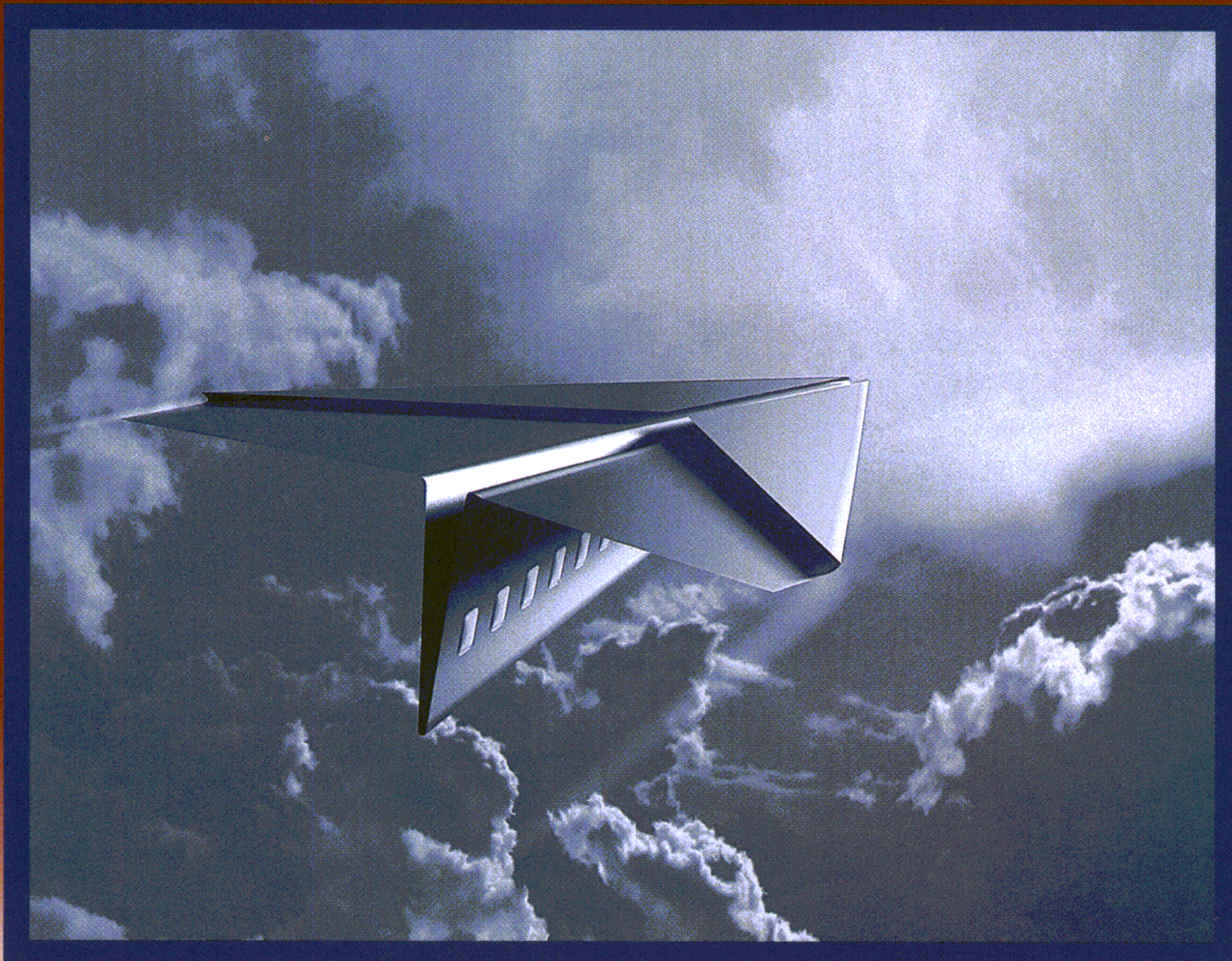

Applied
MICROWAVE & WIRELESS

February 2001
Vol. 13 No. 2

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PRODUCTS & TECHNOLOGIES

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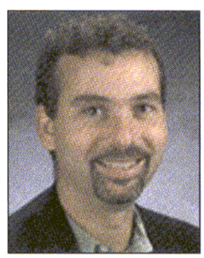

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Editorial

Stimulating Creativity Through Personal Communication

By Gary A. Breed
Publisher

Don't let the title fool you! I'm not going to discuss the personal communication technology of cell phones and wireless PDAs. Instead, I want to remind you of the importance of communication on a personal level. This month's Wireless Symposum in San Jose is a reminder that 'live' events are still a major part of an engineer's professional life. Conferences and trade shows remain a valuable way for ideas to be shared among peers and between suppliers and customers of components, instruments and design tools.

The most valuable activities at most conferences are the conversations held outside the regular program. Discussions with speakers after their presentations, or with an application engineer from a vendor, can shed new light on a problem and lead to successful completion of a design project. Engineers may not be the most outwardly sociable people, but you don't need to be gregarious to have a meaningful conversation on your favorite subject.

Of course, personal communication technology plays a role in continuing these professional discussions. E-mail is especially powerful for fast communications that can carry a lot of information in words, pictures and equations. Internet discussion groups are also valuable places for presenting an engineering problem and getting suggested solutions from experts. My experience is that e-mail and the Internet take nothing away from the old-fashioned technologies of the telephone and regular mail. They simply add the ability to communicate more often and in greater depth.

Fortunately, only a few companies discourage or restrict open communication by their engineers. Certainly, all employees must be sensitive to the need to avoid disclosing any trade secrets, but the benefits of learning new techniques and adopting new ideas far outweigh the risk of unintended disclosure of proprietary information.

Enjoy yourselves while talking with your peers at all the conferences the boss will let you attend. Make sure you know where to find the Internet groups sharing your favorite RF and microwave topics. Take advantage of truly *personal* communication. It will help you get your job done more quickly and with better results. ■


The one constant
Change

Letters

Encouraging kids to become engineers

Editor:

In concurrence with your editorial remarks made in the October 2000 issue, we in the National Laboratory environment have also found recruitment of RF staff difficult. You suggested grooming school children with summer camps and activities, and I would like to mention the role of Amateur Radio in attracting newcomers.

Very few kids are not impressed by a table full of radios, old and new. Nearly every community has active radio clubs, which eagerly participate in civic and scholastic activities. Although the Cold War appears to be over, the technology is alive and well, and is desperately needed to further personal communications, as well as build exotic machines like particle accelerators for basic research.

Very few of my mentors, professors and colleagues were not involved with Amateur Radio in some way or another, and much of my intuition is grounded in my early radio experiences. My most memorable academic moments involved merging mathematical concepts with knowledge gained from operating radios, an "osmosis."

Exposure and mentorship are the lifeblood of the next generation, to which Amateur Radio is passionately designed. I also recommend less television, and support of your local museums!

Keep up the quality publishing.

John Musson
Thomas Jefferson National Accelerator Facility

Corrected article now available

We made a few unfortunate omissions in the preparation of the two-part article, "Unfiltered FQPSK: Another Interpretation and Further Enhancements," published in the February and March 2000 issues of Applied Microwave & Wireless. *Two figures and several equations were left out of the original publication.*

The corrected versions of these articles have been placed on our website. They can be downloaded in PDF format from the appropriate issue listings on the "Archives" page of www.amwireless.com.

We sincerely apologize to authors Simon and Yan for the errors in publication and hope that interested readers will take the time to review the corrected article.

We welcome your comments. Send letters to the attention of "Editor" at the mail or e-mail addresses given on page 8.

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Calendar

CONFERENCES

FEBRUARY

February 12-16, 2001
Wireless Symposium/Portable By Design
San Jose, CA
Information: Penton Media Inc.
Internet: http://www.wirelessportable.com

February 20-22, 2001
14th International Zurich Symposium on Electromagnetic Compatibility
Zurich, Switzerland
Information: Dr. Gabriel Meyer
Tel: +411 632 27 90; Fax: +411 632 12 09
E-mail: gmeyer@nari.ee.ethz.ch
Internet: http://www.nari.ee.ethz.ch/emc

MARCH

March 19-23, 2001
17th Annual Review of Progress in Applied Computational Electromagnetics
Monterey, CA
Information: ACES
E-mail: aces@egr.msu.edu
Internet: http://aces.ee.olemiss.edu

March 20-22, 2001
CTIA Wireless 2001
Las Vega, NV
Information: CTIA
Tel: 301-694-5243; Fax: 301-694-5124
Internet: http://www.ctiashow.com

APRIL

April 17-21, 2001
ICAP — 11th International Conference on Antennas and Propagation
Manchester, UK
Information: ICAP
E-mail: icap@iee.org.uk
Internet: http://www.iee.org.uk/Conf/ICAP

April 21-26, 2001
NAB2001 — National Association of Broadcasters Conference and Exhibition
Las Vegas, NV
Information: NAB
Internet: http://www.nab.org

April 24-25, 2001
Second European Workshop on Conformal Antennas
The Hague, The Netherlands
Information: Wim L. Smith
E-mail: iec@fel.tno.nl
Internet: http://www.tno.nl/instit/fel/felnews/conformal_antennas

MAY

May 8-10, 2001
MIOP — Mikrowellen und Optronik (Microwaves and Optronics), The German Wireless Week
Stuttgart, Germany
Information: Dr. Martin Schallner
E-mail: martin.schallner@marconi.com

May 13-16, 2001
Fourth Conference on Electromagnetic Wave Interaction with Water and Moist Substances
Information: Dr. Klaus Kupfer
Tel: +49 3643 564361; Fax: +49 3643 564 202 or 204
E-mail: klaus.kupfer@mfpa.de
Internet: http://www.mfpa.de

May 13-17, 2001
International Symposium on Electromagnetic Theory
Victoria, British Columbia, Canada
Information: Secretariat
Tel: 613-993-9431; Fax: 613-993-7250
E-mail: URSI-B2001@nrc.ca
Internet: http://www.nrc.ca/confserv/URSI-B2001

May 20-25, 2001
2001 IEEE MTT-S International Microwave Symposium
Phoenix, AZ
Information: LRW Associates
Tel: 704-841-1915; Fax: 704-845-3078
E-mail: lrwassoc@carolina.rr.com
Internet: http://www.ims2001.org

JULY

July 8-13, 2001
2001 IEEE AP-S International Symposium and USNC/URSI National Radio Science Meeting
Boston, MA
Information: Robert McGahan
Tel: 781-377-2526; Fax: 781-377-3469
E-mail: mcgahan@ieee.org
Internet: http://www.ieeeaps.org/2001APSURSI

July 24-27, 2001
ISSSE 2001 — 2001 International Symposium on Signals, Systems, and Electronics
Tokyo, Japan
Information: ISSSE
E-mail: issse01@ee.kagu.sut.ac.jp
Internet: http://issse01.ee.kagu.sut.ac.jp

Calendar

SHORT COURSES

Besser Associates

Advanced Wireless and Microwave Techniques
Mountain View, CA . . . . . .February 12-16, 2001
Phoenix, AZ . . . . .March 19-23, 2001
RF Test Equipment Operation (laboratory course)
Mountain View, CA . . . . . .February 20, 2001
RF Testing for the Wireless Age (laboratory course)
Mountain View, CA . . . . . .February 21-23, 2001
Practical Design of Integrated and Discrete Wireless Circuits
Mountain View, CA . . . . . .February 26-28, 2001
RF CMOS Design
Mountain View, CA . . . . . .March 1-2, 2001
RF Transceiver Design
Mountain View, CA . . . . . .March 5-8, 2001
Nonlinear Device Model Extraction Techniques for Circuit Designers
Mountain View, CA . . . . . .March 8-9, 2001
Wireless Measurements: Theory and Practice
Mountain View, CA . . . . . .March 12-16, 2001
RF and Wireless Made Simple
Phoenix, AZ . . . . .March 19-20, 2001
Boston, MA . . . . .April 24-25, 2001
Electromagnetic Shielding for Wired and Wireless Technology
Mountain View, CA . . . . . .March 19-22, 2001
Applied RF Techniques I
Phoenix, AZ . . . . .March 19-23, 2001
Boston, MA . . . . .April 23-27, 2001
RF and Wireless Made Simple II
Phoenix, AZ . . . . .March 21-22, 2001
Boston, MA . . . . .April 26-27, 2001
Bluetooth: An Introduction
Phoenix, AZ . . . . .March 26-27, 2001
Mobile Commmunications and Wireless Data Networks
Mountain View, CA . . . . . .March 26-29, 2001
CDMA: The Physical Interface (IMT2000 3G WCDMA)
Phoenix, AZ . . . . .March 26-29, 2001
Frequency Synthesis Technology: Wireless Applications
Phoenix, AZ . . . . .March 28-30, 2001
Antennas and Array Design for Wireless Communications
Mountain View, CA . . . . . .April 9, 2001
Filters for Wireless Applications
Mountain View, CA . . . . . .April 16-17, 2001
DSP Made Simple for Engineers
Mountain View, CA . . . . . .April 18-20, 2001
Information: Annie Wong, Tel: 650-949-3300; Fax: 650-949-4400; E-mail: info@bessercourse.com; Internet: www.bessercourse.com.

RTT Programmes Limited

RF Amplifier Design
London, England . . . . . . .February 12-14, 2001
London, England . . . . . . .March 12-14, 2001
RF Oscillator Design
London, England . . . . . . .February 19-21, 2001
RF/IF Processing Design
London, England . . . . . . .March 5-7, 2001
3G Technology
London, England . . . . . . .April 2-4, 2001
Information: Lorraine Gannon, Tel: +44 181 844 1811; Fax: +44 181 751 2616; E-mail: seminars@rttsys.com; Internet: www.rttsys.com.

University of Wisconsin at Madison

Preparing for 3G Wireless Technoloyg: New Wideband CDMA Standards and Systems
Madison, WI . . . . .February 19-20, 2001
Preparing for 3G Wireless Technoloyg: New Wideband TDMA Standards and Systems
Madison, WI . . . . .February 21, 2001
Using and Intalling Fiber Optic Systems for Communications
Madison, WI . . . . .February 26-28, 2001
Fundamentals of Wireless Data Communications
Madison, WI . . . . .March 5-7, 2001
Planning and Implementing Point-to-Point Microwave Radio Systems
Madison, WI . . . . .April 9-11, 2001
Information: Katie Peterson, Tel: 1-800-462-0876; Fax: 608-263-3160; E-mail: custserv@epd.engr.wisc.edu; Internet: http://epd.engr.wis.edu.

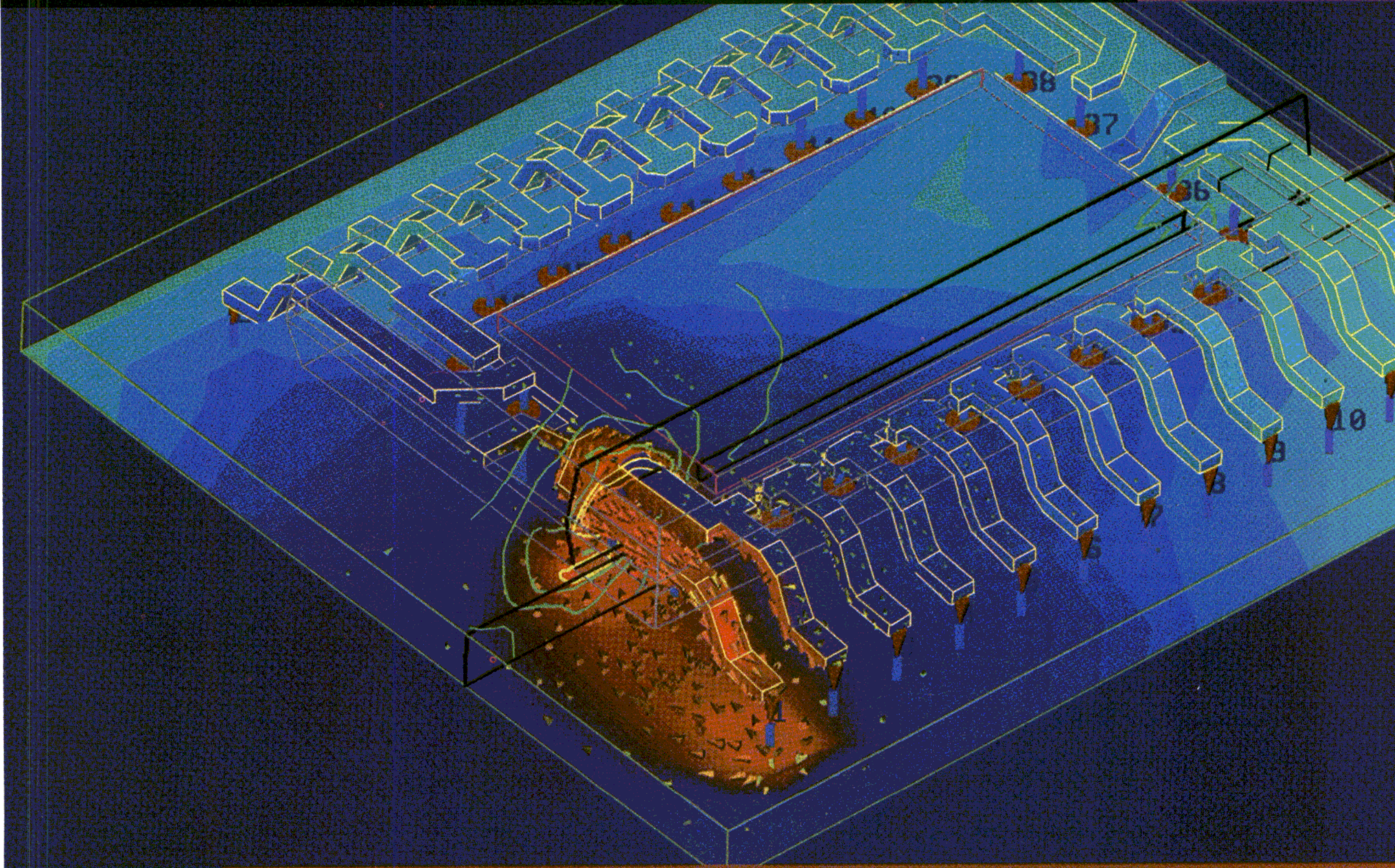
University of California at Los Angeles Extension

CDMA Mobiele Radio Design
Los Angeles, CA . . . . .March 12-15, 2001
Information: UCLA Extension, Short Course Program Office, Tel: 310-825-3344; Fax: 310-206-2815.

TTi Technology Training Initiative (Tustin Technical Institute, Inc.)

Thermal Analysis and Heat Transfer
Huntsville, AL . . . . .February 12-14, 2001
Santa Barbara, CA . . . . . .April 18-20, 2001
Climatic Environmental Testing Procedures
Huntsville, AL . . . . .February 15-16, 2001
Santa Barbara, CA . . . . . .March 22-23, 2001
Fundamentals of Vibration for Design
Santa Barbara, CA . . . . . .March 12-14, 2001
Fixture Design for Vibration and Shock Testing
Santa Barbara, CA . . . . . .March 14-16, 2001
Fundamentals of Vibration for Test
Santa Barbara, CA . . . . . .March 19-21, 2001
Instrumentation for Test and Measurement
Santa Barbara, CA . . . . . .March 26-28, 2001
Metrology Concepts
Santa Barbara, CA . . . . . .March 28-30, 2001
Calibration Processes
Santa Barbara, CA . . . . . .April 2-4, 2001

Burn out
or burn with passion?



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Physical Measurement Techniques
Santa Barbara, CA . . . . . .April 9-11, 2001
Measurement Uncertainty
Santa Barbara, CA . . . . . .April 11-13, 2001
Information: Brian P. Slatery, Tel: 805-682-7171; Fax: 805-687-6949; E-mail: brian@ttiedu.com; Internet: www. ttiedu.com.

Northeast Consortium for Engineering Education

Antennas: Principles, Design and Measurements
San Diego, CA . . . . .March 12-15, 2001
Information: Kelly Brown, Tel: 407-892-6146; Fax: 407-892-0406; E-mail: stcloudof1@aol.com; Internet: www.usit.com/antenna.

University of Missouri-Rolla

Grounding and Shielding Electronic Systems
Dallas, TX . . . . .March 20-21, 2001
Columbus, OH . . . . .April 17-18, 2001
Circuit Board Layout to Reduce Noise Emission and Susceptibility
Dallas, TX . . . . .March 22, 2001
Columbus, OH . . . . .April 19, 2001
Information: Sue Turner, Tel: 573-341-6061; Fax: 573-341-4992; E-mail: suet@umr.edu; Internet: www.umr.edu/~conted.

R.A. Wood Associates

RF Power Amplifiers, Classes A through S: How the Circuits Operate, How to Design Them, and When to Use Each
Baltimore, MD . . . . .April 19-20, 2001
Introductory RF & Microwaves
Baltimore, MD . . . . .April 19-25, 2001
RF and Microwave Receiver Design
Baltimore, MD . . . . .April 23-25, 2001
Information: R.A Wood Associates, Tel: 315-735-4217; Fax: 315-735-4328; E-mail: RAWood@rawood.com; Internet: www.rawood.com.

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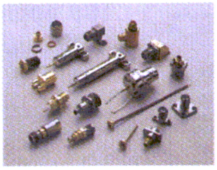

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CALLS FOR PAPERS

IEEE Radio and Wireless Conference (RAWCON2001)

August 19-22, 2001 — Waltham, MA

Topics: Third generation (3G) and fourth generation (4G) cellular; wireless LAN; broadband fixed wireless; Bluetooth/HomeRF/Personal Area Networks; ultra-wideband (UWB) communications; modeling and simulation; active and passive device technology; propagation and channel modeling; signal processing and system performance; and system architecture.

Authors should submit a two-page summary, including figures, electronically through the conference Web site, given below. Submit to:

Dr. Peter Staecker
Tel: 781-861-7643; Fax: 781-863-5751
E-mail: p.staecker@ieee.org
Internet: http://rawcon.org

Deadline: March 9, 2001

16th International Conference on Applied Electromagnetics and Communications (ICECOM'01)

September 24-26, 2001 — Dubrovnik, Croatia

Topics: Antenna theory and techniques, printed and conformal antennas, mobile antennas and vehicle modeling, active and passive phases arrays, smart antennas, multiple beam scanning, electromagnetic fields and guided waves, computational and numerical techniques, indoor and outdoor propagation modeling, remote sensing and polarimetry, microwave and RF devices and circuits, light-wave technology and M/W-optical interactions, mobile and personal communication systems, satellite and ground-based services, high-speed networks (LANs, WANs, optical), biomedical and industrial applications of microwaves, electromagnetic compatibility and EMI, electroacoustics and hydroacoustics and education in electromagnetics and communications.

Authors should submit an original and three copies of the completed paper, limited to four standard pages including text, references and figures, in camera-ready form. Text should be formatted in a 10-point font, single spaced, with 1-inch margins. A cover letter should include the topic area and complete contact information for the corresponding author. Additional information is available at the conference web site (URL below).

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Tel: + 385 1 6129 869
Fax: + 386 1 6129 870
E-mail: korema@fer.hr
Internet: http://www.rasip.fer.hr/korema

Deadline: April 5, 2001

The Applied Computational Electromagnetics Society

Special issue of the ACES Journal on "Approaches to Better Accuracy/Resolution in Computational Electromagnetics"

Topics: Alternative approaches to the method of moments (Nystrom, Boundary Residual Method); new, improved basis functions; use of non-uniform interval selection for collocation and sub-domain basis/testing functions; convergence acceleration methods; comparison of computer cost/time for given levels of convergence and for different basis functions; methods for numerical evaluation of frequently encountered integrals; exponentially converging Green's function summation formulae; and toward "dialable" accuracy in computational electromagnetics.

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Tel: 404-894-4697; Fax: 404-894-5935
E-mail: peterson@ee.gatech.edu
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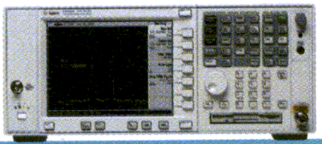
Deadline: May 1, 2001



RIDE THE

BRIEFS

• Vectron International, a subsidiary of Dover Corporation, has added two fully automated assembly lines to its voltage controlled crystal oscillator (VCXO) manufacturing facility in Hudson, NH. The new lines will increase production capacity by more than 60 percent.

• Andrew Insitute, the training arm of Andrew Corporation, has published its RF communications curriculum for 2001. Included are a new VSWR fundamentals course and expanded classes in terrestrial microwave systems installation and connector attachment. Information is available through the company's Web site, www.andrew.com.

• Amkor Technology has announced plans to open its first semicondutor assembly and test manufacturing plant in Shanghai, China. The 115,000-square-foot facility will initially produce processors and controllers.

• EMCORE Corporation has expanded its manufacturing facility in Albuquerque, NM, adding 36,000 square feet of production and cleanroom space.

• GEM Services has opened a 30,000-square-foot power semiconductor assembly and test facility in Shanghai, China. GEM plans to reach full capacity of 1.1 billion units per year by the end of 2001.

• EPCOS has opened a new 130,000-square-meter ferrite core production plant in Sumperk, Czech Republic.

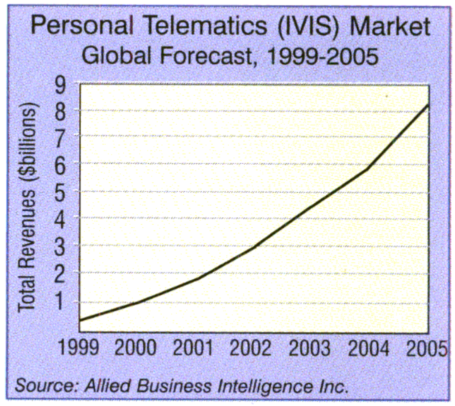
• The International Microelectronics and Packaging Society (IMAPS) has moved into a new world headquarters facility in Washington, D.C.

Companies, organizations and institutions may submit information for our News section to: Shannon O'Connor, *Applied Microwave & Wireless*, 630 Pinnacle Court, Norcross, GA, 30071; Fax: 770-448-2839; E-mail: amw@amwireless.com.

Telematics to lead ITS growth

The telematics industry will be one of the fastest-growing and most lucrative segements of the intelligent transportation systems (ITS) industry over the next five years, according to a new report from Allied Business Intelligence.

The report predicts that the global in-vehicle information system (IVIS) market will grow from about $1 billion in 1999 to more than $8 billion by 2005.



"Many firms were taking a wait-and-see attitude with IVIS," said ABI analyst Frank Viquez. "[But] 2001 will be seen as the year that IVIS takes off."

With approximately 1 million subscribers at the end of 2000, a viable market for IVIS has been established, the report says. Several large companies also plan to enter the market in 2001, further driving the industry.

Allied Business Intelligence is an Oyster Bay, NY-based technology research think tank.

Wireless technology forum set for March

The Wireless and Microwave (WAMI) program at the University of South Florida will host "Wireless and Microwave Technology 2001: An Industry/Government/Education Forum" on Saturday, March 24, 2001, at the Embassy Suites Busch Gardens in Tampa, FL.

This interactive forum is designed for the exchange of information on the status and direction of wireless and microwave technology, and the educational and research challenges now being addressed. Activities will include technical presentations and panel sessions.

Further information is available at http://ee.eng.usf.edu/WAMI.

QUALCOMM introduces new CDMA architecture

QUALCOMM CDMA Technologies, a division of San Diego, CA-based QUALCOMM Inc., has introduced radioOne^{TM}, a new technology for Code Division Multiple Access (CDMA) transceivers that uses Zero Intermediate Frequency (ZIF) architecture for direct conversion of radio frequency signals.

ZIF radios operate by converting incoming RF signals directly to or from baseband analog signals, eliminating a step in the process as well as the need for large IF SAW filters and additional circuitry.

QUALCOMM provides CDMA-based products and services for digital wireless communications.

Nitronex introduces new production capabilities

Nitronex Corporation has announced production of the industry's first gallium nitride-based high electron mobility transistors (GaN HEMTs) on 4-inch silicon wafers.

The company is consistently obtaining room temperature two-dimensional electron gas motilities exceeding 1,600 cm2/Vs on these HEMTs. These results were achieved using Nitronex's proprietary metal organic chemical vapor deposition (MOCVD) growth equipment and deposition process, designed to reduce GaN crystal defects by more than 100,000 times versus conventional techniques.

Nitronex, based in Raleigh, NC, supplies GaN-based materials, electronic devices, and photonic devices.

News

Osicom changes name to Sorrento Networks

Osicom Technologies has changed its name to Sorrento Networks Corporation. The new name comes from the company's flagship optical networking subsidiary.

Sorrento Networks, based in San Diego, CA, supplies intelligent optical networking solutions for telecommunications and other markets.

DragonWave formed for broadband wireless

DragonWave Inc. has been created to develop broadband radios for the fixed wireless market. The new company is based in Kanata, Ontario, Canada.

The company plans to provide high-speed wireless access over broadband to compete with fiber-optics broadband offerings.

Ericsson Cables AB changes name

Ericsson Cables AB has changed its name to Ericsson Network Technologies AB, to reflect a new focus on expertise in telecommunications networks.

The company is a division of Ericsson, a Stockholm, Sweden-based supplier of systems, applications, mobile phones and other wireless products and services.

Wave ID created for RFID

A new company called Wave ID has been formed to put into commercial use proprietary technology developed at Pacific Northwest National Laboratory. Partial financing will be provided by Battelle.

Wave ID, based in Richland, WA, will create wireless communication systems based on radio frequency identification techniques developed at the laboratory. These systems will include RF tags, which are devices that range in side from a grain of rice to a credit card and can be designed to identify, locate or monitor items.

Magnet Applications formed

Following a management buyout, the companies Magnet Applications Inc., Magnet Applications Ltd. and Arelec SA have been incorporated into a new company, Magnet Applications Group Ltd.

Based in Berkhamsted, UK, the company manufactures magnets, magnetic components, materials and catches, as well as complete metalworking and plastic injection capabilities.

Rainbow announces reorganization

Rainbow Technologies has announced a company restructuring, creating four new business units: Spectria, Mykotronx, IVEA Technologies and Digital Rights Management Group. The reorganization is designed to streamline continuing company growth.

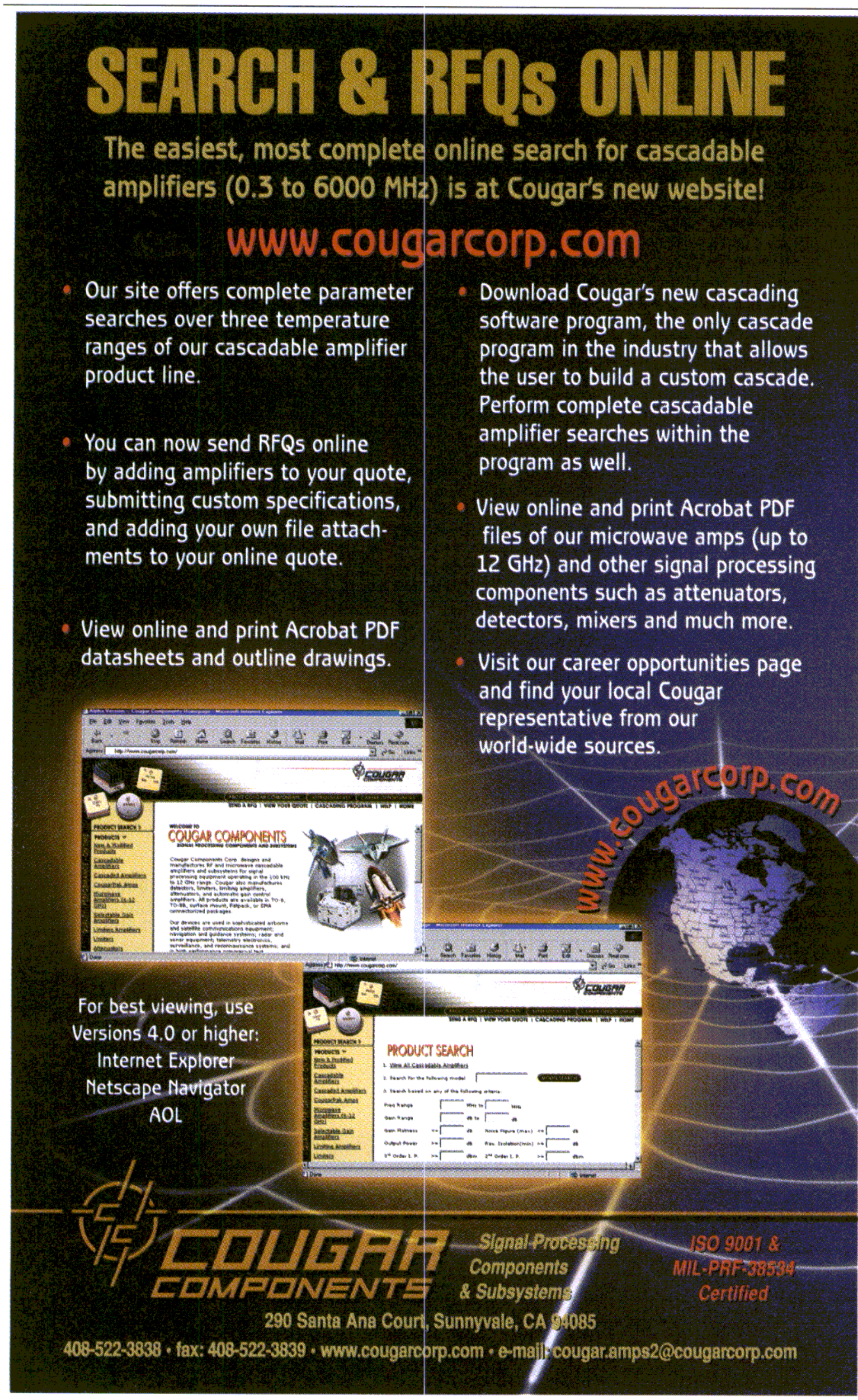
Rainbow, based in Irvine, CA, provides security systems for telecommunications.

BUSINESS AND FINANCE

Tegal receives orders for plasma etch systems

Tegal Corporation has announced two orders for the company's plasma etch systems. Austria Mikro Systeme International has placed a repeat order for Tegal's 6520 high-density plasma etch system, while TELSA SEZAM, the Czech Republic's only semiconductor manufacturer, has ordered three of Tegal's 901e-recertified plasma etch systems. The value of the orders was not disclosed.

Tegal, based in Petaluma, CA, manufactures plasma etch systems used in the production of integrated circuits and related devices.



Merrimac receives contract for GPS retrofit program

Merrimac Industries Inc. has been awarded a $1 million contract from Lockheed Martin Corporation for the Global Positioning System (GPS) Antenna Retrofit Program. The program is designed to enhance the output power of older satellites for better coverage and accuracy.

Merrimac, based in West Caldwell, NJ, manufactures RF microwave components, assemblies and modules for the wireless telecommunications industry.

Motorola awarded contracts for CDMA networks

Motorola Inc. has received two new contracts to provide third generation (3G) Code Division Multiple Access (CDMA) services and equipment for several markets in the United States.

• Under an agreement with Alltel, Motorola will supply and install cdma2000 1x standards-based hardware and software in the Phoeniz, AZ, area. Financial terms were not disclosed.

• A contract with Sprint PCS calls for Motorola to supply CDMA equipment via Interoperatbility Specifications (IOS) and cdma2000 1x high-speed packet data to wireless networks in several Sprint PCS markets. Deployment areas include Chicago, IL, Chattanooga, TN, Richmond, VA, and Cincinnati, OH.

Motorola, based in Schaumburg, IL, provides semiconductors, integrated communications solutions, embedded electronic systems and components.

Summit Partners invests in Hittite Microwave

Private equity investor Summit Partners has announced a $15 million investment in Hittite Microwave Corporation.

Based in Chelmsford, MA, Hittite manufactures integrated circuits for communications systems.

News

BUSINESS AND FINANCE

Amplifier Research purchases Thermo Voltek

Amplifier Research has announced the acquisition of Thermo Voltek Corporation, parent company of Kalmus Corporation. Financial terms were not disclosed.

Under the agreement, Kalmus will operate as AR Kalmus, with its facility remaining in Bothell, WA. Amplifier Research, based in Souderton, PA, manufactures RF power amplifers and other instruments for EMC and RF testing.

Vectron International acquires Cinox

Vectron International, a subsidiary of Dover Corporation, has announced the acquisition of Cinox Corporation of Cincinnati, OH, a supplier of precision quartz crystals. Financial terms were not disclosed.

Vectron, based in Norwalk, CT, manufactures frequency generation and control products.

Spectrian sells UltraRF line to Cree

Spectrian Corp. has sold its UltraRF semiconductor line to Cree Inc. Terms included stock considerations, a supply agreement and a joint development agreement.

Spectrian, based in Sunnyvale, CA, manufactures single-carrier and multi-carrier power amplifiers for wireless infrastructure. Cree, based in Durham, NC, produces semiconductor materials and devices based on silicon carbide, gallium nitride and related compounds.

APW to acquire Lucent operation

APW Ltd. has agreed to acquire the metal, plastics and tooling manufacturing assest of Lucent Technologies' Global Provisioning Center, located in Bydgoszcz, Poland. Financial terms were not disclosed.

APW, based in Waukesha, WI, provides integrated electronic enclosure systems.

IQE acquires Wafer Technology

IQE plc has announced the acquisition of UK-based Wafer Technology Ltd., a supplier of III-V compound semiconductor substrates and high purity polycrystalline materials. Financial terms were not disclosed.

IQE, headquartered in Bethlehem, PA, is an outsource supplier of epitaxial wafers for the compound semicondutor industry.

BUSINESS AND FINANCE

M/A-COM agrees to acquire Stellex

M/A-COM, a unit of Tyco Electronics, has announced an agreement to acquire San Jose, CA-based Stellex Microwave Systems Inc., a manufacturer of subsystems and components for the defense and telecommunications industries. Financial terms were not disclosed.

M/A-COM, headquartered in Lowell, MA, provides RF and microwave semiconductors, components and IP networks for the wireless communications and defense industries.

Cypress, International Microcircuits to merge

Cypress Semiconductor Corporation and International Microcircuits Inc. have announced they have signed a merger agreement. The transaction is valued at $125 million.

Cypress, based in San Jose, CA, provides integrated circuit solutions for telecommunications, industrial control and other markets. International Microcircuits, based in Milpatas, CA, supplies frequency and timing generators and EMI reduction circuits for a range of applications.

Agilent aquires ATN Microwave

Agilent Technologies has completed the acquisition of ATN Microwave Inc., a Billerica, MA-based provider of microwave measurement solutions for wireless and high-speed data communication applications. Financial terms were not disclosed.

Under the agreement, ATN will operate as part of Agilent's component test business, which is headquartered in Santa Rosa, CA.

Agilent, based in Palo Alto, CA, manufactures test, measurement and monitoring solutions, as well as semiconductor and optical components.

Microchip Technology, TelCom Semiconductor merge

Microchip Technology Inc. of Chandler, AZ, and TelCom Semiconductor Inc. of Mountain View, CA, have announced the completion of the companies' merger, in a stock transaction.

Microchip Technology manufactures RISC microcontrollers for 8- and 16-bit embedded control applications. TelCom Semiconductor supplies high-performance linear and mixed-signal integrated circuit solutions.

A Logarithmic Spiral Antenna for 0.4 to 3.8 GHz

This article reviews the design, simulation and measured performance of a broadband antenna

By Jesper Thaysen, Kaj B. Jakobsen and Jørgen Appel-Hansen
Technical University of Denmark

This article discusses a low-cost cavity backed coplanar-waveguide to coplanar-strip fed logarithmic uniplanar spiral antenna that covers a 9:1 bandwidth with a return loss better than 10 dB from 0.4 to 3.8 GHz. The IE3D computer program is used to predict the performance of the spiral antenna in terms of radiation pattern and input impedance. The obtained numerical results are in good agreement with the experimental data. The spiral antenna radiates from an active region that is a frequency-dependent part of the structure. The rotation of the radiation pattern as a function of frequency will also be discussed.

Frequency independent antennas

Frequency independent antennas are antennas whose radiation pattern, impedance and polarization remain virtually unchanged over a large bandwidth [1]. Their electrical dimensions, however, scale with frequency. Ideally, the electrical size of such antennas would remain constant over the entire electromagnetic spectrum. This ideal state requires that the logarithmic spiral antenna be infinite to fulfill the self-scaling and self-complementary conditions. For a certain range of parameters, the logarithmic spiral antenna can be truncated and still retain the properties of the infinite structures over a very wide band. The practical frequency independent structure is truncated, which limits the antenna's upper and lower frequency limits [2, 3].

A broadband antenna could find wide application in many systems. This article is a part of the ongoing research at the Technical University of Denmark in the area of stepped-frequency ground penetration radar (SF-GPR) to detect buried non-metallic Anti-Personnel mines (AP-mines) in a humanitarian mine detection system [4].

In coaxial structures, such as coax cables, the electrical wavelength is changed by a factor of

$$\frac{1}{\sqrt{\varepsilon_r}}$$

due to the dielectric constant of the isolator [5]. In antenna systems, dielectric loading can be used with the same advantageous properties, although the antenna is a radiating structure. The resonant frequency of a patch antenna could be reduced using a piece of dielectric material placed on top of the antenna [5]. It is also possible to reduce the size of the antenna at a given resonant frequency.

Both the permittivity and the thickness of the substrate influence the performance. The loss in the material will alter the performance as well, so material with low loss is preferable. Thick substrates with a low dielectric constant provide better efficiency and larger bandwidth at the expense of larger element size. The same results can be obtained using a thin structure that has a higher dielectric constant, which is a trade-off between efficiency and the physical size of a given structure.

Dielectric loading seems to offer good possibilities for reducing the physical size of an antenna. It is therefore a very important design technique to meet requirements for size and resonance frequency for antenna design. IE3D is used to determine the influence of dielectric loading on the spiral antenna for various permittivity values. A simple 1.5 turn spiral antenna is used to investigate the dielectric loading

effect and the radiation efficiency.

Composite materials offer good mechanical and electrical properties and are available with dielectric constants in the range of 2 to 20. The RT/Duroid is an example of a composite with well-documented electrical, mechanical and temperature stability, and is therefore a popular material in planar antenna design. The disadvantages of this type of material are the availability and the cost.

The substrate used for this article is the FR-4 substrate, which is commonly used for printed circuits. The FR-4 board is less expensive than commonly used microwave substrates, such as RT/Duroid. This feature, combined with the advantages of the uniplanar circuit, makes this configuration suitable as a low cost wideband antenna. Nevertheless, the loss is increasing as a function of frequency for the FR-4 substrate, and the dielectric performance of the FR-4 could change from one manufacture to another, and from batch to batch.

To aid the antenna design, the electromagnetic simulation program (IE3D) computer program developed by Zeland Software [6], was used to predict the performance of the spiral antenna in terms of the radiation pattern and the input impedance. The measured results of the constructed antenna are compared to the simulated results.

Selection of the antenna type

The larger the frequency range, the better the range resolution is when using frequency-domain or time-domain techniques. Thus, the use of a broadband antenna is essential for the signal and image processing in order to improve the detection of non-metallic AP-mines. Because of the signal processing, it is desirable to make the sidelobes as small as possible.

The operating frequency range of the antenna for the SF-GPR is from 0.4 to 3.8 GHz, which yields a bandwidth of 3.4 GHz corresponding to a radar range resolution of about 40 mm in the soil. The lowest operation frequency of 0.4 GHz is a compromise between the ground attenuation and the antenna size, whereas the feeding network sets the limit at the high frequencies.

Although a linearly polarized log-periodic antenna has a wide bandwidth and is an attractive candidate, circular polarization is preferred to linear polarization for several. If a linearly polarized antenna is used, the strength of the reflected wave from an object will depend on the azimuthal position of the antenna relative to the object. Further, if the orientation of the transmitting and receiving antennas is orthogonal, the mutual coupling between the two linearly polarized antennas will be reduced. As a result, the receiving antenna would barely detect the reflected wave from the object. Circular polarization, however, does not have such problems. For this reason, spiral antennas were selected [7].

Another advantage of circular polarization is that the reflected signal from the surface of the soil has the opposite sense of polarization as compared to the incident wave, because the ratio between the permittivity of soil and air is larger than 1. Thus, the antenna does not detect the reflected wave from the surface. However, the reflected wave from the AP-mine has the same sense of polarization as the incident wave because the permittivity ratio between mine and soil is less than 1. This means that the only detected signal is the one reflected from an object having a relative permittivity lower than soil, for example, the AP-mines.

Among the spiral antennas, the conical spiral antenna radiates unidirectionally [8]. This unidirectionality of radiation is required for radar applications. However, for a conical spiral antenna, different frequency waves are radiated from different active regions of the cone. This gives rise to a problem for the range measurement, because the distance to the object depends on the frequency. This problem is reduced with the logarithmic spiral antennas because their shape is not conical but planar. However, the planar antennas have a bidirectional radiation property [2, 3]. By placing an absorbing material in a cavity behind the spiral antenna, the antenna exhibits a unidirectional radiation pattern. The disadvantage is that only half of the input power is transformed into radiated power because of the presence of the absorber. A specially designed reflector may also be used.

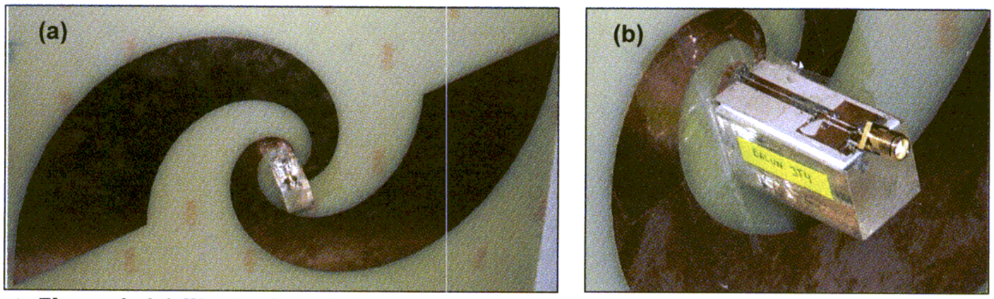
The principle of the spiral antenna

The logarithmic spiral antenna was designed using the equations $r_1 = r_0 e^{a\theta}$ and $r_2 = r_0 e^{a(\theta-\theta_0)}$, where r_1 and r_2 are the outer and inner radii of the spirals, respectively; r_0 and $r_0 e^{-a\theta_0}$ are the initial outer and inner radii; a is the growth rate; and θ is the angular position. To obtain the most frequency independent radiation pattern and the most constant input impedance, the dimensions are r_0 = 2.1 mm, a = 0.5 rad^{-1}, and θ_0 = 1.3 rad = 75 degrees [2, 9].

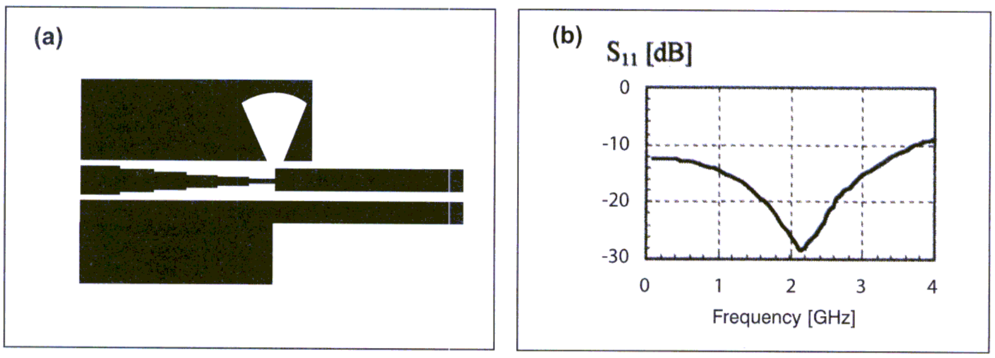
With these design parameters, the width of the arms are the same as the spacing between the arms, and the structure is self-complementary, which gives the most frequency independent parameters [2].

Good radiation patterns can usually be obtained with as few as 1 to 1-1/2 turns of the spiral [2]. Thus, the spiral antenna illustrated in Figure 1 consists of two equal arms, each with a 1.5 turn. The ends of the two spiral arms are truncated to produce the smallest physical antenna for a given lower resonant frequency. Alternatively, the end of the spiral arms can be tapered, which will result in a more constant input impedance of the antenna. The spiral antenna is fabricated on a 220 × 430 mm FR-4 substrate, with a thickness of 1.5 mm and a relative dielectric constant (ε_r) of 4.4.

When the antenna arms are very short compared to one wavelength, the polarization is linear. As the fre-



▲ **Figure 1. (a) Illustration of the spiral antenna placed on the FR-4 substrate. (b) Close up picture of the balun placed perpendicular to the plane of the antenna and soldered directly to the feed terminals on the spiral antenna.**



▲ **Figure 2. (a) Diagram of the coplanar waveguide (CPW) to coplanar strip (CPS) balun. (b) Simulated reflection coefficient S_{11}-parameter for one balun with the CPS terminated by an ideal 80 ohm resistor.**

quency increases, the axial ratio decreases and the polarization becomes elliptical. For frequencies at which the arm lengths of the spiral antenna are greater than one wavelength, or slightly less, the polarization is circular and the input impedance remains almost constant as frequency varies [2, 9]. The spiral antenna is designed to meet conditions where circular polarization is required; thus. the minimum frequency is where the electrical length of the arm is one wavelength. For the antenna discussed here, the arm length is 0.52 m, which results in a resonant frequency of 0.58 GHz. This resonant frequency is likely lowered due to the dielectric-loaded substrate.

Balun configuration

The logarithmic spiral shown in Figure 1 is a balanced antenna and requires a balun to transform the unbalanced coplanar waveguide (CPW) feed line to a balanced coplanar strip (CPS) feed line for the logarithmic spiral antenna. In this article, the transition from CPW to CPS is accomplished by using a wideband balun [12]. This balun is a modified version of that in [11] and has also been used in [3] and [13].

The balun shown in Figure 2 includes a four section Chebyshev impedance transformer with a reflection coefficient of $\Gamma_m = 0.05$, which was designed to transform the impedance from 50 to 80 ohms [12]. The input impedance of the spiral antenna is found to be approximately 80 ohms. The wideband transition from CPW to CPS is accomplished by using a slotted radial patch. The patch represents a very wideband open circuit, which forces the field to be primarily between the two conductors of the coplanar strip feed line. Two bond wires (not shown in the figure) near the discontinuity plane ensure that the potential on the two ground planes is equal [13]. The balun structure was fabricated on a small 16 × 43 mm RT/Duroid substrate with a thickness of 0.785 mm and a relative dielectric constant (ε_r) of 10.2.

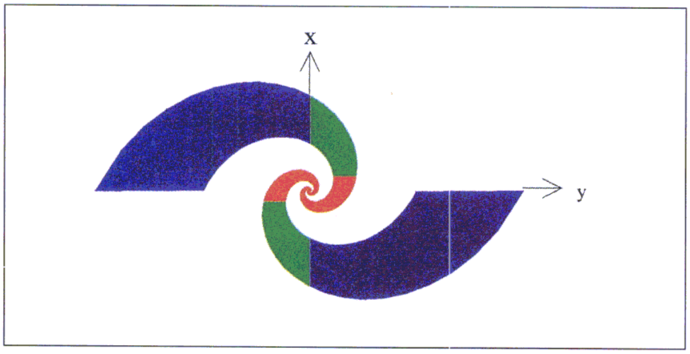
Simulations on the balun structure shown in Figure 2(a) with IE3D are performed by substituting the antenna with an ideal 80-watt resistor. The results show that in the frequency range from 0.1 to 3.85 GHz the simulated reflection coefficient, i.e., the S_{11}-parameter, is better than −10 dB, as shown in Figure 2(b).

A more detailed description of the performance of the balun can be found in [13].

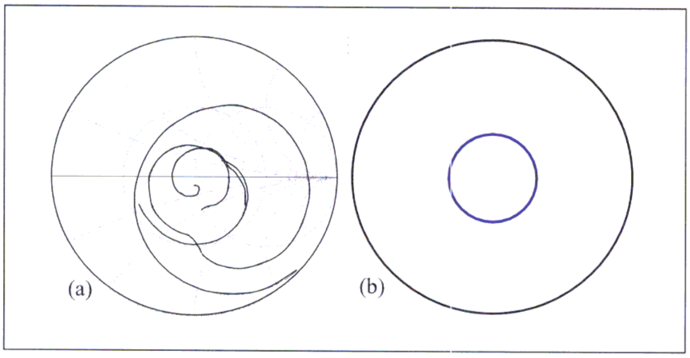
Numerical and experimental results

When the electrical size of the structure becomes too large, the benefits of using the IE3D are reduced. For example, simulations on the entire spiral antenna structure are not possible within a reasonable amount of time because of the limited computer capacity. To solve this problem, the spiral antenna is broken up into smaller parts, as illustrated by the different colors in Figure 3. The electrical size of the spiral increases as the frequency increases. The entire structure is first simulated in the frequency range from 0.1 to 1.4 GHz. Then, the fractions of the two arms, which correspond to the blue parts shown in Figure 1, are removed. Consequently, simulations are performed in a frequency range that is shifted upwards, that is, the influence of the blue parts are neglected. However, inaccuracies can occur when the currents on the arms that are removed are not negligible. A less effective way to solve this problem is to use a faster computer.

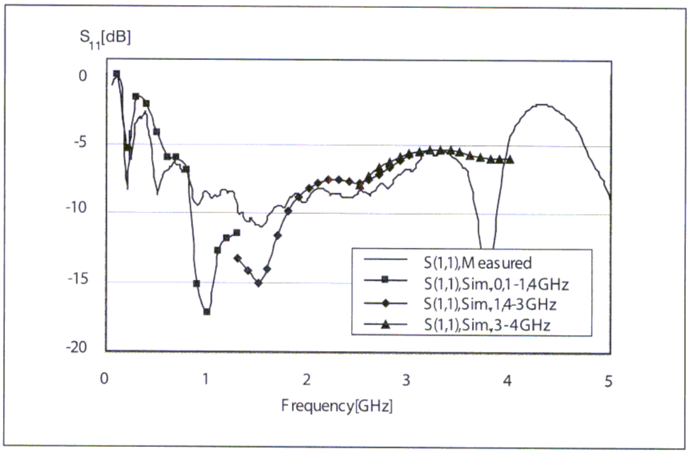
The balun structure was fabricated and connected to the fabricated prototype of the spiral antenna to verify



▲ **Figure 3. The spiral is broken up into smaller parts to reduce the simulation time.**



▲ **Figure 5. Simulated (a) and measured (b) impedances for the CPW-fed spiral antenna. The measured results show several loops in the Smith chart that are caused by a 0.4 m RG316U coaxial cable connected to the CPW on the balun. The circle shown in (b) indicates that every point inside the circle has an impedance match better than –10 dB.**



▲ **Figure 4. Measured and simulated S_{11}-parameter for the CPW-fed spiral antenna on FR-4 substrate material.**

the performance. The reflection coefficient of the spiral antenna, S_{11}, including the CPW- to CPS-feed network is determined using an HP8720D network analyzer. The experimental results are shown in Figure 4 for the CPW-fed spiral antenna.

The measured and simulated results are shown in Figure 4. The best agreement is achieved with large reflection coefficients, i.e., for $S_{11} > -10$ dB. The measured bandwidth for a reflection coefficient better than –6 dB is just slightly higher than the simulated. The measured bandwidth is from 0.45 to 3.1 GHz, and the simulated bandwidth is from 0.6 to 2.9 GHz. At a frequency of 3.8 GHz, the measured impedance matching is much better than the simulated. This may be due to parasitic coupling between the antenna and the balun.

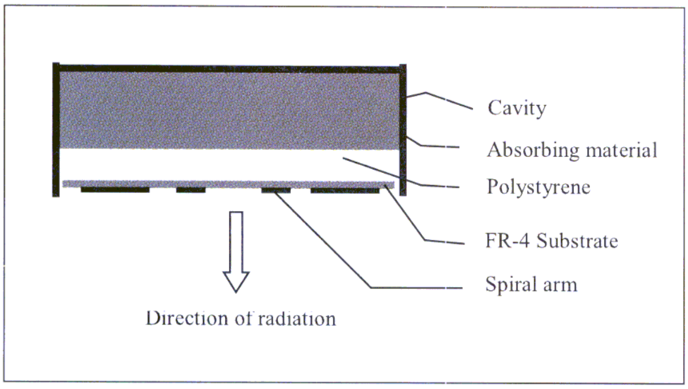
The simulated input impedance of the spiral antenna is 80 ohms. The measured and the simulated impedances for the spiral antenna including the CPW- to CPS-feed network is found to be matched to 50 ohms, as shown in the Smith charts shown in Figure 5. Because of the electrical length of the 0.4 m RG316U coaxial cable, the measured impedance is mapped into the Smith chart as a spiral located around the center of the Smith chart. This indicates that the spiral antenna, including the feed structure, is close to being matched to 50 ohms.

Simulations of two identical spiral antennas placed on two different substrates are made. The substrates used are FR-4 with a relative permittivity of 4.4 and ROHACELL with a relative permittivity of 1.06. The simulated results state that the input impedance for the spiral placed on the FR-4 is 80 ohms, whereas the impedance for the spiral antenna placed on ROHACELL is 130 ohms. This indicates that increasing the relative permittivity on the substrate could lower the input impedance. The result is in agreement with a recently published spiral antenna with an input impedance of 61 ohms. This antenna was placed on a substrate with a relative permittivity of 10.8 [11].

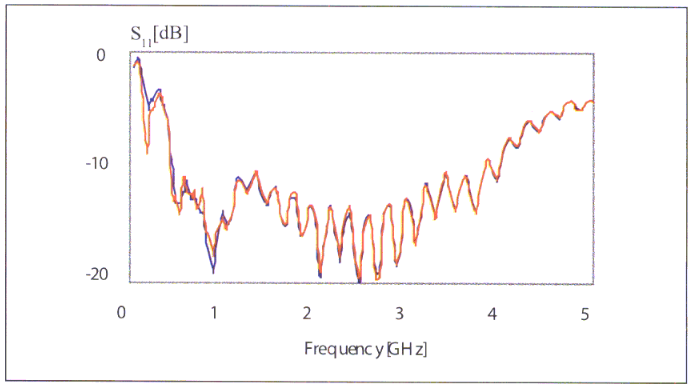
Cavity-backed spiral antenna

The planar logarithmic spiral antenna radiates in two broad lobes whose directions are perpendicular to the plane of the antenna. However, a unidirectional pattern is preferred in order to detect reflections from one direction only; thus, a cavity is needed. An absorbing material and polystyrene foam with thicknesses of 130 mm and 40 mm, respectively, fill up the back of the constructed cavity [14]. Figure 6 shows an illustration of the cavity. The overall size of the cavity is 610 × 290 × 180 mm.

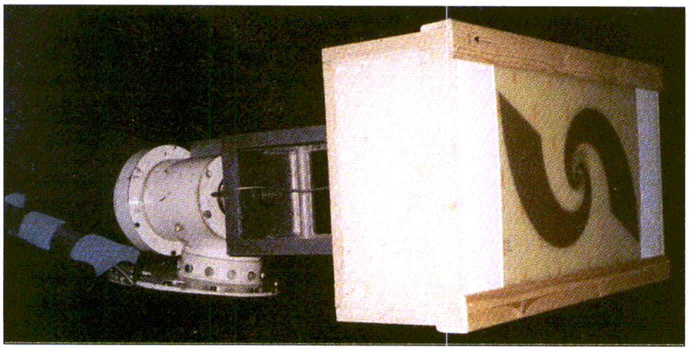
Figure 7 shows the experimental results for the CPW-fed spiral antenna backed with a cavity. The measured bandwidth for a reflection coefficient better than –10 dB is from 0.4 to 3.8 GHz. The main reason for the extended bandwidth as compared to the *S*-parameters shown in Figure 4 is the balun. In this case, the balun approach shown in Figure 2 is applied, whereas the balun used in the experiment in Figure 4 consists of a balun optimized for an approach other than the fabricated spiral antenna. The observed ripple on the measured reflection coef-



▲ **Figure 6. Cross section illustration of the cavity backed spiral antenna.**



▲ **Figure 7. Measured S_{11}-parameter for the spiral antenna with (red) and without (blue) a cavity.**



▲ **Figure 8. Picture of the cavity backed spiral antenna mounted in the anechoic chamber.**

ficient is likely to be caused by a mismatch with the 0.4 m RG316U flexible coaxial cable that is connected to the CPW on the balun.

Limitations on available computer capacity make it too time consuming to simulate the absorbing material using IE3D. In Figure 7, the measured S_{11}-parameter for the cavity backed spiral antenna can be compared to the measured S_{11}-parameter for the spiral antenna without a cavity. Little difference is observed in the S_{11}-parameter for the spiral antenna with or without a cavity. This result indicates that the radiation in the main direction remains virtually unchanged in the presence of the cavity.

Far-field radiation patterns

Anechoic chamber measurements in the Sperical Nearfield Antenna Test Facility (SNATF) at the Technical University of Denmark (DTU) are made in order to measure the far-field radiation pattern and the polarization. A picture of the spiral antenna situated in the anechoic chamber is shown in Figure 8.

Two θ-cuts, measured at 0.8 GHz and 1.5 GHz, are shown in Figure 9. Relevant far-field criterion is fulfilled, so the measured data can be used directly, without a near-field-to-far-field transformation.

From a theoretical point of view, the spiral antenna has a broad bi-directional radiation pattern with the two maxima perpendicular to the plane on which the spiral antenna is located, the *xy*-plane, as shown in Figure 1. Simulations performed without any absorbing material have verified this pattern.

The measured radiation patterns shown in Figure 9 display a minor asymmetry of the main beam at both frequencies. This asymmetry is caused by the asymmetry in the CPS of the balun. At 0.8 GHz, the front to back ratio is 10 dB. This ratio is increased to 15 dB at 1.5 GHz because the attenuation in the absorbing material increases with the frequency.

At a frequency of 0.8 GHz, the simulated directivity is 5.6 dB, which is only 0.2 dB below the measured directivity estimated from the measured HPBW. In the frequency range between 0.5 and 2.6 GHz, the directivity is estimated to be in the range between 2.7 and 6.6 dB, using the measured HPBW.

The antenna is designed to meet conditions where circular polarization is required. A commonly used criterion states that the axial ratio should be less than 6 dB [9]. The axial ratio shown in Figure 10 indicates that the spiral antenna is circular polarized in the $\theta = 0$ degrees direction at frequencies between 0.5 and 2.6 GHz. For frequencies above 2.6 GHz, the axial ratio is not measured, but circular polarization is expected up to a certain frequency point where the size of the feed region, i.e., the center of the spiral, becomes small compared to the balun structure. This frequency limit is expected to be near the frequency limit for acceptable S_{11}-parameter, which is around 3.8 GHz (Figure 7). Above 3.8 GHz, it is likely that the polarization ellipse becomes even more elliptic.

The polarization ellipse is somewhat elliptical at a frequency of 2.6 GHz, as shown in Figure 10(b). This phenomenon can also be observed in Figure 9 as a difference in the amplitude in the $\theta = 0$ degrees direction for the measured θ-cuts.

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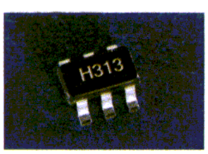
HMC314	0.7 - 4.0	5.0	185	18.0	29.5
HMC315	DC - 7.0 DC - 7.0	5.0 7.0	31 50	12.0 16.5	26.8 31.0
HMC323 & HMC324	DC - 3.0	7.5	57	16.8	30.0
HMC326MS8G	3.4 - 3.6	5	125	24	36.0

HBT Darlington Amplifier

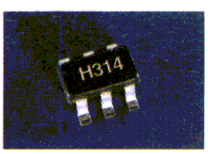
16 dB Gain



HMC313

HBT Amplifier w/ Power Down

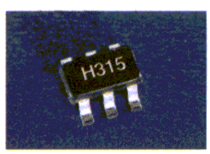
10 dB Gain



HMC314

HBT Darlington Amplifier

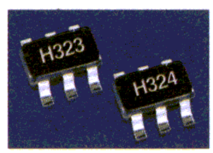
12 dB Gain



HMC315

HBT Driver & Dual Driver Amplifier

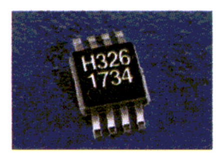
11 - 12 dB Gain



HMC323 & HMC324

HBT Driver Amplifier for WLL

20 dB Gain



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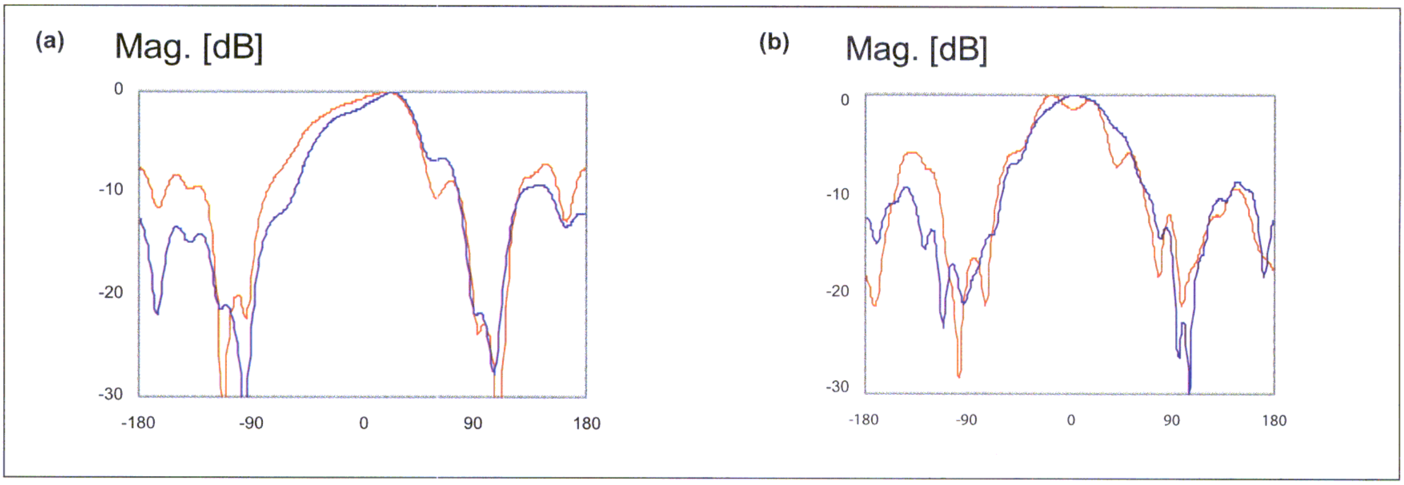

Figure 9. Measured θ-cuts for ϕ = 0 degrees (red) and for ϕ = 90 degrees (blue). Positive values of θ' correspond to $\phi = \phi_0$, and $\theta = \theta$' and negative value of θ' corresponds to $\phi = \phi_0$ + 180 degrees, $\theta = -\theta$', with respect to the polar orientation of the spiral antenna at frequencies of (a) 0.8 GHz and (b) 1.5 GHz.

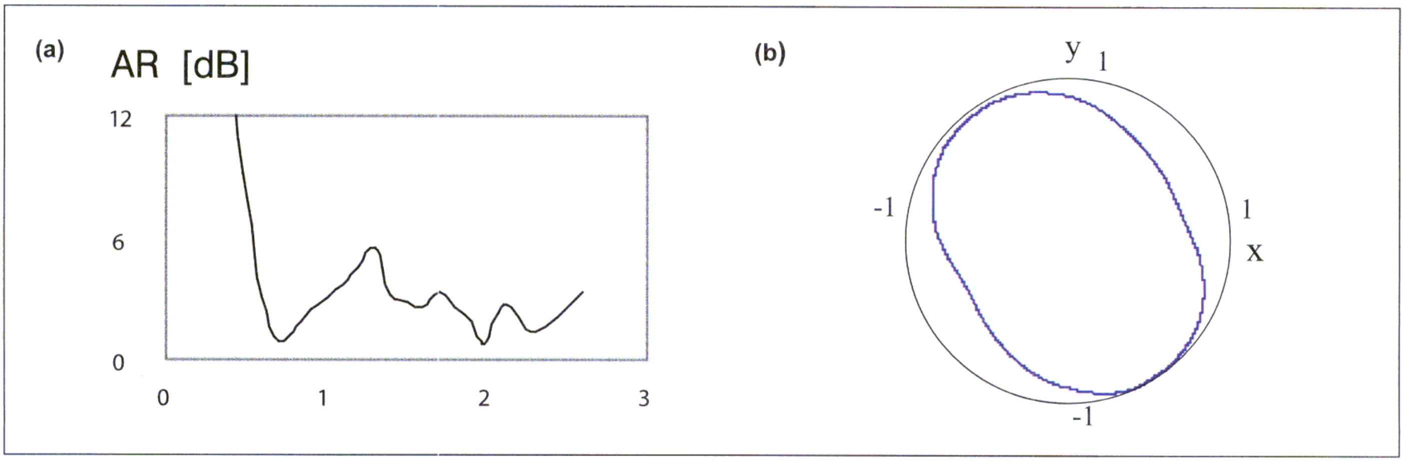

Figure 10. (a) Measured axial ratio vs. frequency and (b) measured polarization ellipse at a fequency of 2.6 GHz.

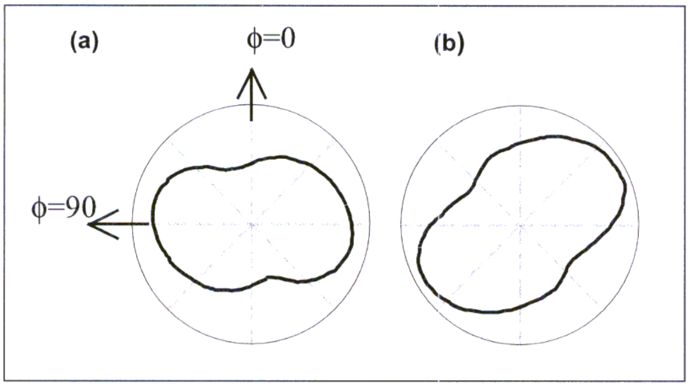

Figure 11. Simulated radiation pattern at a ϕ cut for θ = 60 degrees obtained at (a) 0.8 GHz and (b) 1.3 GHz.

should be unchanged at all frequencies, except for a rotation around the polar axis [2]. The results from the simulations of the radiation at a ϕ cut at two frequencies are shown in Figure 11. The radiation pattern is rotated around the polar axis. The design equations for the spiral antenna state that the relationship for the angle of the rotation, $\Delta\phi$, as a function of the initial upper frequency and the lower frequency, is given by $\Delta\phi = a^{-1} ln(f_{upper}/f_{lower})$ [2], where a is the growth rate. When using this equation, the theoretical angle of rotation between 1.3 GHz and 0.8 GHz is 55 degrees, which is in good agreement with the simulated angle obtained from Figure 11. The simulated angle of rotation is 52 degrees. Therefore, it is very likely that the spiral antenna radiates from an active region that is dependent on the frequency of operation.

Note that for a theoretical model of the structure of the spiral antenna there is no particular feeding point, and the active region should begin at a frequency dependent point on the antenna. The present result seems to contradict the theoretical model. The viewpoint of the theoretical model is in accordance with the

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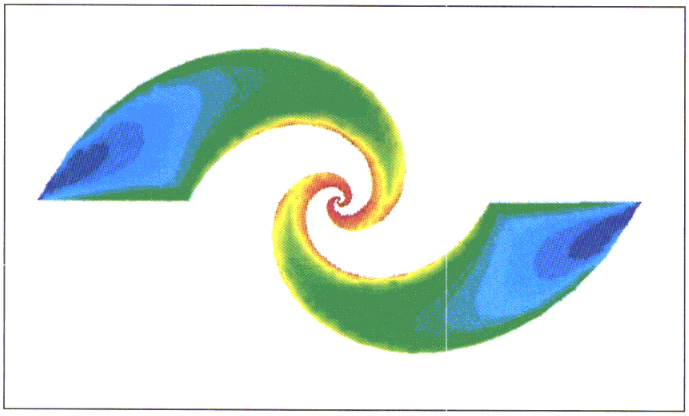

▲ **Figure 12. Illustration of the 1.5 turn coplanar strip fed logarithmic spiral antenna. Simulation of the average current distribution on the metallic surface of the spiral antenna, by using IE3D. Red illustrates the highest current intensity and blue illustrates the lowest current intensity, both at 0.5 GHz.**

rotation of the radiation pattern. The reason may be that the region near the feed point contributes only a little. Note that the measured rotation is 52 degrees and the theoretical is 55 degrees between 1.3 GHz and 0.8 GHz. Since the region near the feed point contributes, the rotation is less than expected from theory. This aspect is examined [15]. There, Thaysen, et al, discuss the relation between the angle of rotation of the radiation pattern and the position of the current active region on the spiral antenna.

If the antenna is constructed so that the initial radius r_0 is only a fraction of a wavelength and is excited from a transmission line connected at the small end, a current wave travels out from the feed point along the arms. As the energy radiates, the amplitude of the current decreases. This effect can be observed in Figure 12, where the average current density is shown for a frequency of 0.5 GHz. This current distribution is calculated using IE3D. The result shows that the average current density is located symmetrically on the two spiral arms with the highest current density at the feed point and at the edges. Also, a slight decrease in current density as a function of the distance from the feed points is observed and is indicated by a shift in color from red to blue.

Conclusion

This article discussed the design and model of a spiral antenna for the FR-4 substrate, and a balun for the RT/Duroid substrate. The electromagnetic simulation program IE3D has been used to simulate the performance of the spiral antenna.

Good agreement was shown between the numerical results and the measured results obtained using a network analyzer in the frequency range from 300 kHz to 5 GHz. The measured reflection coefficient for the fabricated spiral antenna shows a slightly better performance than the simulated results, which is most likely due to computer capacity limitations. The measured reflection coefficient is better than –10 dB, over a 9 to 1 bandwidth from 0.4 to 3.8 GHz.

An inverse proportionality between the input impedance and the relative permittivity of the substrate was found. The simulations and the measurements of the spiral antenna show that the radiation pattern and the input impedance are essentially constant over a bandwidth larger than 9 to 1.

Measurements in an anechoic chamber from 0.8 GHz to 2.6 GHz are made showing an axial ratio of less than 3.3 dB and a directivity in the range between 2.7 and 6.6 dB. At a frequency of 0.8 GHz, the simulated directivity is 5.6 dB, only 0.2 dB below the measured directivity.

From the simulated ϕ cuts at two different frequencies, it is shown that the radiation pattern rotates around the polar axis as a function of the frequency.

The simulated current distribution on the metallic surface of the spiral antenna shows that the average current distribution decreases as a function of the distance from the feeding point.

The constructed uniplanar spiral antenna and the balun are well-suited for use in a stepped frequency ground penetrating radar for humanitarian demining because of the spiral's wide bandwidth, circular polarization, relatively small size, inexpensive cost and uniplanar design. ■

Acknowledgements

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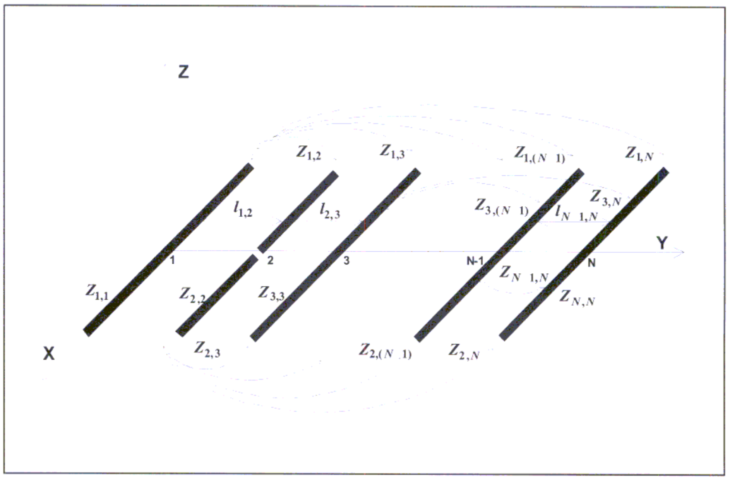
Calculating Multi-Element Antennas Using Mathcad®

Here is a technique for mathematical analysis of antennas, using a Yagi-Uda array of several elements as an example

By Néstor E. Arias
National University of Tucumán

A linear antenna can be combined with another to obtain a new antenna with more gain. If there are more than two or three elements used to build that antenna, estimating the proper and mutual impedances, excitation point and gain becomes complicated.

Estimation of these variables involves the use of exponential integrals, and the number of calculations necessary to form the matrix impedance is equal to the number of elements taken by pairs. For example, in the case of 15 elements, it is necessary to carry out 105 calculations plus 15 more of their proper impedances by means of the integrals as those mentioned above. It is also necessary to estimate the corresponding currents for matrix calculation, which is essential for calculating the gain. Because the number of calculations necessary to estimate the necessary elements can be numerous, and because of the exclusion of any other parameters, the use of a mathematical program is ideal for these calculations.



▲ **Figure 1. Yagi-Uda antenna.**

This article discusses a calculation method using the Mathcad® program (we will use Mathcad 7 Professional), which estimates and plots radiation diagrams on the three planes of space for multi-element antennas. The Mathcad program will also obtain the excitation point impedance.

Multi-element antennas

A thin linear antenna of an approximate length of $\lambda/2$ can be combined with one or more similar antennas to create a new type of antenna with a different characteristic radiation than that obtained with a single element.

Various types of multi-element antennas exist, with groups defined by the way in which the different elements are physically arranged, and how they are excited with regard to phase and amplitude.

A group of two linear antennas of an approximate length of $\lambda/2$ successively distributed in space, in parallel with one another and having a single excited element constitutes the type of antenna known as a Yagi-Uda (Figure 1). This antenna features electromagnetic coupling of all the elements of the system, which can also include nearby physical elements, such as trees, people, roofs and the ground. To simplify calculations, the antenna will be regarded in an empty space free of strange couplings, and only taking into account the antenna elements present in that region.

Array gain

It is necessary to calculate the impedance matrix [$\boldsymbol{Z}$] in order to obtain gain, the radiation patterns for each plane of space and the excitation point impedance. For each disposition of elements of the array, the matrix impedance contains the proper and mutual impedances of the system. The second element normally is excited, so a column tension matrix [$\boldsymbol{v}$] can be defined where that second element is not zero.

$$[v]=[Z]\times[I]\therefore[I]=\left([Z]^{-1}\times[v]\right) \qquad (1)$$

where $\boldsymbol{I}$ = $l$solve ($\boldsymbol{Z}$, $\boldsymbol{v}$).

The matrix [$\boldsymbol{I}$] contains the currents of all the elements and therefore, it is possible to calculate the gain on the space planes and the excitation point impedance. The impedance is

$$Z_i=\frac{v_2}{I_2} \qquad (2)$$

The array gain, relative to a half-wave dipole gain, can be calculated through the currents in each of the elements. To calculate the array gain, it is necessary to determine the actual excitation power of the array.

$$W=R_e\left[(I_2)^2(Z_i)\right] \qquad (3)$$

This power can be used, in turn, to excite a half-wave length dipole which will serve as reference, with a 73-ohm input impedance whose current is

$$i_{00}=\sqrt{\frac{W}{73}} \qquad (4)$$

Using the calculated values and applying the diagram multiplication principle, the array gain in each plane is represented in Equation 5. Equations 5(a) and 5(b) (below) describe the XY and ZY planes. Gain in the XZ plane is

$$G_{plane\ XZ}=\frac{\sum_{1}^{N} I_n}{i_{00}} fd_{\frac{\lambda}{2}} \qquad (5c)$$

where the array gain is relative to the half-wave dipole.

Given two antennas of lengths l_1 and l_2 separated by a distance d, as shown in Figure 2, the mutual impedance $Z_{2,1}$ can be calculated as in Equation (6). There are many references that include methods for calculation of mutual impedances between linear elements of unequal lengths. The method described by Cox [1] has been used for this paper.

By successively applying the calculation in Equation (6), all of the mutual impedances that are involved in a system of N elements are obtained. The same equation can be used to obtain the proper impedances by defining $l_1 = l_2$ and d = 0.02, where d represents the width of an element.

Program sequence

To develop a Mathcad program to perform these cal-

$$G_{azimuth}=\frac{I_1e^{j0}+I_2e^{j\beta 1_{1,2}\cos(\phi)}+\ldots+I_{N-1}e^{j\beta 1_{1,N-1}\cos(\phi)}+I_Ne^{j\beta 1_{1,N}\cos(\phi)}}{i_{00}} fd_{\frac{\lambda}{2}}$$

where,

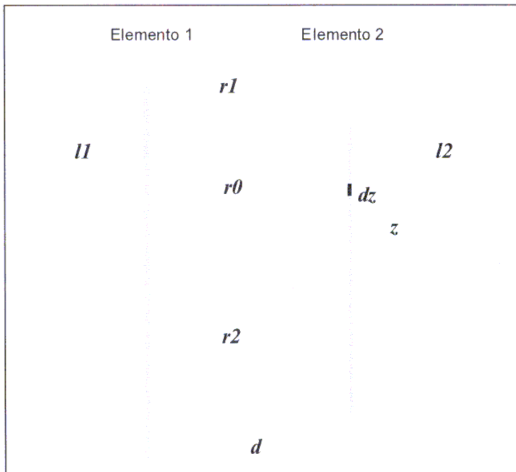
ϕ = Azimuth angle;
β = phase constant;
$l_{1,2}\ldots l_{1,N}$ = distances of the elements to the origin;
$I_1\ldots I_N$ = currents of the elements;
$fd_{\lambda/2}$ = diagram factor of the half-wave dipole;
N = the number of elements.

▲ **Equation 5(a). Gain in the XY plane.**

$$G_{elevation}=\frac{I_1e^{j0}+I_2e^{j\beta 1_{1,2}\cos(\phi)}+\ldots+I_{N-1}e^{j\beta 1_{1,N-1}\cos(\phi)}+I_Ne^{j\beta 1_{1,N}\cos(\phi)}}{i_{00}}$$

θ = elevation angle.

▲ **Equation 5(b). Gain in the ZY plane.**



$$r_0=\sqrt{d^2+z^2}$$
$$r_1=\sqrt{d^2+(l_1-z)^2}$$
$$r_2=\sqrt{d^2+(l_1+z)^2}$$

▲ **Figure 2. Two dipoles of different length.**

$$Z_{2,1} = \frac{30}{\sin(\beta \times l_1) \times \sin(\beta \times l_2)} \left[\begin{array}{c} \int_0^{l_2} \frac{e^{-j\beta(r_1 - l_2 + z)}}{r_1} dz - \int_0^{l_2} \frac{e^{-j\beta(r_1 + l_2 - z)}}{r_1} dz + \\ \int_0^{l_2} \frac{e^{-j\beta(r_2 - l_2 + z)}}{r_2} dz - \int_0^{l_2} \frac{e^{-j\beta(r_2 + l_2 - z)}}{r_2} dz - \\ 2 \times \cos(\beta \times l_1) \int_0^{l_2} \frac{e^{-j\beta(r_0 - l_2 + z)}}{r_0} dz + 2 \times \cos(\beta \times l_1) \int_0^{l_2} \frac{e^{-j\beta(r_0 + l_2 - z)}}{r_0} dz \end{array} \right]$$

▲ **Equation (6). Calculation of mutual impedances.**

culations, the following steps have been followed:

Step 1. Necessary data
- Proper length
- Separation between elements
- Wavelength
- Number of elements N

Step 2.
- The program calculates a matrix (l) that contains the data entered, the distances of the elements to the origin and the proper longitudes.
- The limit conditions for calculation of the integrals with Equation (2).

Step 3.
- Matrix **Z** calculation.

Step 4.
- Definition of matrix $\boldsymbol{v}$ and calculation of the current matrix.

Step 5.
- Calculation of power W_R and current i_{oo}.

Step 6.
- Subroutine for the calculation of equations (5a), (5b) and (5c).

Step 7.
- Graphs on planes XY, YZ and XZ.

Figure 3 shows the radiation patterns relative to the half-wave dipole in a 10-element antenna for the ZY elevation and XY azimuth planes.

The Appendix shows the Mathcad programs used to perform these step-by-step calculations.

Conclusion

Any other value of N, length and separation between elements may be used in the Mathcad calculation. Steps in the optimization of the system to obtain maximum gain and the best situation in the forward-backward relation must be performed manually.

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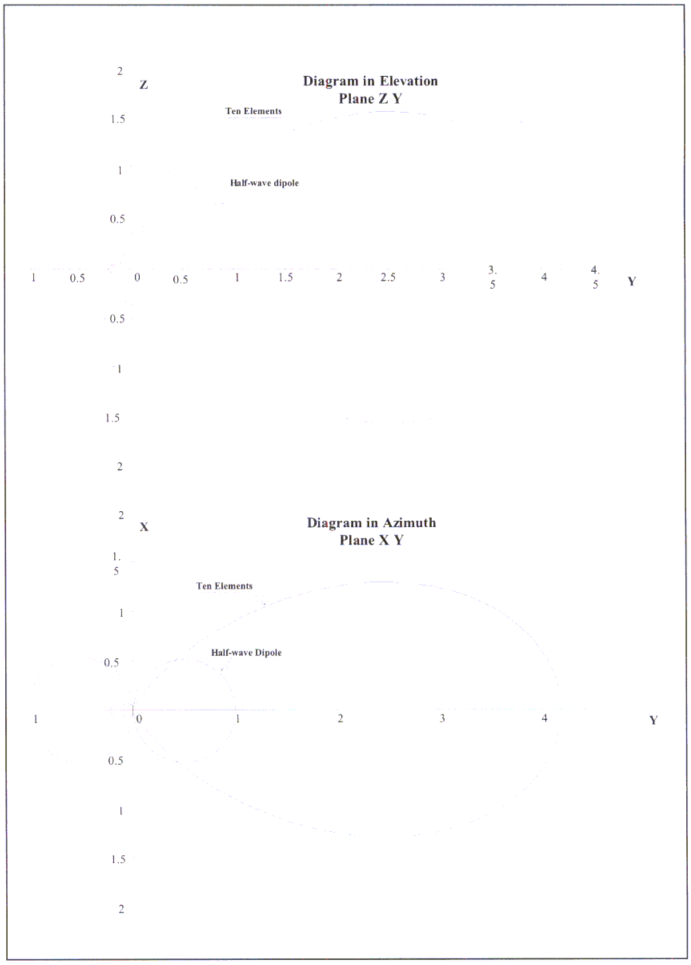

▲ **Figure 3. Radiation patterns relative to the half-wave dipole for a 10-element antenna.**

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4. The first element of a matrix is one. Therefore, it should be modified in Math-Option Array Origin to =1.

Appendix — Mathcad programs

```
:=1        N:=10        :=2 –

l1 1:=.245   l2 2:=.235   l3 3:=.2    l4 4:=.18    l5 5:=.16       <==Proper lengths
                                                                      Data
l6 6:=.15    l7 7:=.15    l8 8:=.15   l9 9:=.15    l10 10:=.15

l1 2:=.17    l2 3:=.15    l3 4:=.15   l4 5:=.15    l5 6:=.15       <== Separation of
                                                                      Elements
l6 7:=.21    l7 8:=.15    l8 9:=.15   l9 10:=.15                      Data
```

▲ **Step 1 of the Mathcad program.**

```
Subroutines set; Matrix l

J:= | j←2
    | while j≤N
    |   | a1 j←0+l j−1 j
    |   | j←j+1
    |   | a
    | i←2
    | while i≤N
    |   |          i
    |   | J1 i←    Σ   a1 n
    |   |        n = 2
    |   | i←1+i
    | J

U:= | for i  1.. N
    |   for j  1.. N
    |     Ui j←J1 j
    | U

UU:= | i←1
     | while i≤N
     |   | j←2
     |   | while j≤N
     |   |   | UUi j←Ui j−Ui i
     |   |   | j←j+1
     |   | i←i+1
     | UU

Q:= | for i  1.. N
    |   for j  1.. N
    |     | Qi j←lj j  if =j
    |     | Qi j←UUi j  otherwise
    |     | Qi j←Qj i  if i≥j
    | Q

l:=Q

L:= | for i  1.. N
    |   for j  1.. N
    |     | Li j←−lj j  if =j
    |     | Li j←0  if i  j
    |     |          lj j−li i
    |     | Li j←  ---------  otherwise
    |     |             2
    | L
```

▲ **Step 2 of the Mathcad program.**

continued on the following page.

Z := for i 1.. N
  for j 1.. N
    $L \leftarrow l_{i\,j}$ if i=j
    $L \leftarrow l_{j\,j}$ otherwise
    $LL \leftarrow l_{i\,j}$ if i≠j
    $LL \leftarrow l_{i\,i}$ otherwise
    $d \leftarrow 0.02$ if i=j
    $d \leftarrow l_{i\,j}$ otherwise

$$b_{i\,j} \leftarrow \left[\int_0^L \frac{e^{-1j\left[\sqrt{d^2+(LL-z)^2}-L+z\right]}}{\sqrt{d^2+(LL-z)^2}}\,dz - \int_0^L \frac{e^{-1j\left[\sqrt{d^2+(LL-z)^2}+L-z\right]}}{\sqrt{d^2+(LL-z)^2}}\,dz\right]\frac{30}{\sin(\ L+.0000001)\ \sin(\ LL+.00000001)}$$

$$c_{i\,j} \leftarrow \left[\int_0^L \frac{e^{-1j\left[\sqrt{d^2+(LL+z)^2}-L+z\right]}}{\sqrt{d^2+(LL+z)^2}}\,dz - \int_0^L \frac{e^{-1j\left[\sqrt{d^2+(LL+z)^2}+L-z\right]}}{\sqrt{d^2+(LL+z)^2}}\,dz\right]\frac{30}{\sin(\ L+.0000001)\ \sin(\ LL+.00000001)}$$

$$f_{i\,j} \leftarrow \left[2\cos(\ LL)\left[\int_0^L \frac{e^{-1j\ \sqrt{d^2+z^2}+L-z}}{\sqrt{d^2+z^2}}\,dz - \int_0^L \frac{e^{-1j\ \sqrt{d^2+z^2}-L+z}}{\sqrt{d^2+z^2}}\,dz\right]\right]\frac{30}{\sin(\ L+.0000001)\ \sin(\ LL+.00000001)}$$

    $Z_{i\,j} \leftarrow b_{i\,j}+c_{i\,j}+f_{i\,j}$
    $Z_{i\,j} \leftarrow Z_{j\,i}$ if i j
    Z
Z

Z =

68.718 27.3i	50.714 − 11.897i	17.767 − 26.218i	−4.56 − 21.028i	−13.17 − 7.22i	−10.714 4.323i	0.741 9.244i	6.811 4.409i	6.775 − 2.455i	1.895 − 6.195i
50.714 − 11.897i	61.039 12.372i	40.92 − 8.264i	17.555 − 21.105i	−1.79 − 18.203i	−10.978 − 7.874i	−8.455 6.039i	−0.393 8.882i	5.937 4.994i	6.707 − 1.516i
17.767 − 26.218i	40.92 − 8.264i	39.815 − 34.457i	29.057 − 9.128i	12.417 − 15.267i	−1.258 − 13.753i	−9.649 − 2.78i	−6.839 4.796i	−0.372 7.125i	4.729 4.038i
−4.56 − 21.028i	17.555 − 21.105i	29.057 − 9.128i	30.716 − 58.342i	22.229 − 9.192i	10.159 − 12.663i	−4.496 − 10.035i	−8.457 − 2.476i	−6.011 4.178i	−0.35 6.237i
−13.17 − 7.22i	−1.79 − 18.203i	12.417 − 15.267i	22.229 − 9.192i	23.275 − 80.655i	17.981 − 9.073i	4.731 − 11.657i	−3.891 − 8.759i	−7.348 − 2.181i	−5.235 3.61i
−10.714 4.323i	−10.978 − 7.874i	−1.258 − 13.753i	10.159 − 12.663i	17.981 − 9.073i	20.08 − 91.295i	13.704 − 8.788i	4.403 − 10.855i	−3.605 − 8.146i	−6.819 − 2.037i
0.741 9.244i	−8.455 6.039i	−9.649 − 2.78i	−4.496 − 10.035i	4.731 − 11.657i	13.704 − 8.788i	20.08 − 91.295i	16.703 − 8.986i	8.235 − 10.417i	−0.839 − 9.838i
6.811 4.409i	−0.393 8.882i	−6.839 4.796i	−8.457 − 2.476i	−3.891 − 8.759i	4.403 − 10.855i	16.703 − 8.986i	20.08 − 91.295i	16.703 − 8.986i	8.235 − 10.417i
6.775 − 2.455i	5.937 4.994i	−0.372 7.125i	−6.011 4.178i	−7.348 − 2.181i	−3.605 − 8.146i	8.235 − 10.417i	16.703 − 8.986i	20.08 − 91.295i	16.703 − 8.986i
1.895 − 6.195i	6.707 − 1.516i	4.729 4.038i	−0.35 6.237i	−5.235 3.61i	−6.819 − 2.037i	−0.839 − 9.838i	8.235 − 10.417i	16.703 − 8.986i	20.08 − 91.295i

▲ **Step 3 of the Mathcad program.**

$$v := \left| \begin{array}{l} \text{for } i \quad 1..N \\ \quad \left| \begin{array}{l} v_i \leftarrow 0 \ \text{ if } \ i \neq 2 \\ v_i \leftarrow 1 \ \text{ if } \ i = 2 \end{array} \right. \\ v \end{array} \right. \qquad v \quad \begin{bmatrix} 0 \\ 1 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ 0 \end{bmatrix} \qquad I := Z^{-1}\, v$$

▲ **Step 4 of the Mathcad program.**

$$Zi := \frac{v_2}{I_2} \qquad Zi \quad 24.625 \quad 18.101i \qquad w := v_2\, I_2$$

$$w \quad 0.026 - 0.019i \qquad wR := \mathrm{Re}(w) \qquad wR \quad 0.026$$

$$i00 := \sqrt{\frac{wR}{73}} \qquad i00 \quad 0.019 \qquad p := 1\ 2..360$$

▲ **Step 5 of the Mathcad program.**

$$b := \left| \begin{array}{l} \text{for } j \quad 1..N \\ \quad \text{for } s \quad 1..360 \\ \quad\quad \left| \begin{array}{l} dr_j \leftarrow \ l_{1\ j} \\ dr_j \leftarrow 0 \ \text{ if } \ j = 1 \\ dr \\ a_{s\ j} \leftarrow \overrightarrow{dr_j \ \cos \ s \frac{}{180}} \\ A_{s\ j} \leftarrow \overrightarrow{\dfrac{I_j\ e^{j\ a_{s\ j}}}{i00}} \\ A_{s\ j} \leftarrow \overrightarrow{A_{s\ j} \dfrac{\cos \ - \sin \ s \frac{}{180}}{\cos \ .0001 + s \frac{}{180}}} \\ \phantom{A_{s\ j} \leftarrow}\ {}^{2} \\ B_{s\ j} \leftarrow \overrightarrow{\dfrac{I_j\ e^{j\ a_{s\ j}}}{i00}} \\ k \leftarrow \displaystyle\sum_{n=1}^{N} I_n \\ C_s \leftarrow \overrightarrow{\dfrac{k}{i00} \dfrac{\cos \ \frac{}{2} - \cos \ s \frac{}{180}}{\sin \ .0001 + s \frac{}{180}}} \end{array} \right. \\ \begin{bmatrix} a \\ dr \\ A \\ B \\ C \end{bmatrix} \end{array} \right.$$

▲ **Step 6 of the Mathcad program.**

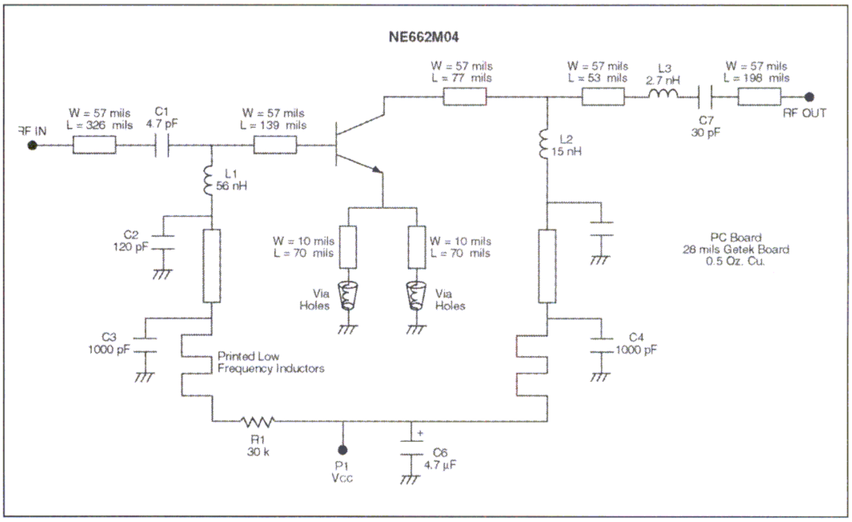
$$E := \left[\sum_{n=1}^{N} \left[b_3 \quad n \right] \right] \qquad B := \left[\sum_{n=1}^{N} \left[b_4 \quad n \right] \right]$$

▲ **Step 7 of the Mathcad program.**

Optimizing a Silicon Bipolar LNA Performance for Bluetooth Applications

By Olivier Bernard
California Eastern Laboratories

This application note describes a low-noise amplifier designed for the specifications of the Bluetooth RF standard. The NE662M04 is NEC's latest generation of silicon bipolar junction RF transistor, using a state-of-the-art UHSO 25 GHz f_T wafer process. It provides low-voltage and low-current performance and is suited for low-noise mobile applications. The emphasis here is on achieving a low-noise, high-gain performance while keeping current consumption at a minimum and a fairly good input and output match. General information, test results, circuit schematic, board layout and billing of material are enclosed as a reference. California Eastern Laboratories also manufactures this circuit as an evaluation board for quick turn-around design cycles.



▲ **Figure 1. NE662M04 Bluetooth LNA schematic.**

LNA design and matching network

The LNA is described in Table 1, and a schematic is provided in Figure 1. In order to achieve the specification and provide the appropriate matching, this design uses an inductive emitter feedback through a high impedance printed transmission line. All other matching elements are lumped components.

The approximately 0.7 nH emitter inductance provides a few advantages: It allows for matching the device for minimum noise figure performance while keeping an acceptable input matching, it enhances the in-band stability performance of the LNA and, to a lesser degree, it improves the linearity of the device. However, the trade-off is reduced gain and the potential for high frequency oscillations. The latter can easily be negated by carefully choosing the out-of-band matching.

The bias network is a simple base resistor that will fix the current to 5 mA. In circuits where temperature stability or reduced sensibility to manufacturing variations are more of an issue, an active stabilization bias can be used with commonly available electronic bipolar transistors.

The input matching network comprises only C_1 and L_1 and is set to achieve Γ_{OPT} from a 50-ohm input load. L_1 is also used as the choke that brings the base current to the transistor and is

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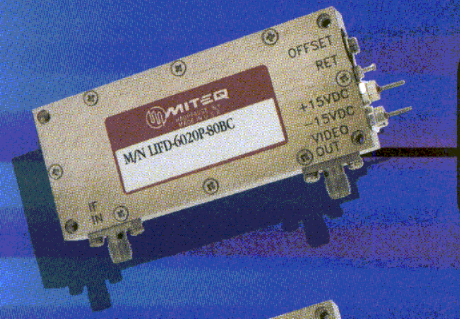
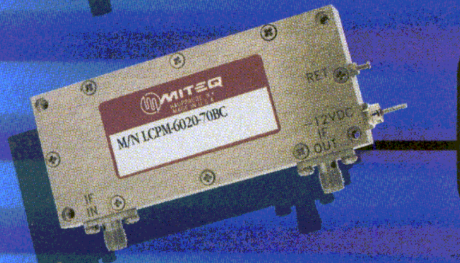


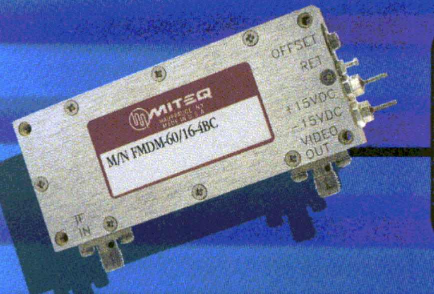
SUCCESSIVE DETECTION LOGARITHMIC AMPLIFIERS

MODEL NUMBER	CENTER FREQUENCY (MHz)	DYNAMIC RANGE (dBm, Min.)	LINEARITY (dB, Max.)	RISE TIME (ns, Max.)	LOGGING SLOPE INTO 93 OHMS (mV/dB, Typ.)
LIFD-3010P-80BC	30	-80 to 0	±0.5	100	25
LIFD-6020P-80BC	60	-80 to 0	±0.5	50	25
LIFD-7030P-80BC	70	-80 to 0	±0.5	30	25
LIFD-16040-80BC	160	-80 to 0	±1.0	30	25
LIFD-300100-70BC	300	-70 to 0	±1.0	20	15



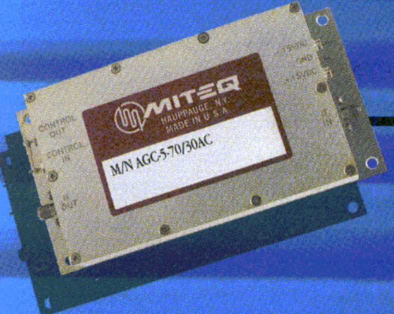
CONSTANT PHASE LIMITING AMPLIFIERS

MODEL NUMBER	CENTER FREQUENCY (MHz)	DYNAMIC RANGE (dB, Min.)	OUTPUT POWER (dBm, Min.)	POWER VARIATION (dB, Max.)	PHASE VARIATION (Max.)
LCPM-3010-70BC	30	-70 to 0	10	±0.5	±3°
LCPM-6020-70BC	60	-70 to 0	10	±0.5	±3°
LCPM-7030-70AC	70	-65 to 5	10	±0.5	±5°
LCPM-16040-70BC	160	-65 to 5	10	±1.0	±3°



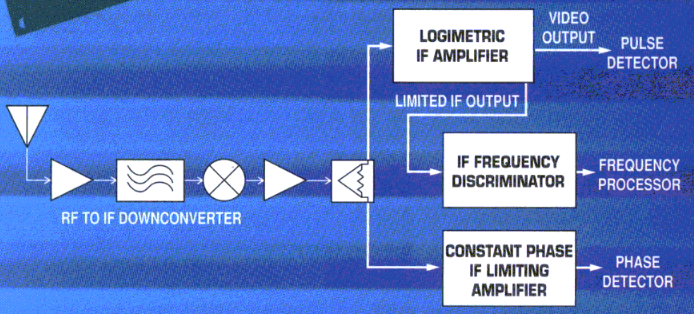
FREQUENCY DISCRIMINATORS

MODEL NUMBER	CENTER FREQUENCY (MHz)	LINEAR BANDWIDTH (MHz, Min.)	SENSITIVITY (mV/MHz, Typ.)	LINEARITY (%, Max.)	RISE TIME (ns, Max.)
FMDM-30/6-3BC	30	6	1000	±3	120
FMDM-60/16-4BC	60	16	250	±3	90
FMDM-70/36-10AC	70	36	50	±2	50
FMDM-160/35-15BC	160	35	100	±2	30
FMDM-160/50-15AC	160	50	40	±2	25
FMDM-750/150-20BC	750	150	20	±3	20
FMDM-1000/300-50AC	1000	300	10	±5	7



AUTOMATIC GAIN CONTROL LINEAR AMPLIFIERS

MODEL NUMBER	CENTER FREQUENCY (MHz)	BANDWIDTH (-3 dB) (MHz, Min.)	DYNAMIC RANGE (dBm, Min.)	OUTPUT POWER (dBm, Min.)	POWER VARIATION (dB, Max.)
AGC-7-10.7/4AC	10.7	4	-70 to 0	10	±0.5
AGC-7-21.4/10AC	21.4	10	-70 to 0	10	±0.5
AGC-5-70/30AC	70	30	-50 to 0	-4	±0.5
AGC-7-160/30AC	160	30	-70 to 0	8	±1.5
AGC-7-300/400AC	300	400	-65 to 0	3	±1.0



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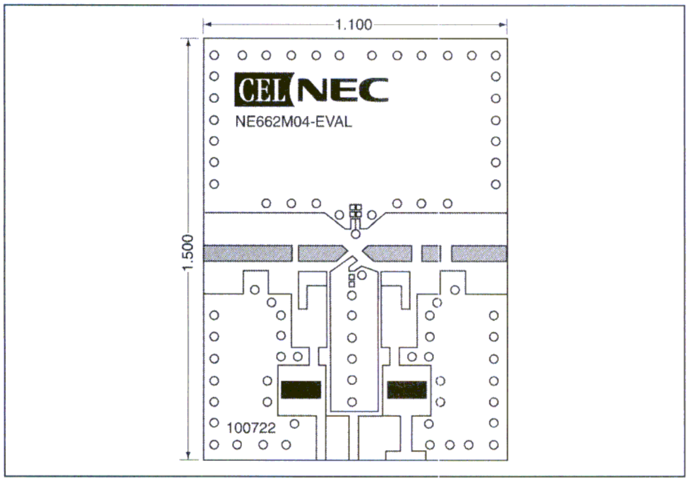
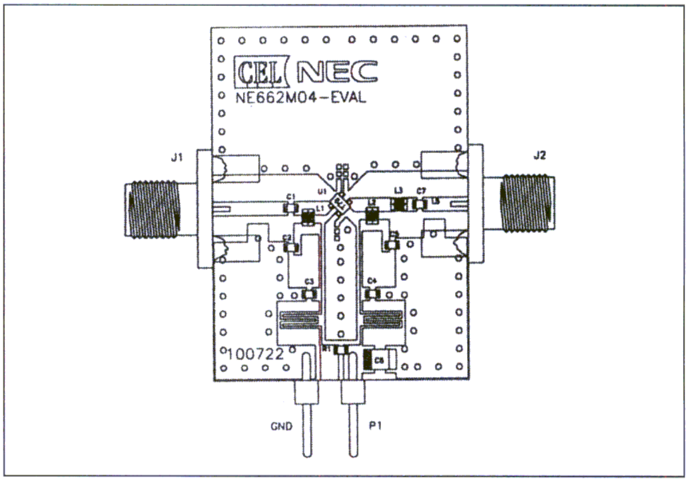
Item	Parameters LNA Section	Specifications	Test	Units Results	Notes
1	Voltage	3	3	V	Low Voltage
2	Current	5	5	mA	Low Current
3	Operating Frequency	2400–2483.5	1930–1990	GHz	US/European Bands
4	Gain	10	12	dB	—
5	NF	1.5	1.3	dB	Low Noise
6	Input IP3	0	3	dBm	High IP3
7	1 dB Compression Point	–5.0	0	dBm	—
8	Input VSWR (50 ohms)	2.5:1 (–9.5 dB)	–10	—	—
9	Output VSWR (50 ohms)	1.5:1 (–14 dB)	–15	—	—
10	Stability at all Frequencies	Unconditionally Stable	Unconditionally Stable	—	—
11	Operating Temperature	–40 to +80	-40 to +80	°C	—

▲ **Table 1. Specifications and test results of the Bluetooth low noise amplifier.**



▲ **Figure 2. Printed circuit board layout.**



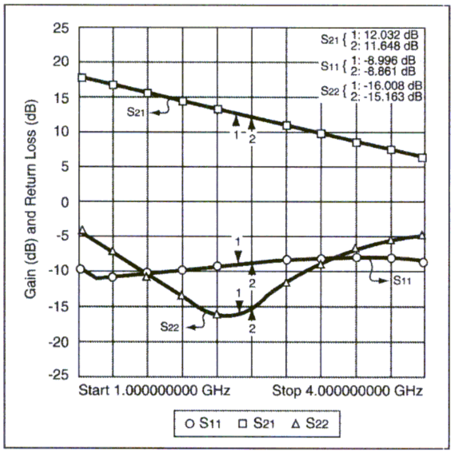
▲ **Figure 3. LNA assembly drawing.**

RF-grounded by C_2. The long transmissions line to and from C_3 are only additional DC chokes to bias the device. The output matching is a simple capacitive match through C_7 that provides a better than 1.5:1 match. L_2 is the RF choke grounded through C_5 and further choking is provided by the printed inductive transmission line. The circuit schematic, layout and assembly drawing are shown in Figures 1, 2 and 3. The bill of materials, including manufacturers' part numbers, is provided in Table 2.

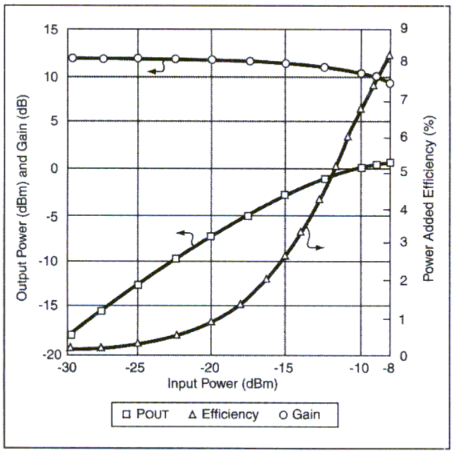
The matching has been optimized for the specific Bluetooth receiving band, but the PCB board can be used in a wide range of applications from 460 MHz to about 3 GHz for a number of devices, providing that an appropriate matching

Qty.	Part or identifying no.	Nomenclature or description	Material/ specification
1	TF-100413	—	NE34018-eval test fixture block
1	LL 1608-FH5N6S	L1	5.6 nH inductor Toko
1	LL 1608-FH15NJ	L2	15 nH inductor Toko
1	LL 1608-FH2N7S	L3	2.7 nH inductor Toko
1	MCR03J303JK	R1	0603 30 K ohm res Rohm
2	MCH185A121JK	C2, C5	0603 120 pF cap Rohm
1	MCH185A4R7CK	C1	0603 4.7 pF cap Rohm
2	MCH185A102JK	C3, C4	0603 1000 pF cap Rohm
1	MCH185C300JK	C7	0603 30 pF cap Rohm
1	881-6116	C6	4.7 mF cap aux
1	NE662M04	U1	IC NEC
2	2340-6111 TG	P1	PIN header 3M
2	2052-1215-00	J1, J2	OSM jack Omni Spectra
1	FD-100722	PCB	NE622M04-eval fab drawing

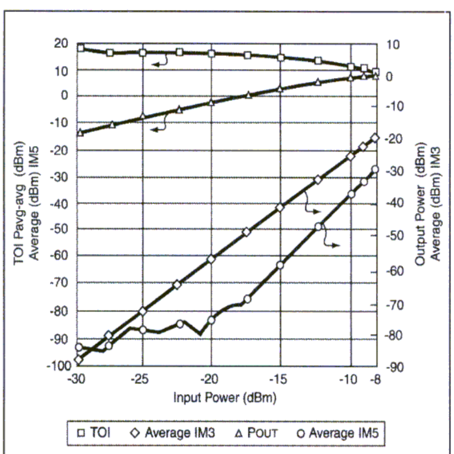
▲ **Table 2. LNA bill of materials.**



▲ **Figure 4. Mesured small signal performance.**



▲ **Figure 5. Output power performance.**



▲ **Figure 6. Linearity performance.**

network is synthesized. The 2.4 GHz LNA performance results are presented in Figures 4 through 6 and Table 3. ■

Author information

Olivier Bernard is the Manager of Strategic Marketing at California Eastern Laboratories in Santa Clara, CA. He received his BSEE and MSEE in RF design from ESIEE, an engineering school in Paris, France, and his MSEE in Electrophysics from the University of Southern California. He can be reached at olivier.bernard@cel.com.

Input Power (dBm)	Output Power (dBm)	Gain (dB)	Average (dBm) IM3	C / IM3 (dBc) Pavg-avg	TOI (dBm) Pavg-avg
–30.00	–18.15	11.85	–88.57	70.50	17.18
–29.00	–17.20	11.80	–87.54	70.42	18.09
–28.00	–16.29	11.71	–80.97	64.59	15.91
–27.00	–15.22	11.78	–78.80	63.50	16.45
–26.00	–14.20	11.80	–75.87	61.58	16.50
–25.00	–13.23	11.77	–72.48	59.16	16.27
–24.00	–12.24	11.76	–69.83	57.67	16.68
–23.00	–11.25	11.75	–66.67	55.33	16.33
–22.00	–10.25	11.75	–63.08	52.75	16.04
–21.00	–9.26	11.74	–60.93	51.66	16.57
–20.00	–8.28	11.72	–57.28	48.91	16.09
–19.00	–7.29	11.71	–54.46	47.09	16.17
–18.00	–6.34	11.66	–51.67	45.16	16.07
–17.00	–5.36	11.64	–48.44	42.83	15.80
–16.00	–4.40	11.60	–45.41	40.67	15.60
–15.00	–3.46	11.54	–42.37	38.59	15.51
–14.00	–2.57	11.43	–39.48	36.34	15.02
–13.00	–1.69	11.31	–35.86	33.58	14.52
–12.00	–0.86	11.14	–31.86	30.25	13.51
–11.00	–0.14	10.86	–27.81	26.83	12.45
–10.00	0.51	10.51	–24.15	23.66	11.34
–9.00	1.04	10.04	–21.04	21.00	10.46

- Frequency: 2.400 GHz
- V_D: 2.999 V, ID = 5 mA
- P_{OUT} at 1 dB: 168 dBm
- Gain at 1 dB: 10.744 dB
- Efficiency at 1 dB: 6.39 %
- IIP3 at PIN = –21 dBm = +4.77 dBm
- Noise Figure at 2.4 GHz = 1.15 dB

▲ **Table 3. Summary of power performance.**

Effects of Parasitics in Circuit Simulations

Simulation accuracy can be improved by including parasitic inductances and capacitances

By Robin Croston
California Eastern Laboratories

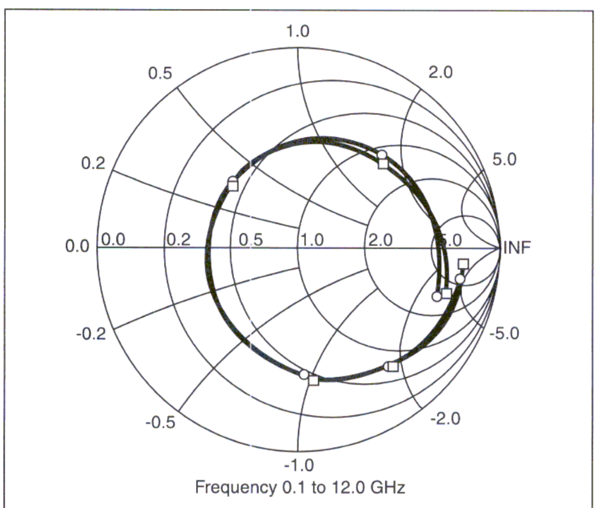
Including the proper parasitics in a nonlinear simulation can make the difference between an accurate prediction of circuit performance with minimal on-bench tuning and a design that requires significant modification after prototyping. This article discusses the inclusion of parasitics in a low noise amplifier designed for the Bluetooth RF standard using the NE662M04, NEC's latest generation of silicon bipolar junction RF transistor. (See "Optimizing a Silicon Bipolar LNA Performance for Bluetooth Applications" by Olivier Bernard, in this issue on pages 56-60.)

California Eastern Laboratories provides this circuit as an evaluation board to its customers. Circuit schematics and measured versus simulated data with varying levels of circuit complexity graphically illustrate the effects of parasitics on the simulated results.

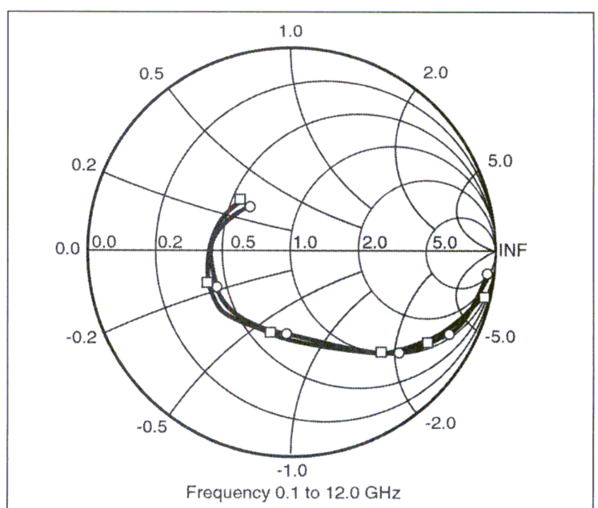
NLM verification

The first step in any nonlinear simulation is to confirm the validity of the nonlinear model at the bias and frequency range of interest. This should be performed to the manufacturer's measured *S*-parameter data. If the particular bias of interest is not available, confirm the model to the nearest available bias.

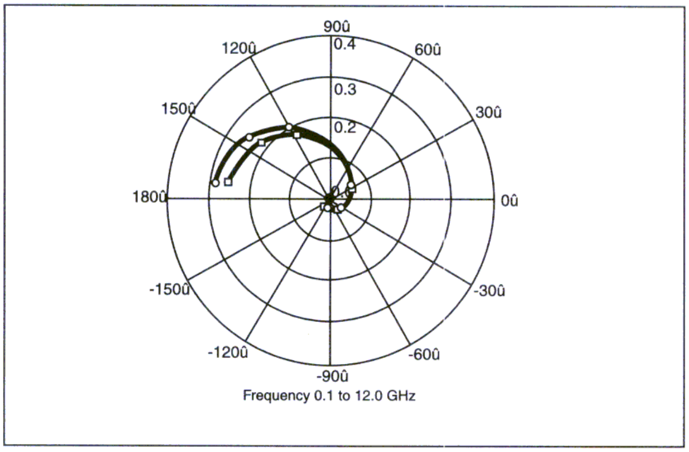
Figures 1 through 4 confirm that the model is appropriate at the circuit design bias of 3 volts, 5 mA and over a frequency range of 0.1 to 12 GHz. The model also matches measured gain performance at 2 volts, 10 mA and 2 GHz to better than 3 percent error and P_{1dB} performance at the same bias to 7 percent error. Modeled NF_{min} deviates from measured data by a 4 percent error at a device bias of 2 volts, 5 mA and the application frequency of 2.4 GHz. This information indicates that the transistor is mod-



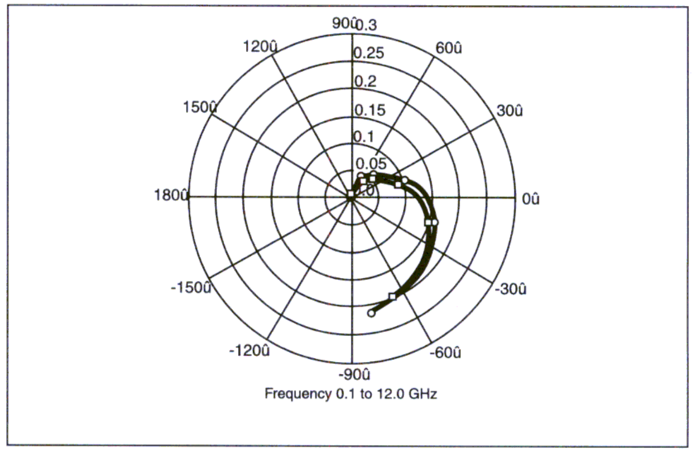
▲ **Figure 1. Measured vs. modeled S_{11} for the NE662M04.**



▲ **Figure 2. Measured vs. modeled S_{22} for the NE662M04.**



▲ **Figure 3. Measured vs. modeled S_{21} for the NE662M04.**

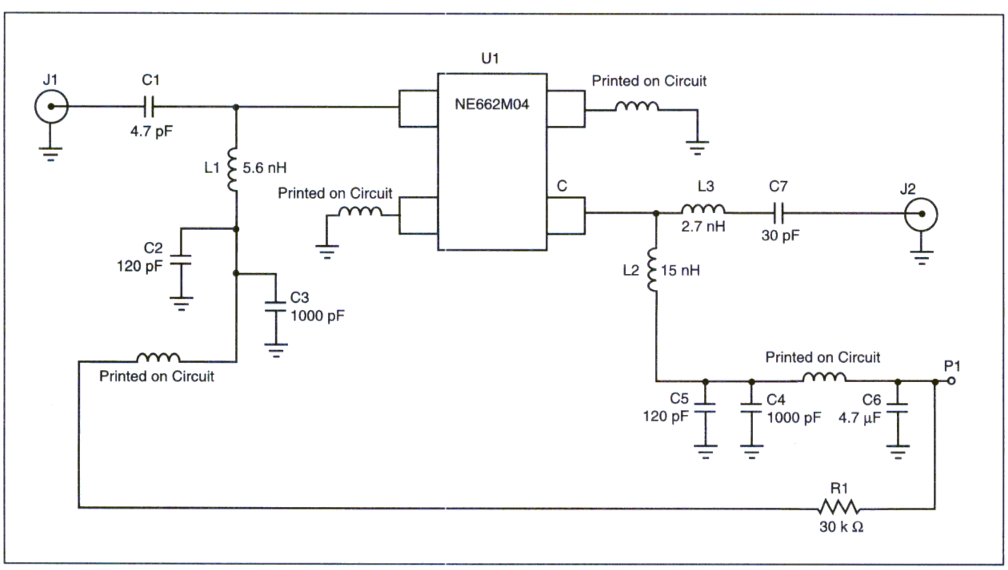


▲ **Figure 4. Measured vs. modeled S_{12} for the NE662M04.**

eled correctly for this application and if the rest of the circuit is properly represented, the goal of minimum tuning of the prototype should be achieved.

Simple circuit simulation

The design used to illustrate this example is a low noise amplifier optimized for the Bluetooth RF standard using the NE662M04 [1]. California Eastern Laboratories provides evaluation boards, a description file, the circuit schematic and a board layout. To demonstrate the effects of including parasitics in circuit simulations, the circuit schematic in Figure 5 as implemented in the HP-EEsof Series IV Libra simulator (Figure 6) was compared to data measured on the evaluation board. Transmission line lengths are measured on the evaluation board. Figures 7 through 9 summarize the results of simulating the circuit without including board parasitics. It is apparent that the actual circuit is not properly represented by the schematic shown in Figure 6.



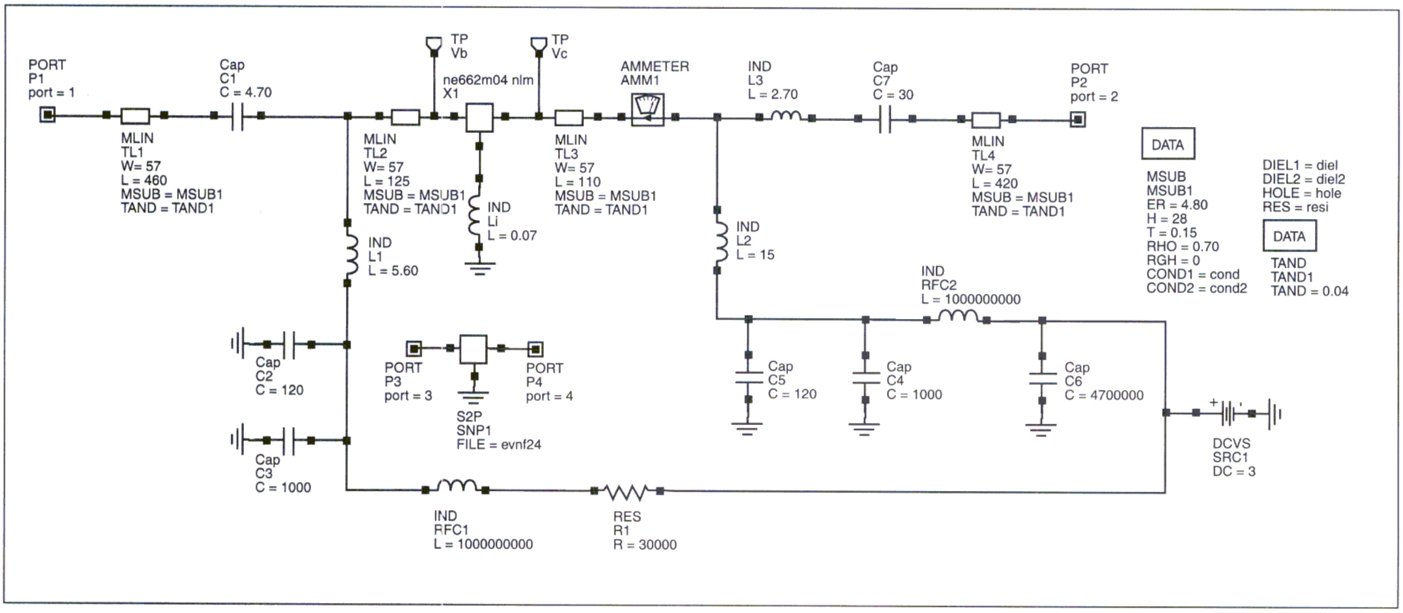
▲ **Figure 5. Evaluation board circuit schematic.**

Advanced modeling

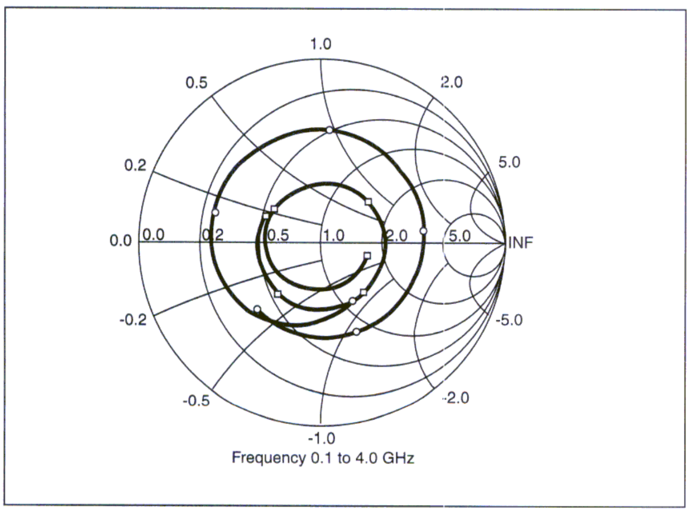
To improve the accuracy of the simulation, various techniques are used. One important parasitic to include is via holes. The via holes take into account the inductive loss through the substrate to ground. While these inductances are small — 40 to 60 pH/mm of substrate thickness [2] — they can significantly alter simulation results when not included. As with all parasitic effects, the higher the operation frequency of the circuit, the greater the effect of the parasitics.

More accurate models of the actual transmission lines are used to account for the discontinuities at the junctions of different width microstrip lines. This discontinuity also occurs where discrete components are soldered to the board. Such junctions have scattering matrix representations that depend on the widths of the microstrip [3] and take into account the width-dependant inductances and capacitance of the junction. The long transmission lines (TL2 and TL3 in Figure 6) on either side of the device are broken into two sections and a t-line is used where the shunt components are placed (TL9, TEE2, TL2, TL3, TEE1, and TL10 in Figure 10). The circuit in this example is fairly simple. For more involved circuits, more elaborate modeling of the transmission lines may be needed. The degree of complexity of all parasitic modeling will depend on the operation frequency of the circuit and how important it is to model the response of the circuit to higher order tones.

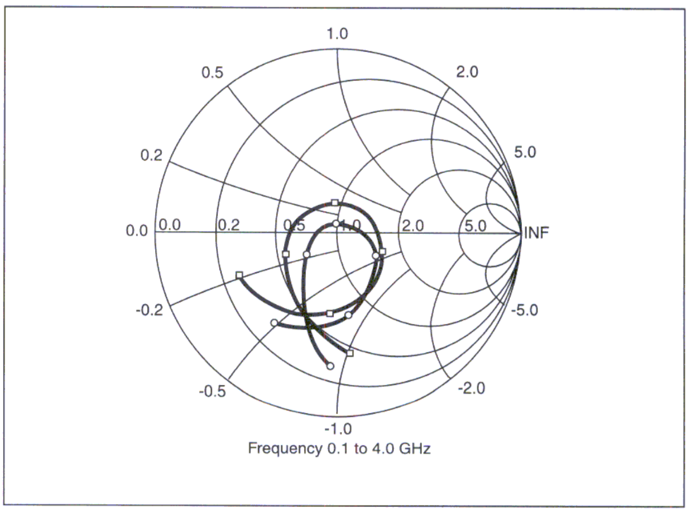
The third step is to model the discrete passive components with more accurate representations. The active



▲ **Figure 6. Simple circuit simulation schematic.**



▲ **Figure 7. Simulation of S_{11} for the simple circuit.**

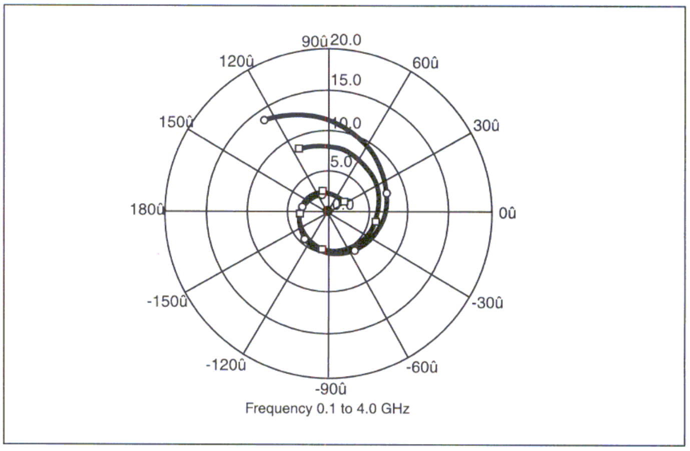


▲ **Figure 8. Simulation of S_{22} for the simple circuit.**

device model has already been confirmed to be appropriate. There are several different choices to model the passives. Where available, actual measured *S*-parameters in the form of Touchstone formatted files (*.s2p) can be used. Information available on the passive devices will vary according to the manufacturer.

If yield analysis needs to be simulated on the circuit, measured data is not a good option. Another choice is a model of the device that includes the parasitics present in the component package. Figures 11 and 12 illustrate the models for the capacitor and the inductor used in the final simulation.

A third option is to use the lossy lumped element models available in the simulator being used. Figure 13 is an example of a capacitor with *Q*, as implemented in



▲ **Figure 9. Simulation of S_{21} for the simple circuit.**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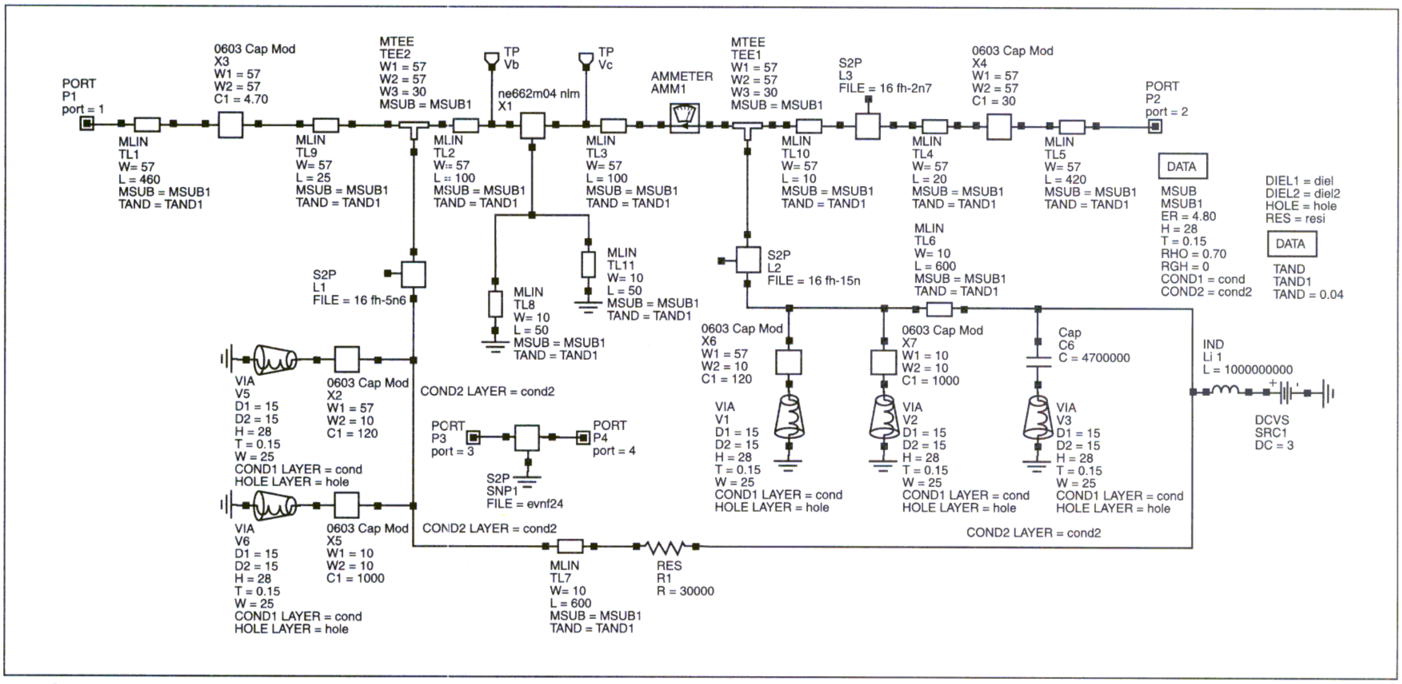
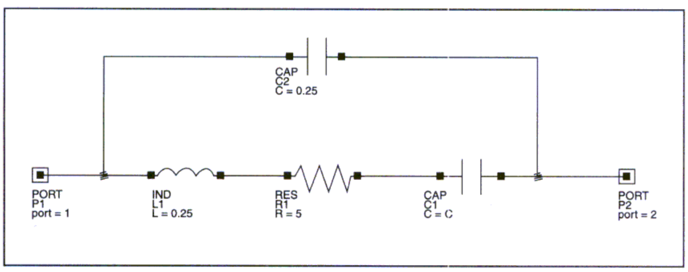
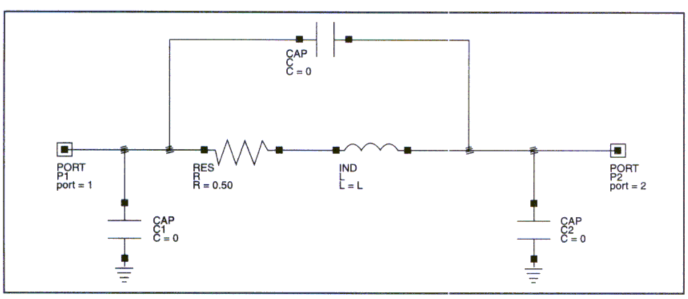


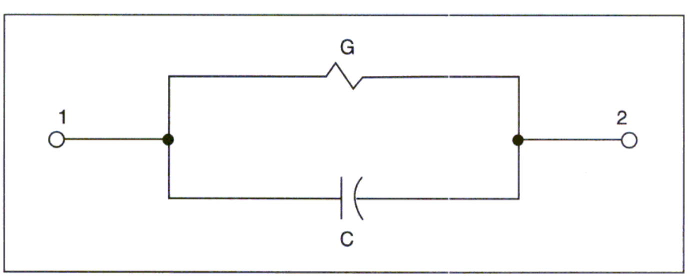
▲ **Figure 10. Final simulation schematic.**



▲ **Figure 11. Capacitor model schematic.**



▲ **Figure 12. Inductor model schematic.**



▲ **Figure 13. Simulator capacitor schematic.**

the Agilent-EESof Series IV Libra simulator [4]. In high-frequency applications where very accurate modeling of passive components is required, passives should be measured on the substrate that will be used and models should be developed from this characterization [5].

Some parasitics are not accounted for in this simulation. Elements not in the RF path, such as the bias resistor R_1 and the DC blocking capacitors are represented by ideal elements. The RF chokes, printed on the circuit board as meander lines, are represented by a straight transmission line. Simulating a more complex circuit does not improve the measured versus modeled performance. The final simulation circuit with added parasitics is shown in Figure 10. The much improved measured vs. simulated results are shown in Figures 14 through 16 and Table 1.

Conclusion

By carefully modeling all of the elements of a circuit, a design can be simulated that can be used to accurately predict circuit performance. This allows circuit designers to gain an edge in trimming the design cycle in both cost and time. ■

References

1. Olivier Bernard, "Optimizing a Silicon Bipolar LNA Performance for Bluetooth Applications," *Applied Microwave & Wireless*, Vol. 13, No. 1, (January 2001): 56-60.

2. Robert A. Pucel, "Design Considerations for Monolithic Microwave Circuits," *IEEE Transactions on Microwave Theory and Techniques*, Vol. MTT-29, No. 6,

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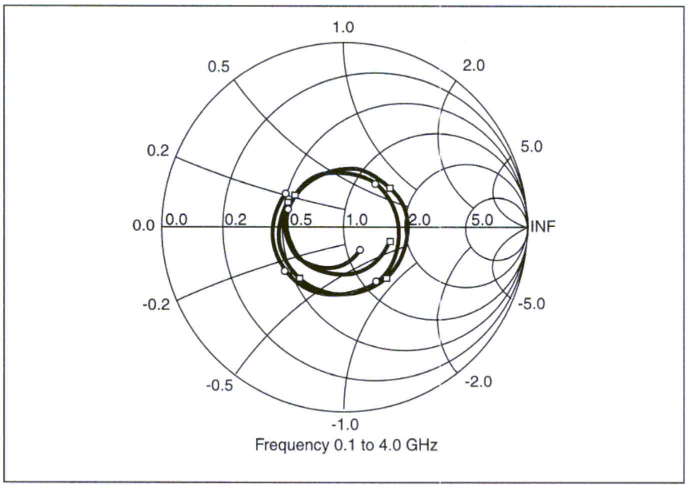
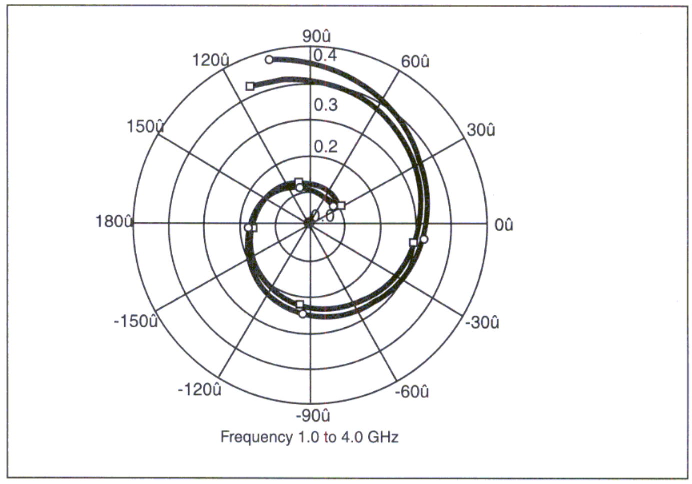
Circle 27



▲ **Figure 14. Simulation of S_{11} for the final circuit.**



▲ **Figure 16. Simulation of S_{21} for the final circuit.**

(June 1981): 513–534.

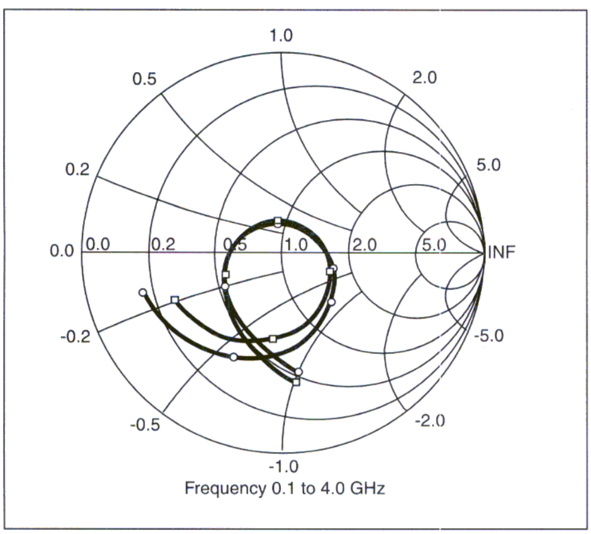
3. K.C. Gupta, et al, *Computer-Aided Design of Microwave Circuits*, 1981.

4. HP-EEsof *Microwave & RF Circuit Design Circuit Element Catalog*: 9–21.

5. Thomas A. Winslow, "Component Modeling for PCB Design," *IEEE Microwave Magazine*, Vol. 1, No. 1, (March 2000): 61–63.

Author information

Robin Croston is a device modeling engineer with California Eastern Labs. She received her BSEE, MSEE and MS in Physics from Montana State University. She has been working in the field of high frequency device modeling for five years. Previously, she was a test engineer at Boeing. She may be reached via e-mail at robin.croston@cel.com.



▲ **Figure 15. Simulation of S_{22} for the final circuit.**

Parameters LNA Section	Specifications	Test Results	Simulation Results	Units
Voltage	3	3	3	V
Current	5	5	5	MA
Operating Frequency	2400–2483.5	2400	2400	MHz
Gain	10	12	13.1	dB
NF	1.5	1.4	1.5	dB
1 dB Compression Point	–5.0	–0.5	–7.6	dBm
Input VSWR (50 Ohms)	2.5:1 (–9.5 dB)	–9.3	–11	dB
Output VSWR (50 Ohms)	1.5:1 (–14 dB)	–15.3	–16.5	dB

▲ **Table 1: Specifications, test results and simulation results of the Bluetooth low noise amplifier.**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Amplifier Linearization Using Adaptive Digital Predistortion

The need for greater linearity can be addressed at the digital coding level

By Shawn P. Stapleton
Agilent Technologies

The emphasis on higher data rates and spectral efficiency has driven the wireless industry towards linear modulation techniques, such as quadrature phase shift keying (QPSK), quadrature amplitude modulation (64 QAM) and multicarrier configurations. Although these linear modulation techniques provide good spectral efficiency, they produce a signal with a fluctuating envelope that generates intermodulation distortion (IMD) at the system's power amplifiers. Most of the IM power appears as interference between adjacent channels, which requires the use of highly linear power amplifiers.

One way to compensate for these nonlinear distortions is linearization of the power amplifier via predistortion. In most applications, linearization is a more desirable approach than backing off a Class A amplifier, which lowers power efficiency and increases heat dissipation. Recently, active linearization has emerged as a critical technology in modern wireless communications systems that addresses this problem.

Adaptive digital predistortion, in which the adaptation mechanism is based on the difference between the desired modulation and the power amplifier's output, is one linearization technique that is well-suited to baseband applications that employ digital signal processors (DSPs). A digital predistorter provides significant IMD reduction over low to moderate bandwidths, while continuously adjusting for component drift and power variations. This predistortion technique operates independently of the chosen modulation scheme.

This article introduces the concepts of power amplifier linearization and digital predistortion, looking in detail at the complex gain-based look-up table technique of digital predistortion. An actual simulated example of this approach is also demonstrated.

Linearization overview

Over the years, a number of linearization technologies have been developed. Predistortion has been the most common approach deployed in new systems today. With predistortion, a nonlinear module is inserted between the modulated input signal and the primary power amplifier stage. The nonlinear module generates IMD products precisely in anti-phase with the IMD products produced by the power amplifier, theoretically removing any out-of-band emissions caused by the power amplifier.

Another common approach is feed-forward linearization, which is the only strategy that simultaneously offers wide bandwidth and good IMD suppression. The price for this performance is high complexity. In addition, automatic adaptation mechanisms are essential for maintaining performance regardless of variables such as temperature and component drift.

Of the two alternatives, RF predistortion is a better choice for many applications. It offers minimal complexity with reasonable IMD reduction over moderate bandwidths. The primary limitation of this method is the difficulty in acquiring accurate RF functional models.

Cartesian feedback, a relatively low complexity approach, also offers reasonable IMD suppression, but stability problems with this line of attack limit bandwidth to a few hundred kilohertz, as well as restrict accuracy.

Linear amplification using non-linear components (LINC) techniques convert the input signal into two constant envelope signals, which

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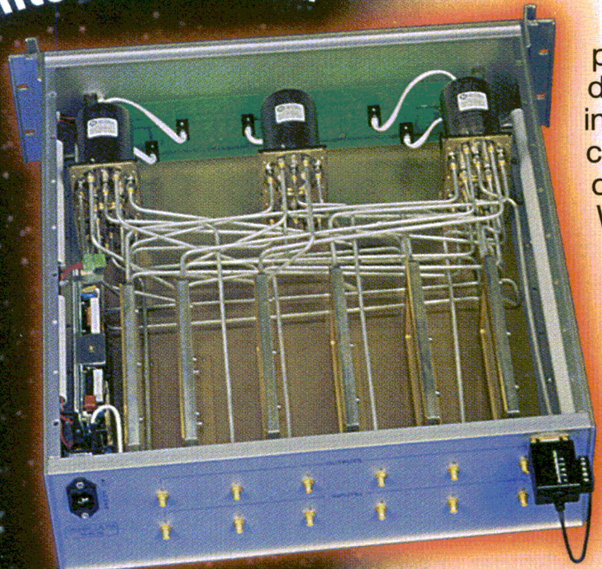

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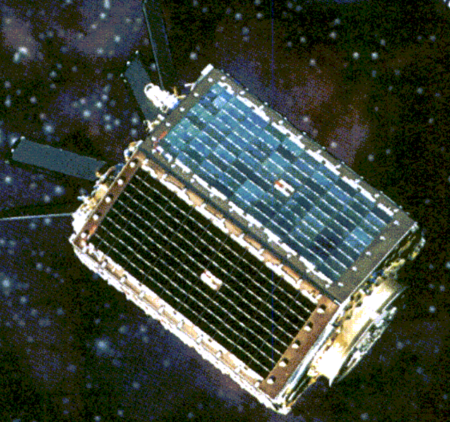

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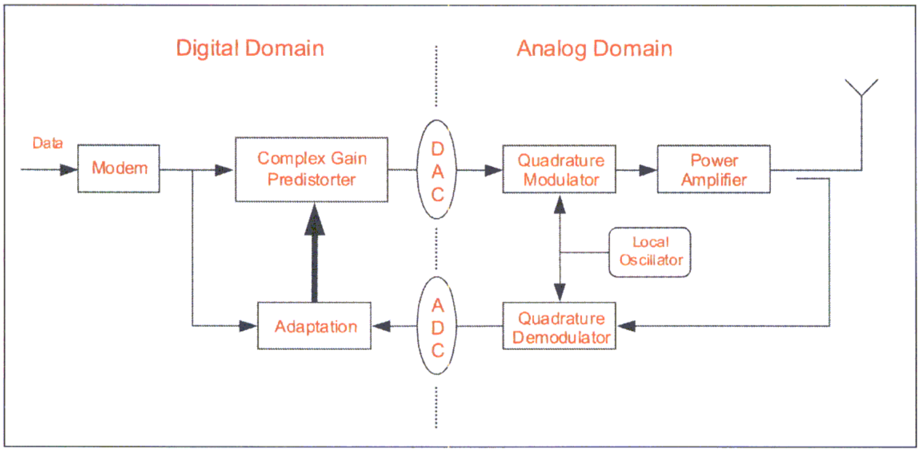
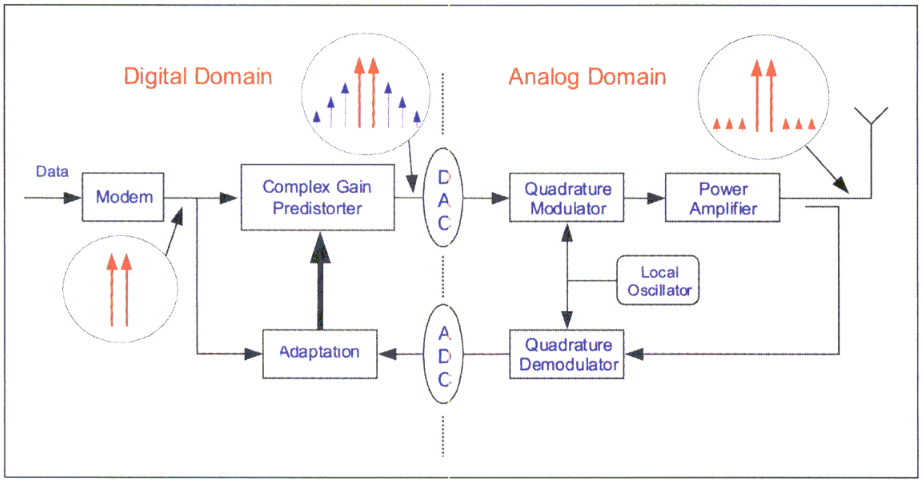


▲ **Figure 1. Design of the digital predistorter.**



▲ **Figure 2. Spectral response of the digital predistorter at various nodes, using a two-tone input signal.**

are amplified by Class C amplifiers and then combined before transmission. In addition to being complex, LINC designs are particularly sensitive to component drift.

Dynamic biasing is similar to predistortion except that the work function is based on the power amplifier's operating bias. Dynamic biasing is notably limited in its ability to suppress adjacent channel interference.

Digital predistortion

Digital predistortion possesses two advantages. First, the correction from the nonlinear module is applied before the power amplifier where high power insertion loss is less critical. Second, significant IMD reductions can be achieved. The primary disadvantages of digital predistortion are its relative complexity and bandwidth limitations tied to the accuracy and computational rate of the specific DSP used in the system.

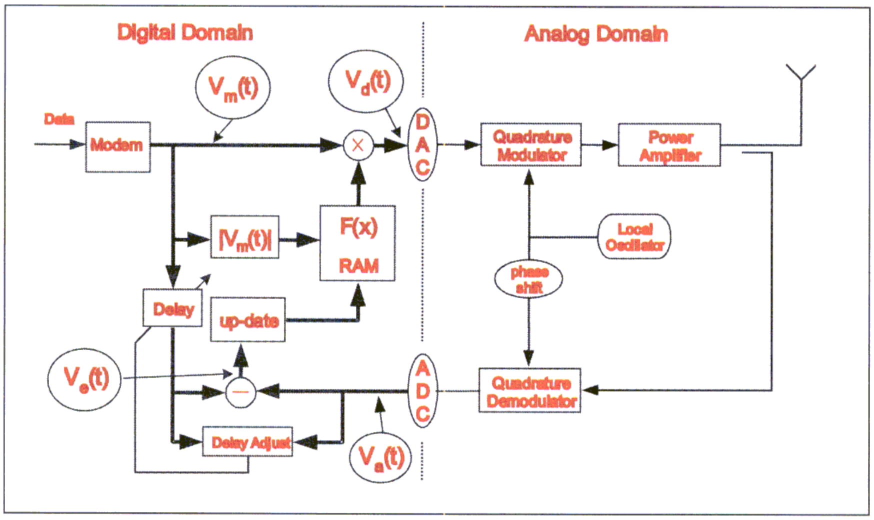
The linearizer circuit in Figure 1 creates a predistorted version of the desired modulation. The predistorter includes a complex gain adjuster that directs the amplitude and phase of the input signal. The amount of predistortion applied is controlled by updating values in a look-up table with the interpolated amplitude modulation to amplitude modulation (AM/AM) and amplitude modulation to phase modulation (AM/PM) nonlinearities of the power amplifier. Note that the inputs to the adaptation function include a delayed version of the output and input signals. The input is delayed and then subtracted from the power amplifier's output signal. Theoretically, the result is only the distortion (that is, the IMD components) added by the power amplifier.

The spectral response of the digital predistorter can be observed at various nodes using two-tone input signal, as shown in Figure 2. Once optimized, the complex gain adjuster (predistorter) should exhibit nonlinear characteristics that are exactly the inverse of the power amplifier. This can be confirmed by observing the spectral growth characteristics of the predistorter at the input node to the power amplifier. Ideally, the IMD products detected at this node will have the same amplitude as the distortion generated when the two tones are passed through the power amplifier, but in anti-phase. The adaptation process quickly adjusts the look-up table entries to minimize distortion.

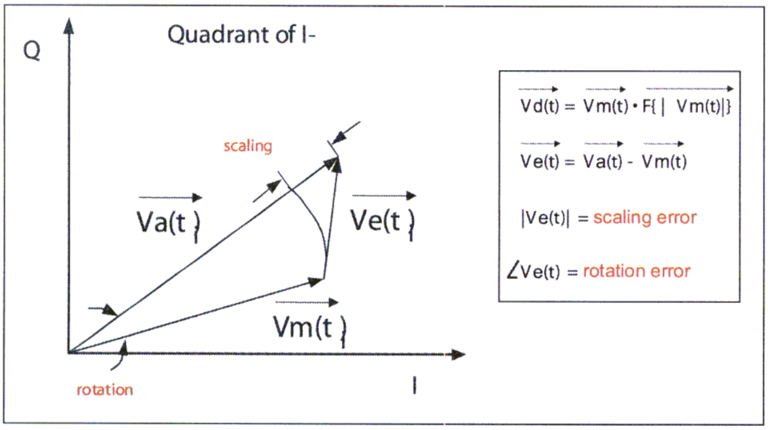
Design techniques

There are three generic approaches to digital predistortion: complex vector mapping lookup, complex gain lookup and Cartesian feedback. The complex vector mapping technique uses interpolated input vectors that are maintained in a look-up table generated by adding an error vector to compensate for AM/AM and AM/PM distortions. The complex gain approach multiplies the input signal by an optimized complex gain vector stored in a look-up table, which is indexed by the envelope of the input signal. Finally, Cartesian feedback is a less complicated approach that does not require a look-up table, but tends to be less stable.

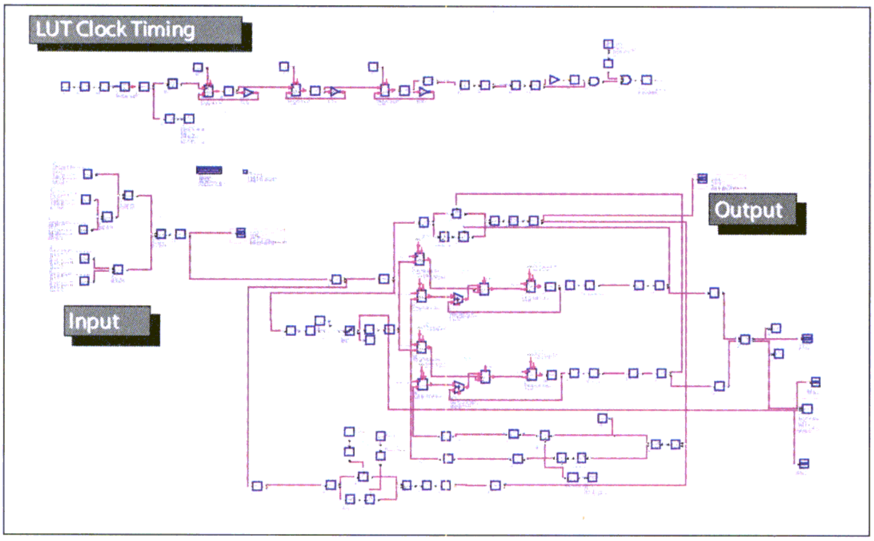
Conventional adaptation techniques for the digital predistorter use a gradient signal that is based on continually computing the gradient of a three-dimensional power surface that represents the difference (i.e., the error signal) between the input signal and the scaled output signal. Adjacent channel interference power is



▲ **Figure 3. The complex gain look-up table.**



▲ **Figure 4. The complex gain function stored in the look-up table is the scaling and rotating error applied to the input vector by the power amplifier.**



▲ **Figure 5. Circuit schematic for the example digital predistorter simulated in the Agilent Advanced Design System.**

minimized when this error signal is completely suppressed. Since the gradient is continuously being computed, no amount of constant misadjustment is required.

Two different gradient estimation methods are commonly used. Linear convergence is a technique that uses a first order feedback loop, while the alternate secant method estimates the gradient using a process based on Newton's classical method.

Complex gain look-up predistorter

An example of the complex gain look-up technique is illustrated by the circuit in Figure 3. The input signal is multiplied by a gain signal stored in RAM. This gain value is dependent on the input signal envelope, which is quantized to a finite number of entries. (64 in this example). These entries are optimized by computing the difference between the input signal and the output of the power amplifier. Provided that feedback delay has been accounted for, the resulting difference will contain only the distortion component. A number of techniques are available to adaptively compensate for the feedback delay, operating in the time or frequency domains. Updating the RAM look-up table is accomplished using either linear convergence or the secant method.

The gain function multiplied with the modulated input signal is a complex quantity that is based on the envelope of the input signal, which is required to compensate for the AM/AM and AM/PM distortion generated by the power amplifier. The look-up table entries are derived from the error vector that remains after subtracting the input signal from the power amplifier output, and can be stored in either polar or rectangular format. The power amplifier's distortion can be expressed as a scaling and rotation of the input vector, as seen in Figure 4. This function may be stored in either polar or rectangular format.

Digital predistorter simulation example

Now we will simulate a real digital predistorter based on the complex gain look-up table technique. Figure 5 presents an example of just such as circuit.

The linear convergence technique is used to adjust the look-up table entries to minimize ACPI (adjacent channel power interference), and an adaptation coefficient value of –0.1 is selected to enable rapid optimization. A 64-entry look-up table is chosen to quantize the input envelope. A rectangular format is used for table entries. Timing clocks are used to read and write to the look-up table RAM. For our input, a 25 MHz wide, 10-tone modulated signal centered at 800 MHz is used. Finally, it is assumed that all passive components, such as power splitters and combiners, are ideal.

The error signal derived from the difference between the input and output signals is scaled by the adaptation constant and the result is latched in data registers (Figure 6). The index for the RAM is established by passing the input envelope through an A/D converter. The in-phase and quadrature signals are stored in their respective look-up tables. The fixed-point summation provides the update for the new table entry based on the previous value at the corresponding index.

The plots in Figure 7 compare the envelope of the input signal with the corresponding look-up table for gain magnitude. Thus, only a nominal amount of gain is required to compensate for the AM/AM compression that occurs due to the power amplifier.

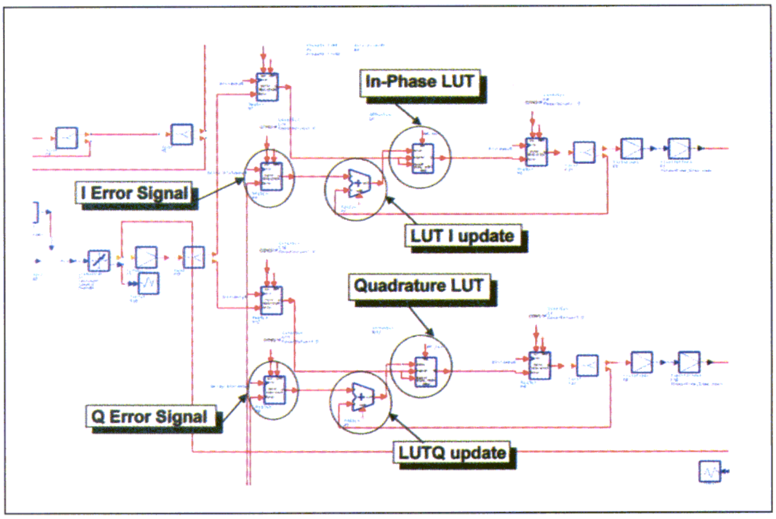
Finally, we can assess the performance of our digital predistortion circuit, as shown in the plots in Figure 8. Observe the spectral growth that occurs using a digital predistorter. Adjacent channel power is spread over a wider bandwidth, but mask requirements can be met more readily.

Conclusion

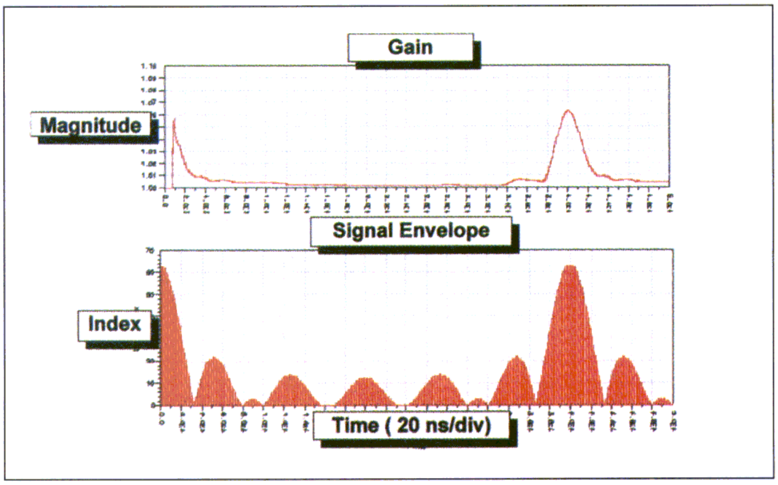
Adaptive digital predistortion is a maturing technology, now making the jump from the realm of research to system development. The digital predistorter example we have discussed in this article exhibits the expected level of performance that can be achieved using linearization. The next stage consists of system-level simulations, which would provide a solid starting point for an actual implementation, where designed components can then be integrated into a system. ■

Author information

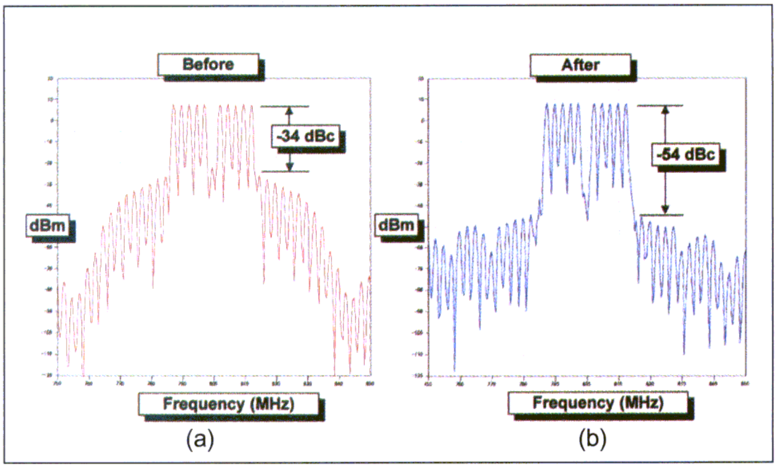
Dr. Shawn P. Stapleton has more than 17 years of experience designing RF and microwave circuits and systems. He is currently a professor of electrical engineering at Simon Fraser University and works as a consultant to the EEsof Division of Agilent. Stapleton has developed GaAs MMIC components, including mixers, amplifiers, frequency dividers and oscillators, and recently worked on projects related to digital signal processing, mobile communications and RF/microwave systems.



▲ **Figure 6. Look-up table detail for the digital predistorter.**



▲ **Figure 7. Optimized look-up table gain for the digital predistorter.**



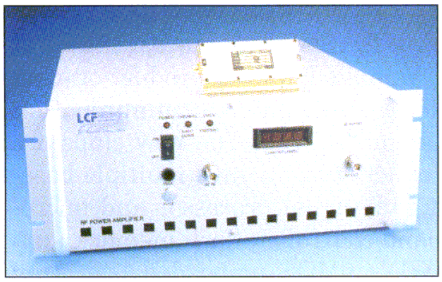
▲ **Figure 8. (a) The driving power amplifier with a 5 dB back-off generates high levels of intermodulation power as well as high levels of harmonics. (b) The output resulting from a circuit using the digital predistorter after the look-up tables are optimized.**

Products

AMPLIFIERS

RF power amplifier

LCF Enterprises has developed a 200 watt RF power amplifier that covers the frequency range from 1

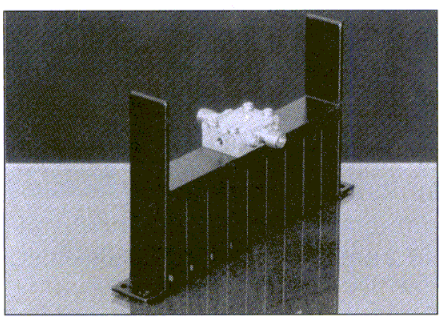


to 30 MHz and 45 dB of gain (Part # 30-1-200-45-R). Features include automatic current limiting, forced-air cooling, current meter, over-current and thermal protection.

LCF Enterprises
Circle #156

High-frequency power amplifier

MITEQ introduces a new high-frequency power amplifier in the 26.5 to 32.5 GHz range. The AMF-



8F-265325-80-29P features an output power of +30 dBm typical (+29 dBm minimum), 8.0 dB typical noise figure and 30 dB minimum gain. The amplifier is available with various electrical options and an assortment of connector combinations.

MITEQ, Inc.
Circle #157

2-watt amplifier

Microwave Solutions has released the MSD-2489701 2-watt amplifier for use in military or commercial test platform applications at 500 to 2000 MHz frequency range. This amplifier has a gain of 30 dB and has been designed with soft board technology and uses discrete hermetically sealed gallium arsenide transistors. All units include an internal voltage regulator and operate at –12 VDC and provide over voltage and reverse-polarity protection.

Microwave Solutions, Inc.
Circle #158

SEMICONDUCTORS

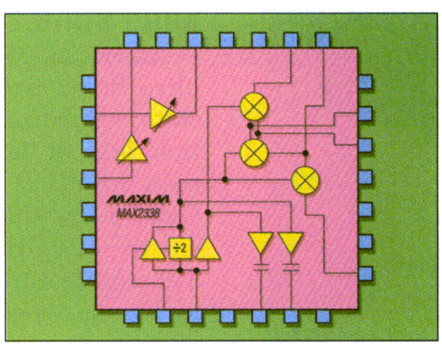
High voltage RF MOSFETs

Advanced Power Technology announces the introduction of a new line of high voltage rugged RF MOSFETs. To address the needs of semiconductor manufacturers and CO_2, laser and MRI equipment manufacturers, APT has developed a new process for making RF amplifier MOSFETs that doubles the available SOA. Typical operating voltages range between 150 and 300 volts. Minimum gain is 13 dBat 65 MHz. Prices for the ARF460A/B is $22 each in quantities of 10,000 pieces or more.

Advanced Power Technology, Inc.
Circle #159

Dual-band cellular phone receiver

Maxim Integrated Products has developed a dual-band cellular phone receiver that is designed for CDMAOne, CDMA2000, CDMA 1X

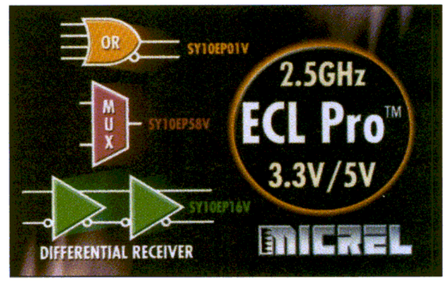


and CDMA 1XEV applications. The unit contains all functional blocks needed for a dual-band receive front end. It also contains the LO divider and dual LO buffer amplifiers to drive the transmitter. Prices start at $3.51 per unit.

Maxim Integrated Products
Circle #160

High-speed communications

Micrel Semiconductor announces the release of three new devices for high-speed communications: the SY10EP16V 3.3 volt/5 volt differential receiver; the SY10EP58V 3.3 volt/5 volt 2:1 multiplexer; and the SY10EP01V 3.3 volt/5 volt 4-input



OR/NOR gate. The new devices are offered in SOIC-8 packaging and in tiny MSOP-8 packaging. They are priced starting at $5.94 in quantities of 1,000 or more.

Micrel Semiconductor, Inc.
Circle #161

Low saturation transistors

Zetex has introduced two series (ZXT10 and ZXT11) of ultra-low saturation transistors in SOT23 and SOT23-6 packages. The transistors combine characteristics of a

matrix design with the power dissipation capabilities of the SOT23-6 package. This enables the devices to be operated with increased current ratings so that short term, high load switching transients can be more easily accommodated. Offered in operating voltages ranging from 15 to 50 volts, the series offers saturation voltages as low as 75 mV at 1A, continuous collector

current ratings of up to 4A, with peak currents to 12A, and typical gains of 300 minimum. Units are sold in 10,000 piece quantities from $0.14 to $0.17 each.
Zetex
Circle #162

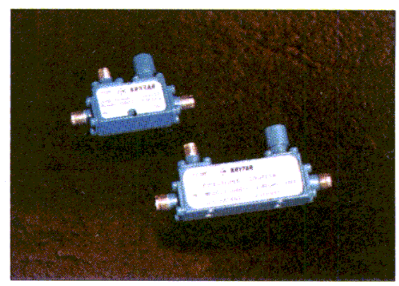
Product-on-chip

Embedded Wireless Devices announces the first members of its e8000 Valhalla family of product-on-chip solutions for wireless multi-point communication devices — the e8720e and e8715e. The new chips are complete wireless products that include multiple basebands, proprietary digital signal processor and microprocessor unit, baseband controller, flash memory, analog interface and EWD's eMOS embedded operating system. These devices handle multiple baseband protocols for a broad range of wireless applications.
Embedded Wireless Devices, Inc.
Circle #163

SIGNAL PROCESSING

Broadband directional couplers

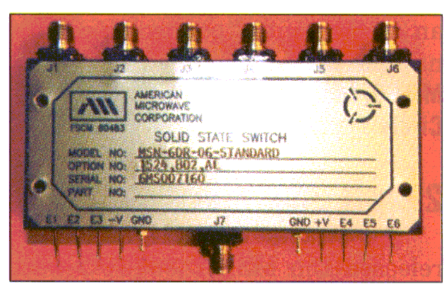
Krytar has made available the 202050010 broadband directional coupler from 0.5 to 50 GHz. Nominal coupling is 13 dB with a frequency sensitivity of around 1.2 dB. Minimum directivity is 14 dB from 1 to 20 GHz and 10 dB from 20 to 40 GHz. Maximum VSWR is 1.5:1 from 1 to 20 GHz and 1.7:1 from 20 to 40 GHz. Couplers are available with the choice of 2.4 mm or K female connectors.



Krytar
Circle #165

Solid state switch

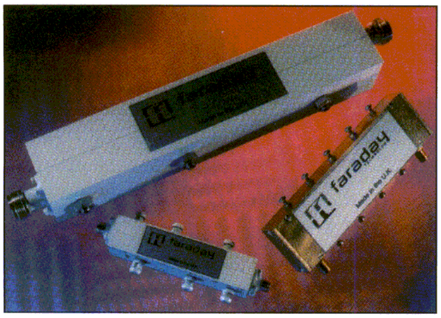
American Microwave Corporation introduces an ultra broad band, low insertion loss, reflective solid state switch that works from 3 to 24 GHz with an insertion loss of less than 1.8 dB at 3 GHz to less than 4.5 dB at 24 GHz. The unit features a switching speed of less than or equal to 45 nS ON and less than or equal to 35 nS OFF, a video transient of less than or equal to 225 mV peak to peak at 300 MHz bandwidth



and less than or equal to 95 mV peak to peak at 20 MHz bandwidth.
American Microwave Corp., Inc.
Circle #164

High frequency filters

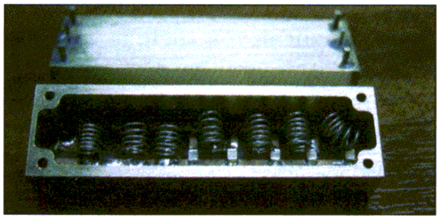
Faraday Technology has recently extended its capability to supply fil-



ters up to 18 GHz. Faraday has solutions for support on rapid filter prototyping or large volume production.
Faraday Technology, Ltd.
Circle #167

PCB mounted filters

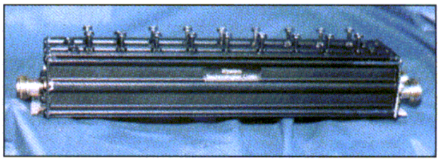
BSC Filters has expanded its range to include lumped element filter design and manufacture.



Frequencies range from below 10 MHz to 3 GHz and attenuation at 80 dB. Delivered as either a PCB, surface mounted or connecterized unit with unit size of 10 mm profile, filters are custom designed to individual specifications.
BSC Filters
Circle #166

PCS diplexer

Wireless Technologies has introduced the Model W1840D IMD-free



PCS diplexer. Standard dual +43 dBm input signals produce less than –140 dBc of IMD. Other features include an insertion loss of less than 1.2 dB, band centers RX/TX isolation greater than 90 dB, a return loss of greater than –16 dB and a power capability of greater than 50 watts.
Wireless Technologies Corp.
Circle #168

TEST EQUIPMENT

Dual range synthesizer

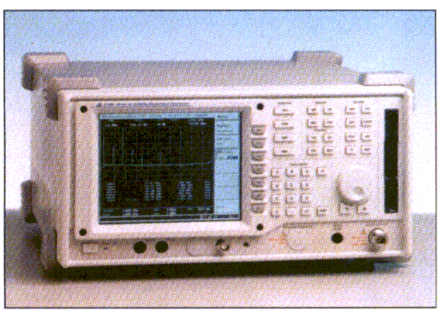
Programmed Test Sources has released a versatile dual-range frequency generator that has a broad coverage from 1 to 250 MHz and allows 5 to 20 μs switching between any two frequencies with a resolution of 1 hertz. Output is 13 dBm

and spurious outputs are –70 dB referenced to +13 dBm. The PTS 250SX-51 is available with remote control and manual controls if desired.

Programmed Test Sources, Inc.
Circle #169

Portable spectrum analyzer

IFR Systems announced the 2399 portable spectrum analyzer, aimed at a wide range of applications including mobile communica-



tion service workshops, base station installation, repair and maintenance, as well as broadcast TV and education. The unit covers a frequency range from 9 kHz to 2.9 GHz and is user-friendly. The analyzer is priced at $9,600.

IFR Systems, Inc.
Circle #170

Field strength analyzer

TEGAM has released the Model 7500 field strength analyzer, a portable instrument that combines the functionality of a graphic RF spectrum analyzer, frequency counter and tunable RF monitor in a single, hand-held instrument. This unit is meant for a wide range of RF signal analysis, detection and verification tests. Applications include mobile, PCS, cellular and cordless phones, antenna tuning, field strength, commercial radio and TV broadcasting, amateur, aeronautical, satellite and marine radio. The list price for this analyzer is $2,195 in the U.S. or $2,415 for export.

TEGAM, Inc.
Circle #171

WIRELESS SYSTEMS

ValuLine antennas

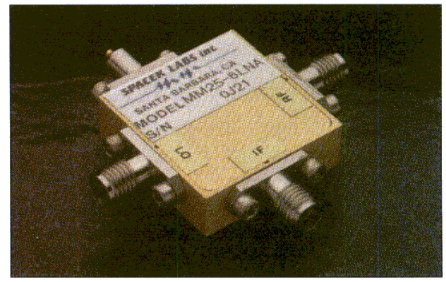
Andrew Corporation has extended its ValuLine range of terrestrial microwave antennas with the release of 1.2 meter and 1.8 meter

diameter antennas available in shielded and unshielded version in the following frequency bands: 5.925 to 6.425 GHz, 6.425 to 7.125 GHz, and 5.925 to 7.125 GHz. The 1.2 and 1.8 meter antennas can be integrated directly with the radio outdoor unit.

Andrew Corporation
Circle #172

MMIC LMDS receiver

Spacek Labs has released the MM25-6LNA integrated receiver consisting of a low noise amplifier, mixer and LO doubler-amp. The RF frequency range is 20 to 30 GHz. The LO frequency range is 10 to 15 GHz at –4 to 0 dBm, and the IF range is DC to 5 GHz. This

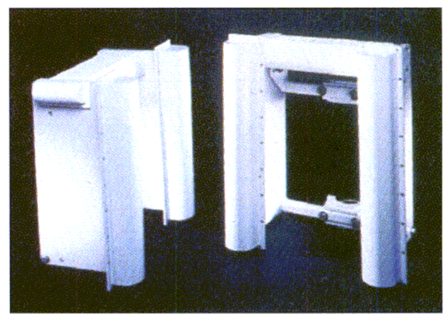


receiver can be used for narrow-band LMDS or full-band, multi-channel radiometer applications.

Spacek Labs
Circle #173

Point-to-multipoint hub

Galleon Corporation introduces a scalable, cost-effective point-to-multipoint hub for broadband wire-



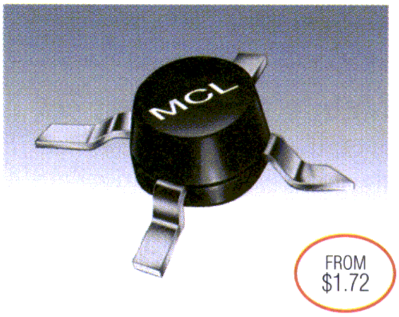
less access communications. This system starts with a single radio transceiver at the hub with coverage flexibility of 2, 4, 8 or 16 sectors.

Galleon Corporation
Circle #174

NEW PRODUCTS

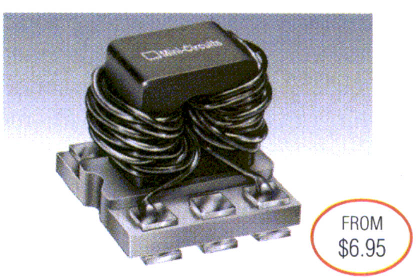
NO. 78

RF/IF MICROWAVE COMPONENTS



DC TO 3GHz MMIC AMPLIFIER HAS GOOD DYNAMIC RANGE

The DC to 3GHz ERA-33SM amplifier is part of a highly reliable family of broad band MMIC amplifiers from Mini-Circuits. Typically at 2GHz (25°C), this low cost GaAs model provides 17.4dB gain, 13.5dBm maximum power output (at 1dB comp.) and 3.9dB NF/28.5dBm IP3 for good dynamic range. The ERA series contains 9 surface mount models covering DC to 8GHz and beyond. S-parameter data, grounding, and biasing techniques available on our web site.



RF TRANSFORMERS HAVE 4:1 IMPEDANCE 10 TO 1900MHz

Broad band TCM4-19 surface mount RF transformers from Mini-Circuits operate in the 10 to 1900MHz band with 4:1 impedance ratio. Referenced to midband loss (1.0dB typ), insertion loss is 1dB from 30MHz to 700MHz, 2dB in the 20 to 1000MHz range, and 3dB band wide when operated within -20°C to +85°C (max.). Open case design has plastic base with solder plated leads, and applications include cellular and PCS. Maximum RF power is 250mW.

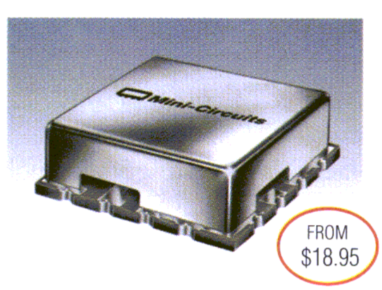
FEATURED PRODUCT

HIGH IP3 MIXER DELIVERS ON COST AND PERFORMANCE

Mini-Circuits has announced a breakthrough low cost high IP3 frequency mixer for the 1819 to 1910MHz (RF) frequency band. Typically, the HUD-19SH boasts very high 38dBm IP3 with extremely high 38dB L-R, 36dB L-I isolation and conversion loss of 7.5dB. This level 19 mixer requires no DC biasing, and the noise figure displayed is the same as the conversion loss. The result is lower overall system noise figure. Ideal for PCS and communications systems.

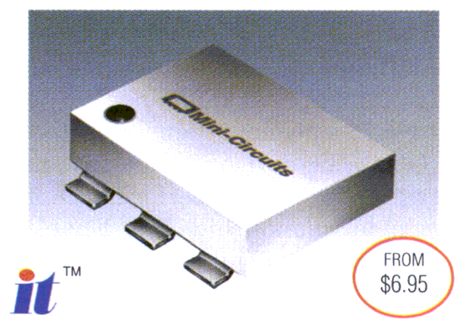
2165 TO 2650MHz VCO EXHIBITS LOW PHASE NOISE

Mini-Circuits new ROS-2650 voltage controlled oscillator features 2165MHz to 2650MHz broad band tuning from a miniature 0.5"x0.5"x0.18" industry standard surface mount package. The VCO targets ISM applications delivering low -101dBc/Hz SSB phase noise typical at 10kHz offset, good 27-36MHz/V typical tuning sensitivity, and 0.5 to 19 minimum to maximum tuning voltage.



750 TO 1300MHz ELECTRONIC LINE STRETCHER SIMPLIFIES VCO TEST

Mini-Circuits has added model ELS-1300 to their family of 110 to 1300MHz electronic line stretchers. This novel unit automates VCO load-pull measurements by providing electronically adjustable phase range of the signal to better than 360 degrees across the entire 750 to 1300MHz band, and is voltage controlled (1-25V) for automated applications. The 3 port device features return loss of 10 to 12dB typical and input power is 10dBm maximum. Patent pending.



BROADBAND 6.3dB COUPLERS COVER 200 TO 1300MHz

Mini-Circuits announces off-the-shelf availability of their new 200 to 1300MHz ADC-6-13 directional couplers. These broad band 50 ohm units provide a nominal 6.3dB±0.5dB coupling value with good ±0.9dB (typ) flatness for cellular and VHF/UHF receiver applications. Housed in a miniature 0.112" (max. height) water washable package, these 1W couplers typically exhibit low 1.8dB mainline loss, good 17dB directivity, and excellent 1.3:1 VSWR. Value priced.

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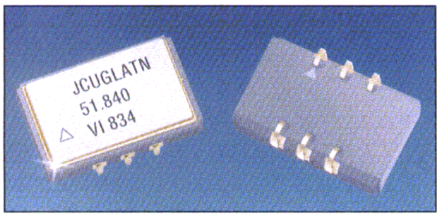
Products introduced at the 2001 Wireless Symposium

Low jitter VCXOs

Vectron International has introduced a series of new low jitter VCXOs that are available at frequencies from 1.02 to 170 MHz. Other features include +3.3 or +5.0 volt options, 14 mm × 9mm J-type package and CMOS or PECL outputs. Typical jitter performance



is less than 0.5 ps rms (12k Hz to 20 MHz) at the output frequency for the CMOS version and less than 1 ps rms for the PECL option.

Vectron International
Circle #175

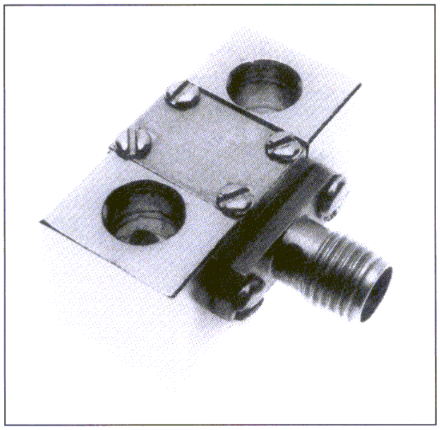
Monolithic chip

Analog Devices has developed the AD8302 monolithic chip, designed to accurately measure gain and phase difference between two independent signals up to 2.7 GHz. The device provides the user with system-level performance in a single monolithic IC. Enabling designers to build into their systems diagnostic capabilities that can monitor performance and diagnose signal purity from within the system, this chip is expected to produce a wide range of applications outside of the cellular basestation market. Prices start at $15.50 in quantities of 1,000.

Analog Devices
Circle #176

25-watt load

Bird Component Products has released the Model 25-CT-FA 25-watt conduction cooled load for volume production applications. This piece is compact, with a 50-ohm load and a frequency range and

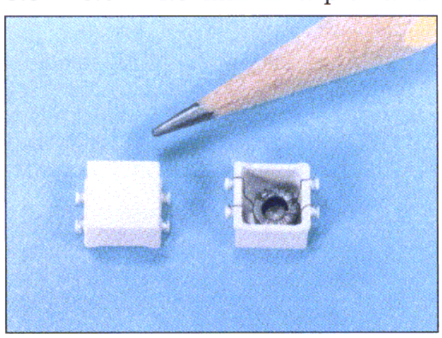


VSWR of DC to 1 GHz at 1.15:1 maximum and 1 to 3 GHz at 1.25:1 maximum. SMA female or male connectors are standard, but customer-specified semi-rigid or flexible cable connections are also available. The 1 to 10 piece price is $70 each.

Bird Component Products
Circle #177

SMT toroidal-core compensated chokes

Sprague-Goodman Electronics has added a new size surface mount toroidal-core transformer to the two sizes already offered in their catalog. The new models measure 8.5 × 6.0 × 4.5 mm in a pick-and-



placeable liquid crystal polymer plastic housing. The series is characterized as current compensated chokes with inductance values from 11 to 4700 μH. The chokes operate from –40 to 125 degrees C. Pricing in quantities of 5,000 for models with inductance values up to 470 μH is $1.43 each; higher inductance value models are $1.53 each.

Sprague-Goodman Electronics
Circle #178

Mobile station test set

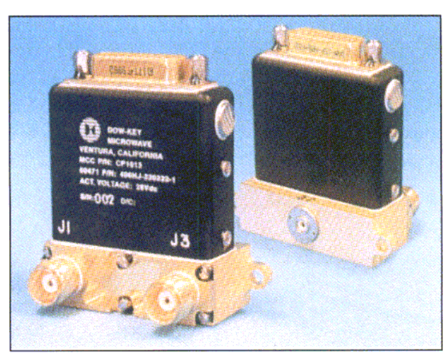
Agilent Technologies has introduced the E6393A mobile station test set for network equipment manufacturers, network operators, and service centers that test and repair faulty mobile phones. The set is upgradeable for cdma2000 capabilities and provides repair and

calibration tools to troubleshoot faults, adjust mobile phones to manufacturers' specifications and help service technicians cope with the growing number of mobile phones coming into repair centers. The mobile station set is available for $19,100.

Agilent Technologies, Inc.
Circle #179

SPDT switch

Dow-Key Microwave introduces the 406HJ-330332-1 high-power



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WJ High Dynamic Range Amplifiers

Product	Frequency (MHz)	IP3 (dBm, typ.)	P1dB (dBm, typ.)	NF (dB, typ.)	Bias current (mA, typ.)
AH1	250-3000	41	21	2.9	150
NEW AH2	50-860	38	20	2.9	150
NEW AM1	250-3000	36	18	2.7	75

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Circle 60

SPDT pulse latching switch that addresses insertion loss, weight and size limitation issues. This switch features an input connector that can be mated directly with a 50-ohm glass feed through and can provide a direct connection between the switch and the hermetically sealed filter assembly. Other specifications include a frequency range of .240 to .320 GHz, a VSWR of 1.20:1, an insertion loss of .10 dB and an isolation of 70 dB.

Dow-Key Microwave Corp.
Circle #180

GaAs HBT linear variable gain amplifiers

RF Micro Devices introduces three new GaAs HBT linear variable gain amplifiers: the RF2376, RF2377 and RF2381. Operating from a single 2.7 to 3.3 volt power supply, all three amplifiers feature a 50 dB linear gain control range, high linearity, high gain and a low noise figure.

RF Micro Devices, Inc.
Circle #181

Hand-held spectrum analyzer

Anritsu Company introduces the MS2711A fully functional handheld spectrum analyzer that provides field engineers and technicians with unprecedented measurement flexibility in field environments and applications requiring mobility. Featuring a rugged, lightweight, battery-operated design, the analyzer enables users to locate, identify, record and solve communication system problems quickly without sacrificing measurement accuracy. The analyzer is priced at $7,000.

Anritsu Company
Circle #182

SiGe front end

Atmel Wireless & Microcontrollers has released a new integrated front end for family radio and remote control applications in the frequency range of 400 to 500 MHz. The T0980 is manufactured in SiGe technology and provides increased efficiency, small size and improved functionality for smaller, high-performance end products with extended operating time. Consisting of a low-noise amplifier and power amplifier that can be switched off to save power, the front end has an LNA with a gain of 19 dB. Pricing starts at $1.80 each in 10,000-piece quantities.

Atmel Wireless & Microcontrollers
Circle #183

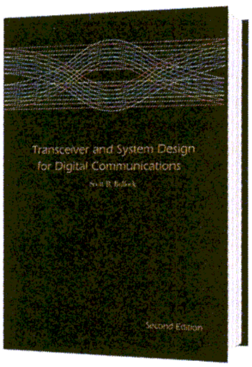
Reverse thread TNC

RF Connectors introduces its new reverse thread TNC. This piece achieves compliance in three ways: reverse, or left-handed threads; reverse polarity, or gender; and the use of metric threads rather than the unified standard. The 50-ohm connector is designed for use with RG-8/X or LMR-240 low-loss cable and has a nickel-plated, machined brass body, Teflon insulation and gold-plated contact.

RF Connectors
Circle #184

Some things were just meant to be over your head . . .

. . . but these won't be !



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This second edition on spread spectrum transceiver design offers updated coverage of the latest advancements involving broadband and home networking systems. **New coverage includes**: power line communications (PLC), phone line network alliance (PNA), and Bluetooth.

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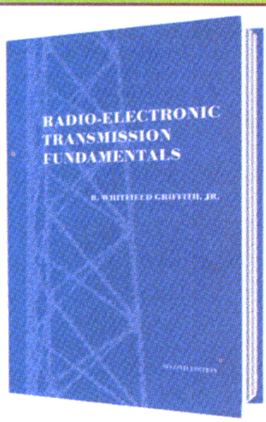

Radio-Electronic Transmission Fundamentals by B. Whitfield Griffith, Jr.

This classic volume, one of the best textbooks of electromagnetic concepts and RF circuits, is reprinted here for the first time since its original publication in 1962. Hailed for its clear and concise explanation of antenna, transmission lines and RF networks from the perspective of electromagnetic field theory, this edition is highly recommended for graduate students and engineers who want the most understandable presentation of radio concepts. The straightforward discussion of the underlying principles, concepts and components remains timeless. Topics include:

- Electrical Networks
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Circle 39

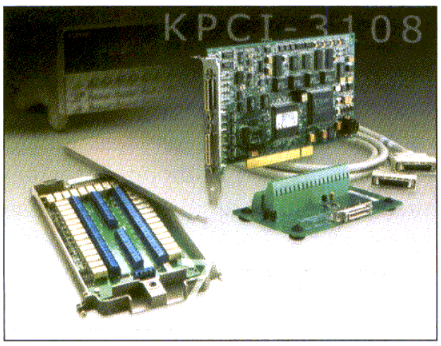
SMB coaxial loads

MECA offers the 468-2 efficient, low power 50-ohm load that is designed for cost-effective WLAN and DCS/PCS/3G applications using SMB connectors for space-limited applications or when rapid connection is desired. The termination is an SMB plug termination that is suited for low power applications that dissipate 1 watt average power. Maximum VSWR is 1.10:1 to 1 GHz, 1.15:1 to 5 GHz and 1.25:1 to 6 GHz.

MECA Electronics, Inc.
Circle #185

Improved performance

Keithley Instruments has introduced two new systems designed to improve measurement performance

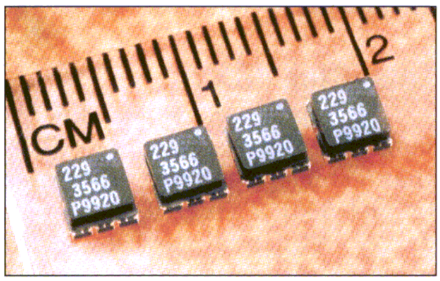


while reducing the cost of environmental stress screening in product development, QA/QC and production applications: Models 7708 and 2700. Both systems provide a low per-channel cost for the acquisition of thermocouple temperature data with automatic cold junction compensation. Both also come with ready-to-use software for easy system configuration, data logging and analysis.

Keithley Instruments, Inc.
Circle #186

W-CDMA power amplifier

Celeritek has announced the availability of a new W-CDMA power amplifier for wireless data transmission at 3.5 GHz. The



CMM3566-LC is intended for use in subscriber units and base stations that operate in the 3.45 to 3.50 GHz frequency range. Typical features include operation at 7.0 volts, 30 dB gain at operating output and +24 dBm linear output power.

Celeritek, Inc.
Circle #187

Power MOSFETs

Siliconix, a subsidiary of Vishay Intertechnology, has released a pair of power MOSFETs (Models Si4894DY and Si4404DY) aimed at core voltage DC–DC converters in high-performance notebook computers. The new devices will allow designers to improve the efficiency of notebook power conversion designs.

Siliconix
Circle #188

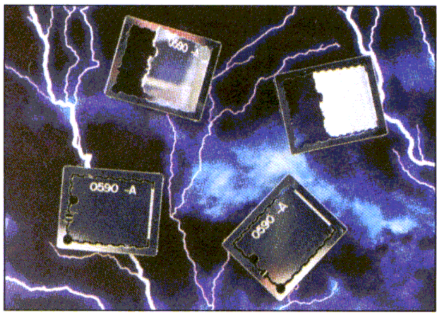
Commercial PLL

Vari-L Company has released the PLL400-915 PLL that generates frequencies from 902 to 928 MHz, with a 200 kHz step size. The unit typically requires 18.5 mA of current from a 5.0 volt supply voltage. Typical phase noise at 100 kHz offset is –131 dBc/Hz, and the typical phase noise at 100 kHz offset is –131 dBc/Hz. Phase detector spurious suppression is typically –82 dBc. Typical output power is 3.7 dBm, and second harmonic suppression is typically –13.3 dBc, and third harmonic suppression is typically –27 dBc.

Vari-L Company, Inc.
Circle #189

RF shielding cans

Photofabrication Engineering has released a photochemically etched RF shielding can with a patented, easy access removable cover. This design allows users to replace or repair internal components without damaging the surrounding shield. An etched line on



the surface is broken to minimize RF leakage. The tab connecting the lid to the RF shield can be snipped and resoldered quickly.

Photofabrication Engineering, Inc.
Circle #190

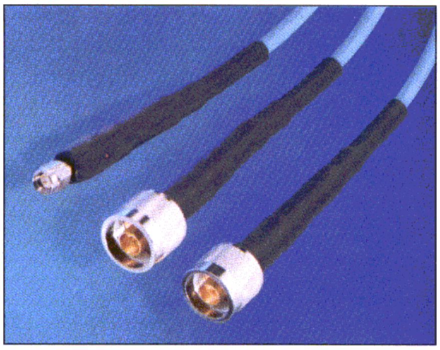
RF power transistor

UltraRF has released its UPF18060 discrete RF power transistor, designed for reliable performance. The transistor has a frequency band of 1.805 to 1.880 and is rated for a minimum 60-watt output power. It is ideally suited for CDMA, TDMA, GSM and multicarrier power amplifiers in Class A or Class AB operation.

UltraRF
Circle #191

Coax test cables

Tyco Electronics announces the availability of TESTLINE 18 coaxial cables that provide cost-effective capability without compromising electrical performance characteristics. Available in SMA to SMA, SMA to type N and type N to type N configurations, the cables can be ordered in .5, 1, 2 and 3 meter

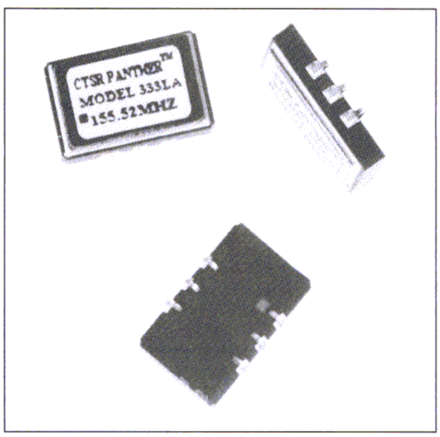


lengths. Shielding effectiveness is greater than 100 dB, and propagation velocity is rated at 76 percent C. The operating temperature range is –55 to +125 degrees C.

Tyco Electronics Corporation
Circle #192

Low jitter VCXOs

CTS announces a new line of miniature surface mount VCXOs that exhibit low jitter properties (Model 333). Packaged in a standard 9 × 14 mm hermetically sealed



SMT ceramic package, the new VCXO operates on +3.3 VDC or +5 VDC. All oscillators are available in frequencies including 77.76 MHz and 155.52 MHz.

CTS Corporation
Circle #193

MMIC power amplifier

Fujitsu Compound Semiconductor announces the expansion of its GaAs MMIC power amplifier product line with Models FMM5803X, FMM5805X, FMM5806X and FMM5807X. All are millimeter-wave, high power MMIC amplifier products covering the 17.5 to 31.5 GHz frequency band with output power ranging between 26 to 31 dBm. These new devices are designed for point-to-point or point-to-multipoint radiolink and LMDS applications.

Fujitsu Compound Semiconductor, Inc.
Circle #194

Triaxial connectors

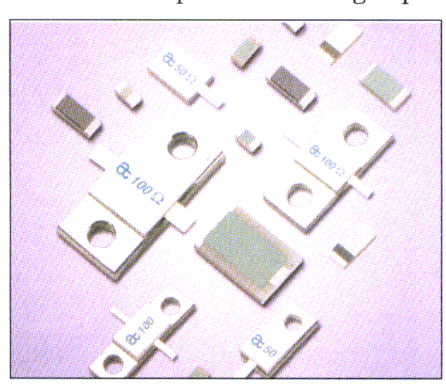
Tru-Connector has released a full line of triaxial connectors that are designed for applications requiring the removal of noise radiation or cross-talk. These connectors are designed for triaxial cables that are essentially coaxial

cables with an added outer shield that lowers groundloop interference and eliminates radiated noise or cross-talk. Prices start at $14.95, depending on style, size and quantity.

Tru-Connector Corp.
Circle #195

Resistors, terminations and attenuators

American Technical Ceramics introduces a new line of resistors, terminations and attenuators designed for microwave and RF applications. The product line can be used in applications from DC to 40 GHz with power handling capa-



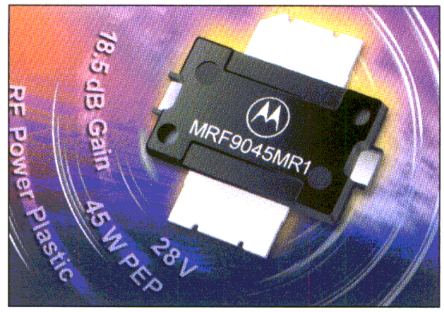
bilities from .1 to 1000 watts. All components exhibit low VSWR and capacitance, are non-magnetic and contain no lead.

American Technical Ceramics
Circle #196

Coaxial surge protectors

CITEL offers a new line of coaxial surge protectors for wireless systems, microwaves, radios, antennas, GPS and satellite systems. They provide protection against lightning surges and electrical transients. Various connectors are offered, including N, BNC, TNC, SMA and 7/16. The protectors are available for any power from 5 to 1000 watts and any frequency from DC up to 4 GHz.

CITEL, Inc.
Circle #197



RF LDMOS transistors

Motorola introduces a new family of RF power plastic LDMOS devices optimized for 1.0 GHz base station applications. MRF9045MR1 is the first product in the family to be unveiled. Capable of handling up to 45 watts PEP, the transistor is a single-ended RF power device operating at 28 volts. The device provides efficiency and linearity performance for linear base station applications up to 1 GHz.

Motorola, Inc.
Circle #198

Linear amplifier

RF Micro Devices introduces the RF2186 high-power, high-efficiency linear amplifier for 3V handheld systems. It is designed for use as the final RF amplifier in handsets, spread spectrum systems and other applications in the 1920 to 1980 MHz band. The amplifier is also compatible with existing second generation (2G) and next generation (2.5G) systems. Operating from a single 3-volt power supply, the device delivers 27 dBm linear output power, 31 dB linear gain and 35 percent linear efficiency.

RF Micro Devices, Inc.
Circle #199

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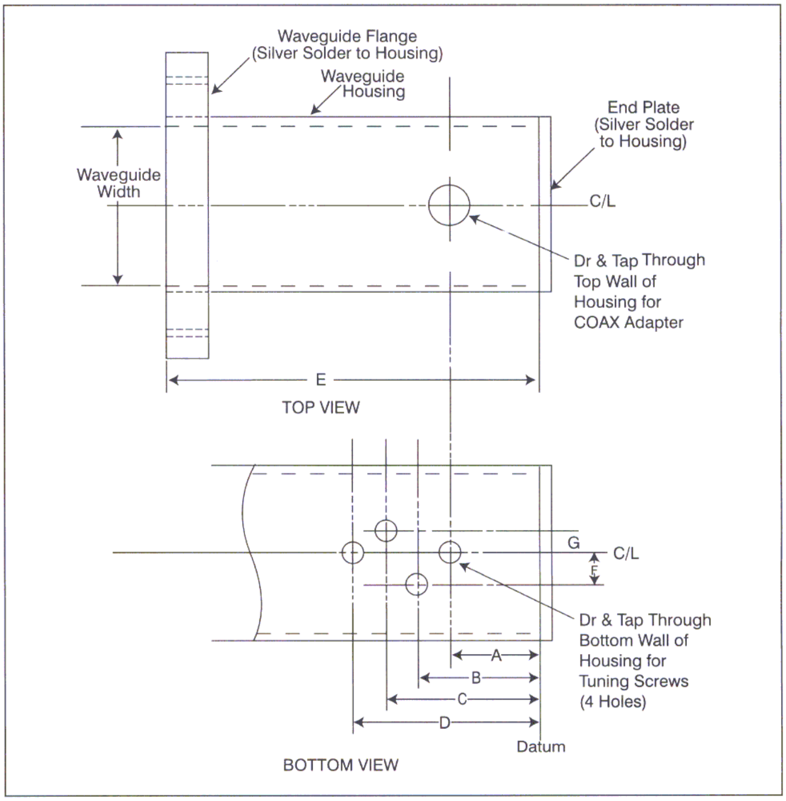
Coax-to-Waveguide Adapters Meet Needs of Communications Equipment

By Richard M. Kurzrok, PE
RMK Consultants

Coax to waveguide adapters without tuning screws were originally developed for full waveguide bandwidths with maximum VSWR specifications of 1.25. Subsequently, adapters with factory-adjusted tuning screws provided full waveguide band coverage with lower VSWR. For some satellite and terrestrial communications equipment, the need for coax-to-waveguide adapters involves smaller usable bandwidths of 500 to 800 MHz with a maximum VSWR of about 1.10. This is usually achievable via the factory-tuned configuration. This article presents approximate information for the development of coax to waveguide adapters suitable for many applications. Note that different approaches will be needed for high-power applications.



▲ **Figure 1. Coax-to-waveguide adapter design details.**

Coax-to-waveguide adapter configuration

The coax-to-waveguide adapters have several significant parts:

- Housing using rectangular waveguide tubing
- Standard waveguide flange
- Metallic end plate
- Coaxial probe assembly with modified coaxial adapter, probe and hardware
- Tuning screws with hardware (except for set screws)

The coax-to-waveguide adapter design details are shown in Figure 1. The coaxial probe assembly for SMA connectors is also used in waveguide bandpass filters with coaxial interfaces [1]. Applicable details will not be repeated here. New probes must be designed when using SSMA connectors. The top broad wall of the waveguide housing is tapped for a modified female to female coaxial adapter. The bottom broad wall of the waveguide contains four tapped holes for metallic tuning screws. For small quantity fabrication, most metallic parts are of copper alloys. Stainless steel hardware is preferred for durability.

Coax-to-waveguide adapter mechanical dimensions

Typical mechanical dimensions for various rectangular waveguide sizes are shown in Table 1. Tuning screw sizes and recommended coaxial connectors are shown in Table 2. For the larger size waveguides, bulkhead feedthrough coaxial adapters are needed to increase the adjustment ranges. These preliminary dimensions might require modifications when design objectives include optimization.

Coax-to-waveguide adapter alignment

Coax-to-waveguide adapter alignment uses a standard coaxial reflectometer of adequate directivity. The swept frequency input is applied to the coaxial port. The waveguide port should be terminated in a load with a VSWR that is less than 1.02. Different size probes and different probe insertion depths provide a cut and try situation when used with the tuning screw array. Typical probe depths are near one half the waveguide height while typical tuning screw penetrations are about

one quarter waveguide height. Sometimes, one or more of the tuning screws are not used. After alignment is completed, the coaxial probe assembly and the tuning screws can be staked to the housing with epoxy.

WG Size	A	B	C	D	E	F	G
WR-229	0.625	0.812	1.000	1.312	2.500	0.438	0.312
WR-137	0.375	0.500	0.625	0.812	2.000	0.281	0.187
WR-112	0.312	0.406	0.500	0.656	1.500	0.219	0.156
WR-90	0.250	0.340	0.475	0.610	1.250	0.187	0.125
WR-75	0.250	0.312	0.375	0.500	1.125	0.187	0.125
WR-62	0.187	0.250	0.312	0.437	1.000	0.187	0.125
WR-51	0.156	0.219	0.281	0.406	1.000	0.156	0.093
WR-42	0.125	0.187	0.250	0.375	1.000	0.125	0.093

Note: For E, tolerances are plus or minus 0.005. All other tolerances are plus or minus 0.002.

▲ **Table 1: Coax to waveguide adapter mechanical dimensions.**

Conclusions

Coax-to-waveguide adapters for low power communications applications can be developed readily using these recommended dimensions as a starting point. There are useful technological overlaps with some waveguide bandpass filters. For manufacturing in substantial quantities, other fabrication methods, such as dip brazed aluminum, aluminum casting, or electroforming can be used. ■

Waveguide Size	Tuning Screws	Coaxial Connectors
WR-229	#10-32	SMA Bulkhead Feedthrough
WR-137	#8-32	SMA Bulkhead Feedthrough
WR-112	#6-32	SMA
WR-90	#4-40	SMA
WR-75	#2-56	SMA
WR-62	#2-56	SMA
WR-51	#1-72	SSMA
WR-42	#0-80	SSMA

Note: SMA adapters have 0.250 to 36 UNS-2A threads, and SSMA adapters have 0.190 to 36 UNS-2A threads.

▲ **Table 2: Coax to waveguide adapter tuning screws and coaxial connectors.**

Acknowledgements

The type of coax-to-waveguide adapter discussed in this article has been developed independently by other microwave engineers. It has also been integrated into various waveguide subassemblies.

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Author Information

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MBA-15L	+4	1.2-2.4	6.95	MBA-25LH	+10	2.2-3.6	6.95
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Amplitude Equalizer Uses Reactively Loaded T-Pad

By Richard M. Kurzrok, PE
RMK Consultants

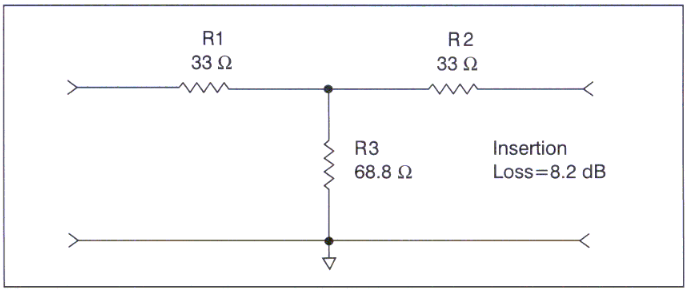
For a small range of amplitude slope, a reactively loaded T-pad attenuator can provide continuous equalization at intermediate frequencies. The circuit can be adjusted for both positive and negative slopes. Return losses are usable and adjustments are quite simple. The equalizer is implemented using low-cost components.

Equalizer circuit description

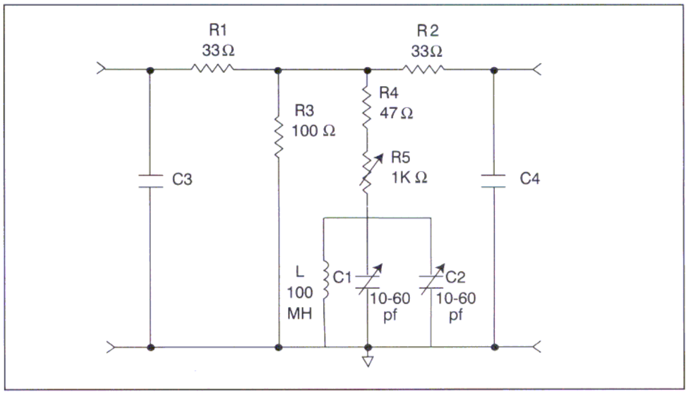
The basic circuit for a 75-ohm T-pad attenuator is shown in Figure 1. By replacing the grounded shunt resistor with a larger resistor in parallel with a lossy resonant circuit, the adjustable amplitude circuit of Figure 2 is obtained. The potentiometer R5 and the variable capacitors C1 and C2 are adjusted for the desired equalization slope around the 70 MHz center frequency.

In the frequency range of 70 MHz plus or minus 18 MHz, the amplitude equalizer is assembled using a singly clad epoxy glass PC board that is 1/16 inch thick. The PC board is mounted on four 1/4-inch diameter × 4-40 male/female standoffs that are fastened to a Bud CU-124 die-cast aluminum box. To adjust R5, C1 and C2 for desired amplitude slope, the cover of the enclosure must be temporarily removed.

The fixed resistors are 1/4-watt composition with 5 percent tolerance. The potentiometer is a single turn Cermet unit. The fixed inductor is a four turn 1/4-inch diameter air core solenoid (mounted parallel to the PC board) using number 18 AWG magnet wire. The capacitors C1 and C2 are adjustable ceramic units with less than one turn of adjustment range. The optional fixed capacitors C3 and C4 are used to improve the return loss of the BNC female connectors and associated wire leads to the PC board.



▲ **Figure 1. T-pad attenuator schematic.**



▲ **Figure 2. Adjustable amplitude equalizer schematic.**

Equalizer performance

The equalizer unit is capable of providing ±2-1/2 dB amplitude slope over the 36 MHz IF bandwidth, centered at 70 MHz. The variable components adjust the magnitude and direction of the amplitude slope. Absolute equalizer insertion losses are typically 6 to 10 dB. Return losses, within the rated range of adjustment, are in excess of 20 dB when the input and output BNC connectors are properly matched. This matching might not be necessary if the 50-ohm BNC connectors are replaced by 75-ohm BNC connectors.

Conclusion

A relatively simple adjustable amplitude equalizer can be realized providing a modest range of amplitude slopes and usable return loss. Equalizer adjustments are continuous rather than discrete. Low cost components are used throughout. ■

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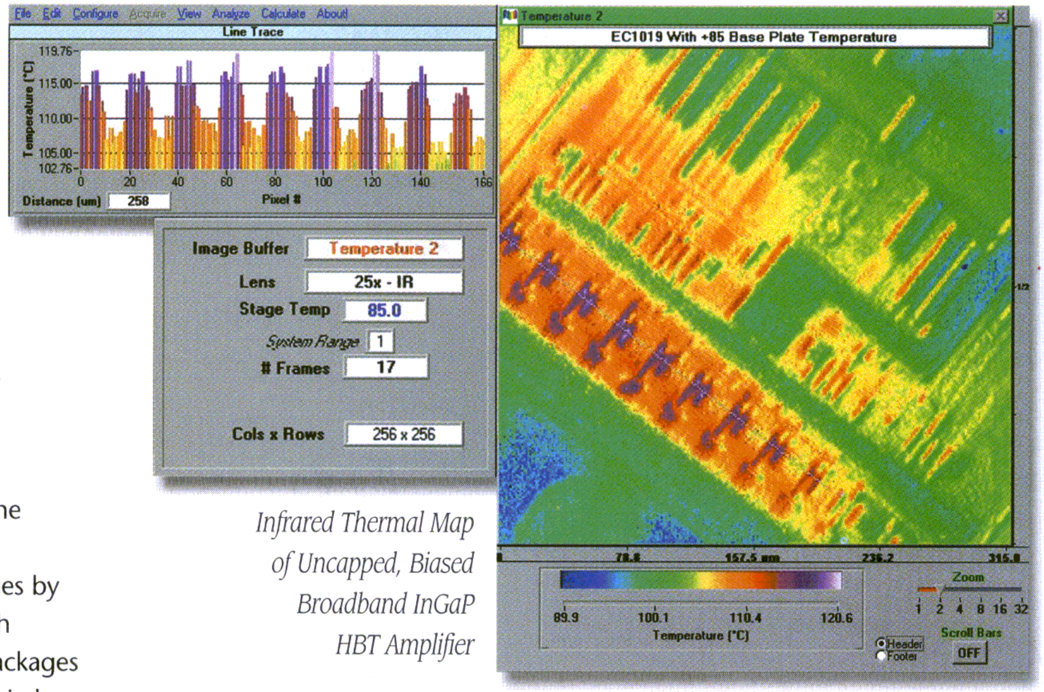

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ECG006	15dB	15dBm	30dBm	278° C/W	50°C	DC-6 GHz
ECG003	20dB	23dBm	39dBm	50° C/W	45°C	DC-3 GHz
ECG008	15dB	23dBm	40dBm	55° C/W	55°C	DC-3 GHz
ECG009	19dB	24dBm	41dBm	85° C/W	65°C	DC-2 GHz
ECG011	20dB	8dBm	20dBm	355° C/W	47°C	DC-6 GHz
ECG012	14dB	20dBm	36dBm	120° C/W	45°C	DC-2.5 GHz
EC-1089	15dB	23.5dBm	>42dBm	~85°C/W	~65°C	DC-2.5 GHz
EC-1019	18.5dB	19dBm	34dBm	120°C/W	40°C	DC - 3 GHz
EC-1078	19.5dB	21dBm	37dBm	120°C/W	60°C	DC - 3 GHz
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Frequency Properties of a Reverse Biased Thick Switching PIN Diode

A review of PIN diode behavior in a large-signal environment

By Lioudmila Drozdovskaia
Villanova University

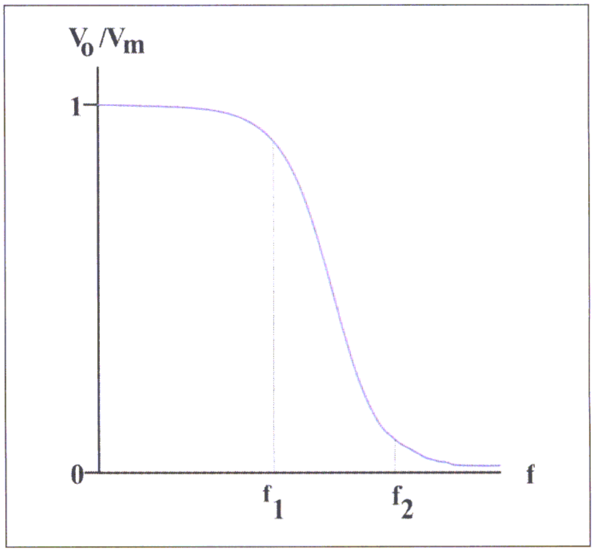
In modern wireless RF and microwave communications systems, PIN diodes and several types of transistors are used as controlled components for different control devices such as switches, attenuators, and phase-shifters [1]. In many cases, Si or GaAs PIN diodes are the devices of choice because of their good power-handling capability. Forward-biased PIN diodes exhibit low impedance; reverse biased ones exhibit high impedance and provide good bandwidth performance.

Although PIN diodes have been used widely in different applications for the last 40 years [2, 3], many of their properties are not well understood. One of these less understood properties is the reverse-biased state in a large-signal environment. The reverse-bias voltage is used to keep a high-impedance PIN diode in a low-distortion state. In this case, the PIN diode has a large RF or microwave voltage and a DC reverse-bias voltage applied to it. This DC voltage prevents any detection effects and, therefore, reduces the harmonic distortions produced by the PIN diode, especially in an RF or microwave high-power regime [2–7]. It is very important to determine correctly the required reverse DC bias voltage.

This article discusses experimental data and theoretical results of the required DC bias voltage estimated by different mathematical and physical models.

PIN diode under DC reverse-bias voltage

Experimental data [2, 3, 5, 8] has shown that the applied DC reverse-bias voltage depends on the frequency of the input RF or microwave power, as shown in Figure 1. The first area ($f < f_1$) is where the minimum required DC reverse bias voltage V_0 is the peak RF voltage across the diode V_m: $V_0 \approx V_m$. In the second area ($f > f_2$), the minimum required DC reverse bias voltage V_0 may be much smaller compared with the RF or microwave peak across the diode: $V_0 << V_m$.



▲ **Figure 1. Minimum Pin diode DC bias voltage versus frequency.**

The low-frequency area ($f < f_1$) is based on the properties of the classical detector [2, 3, 6], because the switching PIN diode behavior is similar to it. However, there are different suggestions regarding the second area ($f > f_2$). The first and simplest explanation is based on an i-region transit time of electrons and holes [2, 3, 6]. This approach may explain the reduction of the reverse DC bias voltage with the frequency increase for the short PIN diodes when the i-layer is less than the depleted region in the non-biased state (above the punch-through).

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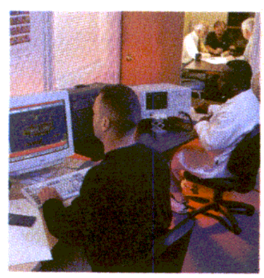

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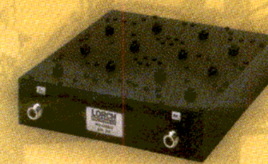

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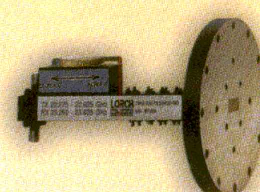

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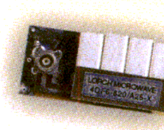

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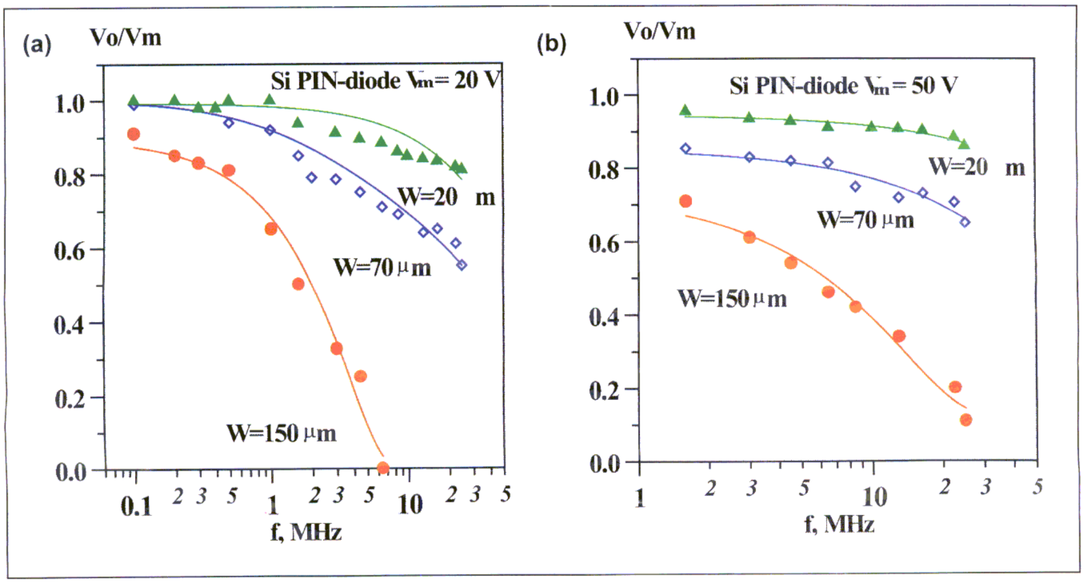


▲ **Figure 2. Required DC reverse bias voltage versus frequency for PIN diodes with different i-layer thickness. (a) V_m = 20 V; (b) V_m = 50 V.**

However, for the thick PIN diodes, where the i-layer is longer than the depleted region, this explanation is not sufficient [7, 9].

Another explanation [7, 9] is based on the multilayer impedance model [10, 11]. This model takes into account the fact that the i-region of the PIN diode below the punch-through consists of undepleted and depleted regions (for details of this model, see the Appendix). Each region can be represented [10, 11] as the shunt configuration of resistance and capacitance. The properties of the frequency-dependent impedance divider formed by these regions should be taken into account to explain the characteristic shown in Figure 1.

The depleted region resistance of the unbiased thick PIN diode is much greater than the resistance of the undepleted region. When the reverse DC voltage is applied to the PIN diode, most of it drops across the depleted region. At the same time, applied AC voltage is divided between two regions according to the properties of the frequency-dependent impedance divider [10]. At low frequencies ($f < f_1$), divider behavior is defined mainly by the depleted region resistance and the AC voltage is across the depleted region. At the high frequencies ($f > f_2$), its behavior is determined by the capacitances of both regions.

In the case of the thick PIN diode, the depleted region capacitance is greater than the undepleted one and the AC voltage is across the undepleted region at high frequencies. AC voltage applied to the depleted region causes detection effects. To prevent it, the DC reverse bias voltage is necessary. The required DC voltage value should be at least equal to the amplitude of the AC voltage applied to the depleted region. But the AC voltage across the depleted region decreases with frequency, so the required DC reverse bias voltage also drops with frequency (Figure 1). The transit time effect provides the full absence of any detection effects in PIN diode with the further frequency increase.

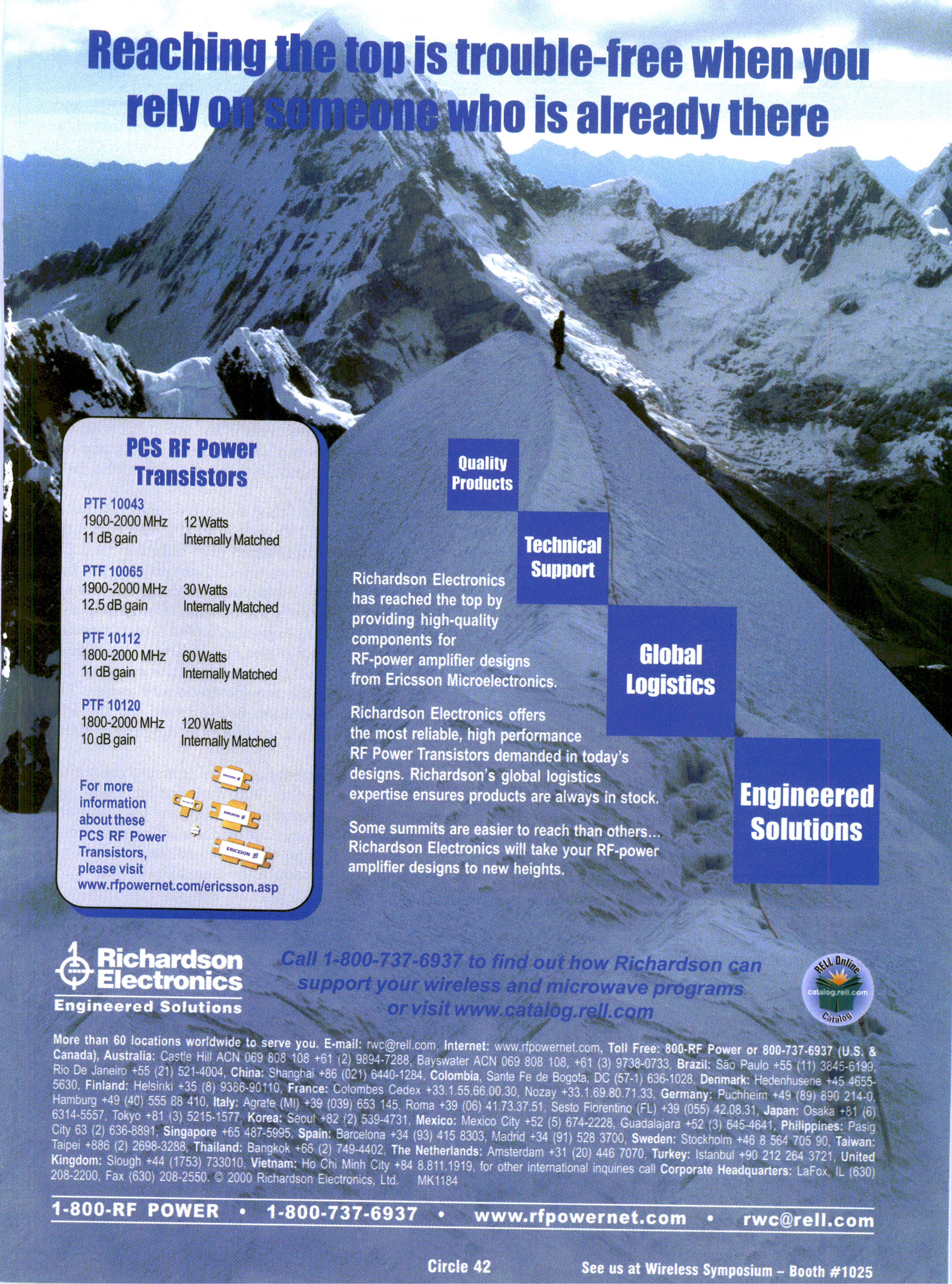
A program based on the isothermal drift-diffusion model of a semiconductor structure was also implemented for simulation and determination of the required DC reverse bias voltage [12]. This program numerically solves one-dimensional continuity equations for electrons and holes, the Poisson equation and the total current equation.

Experimental measurements

The measured required DC reverse bias voltage data were taken using the experimental setup described in [8]. The input RF signal was generated with a power CW transponder that was allowed to reach the amplitude of several tens of volts in the waveband of several megahertz. As a device under tests, the shunting different PIN diodes with the return path were used. The DC current rectified by the diode was monitored as a measure of merit of the diode nonlinearity. The current value of 10 mA was chosen as the sufficient experimental parameter of the low nonlinearity and low harmonic distortion when a reverse bias voltage was applied to the PIN diode.

The silicon PIN diodes used in the experimental measurements had different i-layer thicknesses: W = 20 mm, 70 mm and 150 mm. Figure 2 illustrates the measured minimum necessary DC reverse-biased voltage (V_o), keeping PIN diodes in the low nonlinearity and low harmonic distortion state against frequency. The measurements were performed for all types of diodes, with the RF amplitude across the diode (V_m) of 20 V [Figure 2(a)] and 50 V [Figure 2(b)]. The shapes of the curves are in accordance with the experimental curve shown in [5] for the thick PIN diode and theoretical discussions in [6–9]. Figure 2 shows that when the i-layer thickness increases, the area with the diminished required DC reverse bias voltage shifts to the lower frequencies.

Figure 3 shows the required DC reverse-bias voltage against frequency with a different input RF voltage amplitude. The silicon PIN diode with the i-layer thick-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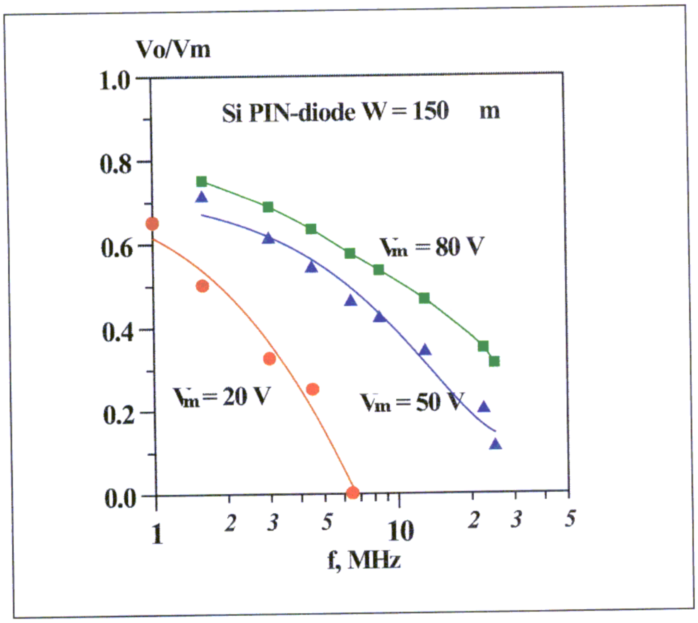
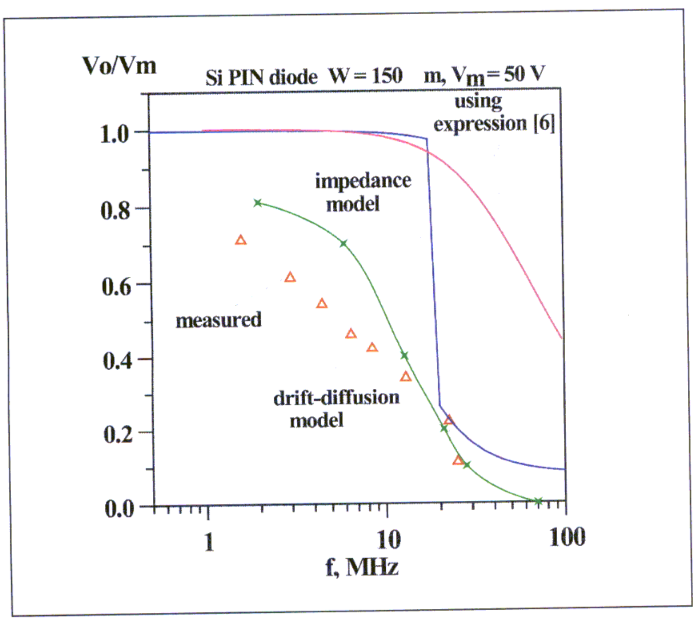


▲ **Figure 3. Required DC reverse-bias voltage versus frequency of a 150 μm PIN diode with a different input RF voltage amplitude.**



▲ **Figure 4. Computed and experimental results [9].**

ness of 150 mm was under test. Figure 3 clearly shows that the area with the diminished voltage shifts to the higher frequencies while the applied RF voltage amplitude increases.

Computer simulation results and discussion

Three different mathematical and physical models were used to estimate the necessary DC reverse bias voltage to keep the PIN diode in a low-distortion state. Computed results based on these models are verified with experimental RF performance data.

The silicon PIN diode with p^+–n–n^+ structure and the i-layer width of 150 μm was used in simulation. The other parameters of the diode were: cross-section of 3.14 × 10^{-2} cm^2; i-layer doping level of 5 × 10^{12} cm^{-3}; and electron and hole lifetimes of 10^{-5} s. The diode doping profile was assumed to be abrupt, with acceptor p^+ and donor n^+ concentrations of 5 × 10^{19} cm^{-3} and 1 × 10^{20} cm^{-3}, respectively.

The distortion factor was used as a nonlinearity and a distortion figure of merit in calculations with a drift-diffusion model [7]. The distortion factor value of 10^{-4} was chosen as the permissible low level of harmonic distortions. The effective threshold voltage $V_{t\,eff}$ of 0.1...0.2 V plays the same role in the multilayer impedance PIN diode model (see the appendix). In both models, it provides a DC current generated by the PIN diode of several microamperes. This is in agreement with the experimental parameter of the low nonlinearity and low harmonic distortion that have been used in the measurements.

Figure 4 compares the measured and computed required reverse bias DC voltage V_o [9]. The computed results based on the isothermal drift-diffusion model of a semiconductor structure (green line) provide the better agreement with the experimental data, but the use of this model requires a lot of time. Computed data based on the analytical expressions (red line) [6], taking into account only the i-region transit time, shows qualitative agreement but overestimates the required DC reverse bias voltage in a wide frequency range. The accuracy of this expression is decreased for the PIN diodes with the i-layer thickness exceeding the depleted region width.

The multilayer impedance model, considering transit time and effects of the field distribution inside the i-layer, produces a compromise between the computational time and accuracy. It provides more precision estimation (blue line) than the model [6] and it does not require significant computation time like the drift-diffusion model [12]. The expressions for the minimum required reverse DC bias voltage estimations, using this model, are shown in the Appendix.

Conclusion

A reverse-bias state of the PIN diodes with different i-layer thicknesses were under experimental consideration in this article. Several mathematical and physical models were used to estimate the necessary DC reverse-bias voltage to keep PIN diodes in a low-distortion state. The multilayer impedance model, considering the electric field distribution inside the i-layer, provides the trade-off between the computational time and accuracy. It allows engineers and designers of microwave control devices to obtain more precise estimations of the required reverse DC bias voltage. ■

Acknowledgements

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Author information

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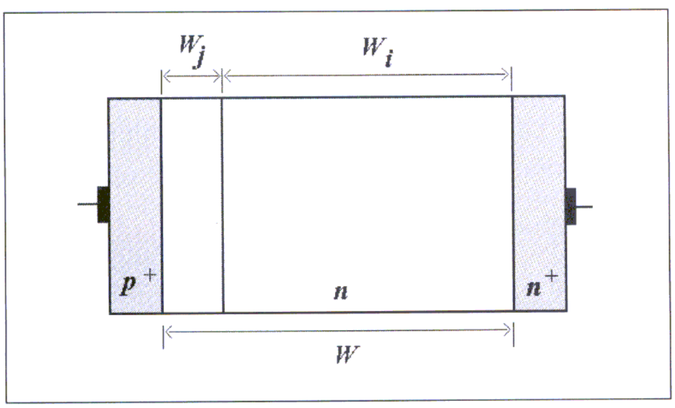
Appendix

The impedance multilayer model of a PIN diode [7, 10] is based on the notion that the i-region of the diode below punch-through consists of undepleted and depleted regions (Figure A1). The depleted region width W_j and capacitance C_j of a reverse-biased PIN diode are defined using the expressions:



▲ **Figure A1. Multilayer model of the PIN diode.**

$$W_j = W_{j0} \times \sqrt{1 + \frac{V_0}{V_k}} \tag{1}$$

$$C_j = C_{j0} \times \sqrt{1 + \frac{V_0}{V_k}} \tag{2}$$

where

$$W_{j0} = \sqrt{\frac{2\varepsilon_i V_k}{eN}} \tag{3}$$

$$C_{j0} = \sqrt{\frac{\varepsilon_i S}{W_{j0}}} \tag{4}$$

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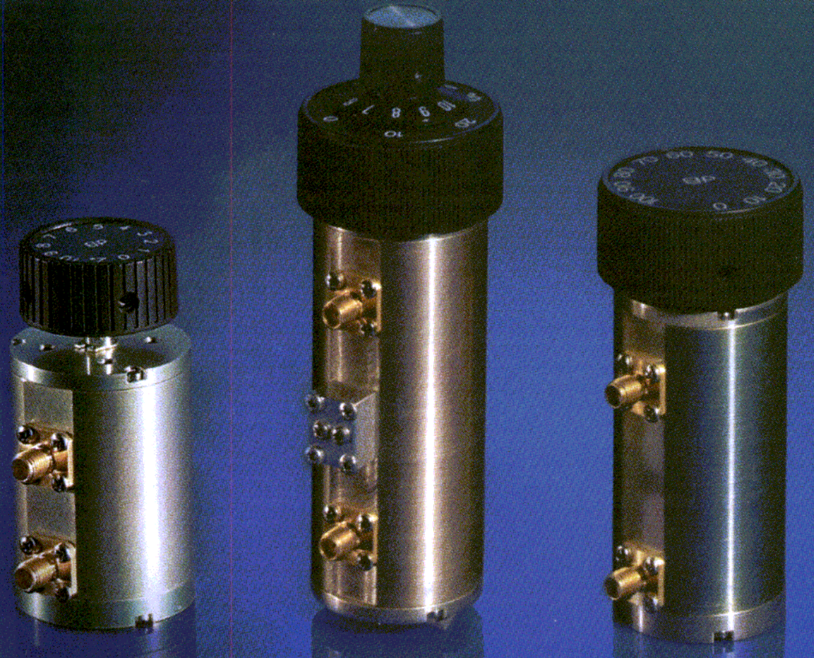


W_{j0} = depleted region width of non-biased diode, V_k = built-in potential, V_0 = DC reverse bias voltage, e = single charge, N = i-layer doping level, ε_i = silicon dielectric permittivity, S = diode cross-section area. The value of depleted region resistance R_j was assumed to be about 1 to 2 Mohms.

The undepleted region capacitance C_i and resistance R_i are

$$C_i = \frac{\varepsilon_i S}{W - W_j} \tag{5}$$

$$R_i = \frac{\rho(W - W_j)}{S} \tag{6}$$

where

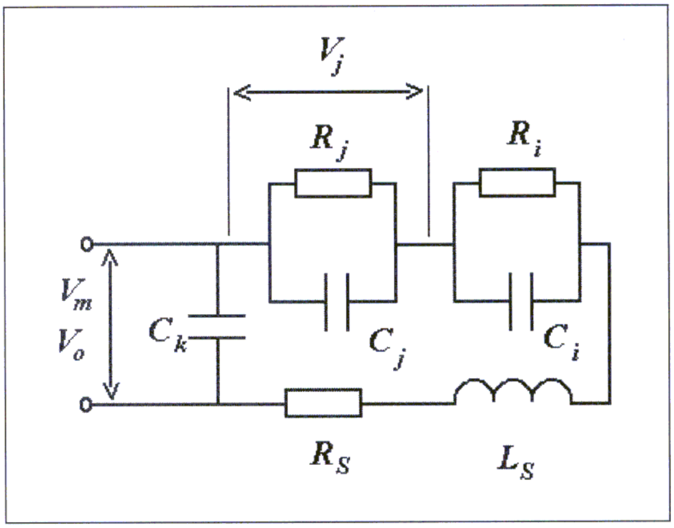
$$\rho = \frac{1}{(e\mu N)} \tag{7}$$

ρ = i-layer bulk resistance, W = i-layer width, μ = the majority carrier mobility.

The equivalent circuit of the diode according to this model is shown in Figure A2. There are also parasitic elements such as inductance L_s, contact or passive regions resistance R_s and package capacitance C_k (in the case of packaged PIN diode).

This model also takes into account the transit time effect. It leads to the reduction of the voltage across the depleted region V_j in $1/F(\theta_j)$ times. Here, $F(\theta_j)$ is the empirical function of the non-zero carrier transit time threw depleted region. This depends on diode geometry, semiconductor material properties, RF or microwave input power and applied DC reverse bias voltage [10].

When the diode is under DC zero or reverse bias voltage, $R_j >> R_i$. Therefore, DC voltage across the PIN diode is applied directly to the depleted region (to p^+–n junction). Applied AC voltage is divided between two regions in an i-layer according to the properties of



▲ **Figure A2. Equivalent circuit of a PIN diode.**

the frequency-dependent impedance divider formed by depleted and undepleted regions. It was assumed that p^+–n junction does not produce harmonic distortions when the total voltage across the depleted region is less than the effective threshold voltage $V_{t\,eff}$. This value was assumed to be about 0.1...0.2 V. This condition is written as [7]:

$$V_j - V_0 \le V_{t\,eff} \tag{8}$$

Finally, we can estimate required reverse bias DC voltage V_0 (for $L_s \approx 0$, and $R_s \approx 0$) as:

$$V_0 = V_m \times F(\Theta_j) \times \left| 1 + \frac{R_i}{R_j} \times \frac{1 + j2\pi f C_j R_j}{1 + j2\pi f C_i R_i} \right|^{-1} - V_{t\,eff} \tag{9}$$

where f is the operating frequency, V_m is the RF or microwave amplitude across the diode. Taking into account Equations (1) and (2), we can see that Equation (9) is the nonlinear equation, with V_0 : $V_0 = F(V_0)$, and should be solved iteratively.

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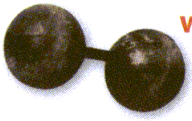

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Food Irradiation and the Microwave/RF Market

RF and microwaves can provide the energy source for eradicating microorganisms in food

By Harold Hansen
Hamilton Sundstrand Space Systems

The issue of food safety is in the forefront of the public consciousness because of recent incidents of food-borne illness that have resulted in product recalls, illnesses and even death. More than 25 years of scientific study has shown that food irradiation is a proven and safe technology that could eliminate many of these concerns. According to recent studies, the public's reluctance to use irradiated food is steadily shrinking [1].

Irradiation and microwaves/RF

What does this have to do with microwave and RF technology? One of the most promising methods of irradiation, electron beam processing, uses high power microwaves to generate the high-energy electrons. The machine that produced high-energy electrons is called an RF Linear Accelerator, or linac. Contained within linacs are familiar RF devices, such as pulse transformers, circulators, waveguides, magnetrons and klystrons. Although the term "linear accelerator" calls up images of large, complex systems, much of a linear accelerator looks similar to the "line type" modulators and RF sources used in 1950s-style radar systems.

Presently, the largest market for linear accelerators in the world is for use in radiation therapy. These machines are complex and extremely expensive. The accelerator systems that are used in electron beam processing are less complicated and significantly less expensive. The potential market for electron beam systems could be large. Under the proposed model of online, in-plant irradiation, a market potential exists for thousands of these systems to be installed in the U.S. In addition, because these systems will operate almost continuously, there will be a large demand for spare parts.

The challenge is in reducing the price of these electron beam systems. This challenge can be met by looking to microwave and RF manufacturers to build a large portion of the systems. Manufacturers with experience in radar and other high-power systems could easily and efficiently build these systems. With cold war budgets greatly reduced, this could represent a significant way of leveraging radar experience into a new commercial market.

This article is divided into three sections. The first section presents an overview of irradiation and the practical issue of specifying an electron beam processor based on product requirements. The second section discusses the linear accelerator. The third section is a cost analysis of a typical application: irradiation of hamburgers.

Food irradiation through the use of electron beam processing, which requires high-power microwaves to generate high-energy electrons, is a potentially large market. The development of commercial competition will result in both the technological and economic innovation that will make this process a reality and, ultimately, lead to a safer food supply.

Irradiation sources

The most useful way of characterizing an irradiation process is by looking at how the radiation is produced. Irradiation can be characterized as either "isotope-based" or "accelerator-based." In isotope-based irradiation, the source of the radiation is a radioactive material called an isotope. An isotope has an unstable nucleus. As the material attempts to achieve a steady state, it releases energy in the form of radiation (alpha, beta and/or gamma particles). The two

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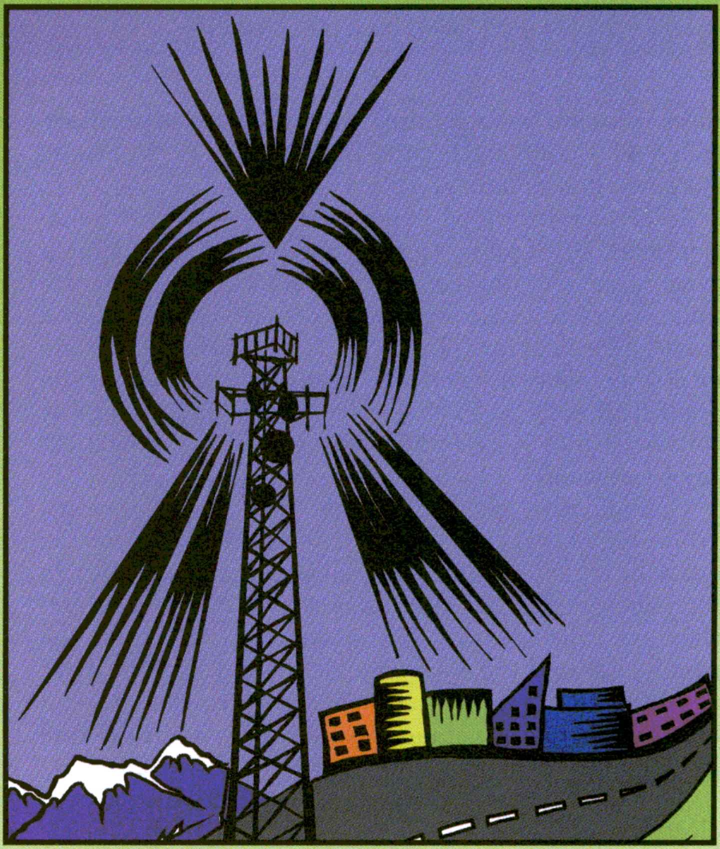

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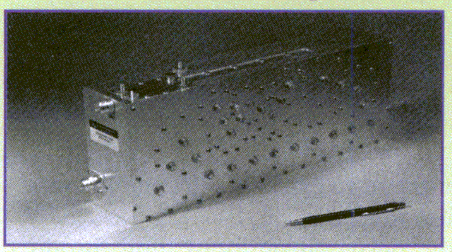

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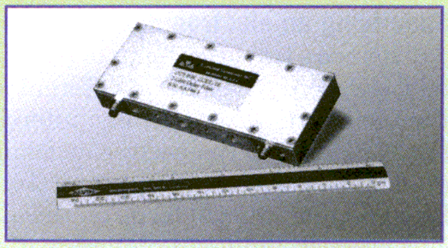

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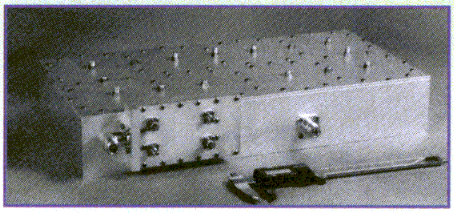

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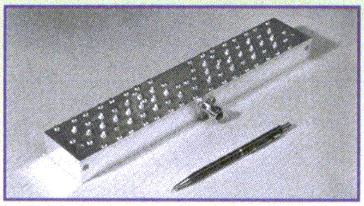

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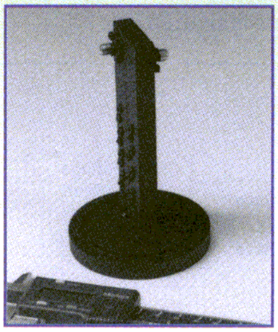

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most common isotopes used in irradiation of food are cobalt 60 (Co-60) and cesium 137 (Cs-137). Both of these isotopes emit gamma rays.

In accelerator-based irradiation, the source of radiation is an accelerator. Again, linac is the most common type of accelerator for food irradiation. Linacs use microwaves to create very high electric fields within a wave guide structure. These fields are used to accelerate electrons. The accelerated electrons can either be applied directly to food (electron beam processing) or converted into X-rays and then applied to the food (X-ray processing)

Gamma rays and X-rays are high-energy photons. When high-energy photons are released as part of the decay of an isotope, they are called gamma rays. When high-energy photons are produced in an accelerator, they are called X-rays. For all practical purposes, the gamma ray and the X-ray are identical. However, there is a difference in the spectrum of the energy for the photon: the energy in gamma rays tends to be in discrete levels or bands, while X-rays are broad spectrum.

The linear accelerator is capable of producing both electrons and X-rays. The primary particle in the linac is the electron, which can be converted into an X-ray through the use of a conversion target. The conversion from electrons to X-rays is very inefficient, with more than 90 percent of the accelerated electron's energy lost in the conversion process as heat. Electrons have a low mass and therefore penetrate only a shallow distance, while X-rays have a very heavy mass and can penetrate a very far distance. Since electrons travel only a short distance, they lose their energy over a short distance and therefore produce very high-dose rates. X-rays, on the other hand, travel over a long distance, releasing their energy over along path resulting in a low-dose rate.

Specifying an electron beam system

There are two basic types of irradiation: gamma/X-ray and electron. They have different characteristics and therefore are not generally interchangeable. Certain applications lend themselves to gamma/X-ray, while others lend themselves to electron. To specify an electron beam system, we will focus on the attributes of the specific product. While this determination may look simple, it is important for determining feasibility and basic machine parameters.

The first step in specifying an e-beam system is the determination of the electron beam characteristics required for that application. The basic determiners are required dose (rads), desired throughput (lbs./hr) and product presentation (primarily thickness). This information may be used to determine the beam energy and the beam power.

Customers will almost always specify required dose levels. This information, combined with throughput requirements, defines how much beam power is required. The dose requirements are defined through rigorous application testing carried out by the customer; however, there are government regulations that specify maximum dose allowable.

The next important piece of information supplied by the customer is the required throughput (lbs/hr). Given this information and the required dose from the previous step, an estimate of required beam power can be made. The required beam power is determined using the fundamental definition of dose [2]. Then Equation (1) can be used to estimate beam power:

$$P = \frac{T \times D}{7920 \times \varepsilon} \tag{1}$$

where P = beam power in watts; T = throughput in lbs./hr; D = total dose in Gy; and ε = efficiency (.33 typically). The efficiency factor, ε, is used to account for the fact that not all the electrons lose all of their energy in the product.

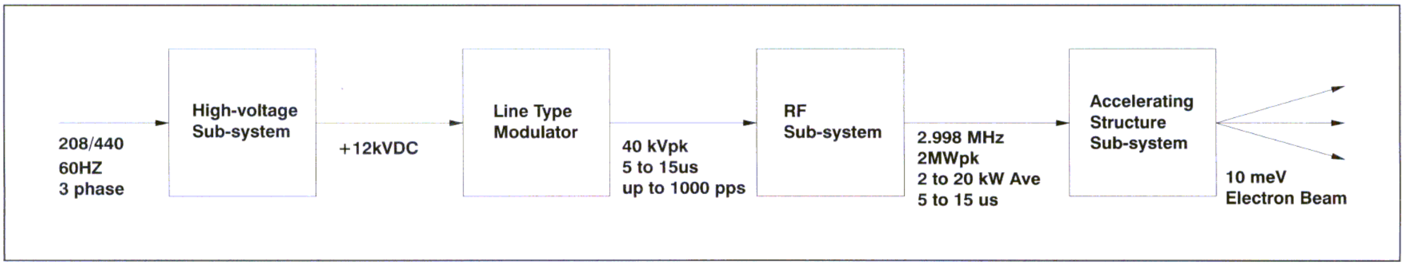
The penetration, or range, of an electron is very shallow. The energy of the system should be designed so that the electrons have enough energy to penetrate the product. If the energy is too high, most of the electrons (and their contribution to the total dose) will pass right through the product and efficiency will drop.

The penetration or range of an electron can be calculated by using Equation (2):

$$E = 1.85 \times R \times \rho + .24 \tag{2}$$

where E = energy of beam in MeV; R = range in cm ρ = density of product in grams/cc.

After determining dose, throughput and product thickness, the basic parameters required of the electron beam system are known. The next step is to look at the



▲ **Figure 1. Block diagram of a basic linear accelerator.**

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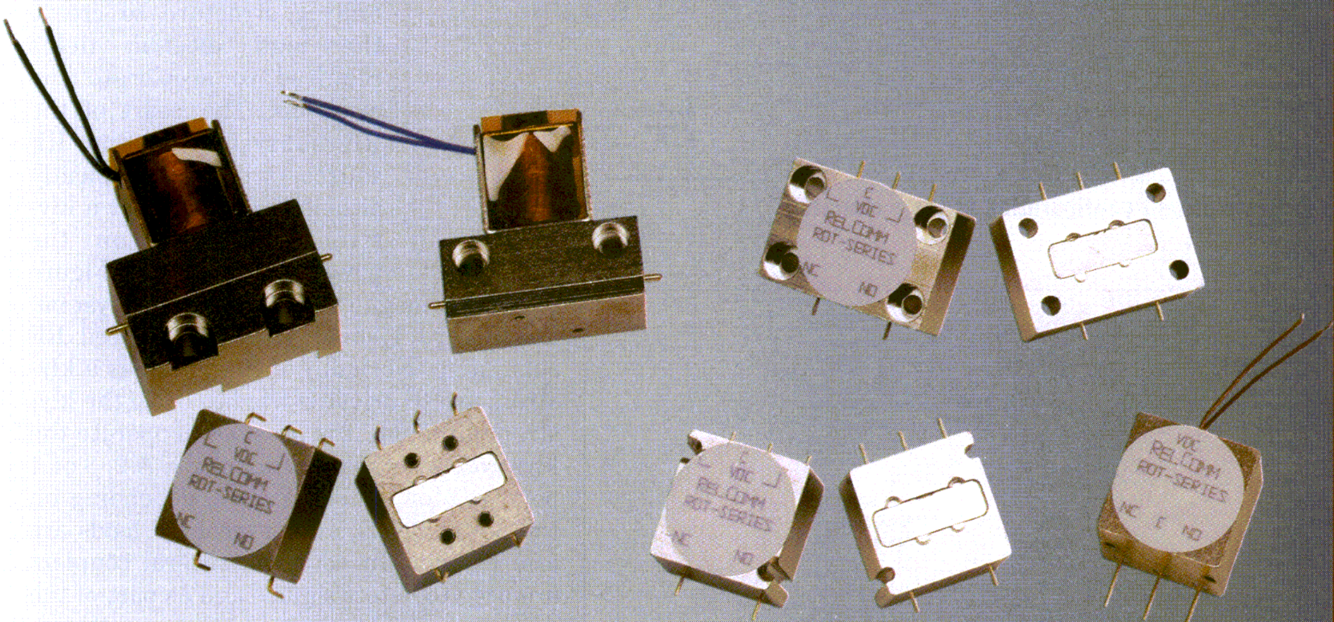

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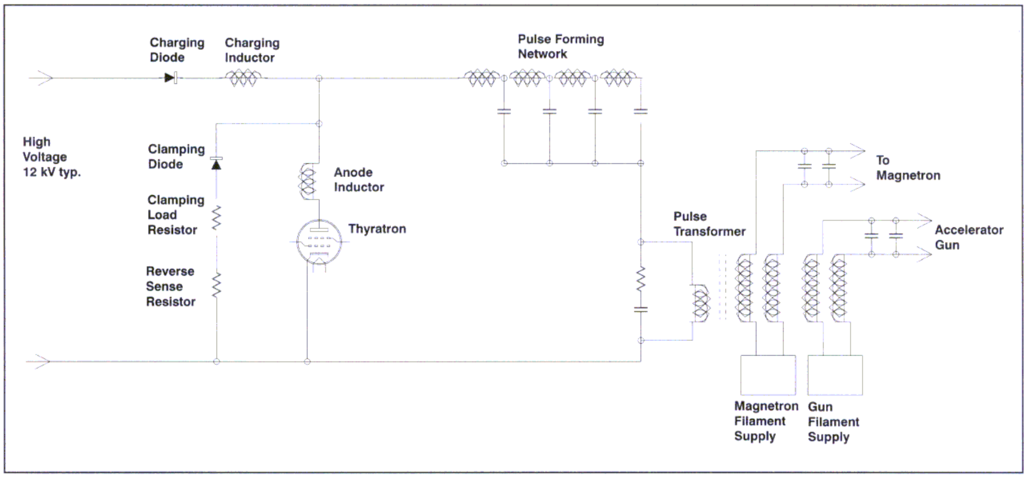
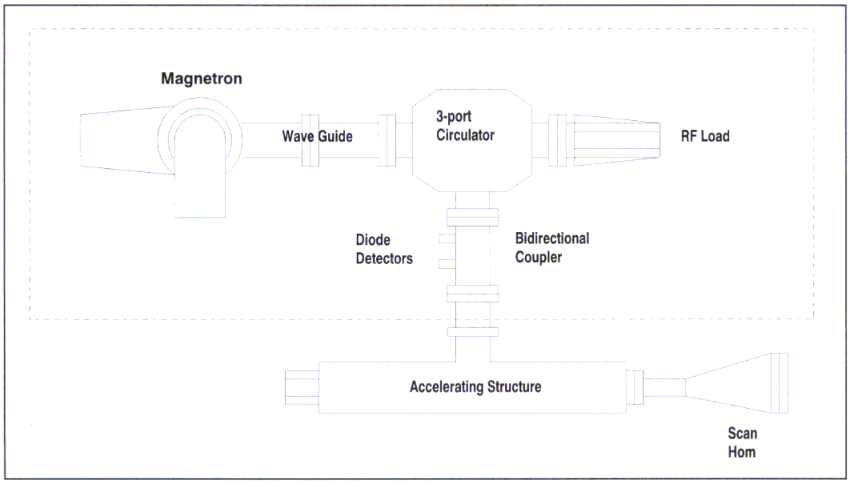


▲ **Figure 2. Basic circuit configuration of the modulator.**



▲ **Figure 3. Magnetron-based RF subsystem.**

chosen electron beam system and develop some simple operating models.

Electron beam accelerators

As with most modulators, high voltage is required. Most commercial accelerators use a brute force method of generating high voltage: step-up transformer, rectifiers and capacitors. Some accelerator designs use switch mode power supplies for the high voltage power supply. Whichever technology is used, the regulated voltage should be in the range of 10 to 20 kV. The supply must be capable of driving the dynamic load, i.e., a line type modulator, and be capable of surviving transient short circuit conditions as part of normal operation.

The modulator design used almost exclusively in every commercial linac in the world is a derivation of the line type modulator. The function of the modulator is to convert high voltage into high voltage pulses (typical pulse width is fixed with a range of 5 to 20 μs). The basic circuit configuration is shown in Figure 2. It contains a resonant charging circuit, a high current switch, a pulse forming network and a pulse transformer.

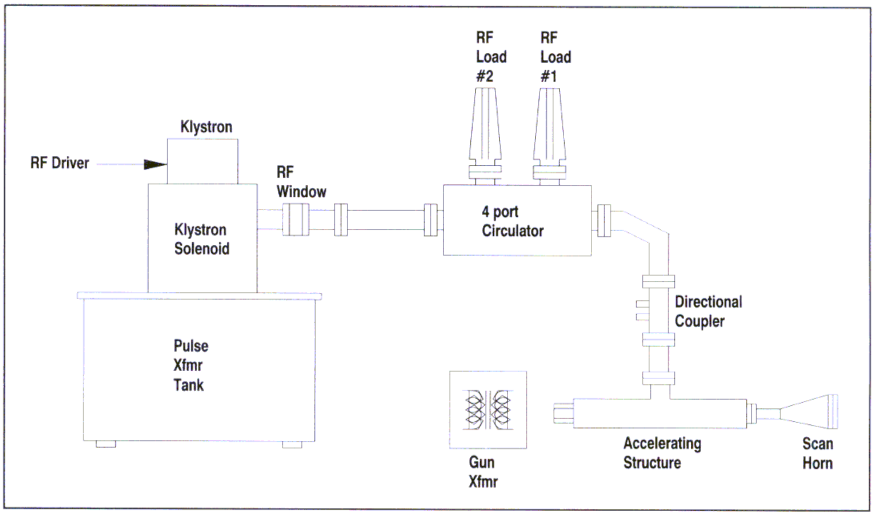
The function of the RF subsystem is to produce a burst of high-energy microwave energy that is applied to the resonant accelerating structure to create the high electric fields required to accelerate the electrons. The heart of the RF subsystem is the RF source, which is typically either a magnetron or a klystron, depending on the power requirements. Figure 3 shows a magnetron-based RF subsystem, and Figure 4 shows a klystron based RF subsystem. Normally, if the source is a klystron, an RF driver is used to drive the klystron, although it is possible to operate the klystron in self-oscillating mode. A 2, 3 or 4 port isolator is always used, depending on the power levels, and several RF loads are also present. In addition, several couplers are used to detect signals and as part of the AFC system (not shown but always there). Most accelerators will operate in an S-band of 2998 MHz, but some high power units (<100 kW beam) operate in L-band.

The design comes together in the accelerating structure subsystem, whose heart is the accelerating structure, consisting of a resonant microwave device with resonant cavities. The microwave energy sets up a very high electric field within this structure. When an electron is synchronized and injected into this structure, it can be accelerated to very high energies. The key elements of an accelerating structure are the electron gun, the RF window(s) [3], the buncher section, the relativistic section and the electron window/target. The electron window is actually a transmission target: electrons pass directly through with only a small energy loss. Often, the electron beam is then scattered, using a scattering foil, or scanned using magnetic deflection.

The required power in the beam is calculated as follows. Going backwards into the system, there is some insignificant power loss in the window [4]. The first heavy hitter is the accelerating structure. For a stable operation, the power in the beam should not exceed the average power in the field. Therefore, we assume that the power beam is equal to the power in the field. The next heavy hitter is the efficiency of the RF source; magnetrons are 50 percent efficient, whereas klystrons are

typically between 30 and 40 percent efficient. The high voltage sub-system is typically 75 to 80 percent efficient. Add power for the other sundry subsystems and we can see that the efficiency defined as power beam/power utility is 10 to 15 percent maximum.

Beam energy is important because it sets penetration. Beam energy is generally a function of the structure design. In the design of an accelerating structure, a compromise is made between the structure's physical length and required peak power (MW) to achieve a specific output beam energy. The availability of high peak power and high average power RF sources (magnetrons and klystrons) usually means that a structure is designed for operation between 500 kW and 2 MW of peak power in the acceleration fields, leaving an equal amount of power left for the beam. Some accelerating structure designs allow for variation of energy [5], which can cause inefficiency. The best results are achieved when a structure is designed to operate at maximum efficiency at the required beam energy.



▲ **Figure 4. Klystron-based RF subsystem.**

Economic analysis

An example: a meat processor has requested an electron beam irradiator to treat hamburger patties. The customer wants to treat 1 million lbs/week to a dose of 1.5 kGy. Hamburger patties (r = 1.5 gram/cc) are 0.5 inches thick (1.27 cm) and will be presented on a flat conveyor belt before packaging.

First, determine the required beam power. Assuming that production is 120 hrs/week (6 days/week, 20 hrs/day), 1 million lbs/week = 8333 lbs/hr. Using Equation (1):

$$P = \frac{8.333 \text{ lbs / hr} \times 1500}{7920 \times .33} = 5260 \text{ watts} \quad (1)$$

Next, we determine required beam energy using Equation (2):

$$E = (1.85 \times 1.27 \text{ cm} \times 1.5 \text{ grams / cc}) + .24 = 3.76 \text{ MeV} \quad (2)$$

Increase this to the next highest integer energy, 4 MeV.

Based on these calculations at 4 MeV, a 5300-watt system would meet the customer's requirements. How much would this process add to the overall cost of the hamburger? The most practical presentation would be to relate these costs as a function per pound. The time value of money is ignored to simplify the analysis.

At 5300 watts, the system could be either a magnetron- or klystron-based system. The cost of such a system would be about $500,000. Assuming a five-year analysis period, this translates to a cost of $100,000 per year. We will assume that the plant operates 120 hours per week and 50 weeks per year, which is 6,000 hours per year. At $100,000 and 6000 hours per year, the hourly cost is $16.60. With a throughput of 8,333 lbs/hr, that translates to a machine cost of $0.002 per pound.

Installation costs would include the radiation shelter and any modifications to the conveyor system. This is estimated at $100,000. Over five years, that is $20,000 per year. The cost per pound is $0.0004. The utility costs are primarily electricity. If the worse case efficiency is 10 percent, for a 5.3 kW beam, we need 53 kW utility power. The yearly consumption is 318 kW hours per year. At a cost of $0.1 per kilowatt-hour, that is $31.80 per hour, or $0.004 per pound.

Spare parts and service will be conservatively estimated at 50 percent of the purchase price of the system. The most costly item in the system is the klystron, which costs about $20,000 and has a lifespan of about 10,000 hours. Other parts will need replacement, but their cost is relatively low. $250,000 over five years is $50,000 per year. At 6,000 hours per year, that translates to an hourly cost of $8.30 and a cost per pound of $0.001. Labor is conservatively estimated at $200,000 per year. This means an hourly cost of $33.30 and a per pound cost of $0.004.

The total cost per pound that this process would add to the hamburger is $0.002 for the system, $0.0004 for the installation, $0.004 for the utility, $0.001 for the spare parts and $0.004 for the labor, a total of $0.011 per pound. This analysis shows that the cost of this process to the hamburger is about $0.01 per pound.

Conclusion

Irradiation is a proven process, and electron beam processing will eventually represent a large portion of the irradiation market. The most common source for electron beams is linear accelerators. Presently, these are being built by medical electronics companies for radiation therapy and are very expensive. In addition, these units are designed for therapy applications, which is significantly different from electron beam processing. For a unit to be viable, it has to be designed with the electron beam application as the end application.

As was shown, a large portion of a linear accelerator is very similar to radar and microwave equipment of the past. Manufacturers from both the commercial and military markets are qualified to build such systems, both from the technical standpoint of high reliability, constant operation and from the economic standpoint of having the infrastructure and production experience. Irradiation represents a potentially large market, and with decline the other high-power microwave applications, this could help lessen the adverse effects. Ultimately, the result will be a safer food supply. ■

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2. Gy is applied energy in joules/kg; watts are joules/sec.
3. One window per microwave feed, units higher than 10 kW usually use two feeds to reduce heating of RF windows. Windows are usually ceramic.
4. If the power loss in the window exceeds even a small fraction of the beam power, its temperature will rise and the window, being a thin foil, will fail.
5. Medical linacs normally have several beam energy settings but since the beam power is low (<500 watts typically), loss of efficiency is not an issue.

Author information

Harold Hansen is pursuing his master's degree in management at the Lally School of Business at Rensealler Polytechnic Institute in Hartford, CT. He is presently employed as a senior engineer at Hamilton Sundstrand Space Systems, a division of United Technologies Corporation. Prior to working at Hamilton, Hansen spent last eight years designing and building linear accelerators for industrial radiography and electron beam applications. He has co-authored several papers on accelerator based irradiation. He is presently writing a book on linear accelerator design.

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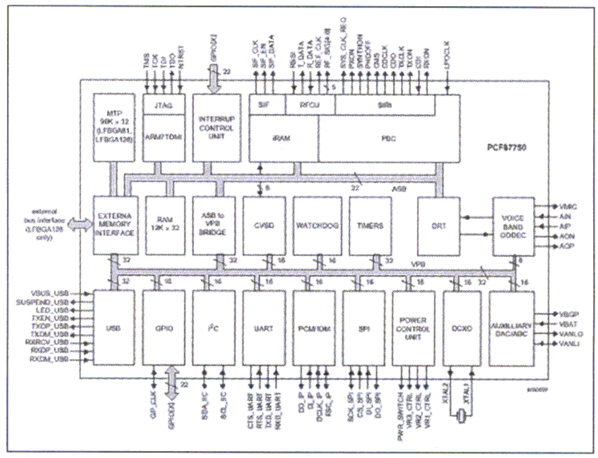
Philips introduces 2G Bluetooth controller IC

The "Blueberry" PCF87750 baseband controller from Philips Semiconductors embodies the company's systems-on-silicon approach to IC design. Blueberry incorporates the Philips Bluetooth Core, based on the Ericsson core, with enhanced voice path, operating as a link controller, including ciphering, scrambling, CRC checking/generation and FEC encoding/decoding. The chip combines the Bluetooth core with an advanced ARM/RISC processor, multiple I/O interfaces and a full analog-to-digital CODEC. The IC's 384 kbytes of Flash memory and 64 kbytes SRAM fill program memory requirements. The RF interface is simplified with on-chip GFSK pulse shaping, RSSI measurements and an 8-bit D/A converter. The Blueberry can be combined with Philips' TrueBlue radio module or with the UAA3558 radio IC to create a complete system. A comprehensive developer's kit (model BByK) is available to help speed product development.

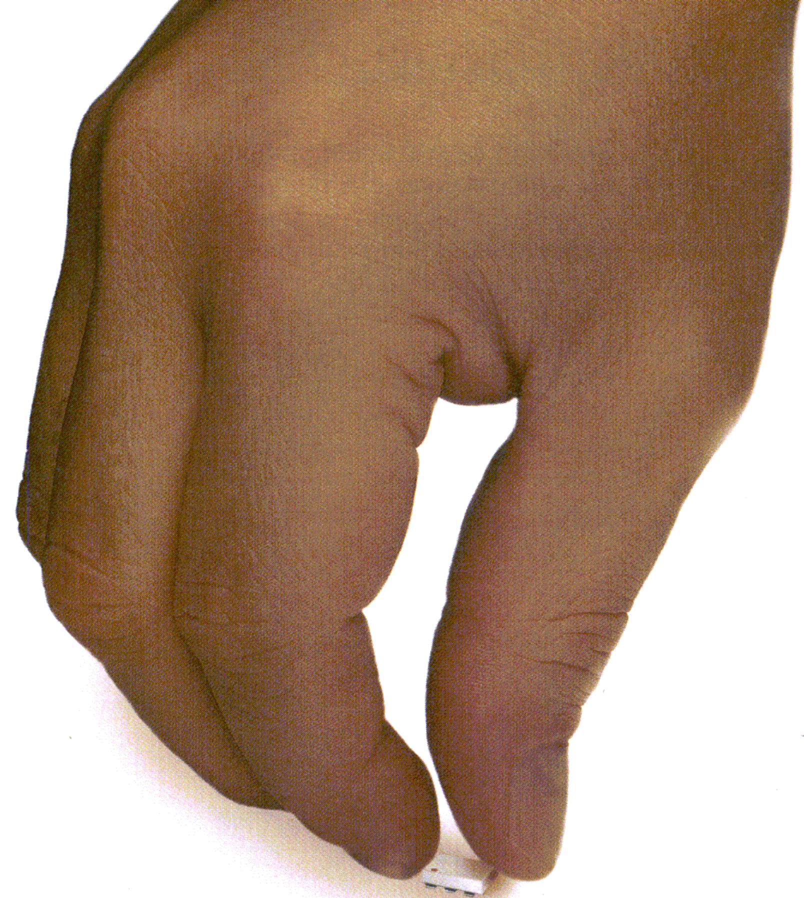


The next generation of Bluetooth products planned by Philips includes dedicated devices for headset/audio or data only and other derivatives of the company's technology.

Philips Semiconductors
Tel: 1-800-234-7381
Internet: www.semiconductors.philips.com
Circle #201

Texas Instruments offers royalty-free licensing for IEEE 802.11g technology

In an effort to satisfy growing demand for standards-based WLAN technology, Texas Instruments will offer royalty-free licenses under TI patents required for implementation of IEEE 802.11g, pending ratification of TI's proposal for this extension of IEEE 802.11b. The new standard proposed by TI doubles the data rate of IEEE 802.11b-based products to 22



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·smaller size ·better performance ·lower cost

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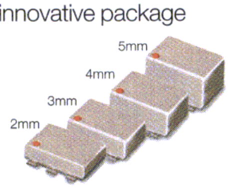

ADE Mixers...Innovations Without Traditional Limitations!

ADE* TYPICAL SPECIFICATIONS:

Model	Height (mm)	Freq. (MHz)	LO (dBm)	Conv. Loss Midband (dB)	L-R Isol. Bandwide (dB)	IP3 (dBm) @ Midband	Price ($ea.) Qty. 10-49
ADE-1L	3	2-500	+3	5.2	55**	16	3.95
ADE-3L	4	0.2-400	+3	5.3	47**	10	4.25
ADE-1	4	0.5-500	+7	5.0	55**	15	1.99
ADE-1ASK	3	2-600	+7	5.3	50**	16	3.95
ADE-2ASK	3	1-1000	+7	5.4	45**	12	4.25
ADE-6	5	0.05-250	+7	4.6	40	10	4.95
ADE-12	2	50-1000	+7	7.0	35	17	2.95
ADE-4	3	200-1000	+7	6.8	53**	15	4.25
ADE-14	2	800-1000	+7	7.4	32	17	3.25
ADE-901	3	800-1000	+7	5.9	32	13	2.95
ADE-5	3	5-1500	+7	6.6	40**	15	3.45
ADE-13	2	50-1600	+7	8.1	40**	11	3.10
ADE-20	3	1500-2000	+7	5.4	31	14	4.95
ADE-18	3	1700-2500	+7	4.9	27	10	3.45
ADE-3GL	2	2100-2600	+7	6.0	34	17	4.95
ADE-3G	3	2300-2700	+7	5.6	36	13	3.45
ADE-28	3	1500-2800	+7	5.1	30	8	5.95
ADE-30	3	200-3000	+7	4.5	35	14	6.95
ADE-32	3	2500-3200	+7	5.4	29	15	6.95
ADE-35	3	1600-3500	+7	6.3	25	11	4.95
ADE-18W	3	1750-3500	+7	5.4	33	11	3.95
ADE-30W	3	300-4000	+7	6.8	35	12	8.95
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ADE-1LHW	3	2-750	+10	5.3	52**	15	4.95
ADE-1MH	3	2-500	+13	5.2	50**	17	5.95
ADE-1MHW	4	0.5-600	+13	5.2	53**	17	6.45
ADE-12MH	3	10-1200	+13	6.3	45**	22	6.45
ADE-25MH	3	5-2500	+13	6.9	34**	18	6.95
ADE-35MH	3	5-3500	+13	6.9	33**	18	9.95
ADE-42MH	3	5-4200	+13	7.5	29**	17	14.95
ADE-1H	4	0.5-500	+17	5.3	52**	23	4.95
ADE-10H	3	400-1000	+17	7.0	39	30	7.95
ADE-12H	3	500-1200	+17	6.7	34	28	8.95
ADE-17H	3	100-1700	+17	7.2	36	25	8.95
ADE-20H	3	1500-2000	+17	5.2	29	24	8.95

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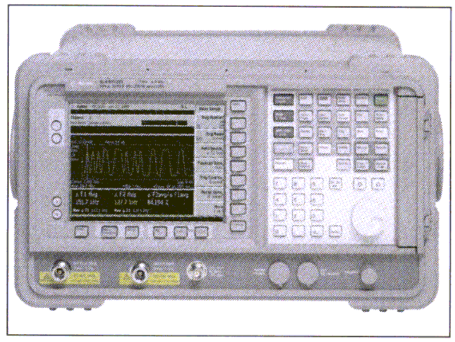
Mbps in a manner that is fully backward-compatible with existing products. The higher data rate will enable higher-performance wireless data products with standards-based interoperability. New applications possible with 22 Mbps include streaming video and HDT.

Texas Instruments Inc.
Internet: www.ti.com
Circle #202

Bluetooth measurement systems aid wireless connectivity

Agilent Technologies has introduced Bluetooth measurement solutions to help designers create new products using this short-range wireless connectivity system. Products in Agilent's Bluetooth measurement family include the Agilent ESA-E series spectrum analyzers, the Agilent E1852A Bluetooth Test Set and the Agilent ESG-D series signal generator with a Bluetooth option. The Bluetooth



package for the ESA-E spectrum analyzers is priced starting at $8,900; the E1852A test set is priced at under $18,000; and the Bluetooth option for the ESG-D signal generators is priced at $14,390. All products were expected to be available by February 1.

Agilent Technologies Inc.
Tel: 1-800-452-4844, ext. 7307
Internet: www.agilent.com/find/bluetooth
Circle #203

Intersil, Silicon Wave to develop dual-mode WLAN system

To address the issue of two major short-range wireless data systems, Intersil and Silicon Wave have announced that development is underway for dual-mode WLAN solutions that allow portable devices to be connected to a company LAN or to companion devices. The initial platforms offered will be Cardbus32 and MiniPCI, containing a fully compliant Bluetooth radio and a WECA Wi-Fi™ certified IEEE 802.11b-compliant radio. An additional goal is to employ dynamic switching to allow both radios to share a common antenna. Teams from both companies are working to develop a reference design that will permit manufacturers to move quickly to implement dual-mode systems on a PCMCIA card or MiniPCI card.

Intersil
Internet: www.intersil.com
Circle #204
Silicon Wave
Internet: www.siliconwave.com
Circle #205

Single-chip Bluetooth solution uses CMOS technology

Cambridge Silicon Radio (CSR) offers a single-chip Bluetooth solution, the BlueCore™01. BlueCore01 combines a fully integrated 2.4 GHz radio, baseband and microcontroller in one CMOS chip. Together with Flash ROM containing the CSR Bluetooth software stack, it provides a fully compliant Bluetooth design solution. The chip uses CMOS technology for the 2.4 GHz radio, a key to one-chip design of radio and baseband circuits. BlueCore01 includes USB interface support. Few external RF component are required, permitting rapid design of a circuit board containing the BlueCore01.

Cambridge Silicon Radio
Internet: www.csr.com
Circle #206

Software alliances assure compatibility of Bluetooth modules

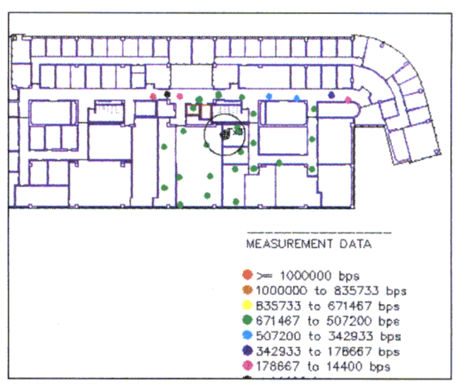
Socket Communications has announced alliances with BSQUARE Corporation, Classwave Wireless Inc. and lesswire AG. BSQUARE's WinDK extension for Bluetooth allows developers to simplify interface issues by using existing Win32 applications. Classwave's Polyphony Servers offer turnkey solutions for network operators by routing information to mobile phones over cellular networks and to

Bluetooth devices over the Internet. lesswire offers the LocalNavigator, an integrated platform of software and hardware components that supports location-aware services. These alliances will allow greater capabilities in Socket's plug-in cards and modules.

Socket Communications, Inc.
Internet: www.socketcom.com
Circle #207

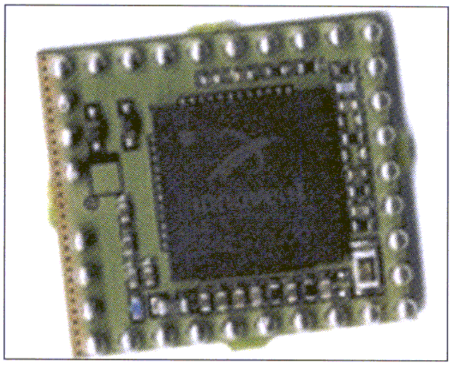
Convenient coverage data collection for IEEE 802.111 systems

Wireless Valley Communications offers the LanFielder, with pushbutton installation and remote performance to evaluate actual coverage



for all IEEE 802.11, 802.11a and 802.11b high speed IP networks. LanFielder runs on a lightweight pen-based computer and interfaces with all popular lightweight receivers and PCMCIA Wi-Fi WLAN modem cards.

Wireless Valley Communications
Tel: 1-540552-8300
Internet: www.wirelessvalley.com
Circle #208



Bluetooth radio module

Bluetronics offers the IRMBB2 fully integrated Bluetooth radio module incorporating Broadcomn's Bluetonium BCM2001B CMOS chip. The module includes an integrated patch antenna, bandpass filter, switch, two baluns, the radio chip and necessary decoupling capacitors. The module offers a cost-effective solution for implementing a Bluetooth-enabled portable device.

Bluetronics
Tel: +46 11 26 41 00
Internet: www.bluetronics.com
Circle #209

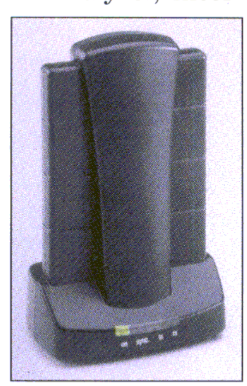
SiGe LNA for Bluetooth and PCS

RF Micro Devices has introduced the RF2472, a SiGe low noise amplifier for 2.4 GHz applications, including Bluetooth. In this band, the part has a noise figure of 1.5 dB and more than 14 dB gain. The device is also characterized for operation at 1.9 GHz and can be used in other applications from DC to 6 GHz. The RF2472 is offered in a SOT-5 package and is priced at $0.39 in quantities exceeding 10,000.

RF Micro Devices
Internet: www.rfmd.com
Circle #210

Bluetooth protocol analyzer evaluates radios and networks

Tektronix announces the BPA100 Bluetooth Protocol Analyzer, incorporating a SIG qualified radio, baseband and host stack. The instrument permits a wide range of performance and compatibility evaluations of Bluetooth radios and networks. The BPA100 was acquired by Tektronix from Digianswer. U.S. pricing is $24,950.



Tektronix, Inc.
Internet: www.tektronix.com/bluetooth
Circle #211

Product Spotlight

How to place Product Spotlight and Classified ads

Our Product Spotlight section allows you to show off your company's products or services in an economical 1/9 page color ad. Ads may be sent camera ready, or we'll make up an ad for you! All you'll need to provide is a product photo, company logo and a description of the product or service you want to advertise.

Our Classifieds section is the place to advertise career opportunities, consulting services, used test equipment and just about anything else. Special Classifieds rates are available and are agency-commissionable if material is supplied. Or, if you prefer, our staff can make up the ad for you. Classified display ads are available in color or black and white in standard ad sizes, or in black and white by column inch.

For more information on advertising in these two sections, please contact Aileen Kronke, *Applied Microwave & Wireless*, 630 Pinnacle Court, Norcross, GA 30071; Tel: 770-449-6774; Fax: 770-448-2839; E-mail: amw@amwireless.com.

Product Spotlight

Classifieds

Applied MICROWAVE & WIRELESS

2001 Editorial Calendar

Issue	Ad Closing	Materials Due	Editorial Emphasis	Special Coverage & Bonus Distribution*
February	January 2	January 9	Wireless Data Antenna Systems Microwave Materials	Wireless Symposium
March	February 1	February 8	Designing for 3G RFICs Receiver Design	CTIA Wireless 2001
April	March 1	March 8	Broadband Wireless Fiber Optics Wireless Test Methods	NAB Convention MTT-S Preview
May	April 2	April 9	Wireless Internet Access Couplers & Combiners IF Circuits	MTT-S International Microwave Symposium
June	May 1	May 8	Part 15 Wireless Devices Cables & Connectors 5 GHz Design Techniques	IEEE AP-S Symposium
July	June 1	June 8	Digital Broadcasting Transistors Filter Design	
August	July 2	July 9	Wireless Consumer Electronics Oscillators Diode Circuits & Tehcnology	RAWCON 2001
September	August 1	August 8	Satellite & Space Systems Spectrum & Network Analysis Using CAD/CAE in Design	PCIA European Microwave
October	September 3	September 10	Microcells & Picocells Capacitors & Inductors Digital Modulation	
November	October 1	October 8	Industrial RF & Microwaves Attenuators & Terminations Baseband Circuits	
December	November 1	November 8	New Wireless Technologies Filters Frequency Synthesizers	

** Additional shows and conferences may be added. We also attend several smaller conferences and meetings, but only the major events are listed above.*

Deadlines for News, Calendar and New Products are two weeks before ad closing dates shown above.

Note that editorial coverage will include many topics not mentioned above! Applied Microwave & Wireless publishes articles on many products and technologies relating to RF design methods, microwaves, wireless systems, and specific product areas.

Applied MICROWAVE & WIRELESS

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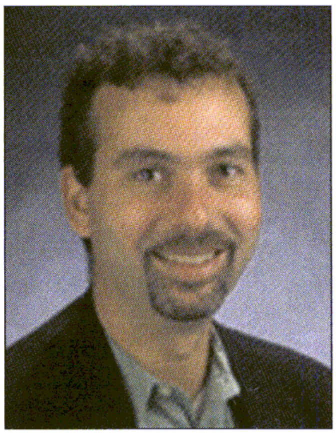
E-Commerce Increases Efficiency of Test and Measurement Purchasing

By Peter M. Ostrow
TestMart

The venerable test and measurement industry has been through dramatic changes in the past few years. Manufacturers are challenged to find more effective and efficient ways to reach and serve their customers. Users of the equipment are following new rules to locate general-purpose and specialized test equipment in a cost-efficient and timely manner. Distributors, brokers and independent manufacturers' sales representatives are redefining themselves and reassessing the value they create and the costs associated with their businesses.

This note describes TestMart, an innovative answer to those needs — a "Net marketplace" where buyers, sellers and all interested members of the test and measurement industry meet to exchange information and conduct business. This new e-commerce marketplace creates a new set of rules for the marketing, purchase, use and sale of test equipment, whether it is new, demo or reconditioned and available for sale, lease or rent.



Peter M. Ostrow is President and CEO of TestMart. Before joining the company, he was Senior Vice President at Narrowline, an Internet advertising transaction and research company. He has held positions at The New York Times and Prudential Securities. He has a BA in Chinese Language from Washington University and an MBA from New York University.

TestMart is located at 851 Traeger Avenue, San Bruno, CA 94066; Tel: 650-624-0525; Fax: 650-624-0535; E-mail: info@testmart.com; Internet: www.testmart.com.

The new rules

Rule #1: Universal — serves all manufacturers and customers. The Internet has become nearly universal. Almost every manufacturer has a Web site and their customers use the Web frequently. TestMart complements manufacturers' other marketing activities with additional information and convenient order processing. Customers have full access to product, price and availability information without needing to be a "major account."

Rule #2: Affordable — lower cost distribution channel. E-commerce is powerful, but maintaining a viable Web site can be expensive, stretching tight budgets in large companies and beyond the reach of many smaller ones. TestMart allows any manufacturer to participate in a best-in-class e-commerce system.

Rule #3: Neutral — reliable, unbiased product comparisons. TestMart presents normalized unbiased data that lends product exposure to manufacturers and gives customers the power to make informed purchasing decisions. Customers who may be skeptical of a manufacturer's own comparisons will have more confidence in this neutral data.

Rule #4: Convenient and secure — price and availability on line. Providing a central online source of reliable ordering information is the primary mission of TestMart. Price and availability data is updated in real time using secure and controlled remote access. A customer can gather the necessary data and complete a secure transaction at one site.

Rule #5: Specialized community. TestMart is also a source of information and ideas from leading publishers and analysts for the entire industry. Customers can visit the site frequently to keep up with news, application notes, white papers and reference libraries. Links to other resources make TestMart a valuable entry point for anyone needing test and measurement information.

Rule #6: A secondary market for asset management. Customers can explore rental and lease options on the site when equipment is only needed for a short time or when they prefer not to own. The site also offers a unique low-cost outlet for used and excess inventory, with a full-service secure ordering system, a significant improvement over most auction and brokerage activities.

A test and measurement marketplace

TestMart has been specifically designed to meet the needs of the test and measurement marketplace. It serves the leaner, more focused manufacturers who must reach large numbers of new customers located throughout the world. It satisfies customers with convenient, reliable access to product data, with accurate price and availability information. It supports wise asset management with a viable secondary market for both buyers and sellers. TestMart is accessible to everyone with an interest in test and measurement. The Web address is *www.testmart.com.* ■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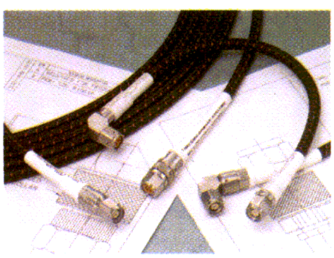

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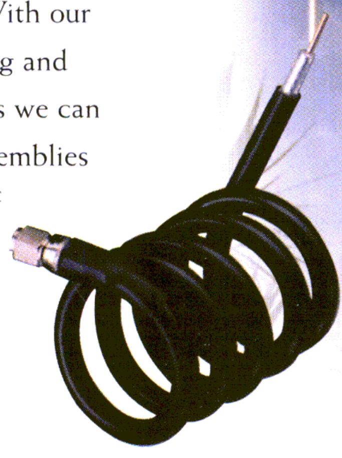

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