October 2000

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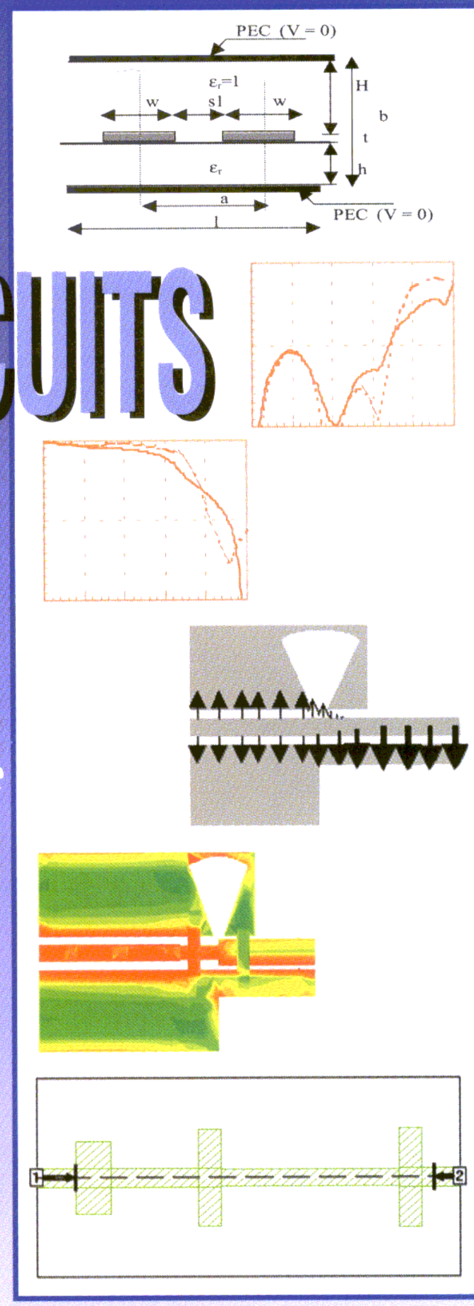
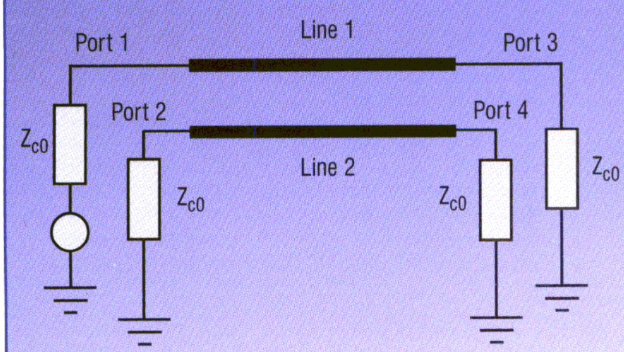
www.amwireless.com



PASSIVE CIRCUITS

Couplers, Baluns and Matching Networks are Keys to Performance



Analytical Techniques

Finite Element Analysis of Planar Couplers

Circuit Design

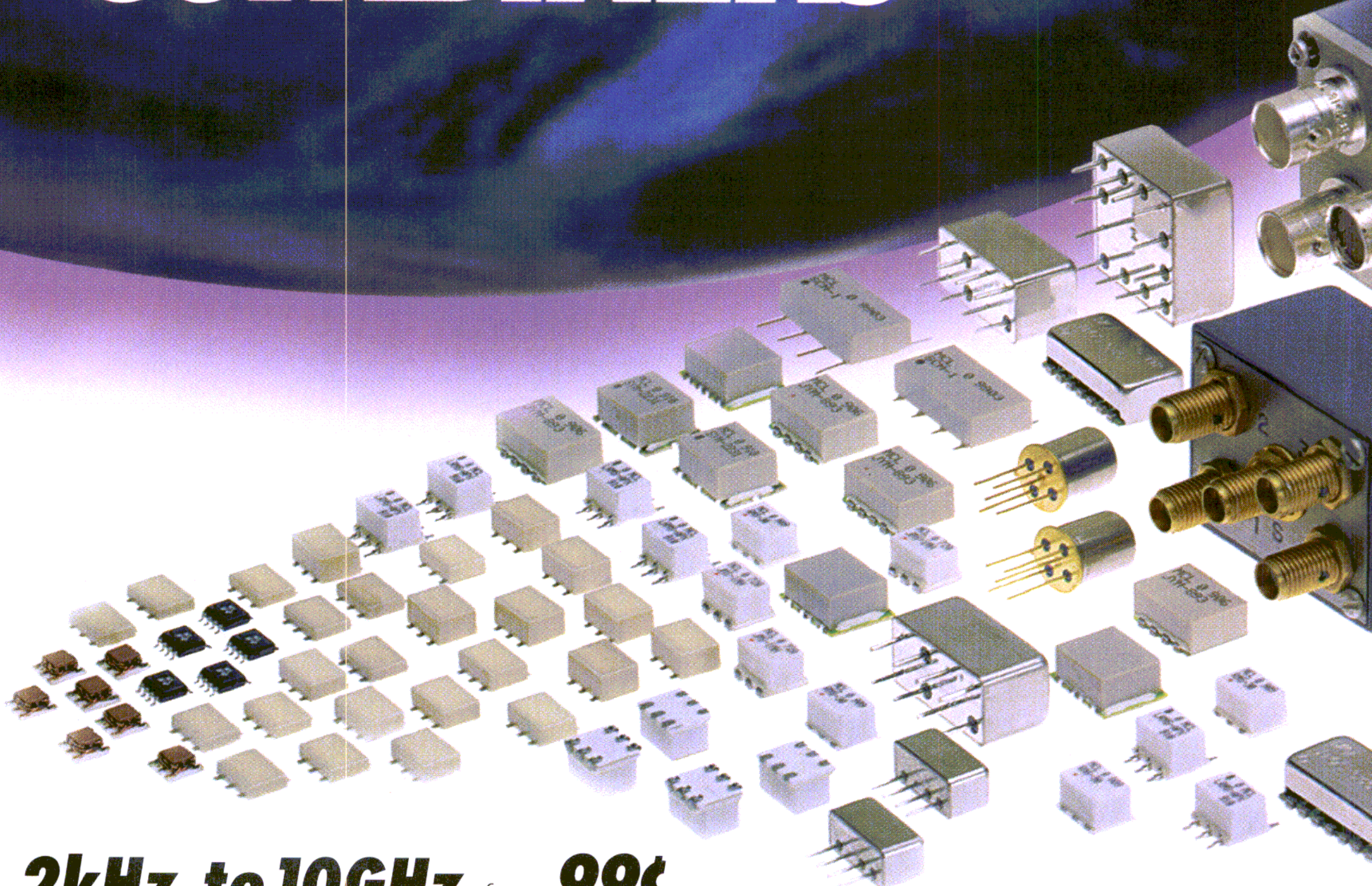
A T/R Diversity RF Switch Design Using PIN Diodes

Guest Editorial

The Gold vs. Aluminum War Revisited

THE WORLD'S LARGEST SELECTION

POWER SPLITTERS/ COMBINERS



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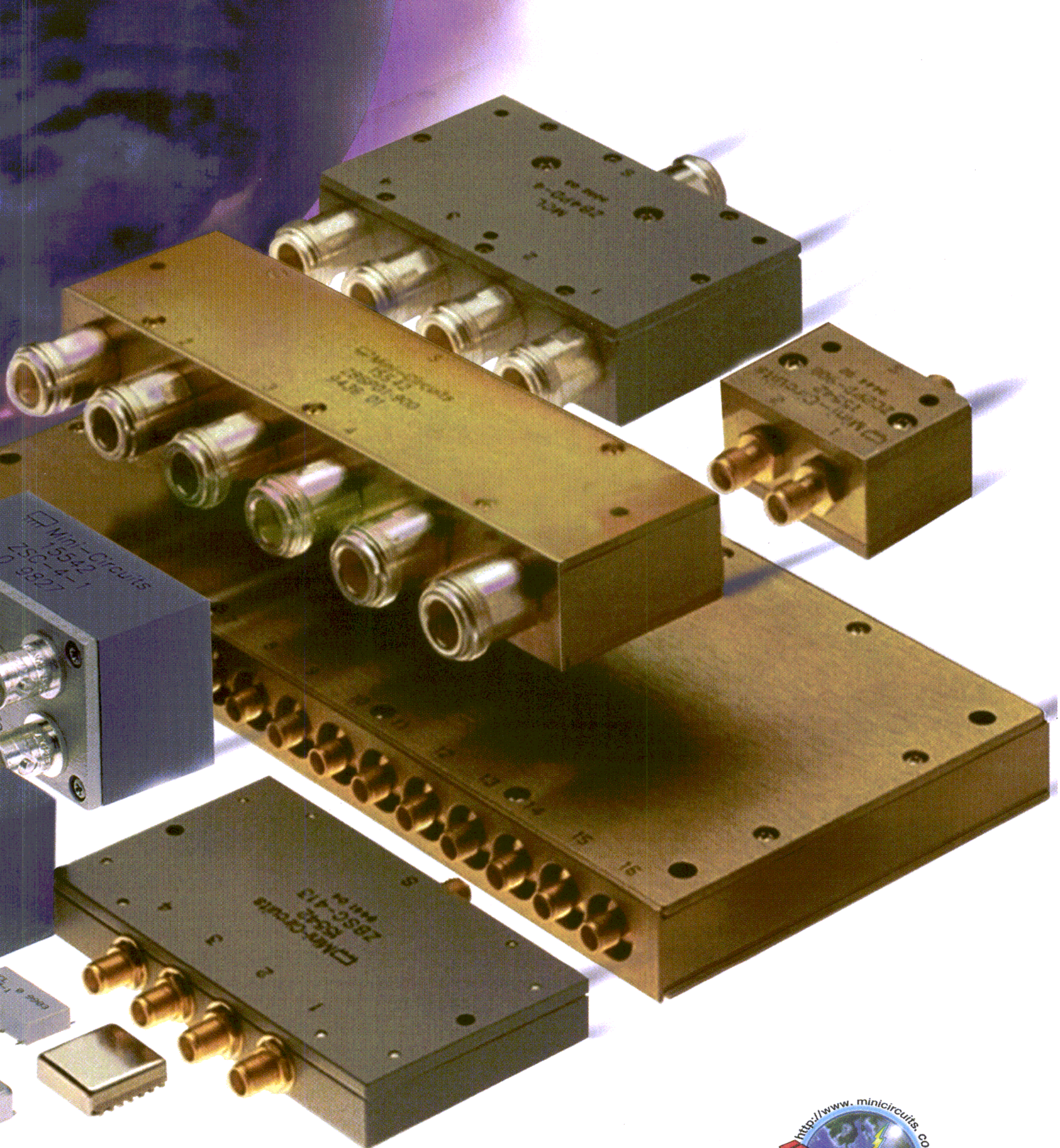
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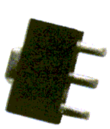

With power to spare, giving you freedom from design constraints, this amplifier was designed and manufactured using state-of-the-art Gallium Arsenide heterojunction bipolar transistor (GaAs HBT) process technology. These devices are ideal for use as a driver stage for power amplifiers installed in cellular PCS infrastructure equipment and conform to TDMA, CDMA and PCS 1900 modulation standards. The SXT-289 operates from a single 5VDC supply and offers exceptional linearity performance in a small form-factor plastic SOT-89 with backside ground.

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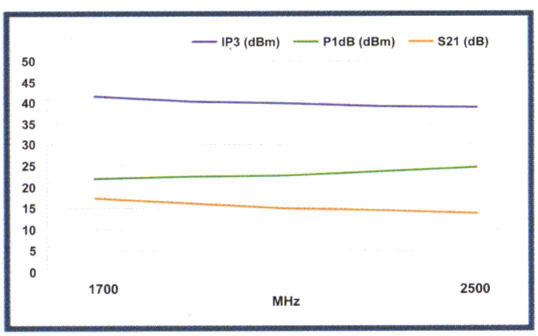

Typical device performance. Bias =5V @ 110mA typ.

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Circle 25

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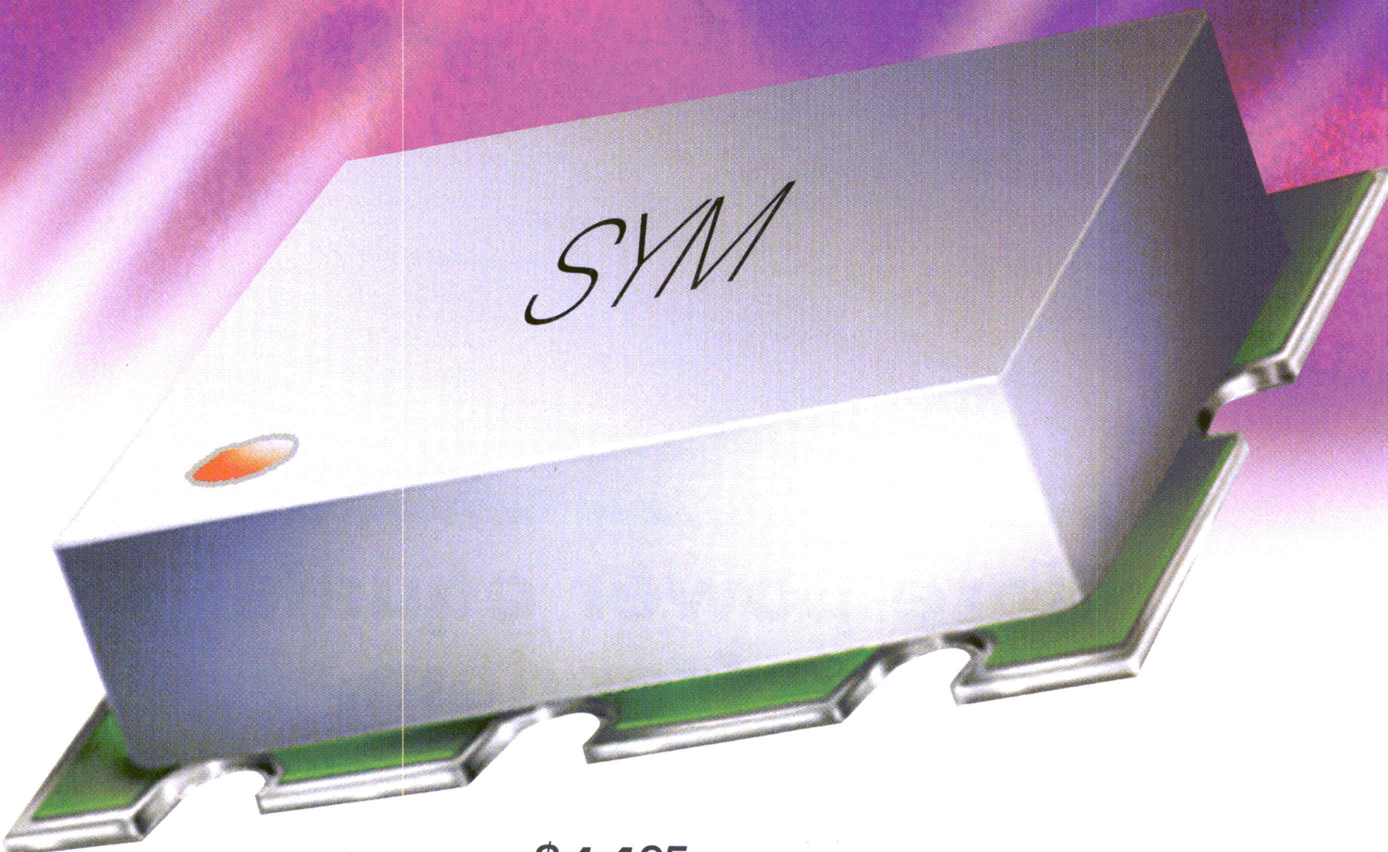

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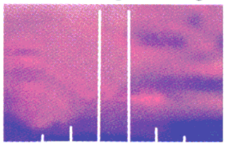

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SYM-14H	100-1370	30	36	30	6.5	14.95
SYM-10DH	800 -1000	31	45	29	7.6	17.80
SYM-22H	1500 -2200	30	33	38	5.6	18.75
SYM-20DH	1700-2000	32	35	34	6.7	14.95

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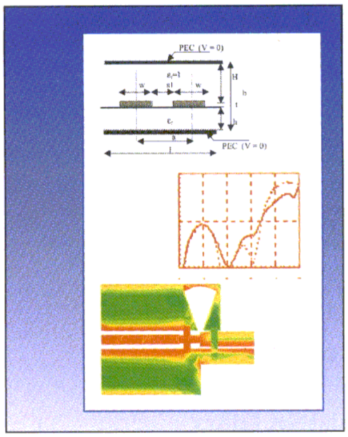


On Our Cover

Passive Circuits

Couplers, baluns and natching networks make wireless products work.

This issue features technical articles analyzing couplers, designing wideband antenna balun feed circuitry and developing accurately-modeled matching circuits for microwave amplifiers.

TECHNICAL FEATURES

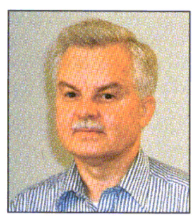
PRODUCTS & TECHNOLOGIES

October 2000
Vol. 12 No. 10

Applied MICROWAVE & WIRELESS

PRODUCTS & TECHNOLOGIES

GUEST EDITORIAL



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Editorial

How Do We Get Kids Interested in Wireless Technology Careers?

By Gary A. Breed
Publisher

According to Marc Holtzman, Secretary of Technology of the State of Colorado, his state has 15,000 unfilled technology jobs. The jobs range from technician to PhD, in many high-tech industries including wireless communications. We can assume that most parts of the U.S. (and around the world) are experiencing the same demand for technical professionals. Where will the industry find all these technicians, engineers, scientists?

Based on the comments of a panel of educators at the recent RAWCON2000 conference, only part of the need will be filled by universities with specialized wireless programs. Their enrollment is growing, but not enough to meet industry's need for engineering talent. While more students are choosing a wireless engineering career path instead of computers and software, all technology industries need many more students to enroll and complete their engineering degrees.

Another contribution to the job pool can be made by fostering higher awareness of the value of a wireless career path at those engineering schools that have traditional degree programs. According to Misty Baker, Director of the Global Wireless Education Consortium (GWEC), these non-specialized engineering programs have a 60 percent dropout rate after the freshman year. That statistic confirms anecdotal evidence that a better focus on future opportunities helps keep students interested.

Getting large numbers of new engineering students will take several years at best, and will only happen if we initiate some dramatic educational and promotional activities in grades K through 12. The state of Colorado has run a pilot Internet Camp summer program, and GWEC is in the early stages of a middle school curriculum assistance program. We can hope that many more ideas will be implemented to instill in today's grade school students an awareness and enthusiasm for high tech as a career path.

World War II, the Space Race and the Cold War combined daily news with technology development, generating a high level of excitement that propelled many present industry leaders into their careers. But there is nothing comparable in the political world of 2000. If we want to replace today's technology geniuses as they retire, we must find other ways to get out the message that a career in RF and microwave technology is exciting, challenging and rewarding. ■

Calendar

CONFERENCES

OCTOBER

October 2-6, 2000
European Microwave Week
Paris, France
Information: Nicole Jedrej
Tel: +44 171 861 6391; Fax: +44 171 861 6251
E-mail: nicola.jedrej@unmf.com
Internet: http://www.eumw.com

October 16-18, 2000
Convergence 2000 — International Congress on Transportation Electronics
Detroit, MI
Information: Carol Lynn
Fax: 313-882-6942
Internet: http://www.convergence2000.org

October 16-19, 2000
ICSPAT — International Conference on Signal Processing Applications and Technology
Dallas, TX
Information: Joan Watkins
Tel: 1-800-789-2223 or 972-906-6800
E-mail: dspworld@mfi.com
Internet: http://www.icspat.com

October 16-20, 2000
AMTA — Antenna Measurement Techniques Association Annual Meeting and Symposium
Philadelphia, PA
Information: Irene Honeycutt
Tel: 215-674-5100; Fax: 215-674-1102
E-mail: irene@orbitfr.com
Internet: http://www.amta.org

October 22-25, 2000
MILCOM 2000 — IEEE Military Communications Conference
Los Angeles, CA
Information: MILCOM
Internet: http://www.milcom2000.org

October 23, 2000
The Fundamentals of EMC: A Tutorial and Exhibition
Research Triangle Park, NC
Information: Jim Marley
Tel: 919-549-1408; Fax: 919-547-6212
E-mail: jim.marley@ieee.org

October 23-24, 2000
Bluetooth Conference
Edinburgh, Scotland
Information: Julie Swanson, Miller Freeman
Tel: + 44 20 7861 6370; Fax: + 44 20 7861 6370
E-mail: jswanson@unmf.com
Internet: http://www.bluetoothconf.com

October 23-24, 2000
New Technologies for RF Circuits
Edinburgh, Scotland
Information: Julie Swanson, Miller Freeman
Tel: + 44 20 7861 6370; Fax: + 44 20 7861 6370
E-mail: jswanson@unmf.com
Internet: http://www.rfconf.com

October 23-25, 2000
33rd Connector and Interconnection Technology Symposium and Trade Show
Orlando, FL
Information: EIA
E-mail: pwalsh@eia.org
Internet: http://www.eia.org

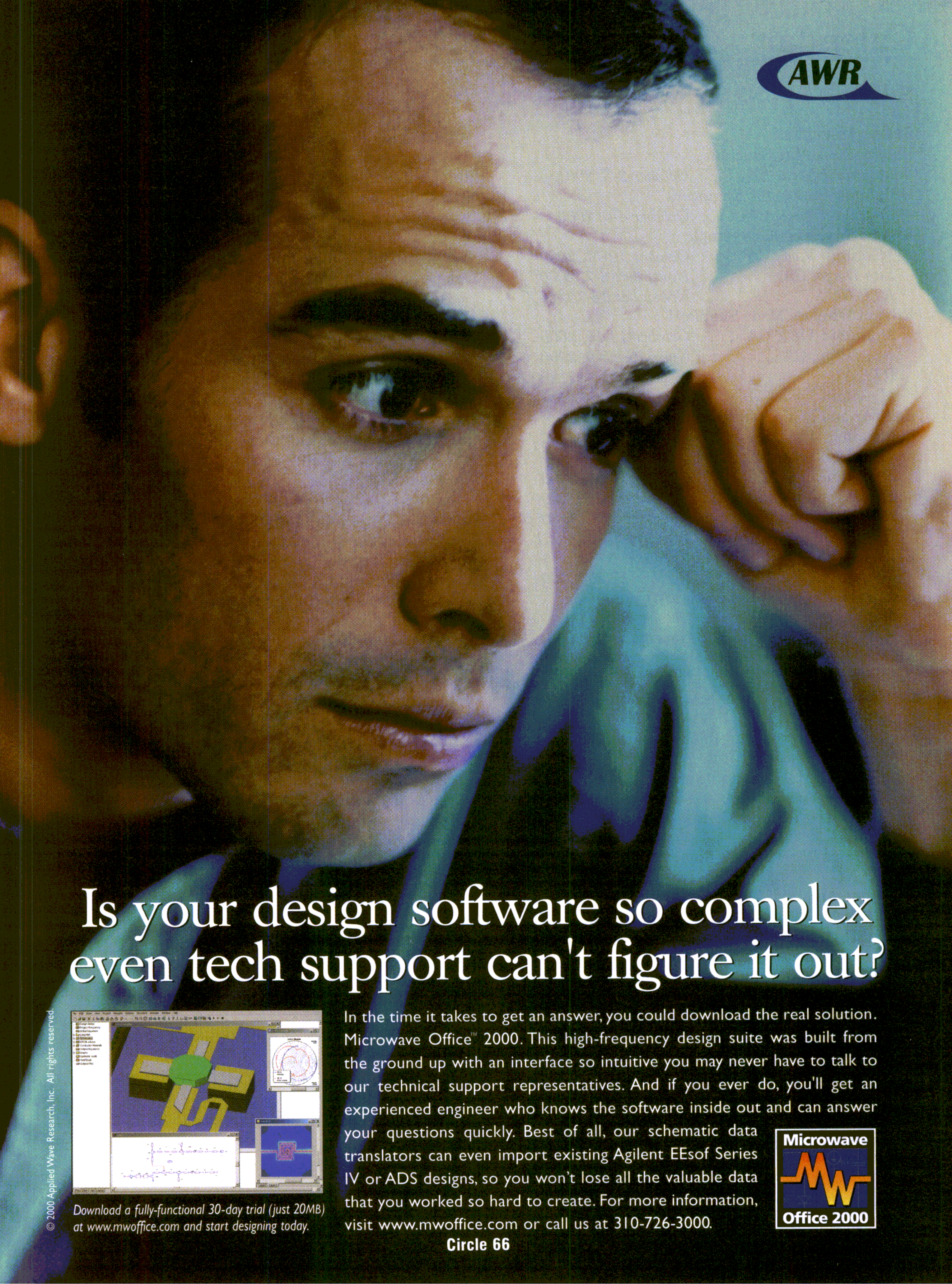
NOVEMBER

November 5-8, 2000
2000 IEEE GaAs IC Symposium
Seattle, WA
Information: Kevin Kobayashi
Tel: 562-627-0765
E-mail: kkobayashi@stanfordmicro.com
Internet: http://www.gaasic.org

November 6-8, 2000
2000 IEEE AP-S Conference on Antennas and Propagation for Wireless Communication
Waltham, MA
Information: Prof. Christos Christodoulou
Tel: 505-277-6580; Fax: 505-277-1439
E-mail: christos@eece.unm.edu
Internet: http://www.eece.unm.edu/apwc2000

November 13-14, 2000
8th IEEE International Symposium on Electron Devices for Microwave and Optoelectronic Applications
Glasgow, Scotland
Information: Iain Thayne
E-mail: edmo@elec.gla.ac.uk
Internet: http://www.elec.gla.ac.uk/edmo2000

November 30-December 1, 2000
56th Automatic RF Techniques Group Microwave Measurements Conference
Boulder, CO
Information: ARFTG
Tel: 303-497-3138
Fax: 303-497-3970
E-mail: dylan@bldrdoc.gov
Internet: http://www.arftg.org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Calendar

SHORT COURSES

Besser Associates

Bluetooth: An Introduction
Mountain View, CA . . . . . .October 2-3, 2001
DSP Made Simple for Engineers
Mountain View, CA . . . . . .October 4-6, 2000
RF and Wireless Made Simple
Mountain View, CA . . . . . .October 10-11, 2000
San Diego, CA . . . . .November 6-7, 2000
RTP, NC . . . . .November 14-15, 2000
RF and Wireless Made Simple II
Mountain View, CA . . . . . .October 12-13, 2000
San Diego, CA . . . . .November 8-9, 2000
RTP, NC . . . . .November 16-17, 2000
Mountain View, CA . . . . . .January 10-11, 2001
Cellular and PCS Design: the Radio Interface
Mountain View, CA . . . . . .October 16-20, 2000
Applied RF Techniques I
Mountain View, CA . . . . . .October 16-20, 2000
San Diego, CA . . . . .November 6-10, 2000
RTP, NC . . . . .November 13-17, 2000
Orlando, FL . . . . .January 22-26, 2001
RF Transceiver Design
Mountain View, CA . . . . . .October 17-20, 2000
All About 3G (Third Generation Wireless)
Mountain View, CA . . . . . .October 23, 2000
RTP, NC . . . . .November 13, 2000
RF Test Equipment Operation (laboratory course)
Mountain View, CA . . . . . .October 31, 2000
RF Testing for the Wireless Age (laboratory course)
Mountain View, CA . . . . . .November 1-3, 2000
Frequency Synthesis and Phase-Locked Loop Design
Mountain View, CA . . . . . .November 6-7, 2000
Wideband CDMA Communications
Mountain View, CA . . . . . .November 8-9, 2000
Advanced Wireless and Microwave Techniques
RTP, NC . . . . .November 13-17, 2000
Short Range Wireless and Bluetooth
RTP, NC . . . . .November 13-15, 2000
Multitone Amplifier Design
RTP, NC . . . . .November 16-17, 2000
RF and High-Speed PC Board Design Fundamentals
Mountain View, CA . . . . . .November 27-29, 2000
Broadband Networking Made Simple
Mountain View, CA . . . . . .November 28-29, 2000
Electro-Optical Fundamentals
Mountain View, CA . . . . . .November 28-29, 2000
Behavioral Modeling
Mountain View, CA . . . . . .Nov. 29-Dec. 1, 2000
RF CMOS Design
Mountain View, CA . . . . . .December 7-8, 2000
Frequency Synthesis Technology and Applications in Wireless Systems
Mountain View, CA . . . . . .December 11-13, 2000
Practical Design of Integrated and Discrete Wireless Circuits
Mountain View, CA . . . . . .December 4-6, 2000
Adaptive Receivers for Wireless Communications
Mountain View, CA . . . . . .December 4-8, 2000
Information: Annie Wong, Tel: 415-949-3300; Fax: 415-949-4400; E-mail: info@bessercourse.com; Internet: www.bessercourse.com.

Georgia Institute of Technology

Adaptive Array Radar Processing
Atlanta, GA . . . . .October 3-6, 2000
Principles of Modern Radar
Atlanta, GA . . . . .Oct. 30-Nov. 3, 2000
Information and Control Techniques for Intelligent Autonomous Vehicles
Atlanta, GA . . . . .October 25-27, 2000
Information: Georgia Tech Distance Learning, Continuing Education and Outreach, Tel: 404-894-2547; Fax: 404-894-7398; E-mail: conted@gatech.edu; Internet: www.conted.gatech.edu.

University of California at Berkeley Extension

Design of Analog Integrated Circuits for Mixed-Signal Integrated Systems
San Francisco, CA . . . . . .October 5-7, 2000
Phase-Locked Loop (PLL) Systems
San Francisco, CA . . . . . .October 19-21, 2000
Low-Power Circuits and Systems for Digital Wireless Communications
Redwood City, CA . . . . . . .October 30-31, 2000
IP/ATM Networks: Switching, Routing, and Internetworking
San Francisco, CA . . . . . .Nov. 29-Dec. 1, 2000
Modern Telecommunications
San Francisco, CA . . . . . .December 4-5, 2000
Wireless Networks and the Evolving Telecommunications Infrastructure
San Francisco, CA . . . . . .December 11-13, 2000
Information: Continuing Education in Engineering, Tel: 510-642-4111; Fax: 510-642-0374; E-mail: course@unex.berkeley.edu; Internet: www.unex.berkeley.edu/enroll.

RTT Programmes Limited

RF Amplifier Design
London, England . . . . . . .October 9-11, 2000
RF Oscillator Design
London, England . . . . . . .October 16-18, 2000
RF/IF Processing
London, England . . . . . . .Oct. 20-Nov. 1, 2000
RF System Engineering
London, England . . . . . . .November 13-15, 2000
3G Technology
London, England . . . . . . .November 20-22, 2000
SMR/PMR Design
London, England . . . . . . .December 4-6, 2000

Information: Lorraine Gannon, Tel: +44 181 844 1811; Fax: +44 181 751 2616; E-mail: seminars@rttsys.com; Internet: www.rttsys.com.

Research Associates of Syracuse, Inc.

ELINT Interception and Analysis
Stockholm, Sweden . . . . .October 10-13, 2000
RF and Microwave Receiver Design
Stockholm, Sweden . . . . .October 24-27, 2000
Radar Fundamentals and FMCS Radar
Stockholm, Sweden . . . . .October 25-27, 2000
Digital Wideband Receivers
Stockholm, Sweden . . . . .Oct. 30-Nov. 1, 2000
Phased Array Radar
Stockholm, Sweden . . . . .November 14-16, 2000
Information: Mary Chamberlain or Richard Wiley, Tel: 315-463-2266; E-mail: seminars@ras.com; Internet: www.ras.com.

University of Wisconsin at Milwaukee

Introduction to Electromagnetic Compatibility Design Practices
Northbrook, IL . . . . .October 12-13, 2000
Information: Loraine Samsel, Program Assistant, Tel: 1-800-222-3623; Fax: 1-800-399-4896; E-mail: samsel@uwm.edu; Internet: www.uwm.edu/dept/ccee.

University of California at Los Angeles Extension

Charge-Coupled Devices/CMOS Imaging Sensors and Cameras
Los Angeles, CA . . . . .November 13-17, 2000
Automatic Test Equipment (ATE) Selection, Design and Programming
Los Angeles, CA . . . . .December 4-5, 2000
Design for Testability and Built-In Self Test
Los Angeles, CA . . . . .December 6-8, 2000
Information: UCLA Extension, Short Course Program Office, Tel: 310-825-3344; Fax: 310-206-2815.

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More events listed at www.amwireless.com

Calendar

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2001 IEEE MTT-S International Microwave Symposium

May 20-25, 2001 — Phoenix, AZ

Topics: Electromagnetic field analysis, computer aided design, passive components and circuits, active components and ICs, millimeter wave and lightwave technologies, package, integration, testing and applications.

Authors should complete a registration form and four-page submission template, both available at the conference Web site, www.ims2001.org. Hard copy and fax submissions will not be accepted.

Deadlines: November 27, 2000 (.doc and .ps formats)
December 4, 2000 (.pdf format)

ISSSE 2001 — 2001 International Symposium on Signals, Systems and Electronics

July 24-27, 2001 — Tokyo, Japan

Topics: Software-oriented radio transmitters/receivers, wireless channel equalization, signal detection and interference cancellation, advanced wireless radio networks, hardware/software cooperation in radio equipment, diversity/RAKE reception, anti-fading techniques, radio-frequency synthesizers, digital signal processing, DSP for smart antennas, optical signal processing, large-capacity optical transmission, integrated modules and elements, hardware-oriented signal processing and compression, advanced technologies for RF circuits, devices for microwaves and photonics, microwave/photonic integrated devices and systems, device modeling and design, numerical and CAD techniques, millimeter-wave applications, cooperation of optical and microwave guides, and full-scale systems on a single chip.

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16th International Conference on Applied Electromagnetics and Communications

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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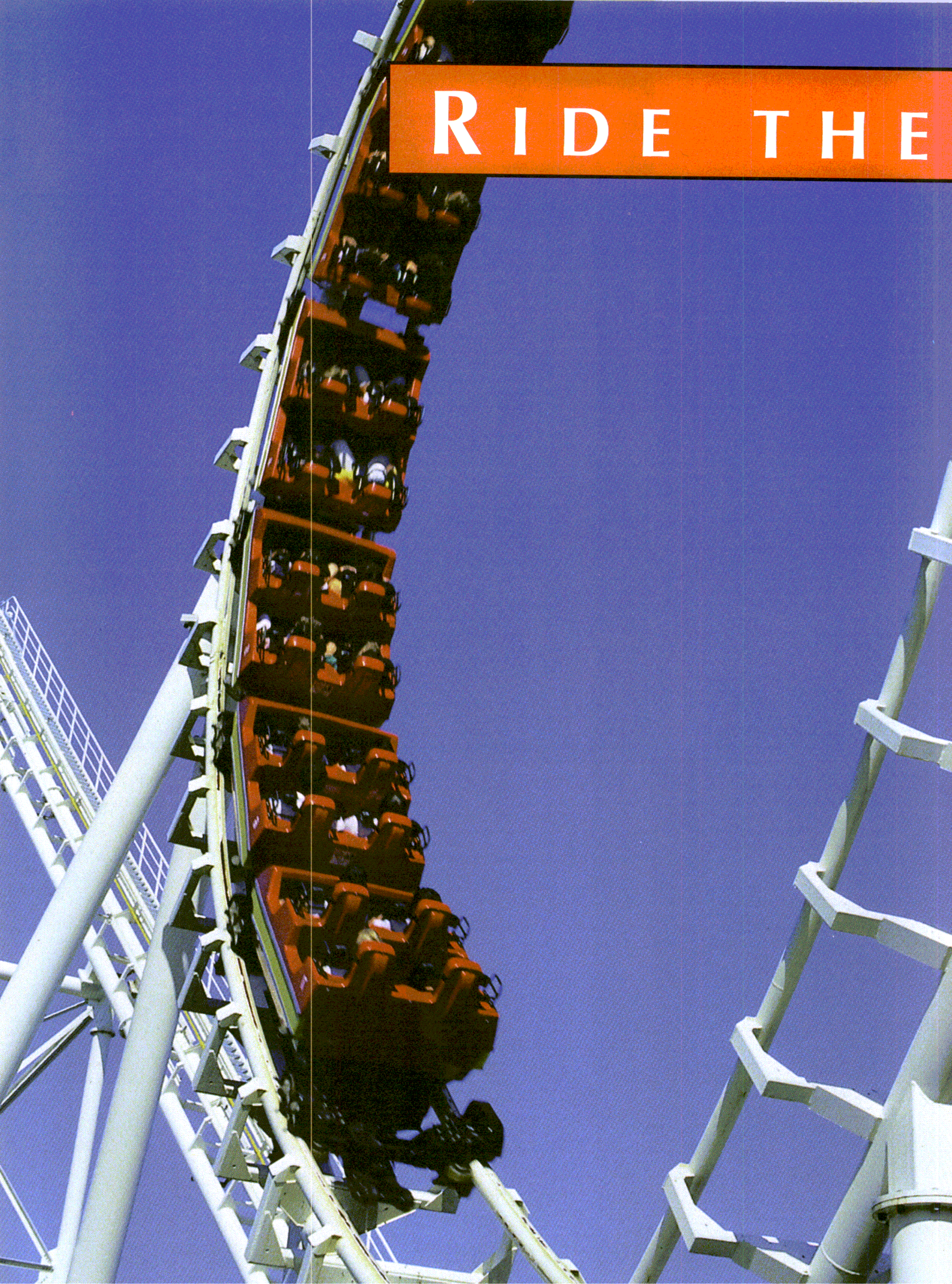


BRIEFS

• Agilent Technologies Inc. has announced the opening of its new R&D Central web site. The site, at www.agilent.com/find/RandD6, offers information about applications, training, consulting services and instrumentation. The site also provides discussion topics on various subjects that allow product developers to share ideas.

• K&L Microwave has expanded its manufacturing facilities with the purchase of a 64,000-square-foot building near its headquarters in Salisbury, MD.

• Gowanda Electronics has broken ground for a new 33,000-square-foot facility in Gowanda, NY, to house the company's manufacturing, engineering, research and administration offices. The company's three current facilities will be moved to the new site.

• Agilent Technologies Inc. will add a dedicated 6-inch wafer fabrication line at its Newark, CA, facility. The company will also add high-volume automated back-end manufacturing capabilities at its facility in Penang, Malaysia.

• Endwave Corporation is expanding two of its manufacturing facilities, in Santa Clara and Diamond Spring, CA, more than doubling the company's previous production capacity.

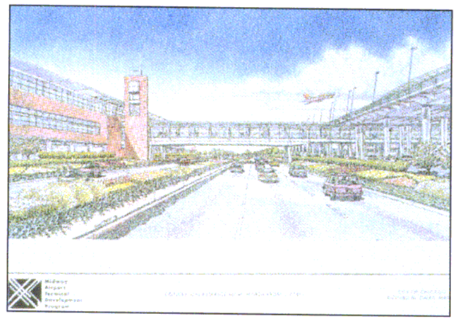
• Tality Corporation has opened a new design site near New Delhi, India. The site's initial focus is system-on-a-chip design.

• TriQuint Semiconductor Inc. has completed the purchase of a wafer fabrication facility in Richardson, TX, from Micron Technology Texas, LLC.

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

Andrew to build wireless network for new Midway airport terminal

Andrew Corporation has won a contract to supply the radio frequency infrastructure for the new Midway Airport terminal in Chicago, IL. The terminal is scheduled to be complete in 2001.



▲ Artist's rendering of the new terminal at Midway Airport in Chicago.

The contract includes design, materials and testing for the network, which will carry 154 MHz communications for the fire department; 460 to 468 MHz communications for police and emergency services; and 476 to 480 MHz trunked radio services for the Chicago Department of Aviation.

The project will deploy more than 10,000 feet of RADIAX® RCT-Series tuned radiated cable, plus supporting passive electronic devices, including couplers and splitters. Engineers from the company's distributed communications services group in Richardson, TX, will design and test the new network.

Andrew, based in Orland Park, IL, supplies communications systems equipment and services for wireless applications.

FCC adjusts rules, deadlines for wireless E911 compliance

The Federal Communications Commission has extended compliance deadlines for wireless E911. Under the new guidelines, the date for carriers to begin selling and activating Automatic Location Identification (ALI) capable handsets has been extended from March 1, 2001, to October 1, 2001.

Carriers also now have until November 9, 2000, to file E911 Phase II implementation reports.

New compliance dates included in the E911 schedule:

• December 31, 2001 — At least 25 percent of all new handsets activated are to be ALI-capable.

• June 30, 2002 — At least 50 percent of all new handsets activated are to be ALI-capable.

• December 31, 2002 — 100 percent of all new digital handsets activated are to be ALI-capable.

The new schedule gives carriers until December 31, 20005, to reach full penetration of ALI-capable handsets in their subscriber bases. Also, the operational definition of full penetration has been modified from "reasonable efforts" to achieve 100 percent penetration of ALI-capable handsets, to a requirement that 95 percent of all handsets in a carrier's total subscriber base be ALI-capable.

RF Micro Devices to expand product line with InP process

RF Micro Devices Inc., based in Greensboro, NC, plans to begin developing and manufacturing integrated circuits using indium phosphide (InP), a next-generation semiconductor process technology. The addition of InP will extend the company's product lines to include five major process technologies — InP, GaAs HBT, silicon germanium, silicon BiCMOS and GaAs MESFET.

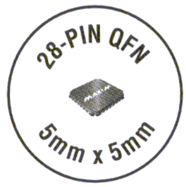
InP is widely viewed as having performance characteristics superior to those of existing technologies. For example, InP has been shown to have better thermal characteristics, higher frequency response and lower threshold voltages.

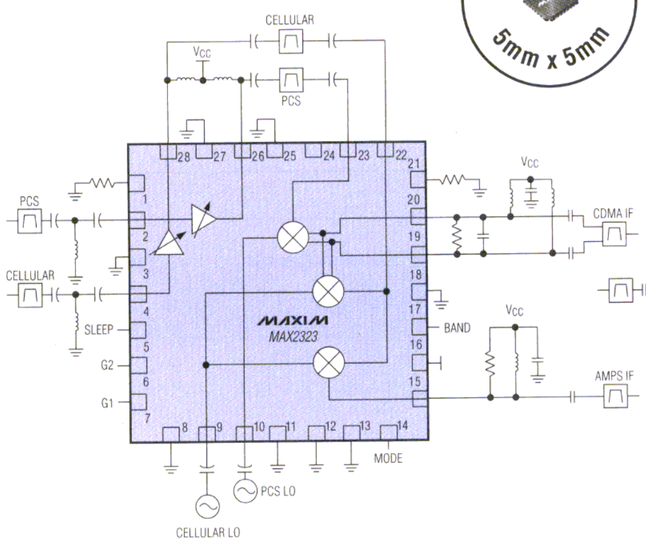
WORLD'S SMALLEST DUAL-BAND CELL PHONE LNA + MIXER IC HAS HIGHEST IP3

SiGe IC in 5mm x 5mm QFN Package Sets New Industry Standard

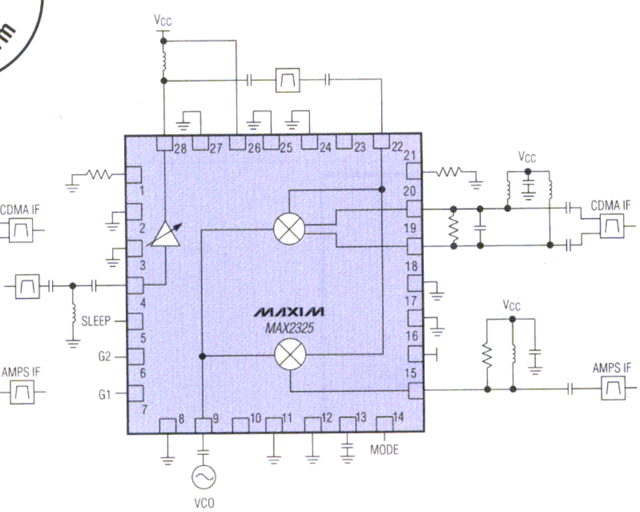
The MAX2323 low-noise amplifier (LNA) plus mixer is designed for dual-band CDMA cellular phones and can also be used in dual-band TDMA, GSM, or EDGE cellular phones. It includes all circuitry needed from the antenna to the IF filter in a miniature QFN package. The LNAs have unprecedented input IP3 and the mixers have very high gain, which is essential to meet sensitivity goals. Both LNAs have switched gain states to conserve power and increase dynamic range. The low-band LNA has three gain states for optimum IP3 margin in CDMA systems. The MAX2323 addresses dual-band, tri-mode applications, and the MAX2325 is a pin-compatible cellular-band version.

- **Over 26dB Overall Gain**
- **<2.6dB Cascade PCS Noise Figure (Includes 3dB Interstage Filter Loss)**
- **Ultra-Small 28-Pin MLF2 Package**
- **Multiple Power-Saving Modes, Including a 17mA Paging Mode**
- **Switchable LNA Gain**
- **>+10dBm LNA IIP3 Adjustable to >+13dBm IIP3**





Very high input IP3 and the industry's smallest package make the MAX2323 unique.



The cellular-band only MAX2325 is pin compatible with the MAX2323.

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BUSINESS AND FINANCE

Space Electronics acquires instruments group

Space Electronics Inc., a part of Maxwell Technologies' Electronic Components Group, has announced the acquisition of Shason Microwave Corporation's Instruments Group. The addition includes a line of automated test equipment and provides Space Electronics with the ability to offer RF/microwave components for satellite applications

Space Electronics, based in San Diego, CA, is a fabless semiconductor company supplying microelectronics for space and hi-rel applications.

Andrew receives initial amplifier order from Lucent

Andrew Corporation has announced an initial $3 million order from Lucent Technologies Inc., for the development and supply of single-channel PCS-TDMA linear power amplifiers. The order is part of a $23 million letter of agreement between the companies.

Under the order, Andrew will deliver the first consignment of prototype amplifiers in the fall of 2000 and will move into full production at the beginning of 2001. Lucent will use the amplifiers in its TDMA PCS Minicell systems, used in wireless networks to increase the efficiency of cellular telephone systems.

Andrew, based in Orland Park, IL, supplies communications systems equipment and services for wireless applications.

Radio Frequency wins contract in Hong Kong

RFS Radio Frequency Systems has received a contract to add third generation mobile wireless services to the MTRC metro transportation system in Hong Kong. The contract's value was not disclosed.

RFS, based in Charlotte, NC, manufactures antenna and cable systems and RF subassemblies.

Wi-LAN signs network supply agreement

Wi-LAN Inc. has signed an 18-month supply agreement with Wenatchee, WA-based Northwest Telephone Inc., calling for the construction of a new broadband wireless network in central Washington state. The agreement is valued at $1.75 million.

Wi-LAN, based in Calgary, Alberta, Canada, supplies high-speed wireless data communications services.

BUSINESS AND FINANCE

Motorola announces contracts, agreements

Motorola has announced several new contracts and agreements to provide of wireless networks and equipment worldwide, as well as a partnership designed to improve its wireless internet services.

• A five-year, $45 million agreement with GoAhead Software Inc. will include the incorporation of GoAhead's SelfReliant software into Motorola's Aspira™ end-to-end solutions for wireless internet. The partnership will offer advanced reliability and availability of wireless networks for the internet.

• As part of an agreement with the Cape Metropolitan Council of Capetown, South Africa, Motorola will provide equipment for the country's first digital two-way radio communication infrastructure. Alcom, a subsidiary of the Altech group, will supply the complete TETRA (Terrestrial Trunked Radio) network.

• Seven separate contracts with China Unicom, valued at a total of $258 million, call for the expansion of China Unicom's GSM 900 networks in five of the country's major provinces: Guangdong, Fujian, Jiangxi, Shaxi and Xinjiang.

• Under a contract with BPL Mobile of Mumbai, India, Motorola will deploy General Packet Radio Service (GPRS) high speed mobile data capability on the existing BPL Mobile GSM network. The value of the contract was not disclosed

• A $20 million contract for Motorola's Global Telecom Solutions Sector will provide a Code Division Multiple Access (CDMA) network in Uruguay. The agreement is with South American cellular provider Movicom BellSouth Uruguay.

• A contract with Kuwait GSM operator Mobile Telecommunication Company calls for the expansion and enhancement of the existing GSM network, increasing capacity by 30 percent. The agreement is valued at $28 million.

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

Superconductor Technologies receives systems order

Superconductor Technologies Inc. has received a $7.8 million purchase order for SuperFilter® Systems from United States Cellular Corporation, to be deployed over the next nine quarters.

The SuperFilter System uses high-temperature superconducting technology and proprietary cryogenic cooling to create a front-end filter and amplifier system designed to enhance performance in wireless base stations to enhance their performance.

Based in Santa Barbara, CA, Superconductor Technologies manufactures superconducting products for wireless communications and wireless Internet access. United States Cellular Corporation, headquarters in Chicago, IL, manages cellular systems throughout the United States.

M/A-COM to provide public safety system

M/A-COM Inc. of Lowell, MA, has signed an agreement with the Board of Commissioners of Pennsylvania's Cumberland County to provide a state-of-the-art public safety communications system. The value of the contract was not disclosed.

M/A-COM will provide its OpenSky® Wireless Private Network (WPN), which integrates digital trunked voice and packet data over an IP-based network. The network allows a scalable system that translates to cost-effective design and expansion. It also provides interoperability with analog radios, making it possible to preserve existing public safety equipment while offering upgrades and migration to digital equipment.

M/A-COM, part of Tyco Electronics Corporation, supplies radio frequency, microwave and millimeter-wave semiconductors, components and IP networks.

Lindgren, ETS to combine operations

Lindgren RF Enclosures Inc. and EMC Test Systems, both subsidiaries of ESCO Technologies Inc., are integrating their EMC test and measurement operations into the new ETS-Lindgren™ group. Also, Lindgren's shielding-only business will be expanded to include shielding-only products being transferred from ETS.

ETS-Lindgren will be based in Austin, TX, while Lindgren's shielding business will remain at its headquarters in Glendale Heights, IL.

Lindgren manufactures radio frequency shielding products for EMC, medical, industrial and government markets. ETS produces anechoic absorbers, test chambers, antennas and other test equipment for the EMC, RF/microwave and wireless markets.

Spirent acquires InfoSOFT Technologies

Spirent Communications, formerly known at Telecom Analysis Systems (TAS), has acquired InfoSOFT Technologies Inc. of Ft. Worth, TX, a supplier of CDMA handset test and deployment solutions. Terms were not disclosed.

Spirent, based in Eatontown, NJ, provides a range of test and measurement systems and solutions.

Spacek completes Sierra Wireless acquisition

Spacek Labs Inc. has announced the completion of its acquisition of millimeterwave amplifier manufacturer Sierra Wireless Systems. Financial terms of the cash agreement were not disclosed.

Based in Santa Barbara, CA, Spacek Labs is a supplier of millimeterwave amplifiers.

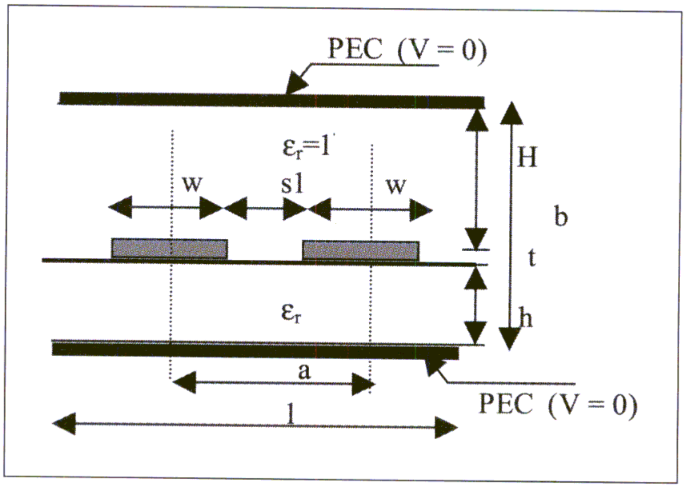
Finite Element Analysis of Planar Couplers

A technical review and analysis of electromagnetic coupling between microstrip lines

By N. Benahmed, M. Feham, and M. Kameche
University of Tlemcen

In this paper, we present the analysis of electromagnetic coupling between two lossy and inhomogeneous microstrip lines. This analysis is based on a numerical resolution of Laplace's equation by the finite element method (FEM), because this complex configuration of lines does not have rigorous analytical resolution. The modelling of this structure is designed to yield the primary parameters (L, C, R, G) of the equivalent electronic circuit and the secondary parameters described by the impedances (Z_{0e}, Z_{0o}) of the even and odd modes and the coupling coefficient k.

As an application, we present the results of the conception of a microwave coupler at the reference frequency 5 GHz. The numerical model developed remains valid to all configurations of structures that propagate the fundamental TEM mode or the quasi-TEM mode.



▲ **Figure 1. Cross section of a coupled line with inhomogeneous microstrip.**

Introduction

The analytical characterization of coupling between two coupled lines with losses and inhomogeneous microstrip is a difficult task. Numerical methods are appropriate to solve this problem. To reach this objective numerically, it is necessary to evaluate the characteristic impedances of the even and odd modes and then the primary parameters L, C, R and G of the equivalent electronic circuit of the structure.

We are interested in the numerical characterization of the coupling coefficient k and the primary parameters of two coupled lines with inhomogeneous microstrips by using the finite element method. Using this technique, we determine the influence of the electrical and the geometric parameters of the analyzed structure. Finally, we present the design of a microwave coupler.

Coupled microstrip lines

A microwave coupler has been analyzed and tested with a coupled line composed of two inhomogeneous microstrips [1]. The electrical properties of the coupler with a TEM mode or a quasi-TEM mode coupled line can be described in terms of even (Z_{0e}) and odd (Z_{0o}) mode impedances. Various numerical techniques can be used to determine the accurate characteristic impedances of the TEM and quasi-TEM coupled lines [2, 3].

In this article, we present a numerical model that analyzes a more complex coupled line. For example, we analyze Figure 1, a structure formed by two inhomogeneous microstrip line without shielding.

Figure 1 shows the cross section of a coupled line with an inhomogeneous microstrip shielded with two parallel plates considered as a perfect electric conductor (PEC). Each line consists of a conductor with the width w and the depth t. A dielectric material with permittivity ε_r ($\mu_r = 1$) fills the inside of the lines. The two lines are separated by the distance s_1.

Numerical resolution

The study in the electrostatic domain of the structures shown in Figure 1 is based on the solution of Laplace's equation in two dimensions for the two modes (even and odd):

$$div\left[\varepsilon_r \nabla_t V(x,y)\right] = 0 \tag{1}$$

Even mode:

$V = 1\ V$ on the two conductors.
$V = 0$ on the shield.

Odd mode:

$V = 1\ V$ on one conductor.
$V = -1\ V$ on the other conductor.
$V = 0$ on the shield.

The solution for this equation is found by using the finite element method (FEM) [4]. The solution for Figure 1 represents the distribution of the potential V at the different mesh nodes of the structure. When the potential V is known, we can calculate the even and odd mode impedances.

Odd and even mode characteristic impedances

The lossless lines theory allows us to determine the electrical field and the magnetic field from the potential V. The electrical energy W_{em} accumulated in the structure is calculated from the electrical field. All of the characteristic impedances (for the two modes) are deduced easily from the electrical energy W_{em}. Consequently, it is important that the exact potential V is found [4]. In the following paragraphs, we show how to study a given mode.

Electrical field — The electrical field is found through simple derivation from the potential V, using the expression

$$\vec{E}_t = -gr\vec{a}d_t(V) \tag{2}$$

The subscript t indicates the cross section of the structure.

Electrical energy — When the structure is composed of several materials of permittivity ε_{ri} (i = 1 to n), the electrical energy is deduced from the electrical field through the relationship

$$\overline{W}_{em} = \frac{1}{4}\sum_{i=1}^{n}\left(\iint \varepsilon_0 \varepsilon_{r_i} \times \vec{E}_{t_i} \times E^*_{t_i} dxdy\right) \tag{3}$$

Capacitance per unit length — This is deduced directly from the electrical energy:

$$C = \frac{4\overline{W}_{em}}{\left(V_1 - V_2\right)^2}(F/m) \tag{4}$$

where V_1 and V_2 represent the fixed potential of the conductors.

Effective permittivity — This is calculated by obtaining the ratio of the electrical energy per unit length accumulated in the inhomogeneous structure and the energy accumulated in the same empty structure.

$$\varepsilon_{eff} = \frac{\iint \varepsilon_0 \varepsilon_r \vec{E}_{t0} \times \vec{E}^*_{t0} dxdy}{\iint \varepsilon_0 \vec{E}_{t1} \times \vec{E}^*_{t1} dxdy} \tag{5}$$

Characteristic impedance — This is calculated using

$$Zc = \frac{1}{v\varphi C}(\Omega) \tag{6}$$

where

$$v_\varphi = \frac{3\times10^8}{\sqrt{\varepsilon_{eff}}} en(m/s)$$

Coupling coefficient — When the even and odd mode characteristic impedances are known, we calculate the coupling coefficient k, using the relationship:

$$k = \frac{Z_{0e} - Z_{0o}}{Z_{0e} + Z_{0o}} \tag{7}$$

Primary parameters

The coupling capacitance γ and the mutual inductance M are deduced from the coupling coefficient k.

$$k = \frac{\gamma}{C_0} = \frac{M}{L} \tag{8}$$

Equation 8 allows us to determine the matrices L, C, R and G using the formulas:

$$[L] = \begin{bmatrix} L & M \\ M & L \end{bmatrix},\ [C] = \begin{bmatrix} C_o + \gamma & -\gamma \\ -\gamma & C_o + \gamma \end{bmatrix} \tag{9a}$$

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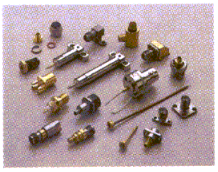

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and

$$[R]=\begin{bmatrix} R & O \\ O & R \end{bmatrix}, \quad [G]=\begin{bmatrix} G & O \\ O & G \end{bmatrix} \tag{9b}$$

L, C_o, R and G are the isolated line parameters calculated numerically using the finite element method.

FEM results

Using the presented theory, we established CAO programs to calculate the coupling coefficient and the matrices $[L]$, $[C]$, $[R]$ and $[G]$ of the coupled lines. All parameters depend on the features and properties of the dielectrics in which the line conductors are embedded.

When the matrices $[L]$, $[C]$, $[R]$ and $[G]$ of the structure are determined, we analyze each structure for a coupler using an adapted numerical model.

Inhomogeneous and coupled microstrip lines between two parallel plates

In order to validate our numerical results, we first studied the inhomogeneous structure presented in Figure 1, already analyzed by the moment method (MM) [1]. The features of this structure are:

• Strip width	$w = 0.85$ mm
• Strip thickness	$t = 0.0254$ mm
• Cover height	$b = 5$ mm
• Substrate thickness	$h = 0.85$ mm
• Distance	$s = 2$ mm
• Separation width	$s_1 = 0.25$ mm
• Substrate permittivity	$\varepsilon_r = 9$
• Substrate loss tangent	$t_{gd} = 0.0001$
• Reference frequency	$f = 5$ GHz
• Conductivity	$s = 5.65 \times 10^7 (\Omega m)^{-1}$

The equipotential lines of the even and odd mode are illustrated in Figure 2.

The primary parameter values obtained are:

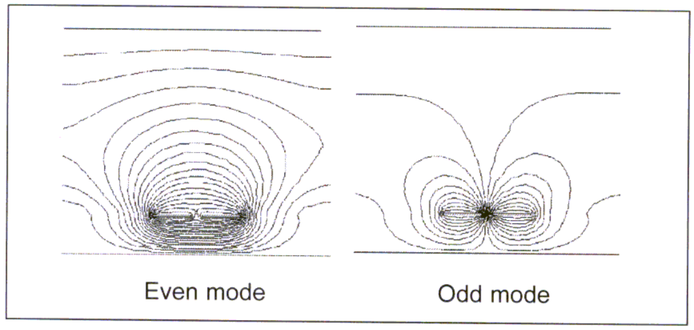
$Z_{0e} = 64.4893\ \Omega$
$Z_{0o} = 34.9464\ \Omega$
$k = 0.297105$

$$[L]=\begin{bmatrix} 419.521 & 124.642 \\ 124.642 & 419.521 \end{bmatrix}(nH/m)$$

$$[C]=\begin{bmatrix} 205.916 & -47.1655 \\ -47.1655 & 205.916 \end{bmatrix}(pF/m)$$

$$[R]=\begin{bmatrix} 19.54 & 0 \\ 0 & 19.54 \end{bmatrix}(\Omega/m)$$

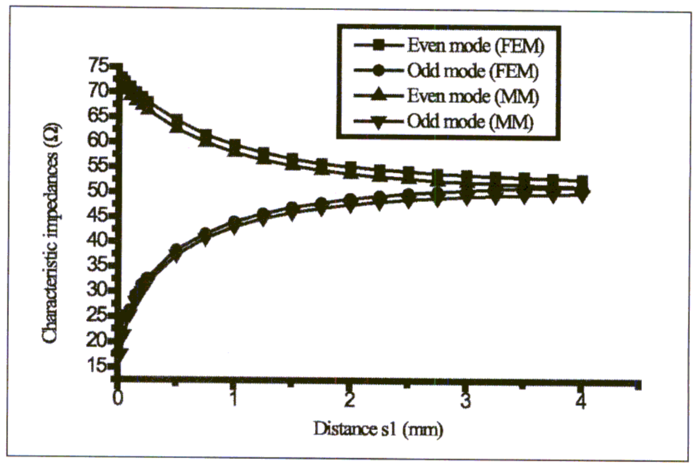
$$[G]=\begin{bmatrix} 498.729 & 0 \\ 0 & 498.729 \end{bmatrix}(\mu S/m)$$



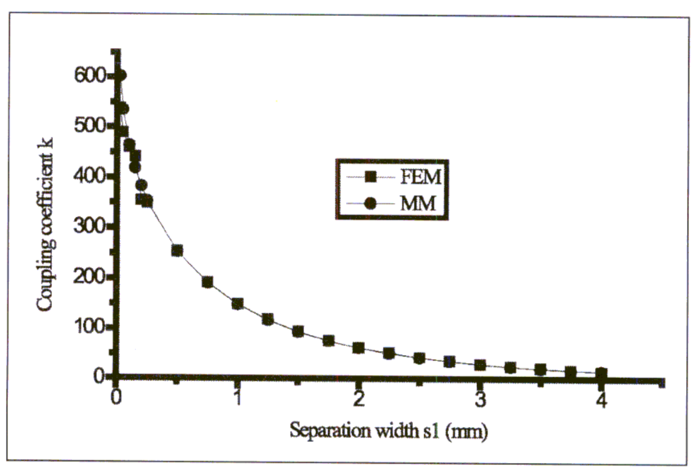
▲ **Figure 2. Equipotential lines of the even and odd modes.**

For the same geometrical and electrical parameters, we obtained the following results using the moment method (MM) [1]:

$Z_{0e} = 63.86\ \Omega$
$Z_{0o} = 34.71\ \Omega$
$k = 0.295729$



▲ **Figure 3. Impact of the separation width on the even and odd mode characteristic impedances.**



▲ **Figure 4. Impact of the separation width on the coupling coefficient.**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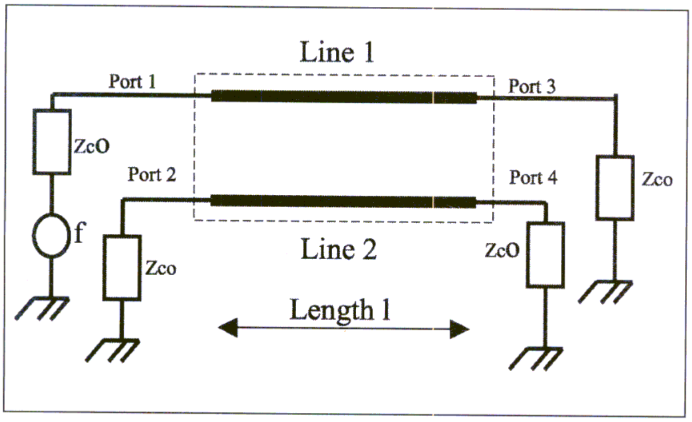
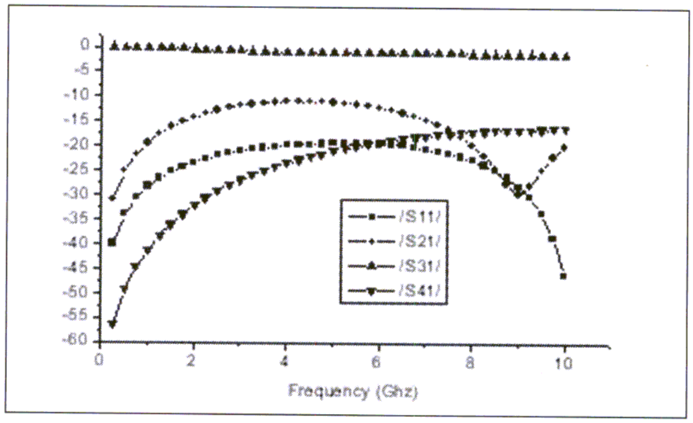


▲ **Figure 5. Structure of a four-port coupler.**



▲ **Figure 6. Scattering parameters of the coupler.**

$$[L]=\begin{bmatrix}390.1 & 138.6\\ 138.6 & 390.1\end{bmatrix}(nH/m)$$

$$[C]=\begin{bmatrix}170.1 & -40.03\\ -40.03 & 170.1\end{bmatrix}(pF/m)$$

$$[R]=\begin{bmatrix}23.13 & 0\\ 0 & 23.13\end{bmatrix}(\Omega/m)$$

$$[G]=\begin{bmatrix}585.9 & 0\\ 0 & 585.9\end{bmatrix}(\mu S/m)$$

Comparing the results, we find that our results correlate with those already published.

Figure 3 shows the dependence of the separation width on the even and odd mode characteristic impedances. Figure 4 illustrates the influence of the separation width on the coupling coefficient. It demonstrates an excellent agreement between our results and those obtained by the moment method.

Design of a directional coupler at 5 GHz

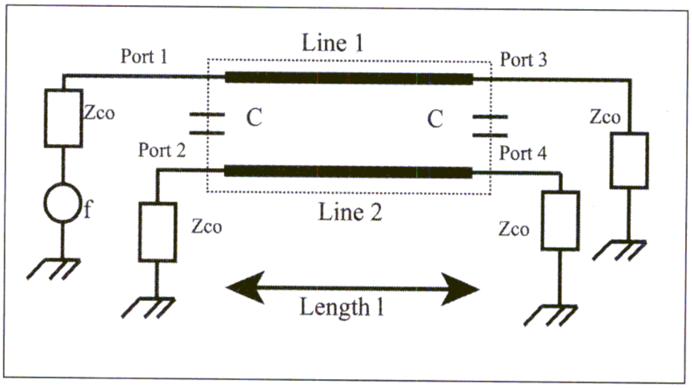
Figure 5 presents the structure of a four-port coupler. All the ports of the coupler are matched with Z_{co} = 50 Ω. For a length l = 6.4 mm, the resulting scattering parameters (with respect to 50 Ω) are plotted in Figure 6 in the band frequency (250 MHz, 10 GHz).

At 5 GHz, the coupling (–20 log IS12I) is 11.01 dB and the isolation (–20 log IS14I) is 20.35 dB.

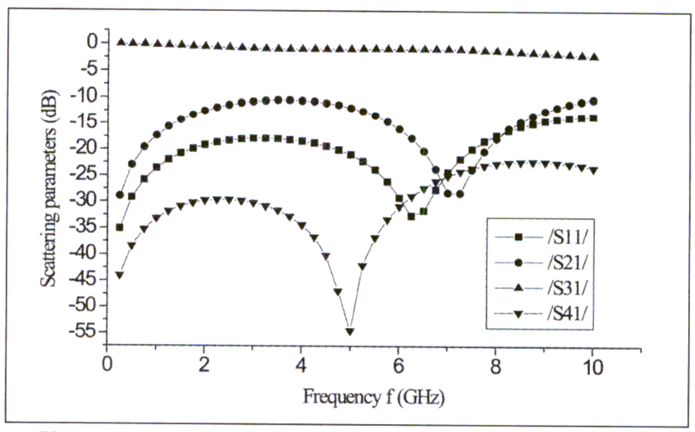
In order to modify the response of this coupler, we connect the two signal conductors at each end of the coupler with discrete capacitors (C = 0.1 pF), as shown in Figure 7. The response of this coupler is shown in Figure 8. Much better isolation is seen than in the previous case (Figure 6), where at the reference frequency 5 GHz the isolation is 54.58 dB.

Shielded and coupled microstrip lines

We analyzed the coupled line (Figure 9), composed of two inhomogeneous and shielded microstrips, with the same geometrical and electrical parameters as in the last case. The analysis of this structure with the finite element method gives the following results (the primary parameters values obtained are displayed below):



▲ **Figure 7. Coupler with compensating capacitances.**



▲ **Figure 8. Influence of the frequency on the response of the coupler.**

Z_{0e} = 62.8792 Ω
Z_{0o} = 34.8038 Ω
k = 0.287414

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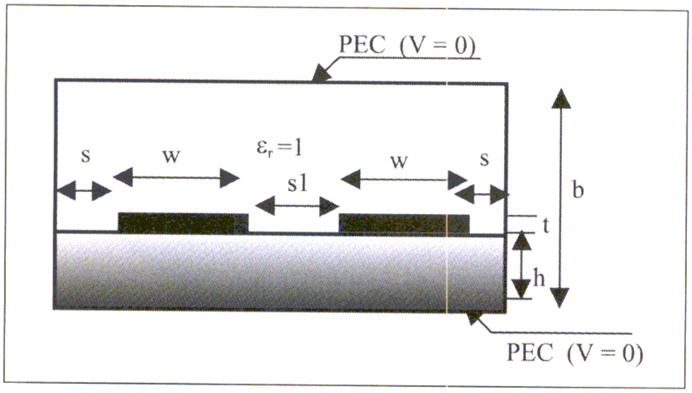
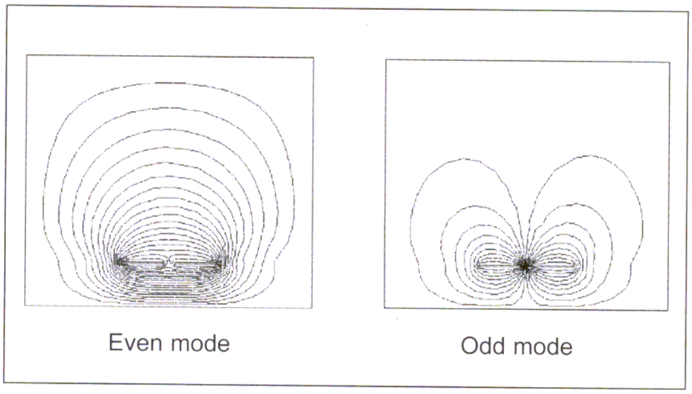
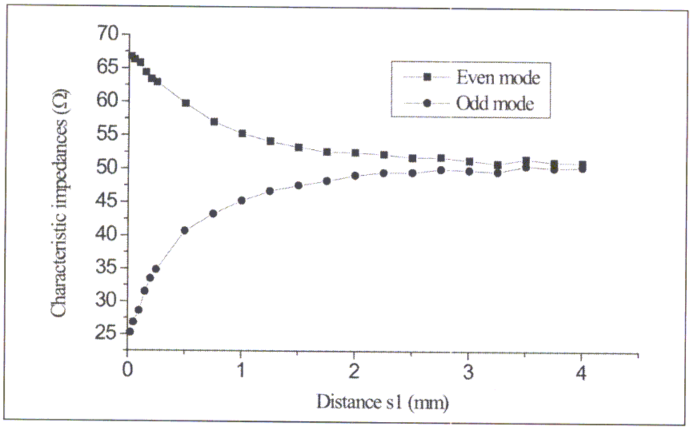


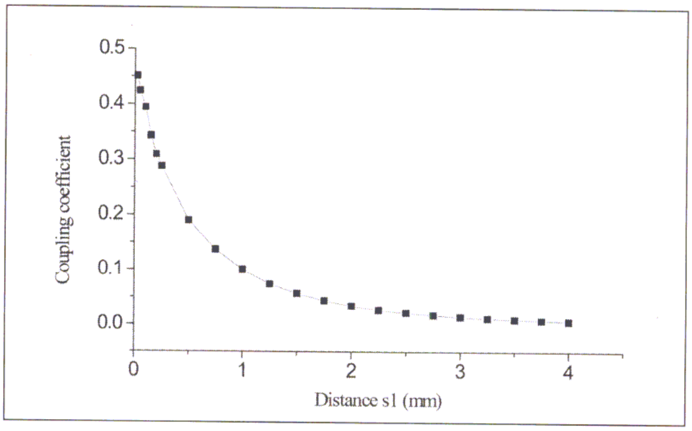
▲ **Figure 9. Cross section of the coupled line.**



▲ **Figure 10. Equipotential lines of the even and odd modes.**



▲ **Figure 11. Impact of the separation width on the even and odd mode characteristic impedances.**



▲ **Figure 12. Impact of the separation width on the coupling coefficient.**

$$[L]=\begin{bmatrix}397.016 & 114.108\\ 114.108 & 397.016\end{bmatrix}(nH/m)$$

$$[C]=\begin{bmatrix}205.518 & -45.8818\\ -45.8818 & 205.518\end{bmatrix}(pF/m)$$

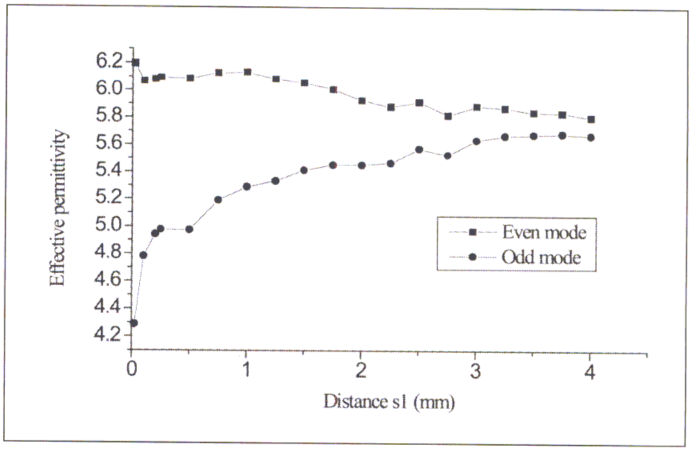
$$[R]=\begin{bmatrix}18.209 & 0\\ 0 & 18.209\end{bmatrix}(\Omega/m)$$

$$[G]=\begin{bmatrix}501.513 & 0\\ 0 & 501.513\end{bmatrix}(\mu S/m)$$

The dependence of the separation width on even and odd mode characteristic impedances is shown in Figure 11. Figure 12 illustrates the impact of the separation width on the coupling coefficient. The dependence of the separation width on the effective permittivity of the structure is shown in Figure 13.

Finally, we analyzed the same directional coupler at the reference frequency 5 GHz, this time using two shielded and inhomogeneous microstrip lines.

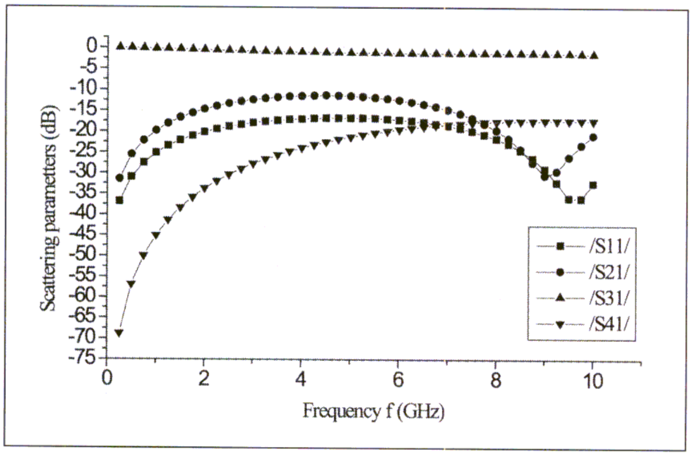
For a length l = 6.4 mm, the resulting scattering parameters (with respect to 50 Ω) are plotted in Figure 14 in the band frequency (250 MHz, 10 GHz).



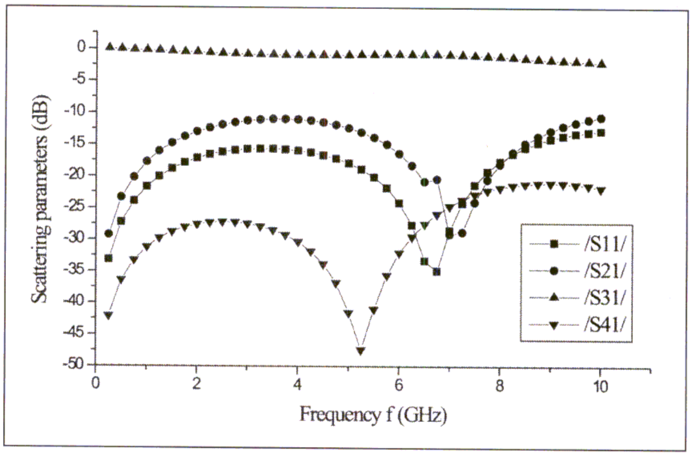
▲ **Figure 13. Influence of the separation width on the substrate permittivity.**

At 5 GHz, the coupling (–20 log IS12I) is 11.44 dB and the isolation (–20 log IS14I) is 21.31 dB.

Figure 15 shows the response of this coupler when a discrete capacitor (C = 0.1 pF) is connected at each end of the coupler. In this case the isolation improves to 41.57 dB.



▲ **Figure 14. Influence of the frequency on the response of the coupler.**



▲ **Figure 15. Influence of the frequency on the response of the coupler.**

Conclusion

This article discussed the design of a microwave coupler using lines with inhomogeneous microstrips. In order to achieve this design, it was necessary to determine the electromagnetic parameters of the system. In the band of frequency (250 MHz to 10 GHz), the problem is approximated by the resolution of Laplace's equation. Its solution was made by the finite element method, which allows the determination of the electromagnetic parameters (for example, coupling coefficient, characteristic impedances) of the structure composed of two inhomogeneous microstrip lines. The result of the CAO program shown here correlates to those that have previously been published.

The numerical model developed is valid for analyzing all configurations of multiline structures propagating the fundamental TEM or quasi-TEM modes.

All the curves presented in this paper, taking into account the influence of the electrical and geometrical parameters, prove the interest of the CAO program developed. The association of analytical functions to these curves remains possible. ■

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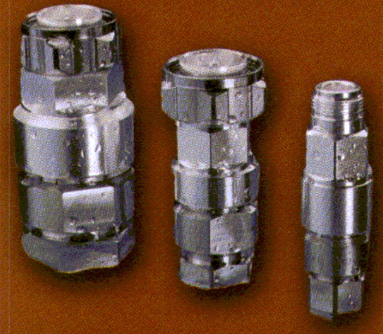
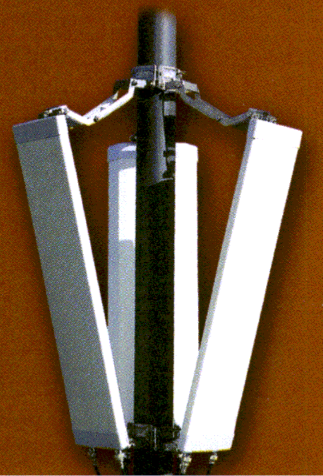




A Wideband Balun — How Does it Work?

Antennas often require a balanced feed system like the circuit described in this article

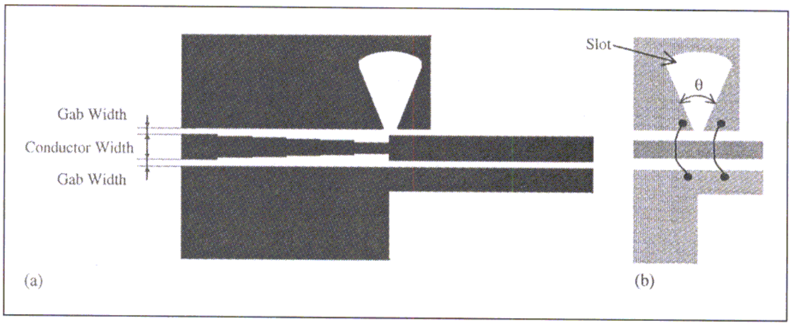
By J. Thaysen, K. B. Jakobsen and J. Appel-Hansen
Technical University of Denmark

Whenever a balanced antenna (i.e., dipole, loop or spiral) is used, the issue of how to feed the antenna becomes relevant. Because a balanced antenna requires a balanced feed, a balun is needed.

A balanced antenna fed by a two-wire transmission line is a balanced system with respect to the lines, provided that the two feed points on the balanced antenna have the same orientation and placement with respect to the lines. If the balanced (symmetrical) antenna is connected to a coaxial transmission line, the transition from the feed line to the balanced antenna is an unbalanced (asymmetrical) driven system. The balun is inserted between the feed line and the antenna in order to provide a transition between the coplanar waveguide (CPW) and the coplanar strip line (CPS). The balun produces a symmetrical radiation pattern however, this article does not deal with the radiation from a connected antenna.

In some applications, it is necessary to connect the feed terminals on the balanced antennas to an unbalanced coaxial cable that requires not only a balanced-to-unbalanced transformation circuit, but also an impedance match due to the different characteristic impedances of the antenna and the cable.

In the literature, different types of baluns are described [1, 2]. In this article, the focus is solely on a planar balun [3] because it has some very good qualities, such as a low insertion loss and wide bandwidth. The balun is planar, which makes it realizable using conventional photo techniques.



▲ **Figure 1. Balun for use with a spiral antenna.**

To aid the balun design and characterization, the electromagnetic simulation program IE3D Version 6.03, a method of moment computer program developed by Zeland Software [4], was used to predict the performance of the balun structure. The measured results of the constructed balun structure were then compared to the simulated results obtained from IE3D.

Baluns whose characteristics remain virtually unchanged over an exceptionally large bandwidth have a multitude of uses. For example, they can be used for ground penetrating radar where the use of a wideband antenna is necessary [5, 6].

The principle of the balun

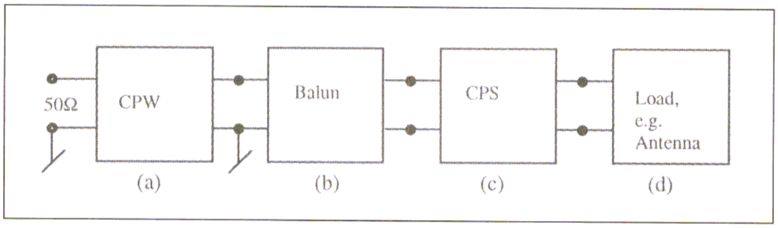
The wideband coplanar-waveguide to coplanar-stripline (CPW-to-CPS) balun was designed to transform the unbalanced CPW feed line to a balanced CPS feed line. The balun shown in Figure 1 is intended to be used with a spiral antenna. The CPW-to-CPS balun is a modified version of Li's [7], and a somewhat similar

balun configuration was successfully used in [8]. To characterize the balun, a model of the structure is shown in Figure 2. This model is used to explain the different parts of the balun.

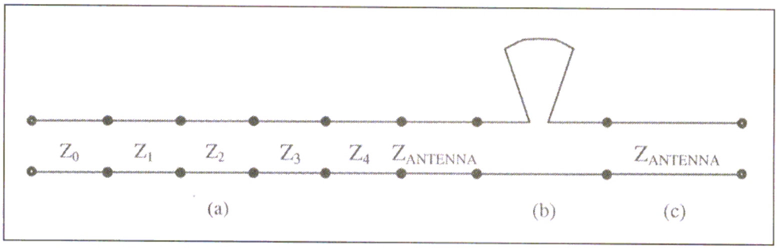
Coplanar waveguide (a) — Due to the difference in the impedance between the two-arm spiral antenna and the unbalanced coaxial cable feedline, it is necessary to provide an impedance match as well as balanced-to-unbalanced transformation.

The impedance match between the cable and the balun is obtained by using a Chebyshev multi-section impedance transformer circuit. Depending on the ratio between the two impedances, the number of sections can be numbered. In this case, a four section Chebyshev impedance transformer with a reflection coefficient of Γ_m = 0.05 was designed to transform the impedance from 50 ohms to 80 ohms.

Balun (b) — The actual transformation between the unbalanced and the balanced mode is achieved by the balun. The coplanar waveguide (a) and the coplanar strip line (c) are only needed to connect the sender/receiver and the antenna.



▲ **Figure 2. Balun consisting of a coplanar waveguide (a); balun (b); coplanar strip line (c); and the coplanar strip line connected to a balanced antenna (d).**



▲ **Figure 3. Transmission line model of the balun consisting of the coplanar step impedance waveguide (a), the balun (b), and the coplanar strip line.**

The balun is quite complex by itself. The basis of it is the wideband transition from the CPW-to-CPS, which is accomplished through a radial slot. This slot represents a very wideband open circuit, which forces the electrical field to be mainly between the two conductors of the CPS, as illustrated in Figure 8. The two bond wires near the discontinuity plane ensure that the potential on the two ground planes is equal, as shown in Figure 1(b).

Coplanar strip line (c) — The balanced output from the balun is connected to the antenna using the balanced coplanar strip line. The distance between the balun and the antenna determines the length of the strip line. Because a short distance is preferred, the balun is placed as close to the antenna as possible.

Transmission line model of the balun circuit — A transmission line model of the entire balun structure, including a four section Chebyshev section impedance transformer, is shown in Figure 3.

Balun configuration

Choosing a substrate is a bottleneck when designing the CPW and CPS, because both the CPW and the CPS have to be designed on the same substrate and have the same impedance, namely $Z_{ANTENNA}$. Using classic equations for CPW and CPS [2], it is found to be feasible to use a substrate material having a relative permittivity of 10.2 and a thickness of 0.785 mm. These parameters are chosen as a trade-off between the permittivity and the thickness, so that standard substrate parameters can be used.

The next step involves the design of the impedance transformer, which is accomplished using the classic equations for Chebyshev multi-section matching transformers [9] when deciding how many sections the transformer should have.

Preliminary simulations found that the length of the ground plane is proportional to the upper usable frequency of the balun. Also, an inverse proportionality was found between the length of the sections in the impedance transformer and the reflection coefficient (S_{11}). Choosing the number and the length of the sections is therefore a trade-off between a high bandwidth and a low reflection coefficient. To optimize the balun with respect to the bandwidth, it is necessary to reduce the overall length of the balun. By choosing four sections, each is minimized to 3 mm.

The design impedance is found by using the equations for the CPW and the CPS. Next, the physical dimensions are found and are adjusted with respect to production tolerances. The adjusted results are then used to calculate the impedance with the use of IE3D.

The width of the ground plane of the CPW is 15 mm, due to symmetry. It is, however, desirable to minimize the overall size of the structure, resulting in a cheaper and smaller solution. Thus, simulations are carried out in order to decrease the width of the CPW.

By changing the width of the CPW from 15 mm to 6.5 mm, the balun is optimized with respect to the bandwidth — that is, the upper frequency for a reflection coefficient higher than 10 dB is increased 20 percent. In [4], it was found that the optimal width for the ground plane of the CPW is about 2.5 times the distance between the two ground planes, which is in accordance with the results presented in Table 1.

Four-section Chebyshev step impedance transformer having an impedance ratio Z_0/Z_L = 80/50 = 1.6	CPW						CPS
	Z_0	Z_1	Z_2	Z_3	Z_4	$Z_{ANTENNA}$	$Z_{ANTENNA}$
Design impedance [Ω]	50	55.4	60.5	66.7	72.6	80	80
The physical parameters are adjusted with respect to production tolerance Strip width [mm] Gab width [mm] Length of the sections [mm]	1.7 0.45 3	1.5 0.55 3	1.3 0.65 3	1.1 0.75 3	0.9 0.85 3	0.7 0.95 30	1.8 0.4
Calculated impedance of the adjusted parameters using IE3D [Ω]	49.8	54.4	59.4	64.7	71.6	79	81

▲ **Table 1. Calculated results for the four sections Chebyshev step impedance transformer.**

Simulations with and without bond wires are performed. To ensure proper operation of the balun, it is necessary to state that the two bond wires are located near the discontinuity plane. The simulated as well as the prototyped bond wires have a diameter of 0.15 mm.

With the bond wires, the bandwidth for a reflection coefficient better than –10 dB of the balun is 0.1 to 3.45 GHz, whereas the bandwidth for the balun without any bond wires is as low as 0.1 to 0.5 GHz and from 1.95 to 2.4 GHz. From 0.5 to 1.95 GHz and 1.95 to 3.45 GHz, the reflection coefficient is simulated to be below –5 dB, as shown in Table 2. In conclusion, we find that the bond wires have to be used. This result has been verified through measurement.

Recently, a description of a balun was published based on the same principle, except that it did not benefit from a radial slot [10]. That balun is rather narrow-banded compared with the balun discussed in this article.

The advantages of a radial slot over a rectangular slot are smaller resonance length and wider bandwidth [11]. Based on experimental investigation, it is found that the optimized angle, θ, of the radial slot is 45 degrees [5]. In this article, the simulated dimensions of the radial slot are a radius of 6 mm and an angle of 45 degrees [12].

Kolsrud, et al. [13] argue that a balun with a circular slot provides a larger bandwidth compared to a balun with a radial slot. Two back-to-back balun structures (one having radial slots, the other circular slots) were simulated and prototyped in order to find the most broadband configuration. The simulated as well as the measured results for the back-to-back coupled balun are shown in Table 2.

The simulated upper frequency for a reflection coefficient higher than –10 dB is 3.6 and 3.45 GHz for the balun structure with circular and radial slots, respectively. The measured upper frequency for a reflection coefficient higher than –10 dB is 3.3 and 3.4 GHz for the balun structure with circular and radial slots, respectively. The balun with the radial slot yields the best solution because of a somewhat lower insertion loss and a better reflection coefficient though the band of operation frequencies.

The simulations carried out are not exhaustive and it is possible that different results can be obtained by changing some parameters.

Numerical and experimental results

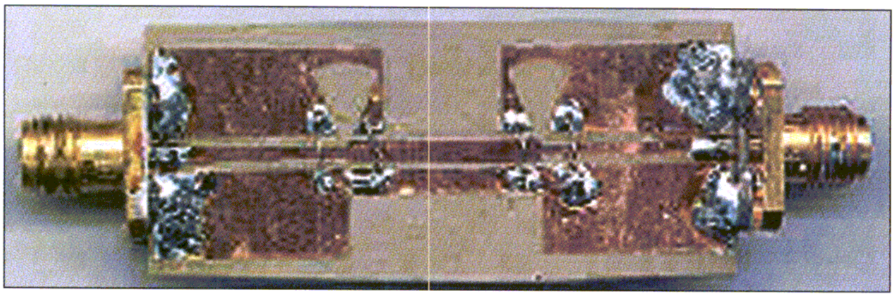
To analyze the balun, IE3D was used to simulate the performance of the printed balun in terms of the *s*-parameters, as well as the current distribution on the metallic surface of the balun structure. Twenty cells per wavelength and edge cells are used to model the balun structure [4].

The balun structure shown in Figure 4 was fabricated on a Rogers RT/Duroid® substrate with a thickness of 0.785 mm and a relative dielectric constant ε_r of 10.2.

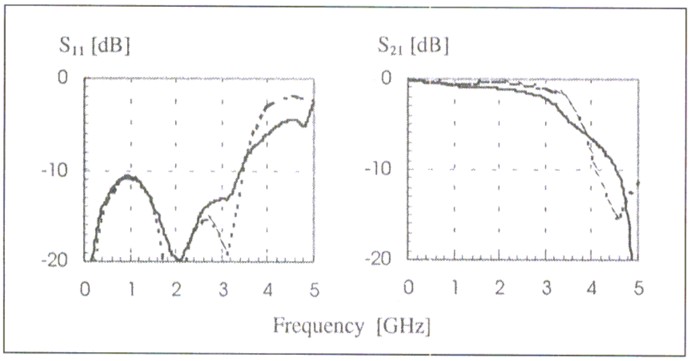
Two back-to-back CPW-to-CPS transitions were simulated and optimized using IE3D. The baluns were connected so that the unbalanced port of each balun was accessible externally. The SMA connectors and the wire bonding were soldered directly on the substrate. The

	Frequency range with a return loss better than 10 dB			
	Bandwidth for S_{11}<–10 dB [GHz]		Insertion loss for S_{11}<–10 dB [dB]	
	Simulated	Measured	Simulated	Measured
Without bondwires, with radial slot	0.1–0.9	300 kHz–0.75	< 1	< 1
With bondwires, without slot	0.1–2.5			
With bondwires, with circular slot	0.1–3.6	300 kHz–3.3	< 1.3	< 2.5
With bondwires, with radial slot	0.1–3.45	300 kHz–3.4	< 1	< 2

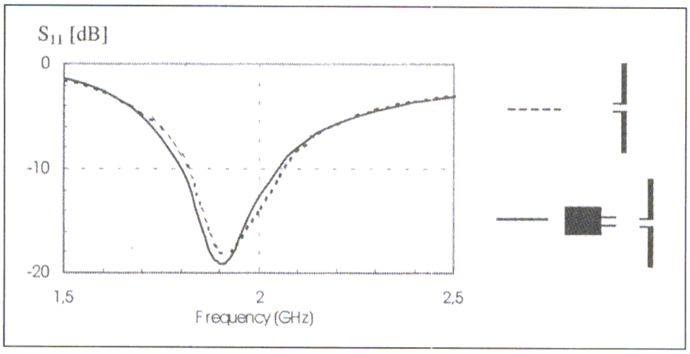
▲ **Table 2. Measured and simulated results for the back-to-back coupled balun.**



▲ **Figure 4. Back-to-back coupled balun placed on RT/Duroid substrate.**



▲ **Figure 5. Measured (solid) and simulated (dashed) *S*-parameters for the two baluns mounted back-to-back.**



▲ **Figure 6. Reflection coefficient for the ideal feed dipole antenna (dotted) and for the dipole feed via the simulated balun (solid).**

size of the prototype is 16 × 46 mm. The structure was fabricated to verify the bandwidth and insertion loss. In terms of scattering parameters, the numerical results can be compared with the experimental results in Figure 5. The result for S_{11} seems to be the most accurate; however, the agreement is good for frequencies up to 2.5 GHz. The insertion loss (i.e., the S_{21}), the numerical and the experimental result similarities are obtained. The simulated as well as the measured results show an upper frequency limit where the reflection coefficient is 10 dB at 3.4 GHz. Good agreement is obtained between the simulated and measured balun for frequencies up to 4.5 GHz.

The structure was tested on an HP 8720D and an HP 8752A network analyzer to determine the reflection coefficient and the insertion loss of the balun. The two back-to-back CPW-to-CPS transitions provide a measured insertion loss of less than 2 dB from 300 kHz to 3.4 GHz with a reflection coefficient higher than –10 dB. Although the balun characteristics are measured starting from 300 kHz, the lower limit of the frequency bandwidth is nearly DC.

Test of the balun

Additional simulations of the discussed balun were carried out to verify the validity of the balun and to test the validity of the simulation.

The simulations and the measurements on the constructed prototypes verify the principle for the balun. In order to verify the balun it is necessary to investigate a single balun.

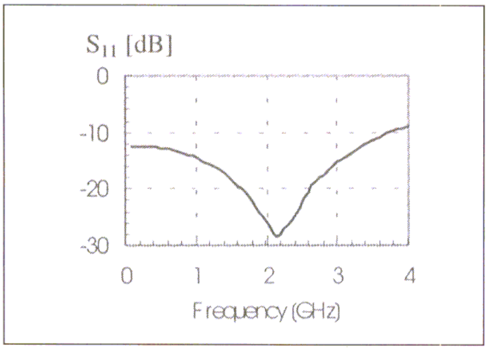
In IE3D, it is possible to merge two individually simulated structures (e.g., a balun concatenated to a dipole antenna). This setup can be made by simulating the balun, then simulating the dipole, and finally merging the two resulting files to end up with the result for the concatenated circuit [4] — for example, the scattering parameters of the dipole feed by the balun.

An ordinary planar half wavelength dipole with a length of 50 mm placed on the same substrate was used for the balun, and is expected to have constant input impedance around the resonant frequency. The planar dipole is a small and simple structure, causing a simulation time of only a few seconds. The simulated reflection coefficient for the planar dipole is shown in Figure 6.

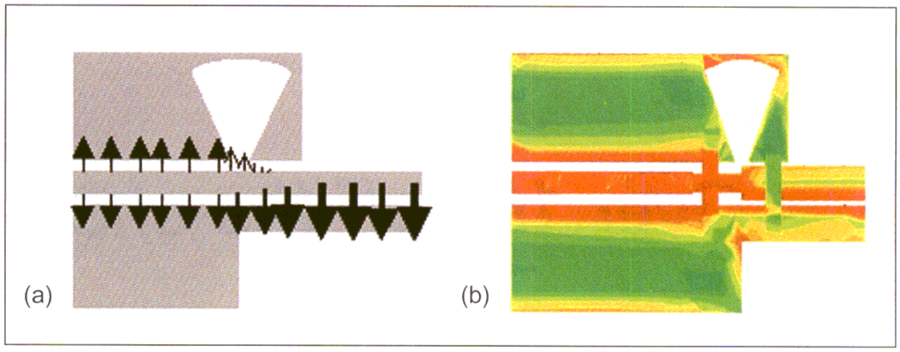
In Figure 6, the comparison of the simulated reflection coefficient for the ideal feed dipole antenna and the simulated reflection coefficient for the dipole feed via the simulated balun shows that there are virtually no differences. In both cases, the resonant frequency is 1.9 GHz and the bandwidth is 250 MHz, for a reflection coefficient higher than –10 dB. The result indicates that the balun actually works as a balun and that IE3D can be used to concatenate two individually simulated structures. It should be noted that the free space resonant frequency for the dipole is 3 GHz, but due to the dielectric loading (substrate), the resonant frequency is lowered to 1.9 GHz only.

The balun is intended to be used as the feed network for a frequency independent antenna (e.g., a spiral antenna). In order to verify how frequency independent the balanced port on the balun are with respect to the input impedance, simulations on the balun using IE3D are performed by terminating the CPS with an ideal 80-ohm resistor. The result is shown in Figure 7.

In the frequency range from 0.1 to 3.85 GHz, the sim-



▲ **Figure 7. Simulated reflection coefficient s_{11}-parameter for the balun where the CPS is terminated by an ideal 80-ohm resistor.**



▲ **Figure 8. Electrical field pattern for the CPW-to-CPS balun (a). The average current density is shown in (b) where the density at each location is illustrated by IE3D where red indicates the highest and green the lowest current density.**

ulated reflection coefficient is better than –10 dB. For the balun loaded with an ideal 80 ohm resistor, the simulated bandwidth within which the reflection coefficient is better than –10 dB is higher than the bandwidth obtained for the back-to-back balun, because of the interaction between the two baluns. The actual balun is used with a spiral antenna, where the measured upper frequency limit is 3.8 GHz for a reflection coefficient better than 10 dB [3].

Simulations are carried out in order to investigate the physical operation of the balun. A sketch of the electrical field pattern is shown in Figure 8. The sketch shows how the slot forces the field to be transformed from a symmetrical pattern between the center conductor and the two ground planes of the CPW to be mainly between the two conductors of the CPS. Furthermore, the average current density, calculated using IE3D, is shown in Figure 8. The figure shows that the highest current densities for the CPW are in the vicinity of the center conductor of the CPW and at the edges on the two ground planes of the CPW. On the CPS, the highest current density is symmetrical in the vicinity of the two conductors' edges.

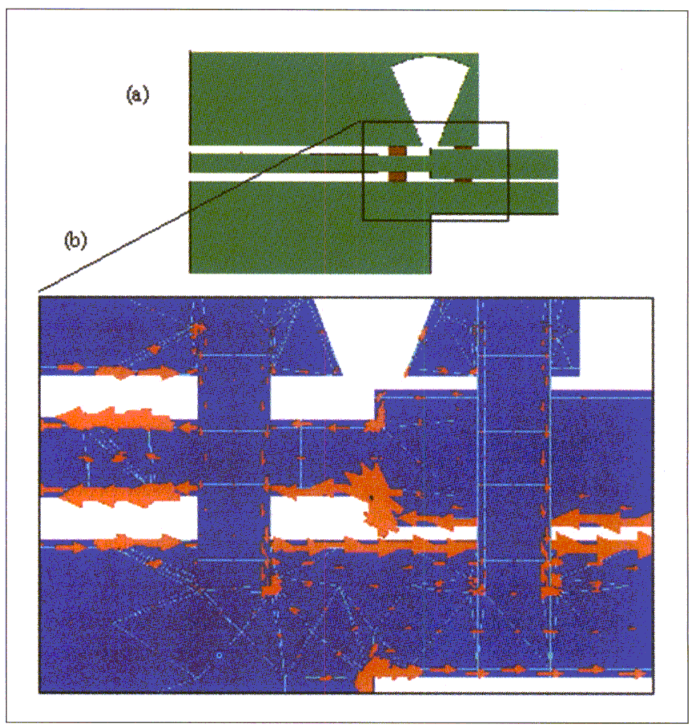
The arrows shown in Figure 8 indicate the direction and the density of the current at a specific location. The current is flowing from the CPW to the CPS through the ground plane on the CPW to one of the CPS conductors' edges. The current is flowing in the opposite directing (i.e., from the opposite conductor on the CPS via the center conductor of the CPW).

The grid shown in Figure 9 is the result of the automatic grid in IE3D. Near the edges, the modeled structure consists of edge cells which leads to a higher accuracy of the simulated results, compared with the measurements performed on the constructed prototype [4]. Also, this feature yields a 3 to 5 times longer simulating time, which means the user can use it only where high precision of the simulated results is required. For further discussion, see Zeland Software's IE3D [4].

Conclusions

The use of the electromagnetic simulation program IE3D to verify the theoretically expected performance with a measured prototype has been demonstrated. The scattering parameters can be predicted with good accuracy using the simulator. Howevcr, due to limitations such as infinite dielectric layer and non-homogeneous metal thickness due to the solder and connectors, everything that is measurable cannot easily be simulated. Generally, the simulations offer a fast and reasonably accurate way of investigating the balun structure.

A wideband coplanar waveguide to coplanar strip transition, which covers a frequency range from practi-



▲ **Figure 9. The CPW-to-CPS balun (a). A Close-up picture of the current density where the arrows show the direction and the density of the current at a specific location (b).**

cally DC to 3.85 GHz with a reflection coefficient better than –10 dB, was presented. Good agreements were obtained between the numerical results and the measured results in the frequency range from 300 kHz to 5 GHz using an HP 8753 and an HP 8720 network analyzer. IE3D simulations have been used to verify that the balun structure operates as a balun, with the balun structure showing that the input impedance is essentially constant over a bandwidth from 0.1 GHz to 3.85 GHz.

By using a Chebyshev impedance transformer, the balun design allows impedance transformation in addition to providing the necessary transition from the unbalanced feed line (e.g., coaxial cable, microstrip and coplanar waveguide) to the balanced feed line (e.g., coplanar strip line).

The balun described here, which is as small as 16 × 46 × 0.8 mm, is designed and prototyped for the RT/Duroid 6010 substrate having a relative dielectric constant ε_r of 10.2. Also, the proposed design is uniplanar and thus feasible to realize using ordinary photographic fabrication techniques without the need of additional components, such as a wire wound transformer. However, this article has shown that wire bonding near the discontinuity plane improves the performance of the balun.

The balun has several advantages, including very wide bandwidth, relatively small size and low profile. The balun could easily be integrated as part of a microstrip circuit, which furthermore yields a low-cost solution. Thus, the balun should find many applications in broadband systems. ■

Acknowledgements

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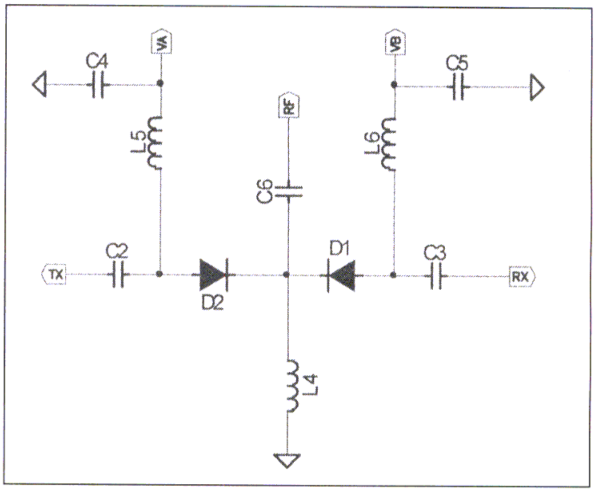
Circle 79

A T/R Diversity RF Switch Design Using PIN Diodes

This design combines low-cost implementation with good RF performance

By Louis Fan Fei
Lucent Technologies

TDMA is a very popular standard for today's wireless communication equipment. It is widely adopted in TDMA based cellular phones, GSM handsets and wireless LANs. One of the most important building blocks for TDMA is a high performance RF switch. The RF switch's main function is to switch between the transmitter (Tx) and the receiver (Rx). Typical design requires low insertion loss (IL), low intermodulation distortion (IMD), high isolation between Tx and Rx, fast switching time, and low current consumption. In mobile phone and outdoor wireless LAN applications, the environment also presents the fading problem to the system designer. One low-cost and effective way to combat fading is diversity. This article discusses a T/R switch with inherent space diversity at 2.45 GHz.



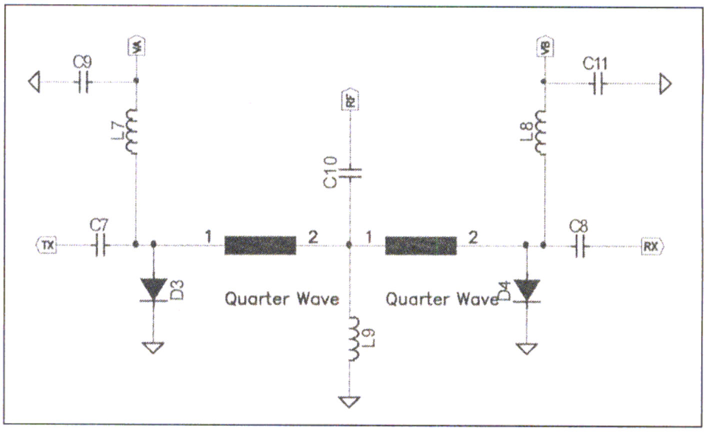
▲ **Figure 1. A typical SPDT switch topology with series diodes.**

PIN diode switch design

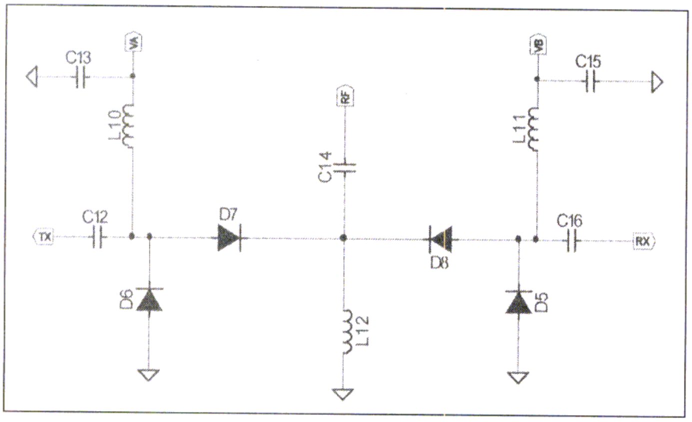
An RF switch can be implemented with either a MESFET or PIN diode. The MESFET-based RF switch is good for MMIC design, while PIN diodes are good for low-cost discrete design. This paper focuses on the design using a PIN diode. Typical single pole double throw (SPDT) topologies are shown in Figures 1, 2 and 3.

In Figure 1, a pair of series PIN diodes is used. The incoming RF signal can be switched to either Tx or Rx by turning on one of the PIN diodes while the other is off. For example, if V_a is a 5 V and V_b is –5 V, then D2 is on. The RF port is connected to Tx. If V_a and V_b are reversed, then D1 is on. The RF port is connected to Rx.

In Figure 2, a pair of shunt-configured PIN diodes is used. The shunt-configured PIN diodes are combined with quarter-wave transmission



▲ **Figure 2. Switch with λ/4 wave sections to permit the use of shunt diodes.**



▲ **Figure 3. Series-shunt switch with high isolation.**

line. The incoming RF signal is switched to either Tx or Rx, depending on which diode is on. If D4 is on and D3 is off, the RF port sees open circuits on the side of D4, so the RF signal is connected to Tx. If the bias voltage is reversed, the RF signal is connected to Rx.

Figure 3 features a design using a combination of series and shunt-configured PIN diodes. This design will have a very high isolation between Tx and Rx. When D7 and D5 are on, the RF signal is connected to Tx. When D6 and D8 are on, the RF signal is connected to Rx.

All the above topologies are widely used in RF and microwave design, and all give good performance. All are also are symmetric. They are good for a general purpose application, but that could also be to their disadvantage.

First, a pair of digital control signals is needed. Typically, a cascade logic inverter is needed to control the general purpose SPDT switch. Second, bias current is always needed for either the Tx or Rx mode, even when the system is in stand-by or sleep mode. No current consumption is desired when the system is in stand-by mode. Finally, the general purpose SPDT switch is symmetric. The Rx and Tx can be reversed without seeing any performance difference, so it does not have any diversity capability.

An even better RF switch topology is a T/R switch. It offers performance similar to the above general purpose SPDT RF switch, but without the disadvantages associated with the SPDT switch.

T/R switch design

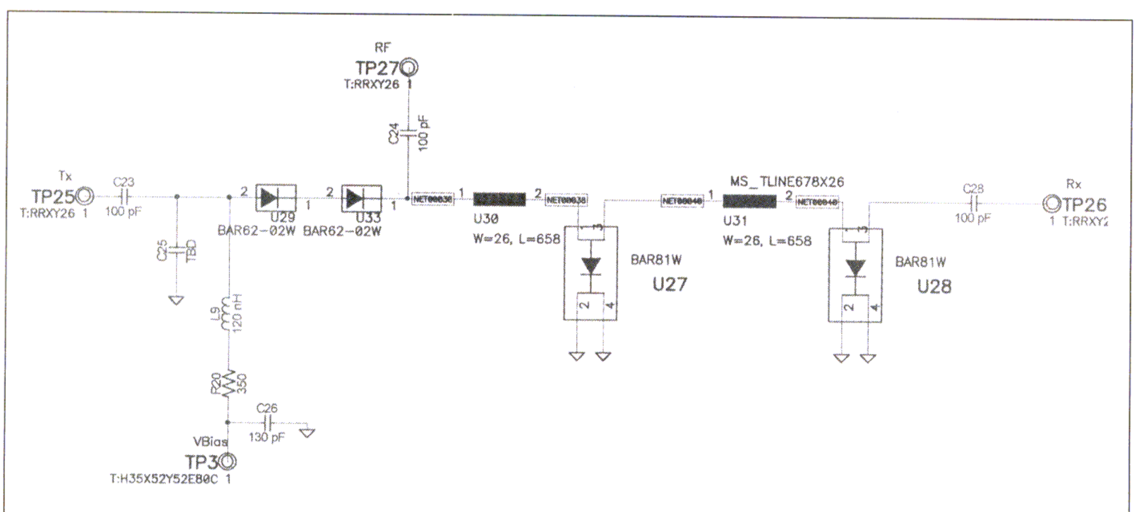
The schematic and layout of a T/R switch design at 2.45 GHz are shown in Figures 4 and 5. The basic circuit consists of a pair of series PIN diodes on one side and a pair of shunt PIN diodes combined on the other side. The extra diode is added on each side to provide more isolation and diversity.

The left half of the T/R switch is similar to half of the circuit in Figure 1, while the right half of the T/R switch is similar to half of the circuit in Figure 2. The Tx port is placed on the left and the Rx port is placed on the right. Notice that only one control signal is needed for this design. When V_{bias} goes to 5 V, all four diodes will be on at the same time. The two series diodes provide the low insertion loss path from the Tx to RF port. The two shunt diodes are placed after the quarter wave TL.

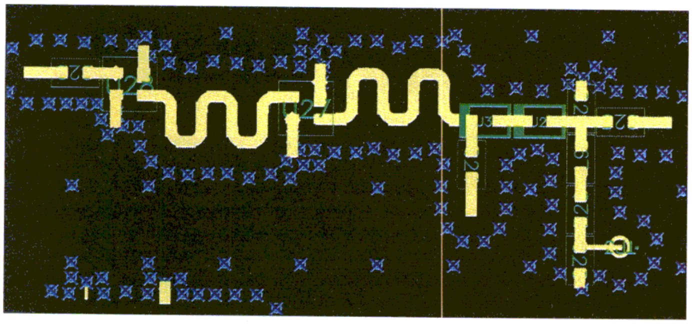
When the two shunt diodes are on, it presents a very high impedance state from either RF or Tx to Rx. Thus, all the RF signals coming in from Tx will arrive to the RF port with minimum leakage to Rx. During the Rx mode, the control voltage is 0 V. All four diodes are in off states. The RF comes into the RF port, and sees a transmission line (TL) one half a wavelength long on the Rx side. On the Tx side, the RF signal will see an open circuit since the two series PIN diodes are in the off state. Thus, the majority of the RF signal arrives at the Rx port with minimum leakage to Tx.

The above description gives a brief introduction to how a T/R switch works. The careful reader might observe that no bias current is needed during the Rx mode. The Rx mode is typically stand-by mode for a receiver. Minimum current consumption will conserve the battery power and increase the stand-by time.

The basic T/R switch circuit shown in Figure 4 was built and tested. The major design parameter was tested, and the result is shown in

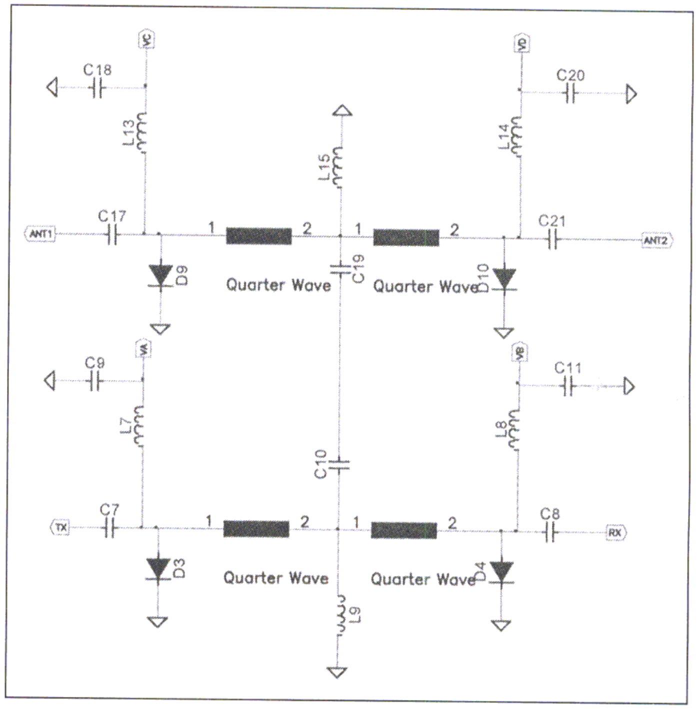


▲ **Figure 4. Schematic of a 2.45 GHz T/R switch with series diodes on the Tx port and shunt diodes on the Rx port.**



▲ **Figure 5. Circuit board layout for the switch circuit shown in Figure 4.**

Figure 8. The IL from Tx to the RF port is –1.2 dB, with about 0.3 dB cable and connector IL loss. The IL from the RF port to Rx is 1.3 dB. Again, the IL includes cable and connector loss. The isolation from Tx to Rx is 44 dB during the Tx mode. During the Rx mode, the isolation is 24 dB from Rx to Tx, but isolation is not critical in the Rx mode because the Rx signal typically is very small, and the Tx chain does not have any significant reverse gain. The return loss (RL) is better than –14 dB under all conditions.

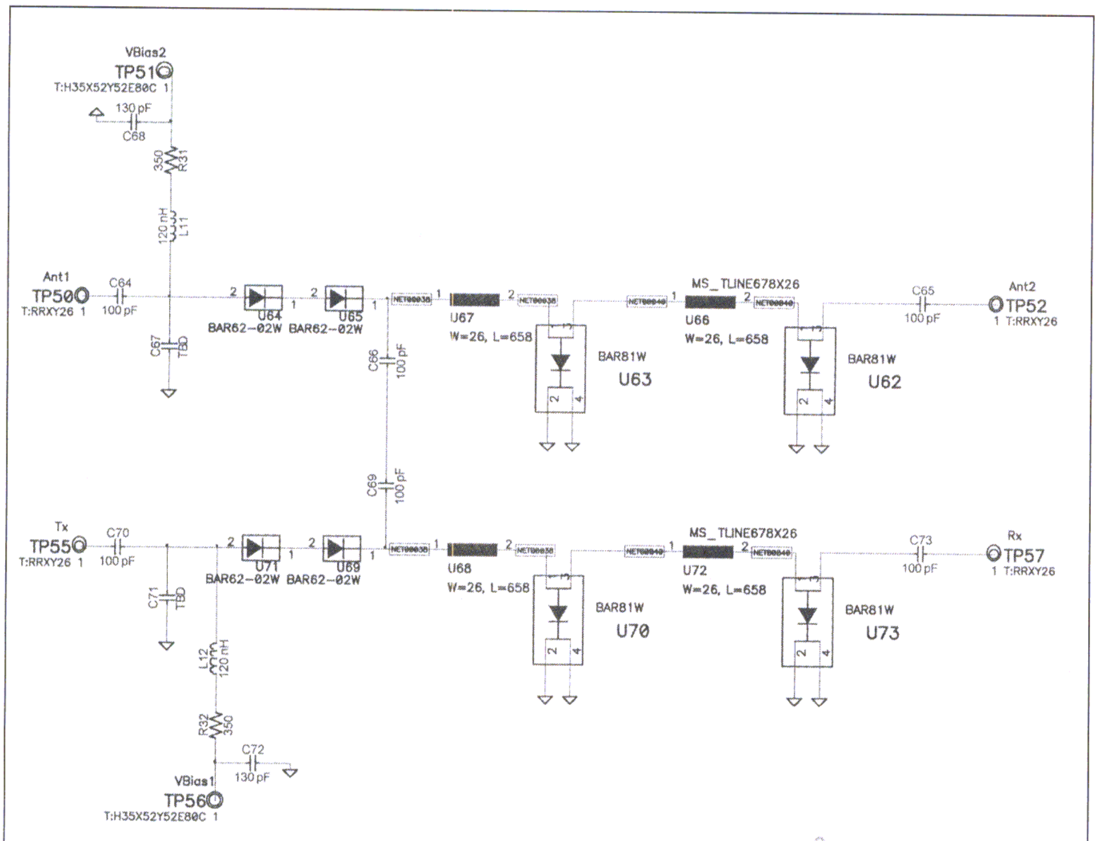


▲ **Figure 6. Two shunt switches provide both T/R and diversity antenna switching.**

Diversity

As mentioned before, diversity is a low-cost and effective way to combat fading. In mobile and outdoor urban environments, multiple paths are available for the signal to reach its destination. The signal might arrive in-phase or out-of-phase. It is typically modeled as Rayleigh fading. In the case that the signal arrives out-of-phase, the RF signal will experience high signal loss. The purpose of the diversity is to provide several different paths to receive the signal. The strongest signal will be picked as an incoming RF signal. Commonly used diversity schemes are space, angle, polarization, field, frequency and time. In this design, space diversity is used because of the simplicity. The two antennas are spaced at a minimum of a half wavelength apart, which will work well for space diversity.

The T/R switch cir-



▲ **Figure 7 . Circuit diagram of a diversity switch using two shunt switch sections.**

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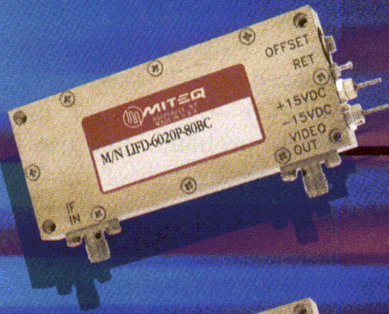

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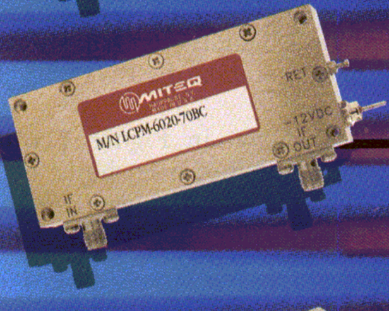
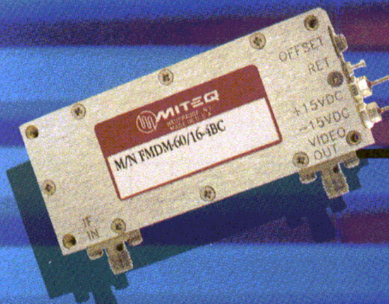
LIFD-3010P-80BC	30	-80 to 0	±0.5	100	25
LIFD-6020P-80BC	60	-80 to 0	±0.5	50	25
LIFD-7030P-80BC	70	-80 to 0	±0.5	30	25
LIFD-16040-80BC	160	-80 to 0	±1.0	30	25
LIFD-300100-70BC	300	-70 to 0	±1.0	20	15



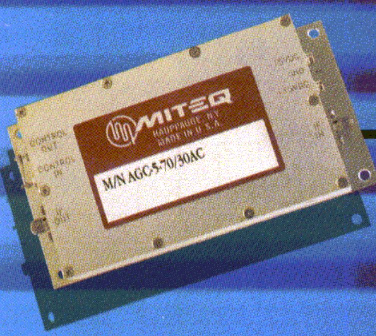
CONSTANT PHASE LIMITING AMPLIFIERS

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°



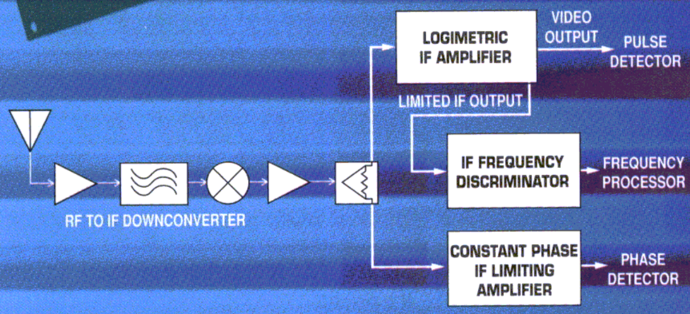
FREQUENCY DISCRIMINATORS

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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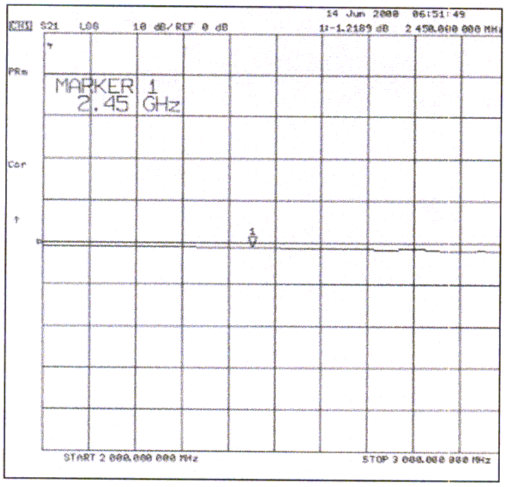
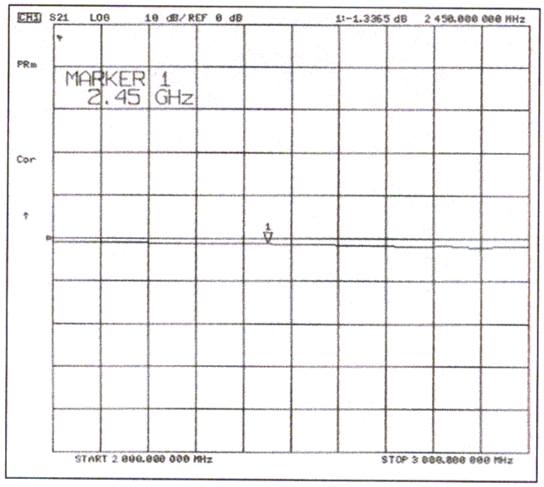


▲ **Figure 8. Tx mode IL including cable loss.**



▲ **Figure 9. Rx mode IL including cable loss.**

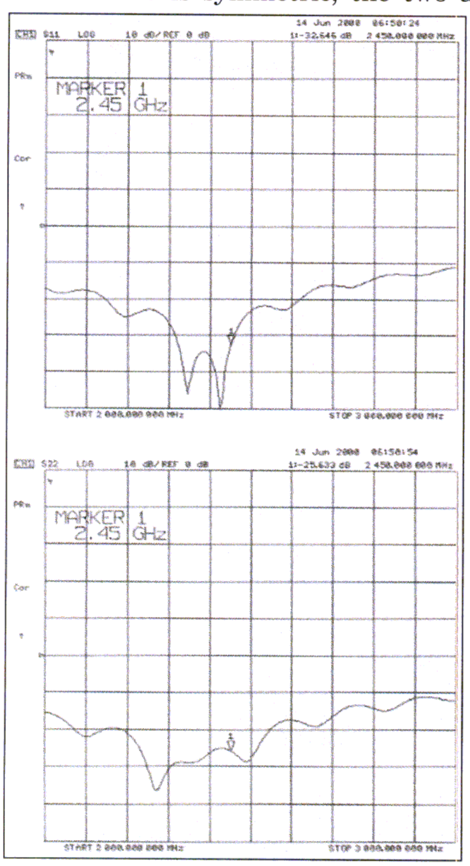
cuit presented in Figure 4 is a three-port building block containing Rx, Tx and the RF port. The standard SPDT has three ports with one antenna port. To accommodate two antenna ports, two basic RF switches are connected at the antenna port, as shown in Figure 6.

The circuit shown in Figure 6 uses two shunts configured as a SPDT switch. All the other SPDT switches can be used to build the diversity switch. Because the basic SPDT is symmetric, the two antenna paths are exactly the same.

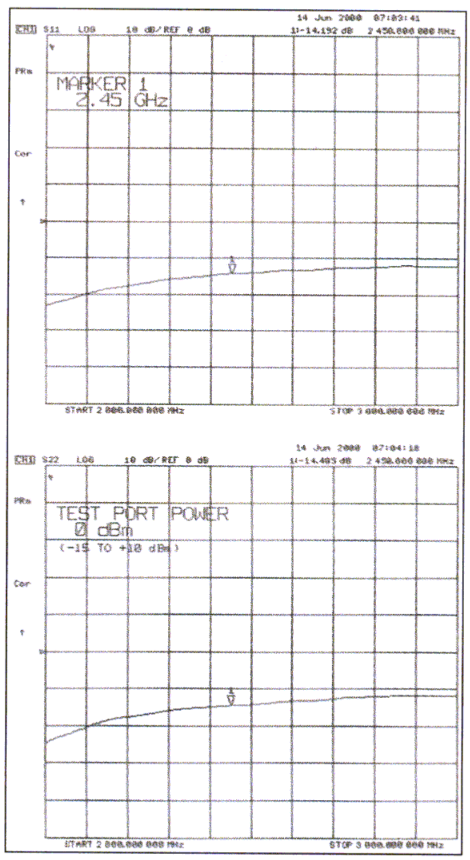
To achieve space diversity, a minimum of a half wavelength TL has to be added intentionally in one of the antenna paths. The half wavelength TL will require more space and increase IL.

A diversity switch based on a T/R switch (see Figure 4) does not have that problem. Because in the Rx branch of the T/R switch, there are already two quarter wavelength TLines, there is already about a half-wave phase difference between Rx and Tx. No additional half wavelength TL is needed as in the standard SPDT based diversity switch.

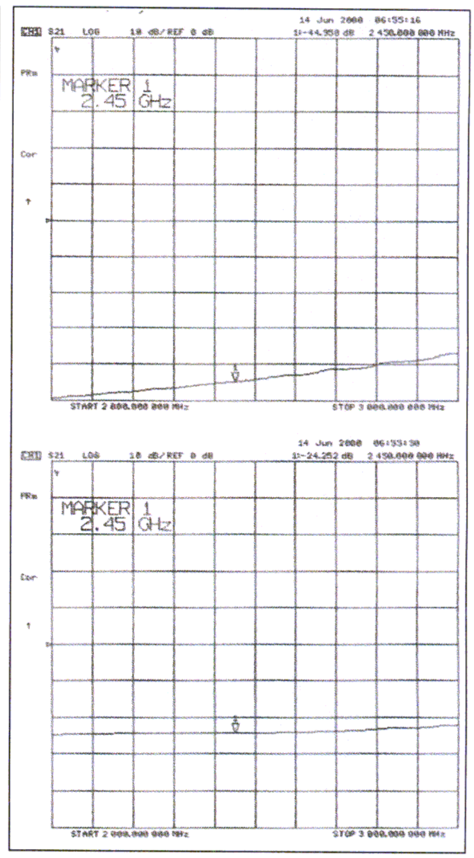
The T/R diversity switch shown in Figure 7 will require another board spin. It has not been built, but the performance of the T/R diversity switch can be easily deduced from the single stage T/R switch. The IL will be about 2 dB in both the Tx and Rx modes, while isolation between Tx and Rx remains the same at 44 dB.



▲ **Figure 10. Tx mode matching performance.**



▲ **Figure 11. Rx mode matching performance.**



▲ **Figure 12. Isolation between Tx and Rx ports.**

Conclusion

A low cost diversity RF switch at 2.45 GHz has been presented. It features low insertion loss, low bias current, high isolation and good matching, as shown in Figures 8 through 12. Compared to the standard SPDT switch, it only needs two control signal lines and no bias current consumption during Rx mode. The low-cost RF switch has built-in diversity. ■

Acknowledgments

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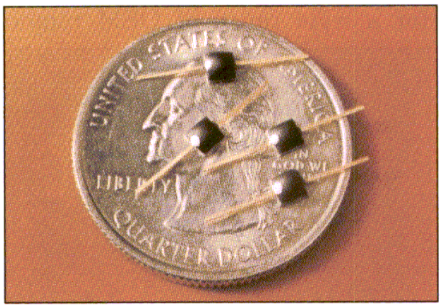

megaohm. Additionally, it handles an input range of 0 to 20 dBm over a temperature range of –55 to +125°C.

Signal Technology Corporation
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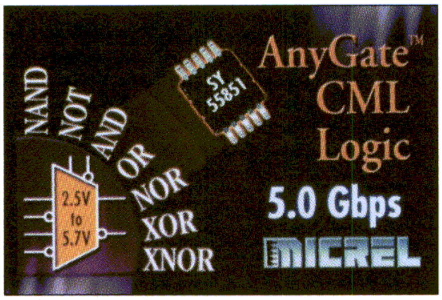
2.5 GHz CML IC for telecom applications

Micrel introduces the SY55851 AnyGate CML IC for designers of ultra-high speed systems. The SY55851 operates from a wide voltage range of 2.3 volts to 5.7 volts and has outputs that are guaran-



teed to toggle at least 2.5 GHz. The part uses CML logic levels for all I/Os and is compatible with legacy PECL and LVDS interfaces.

Micrel, Inc.
Circle #165

High-frequency, low-power prescaler

SiGe Microsystems introduces the D602, a high-frequency, low-power prescaler capable of performing a divide-by-2 function for low phase noise frequency synthesizers up to 6 GHz. The D602 operates at 16 mA from a 3-volt power supply and extends the battery life of portable products through its power-down mode. Typical input sensitivity is –27 dBm, which can help to reduce design cost and complexity.

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180 MHz amplifiers offer fast settling time

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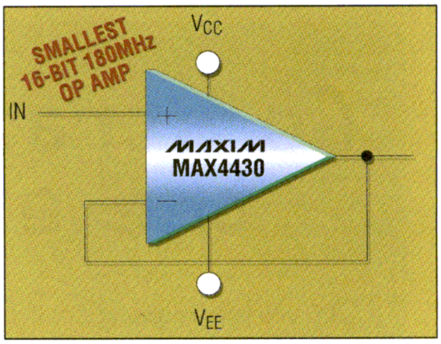

.0015 percent settling time and 110 dB minimum open-loop gain make these amplifiers ideal for buffering high-speed, high-performance ADCs for precision instrumentation and industrial applications. The devices have a –3 dB bandwidth of 45 MHz, and a slew rate of 145 V/µs. Prices start at $1.75 each in quantities of 100,000, or more.

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Amplifier targets fiber optic systems

Anadigics introduces the ATA7600D1, an InGaP HBT transimpedance amplifier (TIA) for OC-

192 fiber optic systems. Operating at +5 VDC, this amplifier provides low power consumption, low group delay, high gain and low noise performance to enable high-speed, broadband fiber optic systems. This product is ideal for 10 Gb/s and DWDM applications, dissipating a maximum of 500 mW and offering the additional advantage of good gain flatness, –20 to +20 ps group delay, and low noise. The ATA7600D1 is available for $240 in quantities of 1,000 units.
Anadigics
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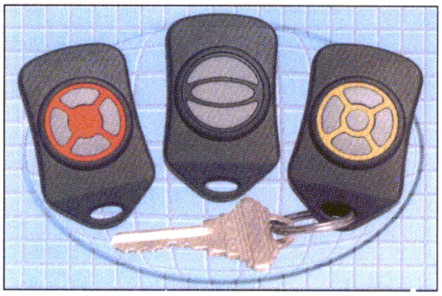
WIRELESS SYSTEMS

OEM transmitter modules

RF Monolithics announces a new series of OEM transmitter modules for use in wireless data communications. The DR4000 series is designed for short range wireless control and telemetry applications. These modules are less than three-quarters of an inch square and can be operated from small 3-volt type batteries. They include provisions for on-off keyed and amplitude-shift keyed modulation.
RF Monolithics, Inc.
Circle #169

Keyfob-style transmitters

Linx Technologies has released ultra-compact keyfobs, well-suited for a wide range of OEM remote



control and command applications. The fobs generate a secure encoded transmission that can be received and decoded using a variety of receiver products. They operate from a standard 3-volt lithium cell, allowing for several years of use under normal conditions. The fob is priced at $9.25 in volume product quantities.
Linx Technologies
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Low-profile antenna

Xertex Technologies offers the omni-directional Sphere Diversity 1900, an antenna that provides improved signal reception in areas where scattering and reflection can cause multipath problems. It provides 3 dBi gain and a maximum VSWR of 2.0 across the PCS band.
Xertex Technologies, Inc.
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Products

SIGNAL PROCESSING

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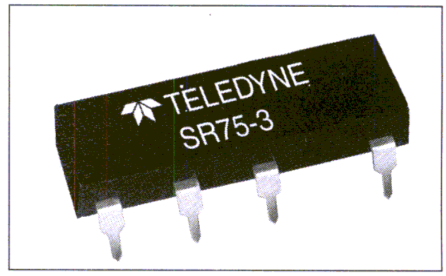
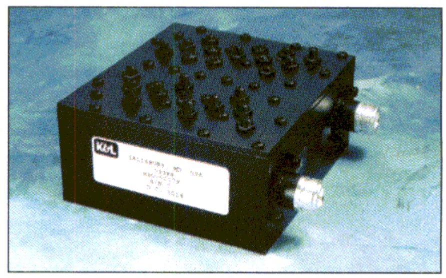
Teledyne Relays' offers the SR75-3 solid state relay for military and aerospace applications. The relay provides short circuit protection, optical isolation, low off-state leakage, high noise immunity and high dielectric strength. The SR75 is priced at $29 each in quantities of 500 or more.



Teledyne Relays
Circle #172

Japanese band CDMA duplexer

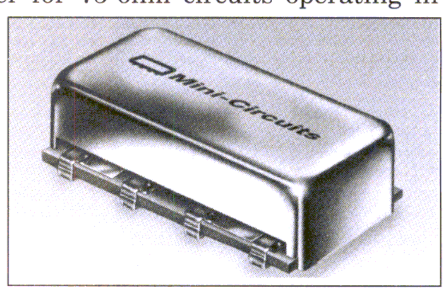
K&L Microwave has released a high-performance Japanese band CDMA duplexer offering low insertion loss and excellent selectivity in a small size. General purpose handset or equipment test systems will benefit from the use of this new duplexer. It is configured with a transmit passband of 832 to 870 MHz and a receive passband of 887 to 925 MHz and features 1.5 dB maximum insertion loss and 15 dB minimum return loss within each passband.



K&L Microwave
Circle #173

4-way splitter/combiners

Mini-Circuits has released a 4-way, 0 degree power splitter or combiner for 75-ohm circuits operating in the 5 to 750 MHz band. With solder plated J leads for solderability and strain relief, this miniature surface mount unit exhibits 35 dB isolation, matching with a typical VSWR of 1.20:1 in/1.15:1 out, and .3 dB maximum amplitude unbalance at midband. The splitter/combiners are priced at $18.95 each in quantities of one to nine pieces.

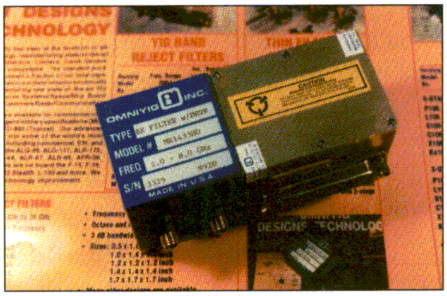


Mini-Circuits
Circle #174

Products

Band reject filter

Omniyig has released a multi-octave band reject filter with an integrated digital driver covering a frequency range of 2.0 to 8.0 GHz. At 2.0 GHz, the 30 dB rejection is 9 MHz, and at 8.0 GHz, the 60 dB rejection is 30 MHz. This unit is integrated with a 12-bit digital driver to tune the full frequency range. The operating temperature ranges from –54 degrees Celsius to +85 degrees Celsius, and the filter is built to mil-e-5400 specifications.



Omniyig, Inc.
Circle #175

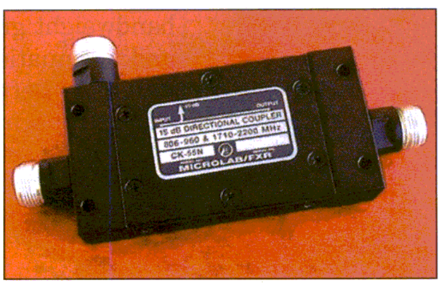
Switch filter bank

Planar Filter Company announces the release of a micro-miniature switch filter operating from 50 MHz to GHz. Insertion loss is –2.5 dB, and VSWR is typically 1.5:1. The unit offers 60 dB isolation per channel and operates from a single +5 VDC at 200 mA.

Planar Filter Company
Circle #176

Directional coupler

Microlab/FXR announces an addition to its range of directional couplers for the 800 to 2200 MHz wireless frequency bands used for in-building applications. The model CK-55N 15 dB directional coupler offers passive intermodulation below –140 dBc because it uses an air dielectric between silver-plated conductors to minimize main line loss.



Microlab/FXR
Circle #177

Miniature cavity diplexer

Wireless Technologies Corpor-ation has introduced a miniature cavity diplexer that supplies both antenna signals, Rx and Tx, to one antenna. It has a return loss of greater than –18 dB at all ports; a power capability greater than 10 watts at both bands, and a type SMA or N weatherproof connector.

Wireless Technologies Corporation
Circle #178

Switched-combiner module

Signal Technology Corporation's Model O/N-SDU-2146 is a full function 4-way switched-combiner module with an integral power divider. It operates in the cellular band of 869 to 894 MHz over a temperature range of 0 to 80 degrees Celsius. The combiner is optimized for operation in the 3 and 4 port active mode and features an insertion loss of less than 0.5 dB with a VSWR of less than 1.25:1 at the combiner port in these conditions. Loss remains under 0.8 dB with a VSWR of better than 2.0:1. A phase balance of higher than ±2 degrees is maintained in all of the above modes. The module is priced at $1,495 each in quantities of five to nine.

Signal Technology Corporation
Circle #179

High-frequency filters

Faraday Technology has recently extended its capability to supply filters up to 18 GHz to the video industry. Faraday can provide support on rapid filter prototyping and large volume production.

Faraday Technology Limited
Circle #180

Cable loads allow heat reduction

RF Power, a subsidiary of Anaren Microwave, introduces cable loads that assist RF designers in the reduction of thermal gradients of sensitive components. This unit allows designers to move heat generating RF terminations away from more sensitive components such as transistors, filters and ferrite devices.

RF Power
Circle #181

Duplexer for GPS

Temex Components introduces duplexers for GPS military airborne applications at 1227 and 1275 mHz. With loss at lower than 1 dB, these duplexers offer minimum 0.5 dB bandweight greater than ±10 MHz and attenuation in the other channel is better than 40 dB. Dimensions are approximately 1.1 by 0.55 by 0.275 inches. SMD and PC board packages are available.

Temex Components
Circle #182

SOFTWARE

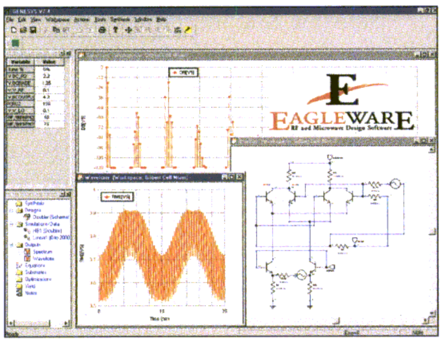
Radio link design software

Radio Engineering Services has released its HERALD Professional Edition, a Windows-based PC program developed to assist the radio engineer in designing point-to-point radio links. The software allows designers to use a single command to initiate a search for all the significant interference elements throughout the entire network. Additional features include link budget, profile clearance and obstructions, reflections, prediction of unavailability due to rainfall, potential for multipath outage, diversity reception, and the impact of incorporating passive repeaters.

Radio Engineering Services
Circle #183

Harmonic balance simulator introduced

Eagleware announces the release of a new harmonic balance simulator (HARBEC), a fast and easy way for designers to co-simulate EM,



linear and non-linear circuits. Users can specify any number of fundamental tones for analysis, creatng an arbitrarily complex spectrum limited only by computer resources. The engine has been developed with artificial intelligence techniques to find and use the best convergence strategy. HARBEC is priced at $4,990 and is fully and transparently integrated into the GENESYS environment.

Eagleware
Circle #184

Network planning solution

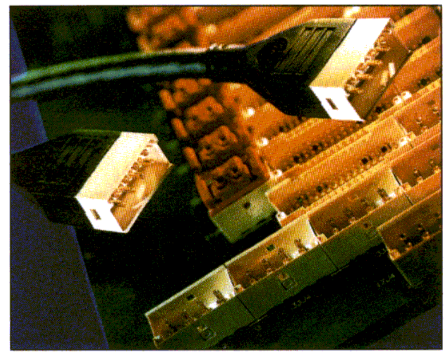
Freshfield Communications offers ORTHIA, an advanced solution for broadband wireless access, LMDS, and MMDS operators that enables them to manage and plan their network anytime, anywhere, using state-of-the-art technology at low cost. This software supports rapid network roll-out and eliminates the need to purchase additional tools or resources.

Freshfield Communications
Circle #185

CABLES AND CONNECTORS

Mini coax connector system expanded

Harting has expanded the scope and performance of its har-pak® mini coax connector system. The system is flexible and can accommodate a wide variety of cable assemblies, including SMA, SMB and type

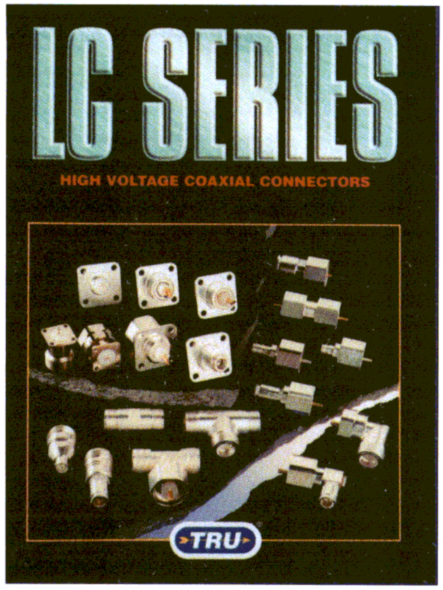


N. The modules are available in various sizes with 2 to 10 coax lines per module. Straight and angled connectors feature blind mating board-to-backplane RF connection and are press-fitted with flat rock tooling. The system also features board-to-cable RF interconnection and Float Plate solutions for blind mating of heavy sub-assemblies. The system is priced at $28 each in quantities of 100 pieces or more.

Harting, Inc.
Circle #186

RF coaxial connectors

Tru-Connector Corporation offers a new 12-page catalog featuring large-sized, high voltage 50-ohm coaxial connectors designed for cables from .393 to 1.00 inches O.D. The catalog provides specifications, engineering drawings and parts numbers. Additional products



offered in the catalog include plugs, jacks, receptacles, in- and between-series adapters and complete cable assemblies and caps.

Tru-Connector Corporation
Circle #187

Standard and custom semi-rigid assemblies

RF Connectors, a division of RF Industries, now offers both standard and custom semi-rigid cable assemblies. RF Connectors provides RG-405, RG-402 and custom assemblies manufactured to customer specifications. Value-added testing services are available for all cable assemblies.

RF Connectors
Circle #188

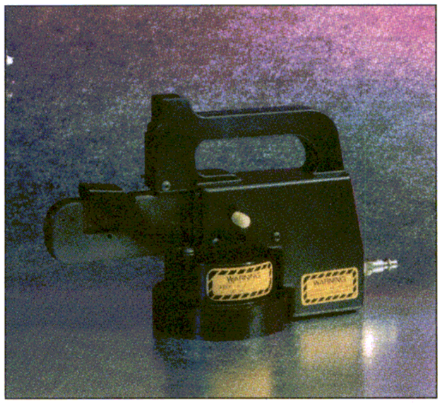
Bulk cables released

Semflex has introduced the Vital Links line of bulk cable products. The DB and SI series are an economical performance upgrade over standard RG178, RG316 and RG142 cables. The SW series is a cost-effective, microporous PTFE cable group offering more flexibility and performance than solid PTFE cables. Flexible and high-powered, the KW series falls in an ultra-low loss cable group. Completing the Vital Links family is the HP series, offering excellent VSWR performance and low loss for applications up to 50 GHz.

Semflex, Inc.
Circle #189

Pneumatic BNC crimper

Trompeter has released a pneumatic hex die benchtop crimp tool for the installation of BNC coax connectors. The tool produces quality crimps while speeding installation efficiency and reducing fatigue and risk of repetitive motion injuries, such as carpal tunnel syndrome. Operating at 32 cycles per minute, the crimper can be actuated by hand or via a foot pedal. Including the foot activation switch

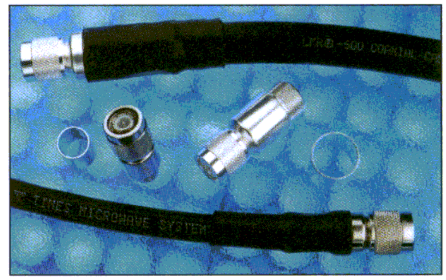


and one die-set, the crimper is priced at $2,800.

Trompeter
Circle #190

Non-solder TNC connectors

Times Microwave Systems has announced the availability of non-solder "EZ" TNC male connectors for its LMR-400 and LMR-600 flexible low loss coaxial cables. The connectors are designed to operate at frequencies up to 6 GHz. Suitable for use as antenna feeders, system jumpers and interconnects, the



flexible, non-kinking, low-loss RF transmission line cables use easy-to-install connectors.

Times Microwave Systems
Circle #191

Solutions for modular space applications

The Phoenix Company of Chicago introduces the PkZ 56 series, a blindmate, size 12 contact that offers high frequency performance, light-weight construction and dense packaging. The PkZ's patented design allows for a "Z"-axis mating tolerance of up to .070 inches without compromising electrical performance while maintaining constant impedance. The series is available in a variety of styles and configurations.

The Phoenix Company of Chicago, Inc.
Circle #192

TEST EQUIPMENT

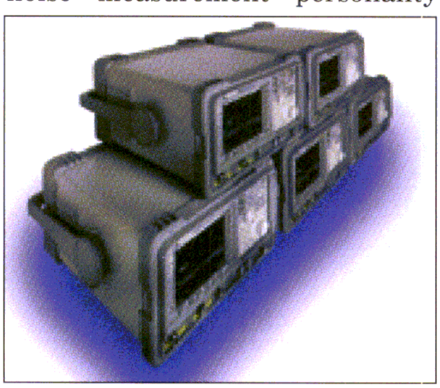
VNA line expanded

Anritsu Company's Microwave Measurement Division has expanded its Scorpion family of Vector Network Measurement Systems with the MS462xC Direct Receiver Access models. The MS4622C provides measurement capabilities from 10 MHz to 3 GHz, while the MS4623C operates from 10 MHz to 6 GHz. Pricing is $28,5000 and $33,500 each, respectively.

Anritsu Company
Circle #193

ESA spectrum analyzer

Agilent Technologies has enhanced features and performance of its the ESA spectrum analyzer for design engineers. The enhancements include a phase-noise measurement personality

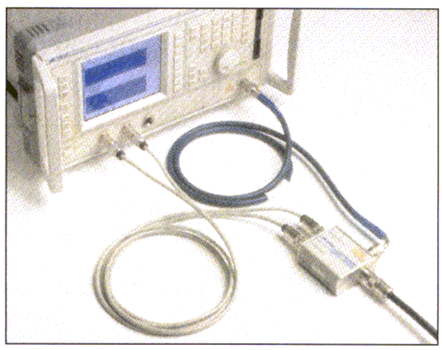


option that provides one-button phase-noise measurements, significantly reducing the amount of time it takes to make the same measurement manually. Also included is a segmented sweep feature with which users can choose to display or remotely transfer between 2 and 8,192 sweep points in a single sweep to optimize measurements for speed or resolution. The cost of a basic model starts at $12,000. Units with greater frequency and performance options begin at $25,000.

Agilent Technologies
Circle #194

Guided scalar measurement software

IFR Systems announces the availability of guided scalar measurement software that allows the 6800 series of scalar and system microwave analyzers to be config-

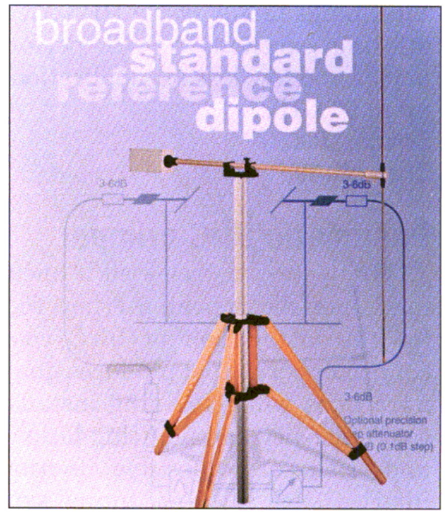


ured so that unskilled operators can simply and precisely perform complex or repetitive measurements. The software is designed for short-run production measurements on components, assemblies and subsystems; for cable and wave guide installation measurements in aircraft and ships; and for radio link feeder testing. Measurements can be customized and changed only with the use of a password. Available with a user manual, the application is priced at $990.

IFR Systems, Inc.
Circle #195

Site calibration system

Schaffner EMC has developed a site calibration system that significantly reduces the time necessary for the verification of EMC test site calibration. Calibration test sites can themselves be accurately calibrated using the new hardware and software combination. It is necessary to use only four dipoles to



cover the range of 30 MHz to 1000 MHz, which can reduce verification times by as much as 80 percent compared to conventional 24 spot measurement techniques. The software ensures that EMC test sites can be calibrated with an uncertainty better than ±0.2 dB in the 30 MHz to 600 MHz range, and ±0.3 dB in the 600 MHz to 1 GHz range. Prices start at $8,500 for a single antenna package.

Schaffner EMC
Circle #196

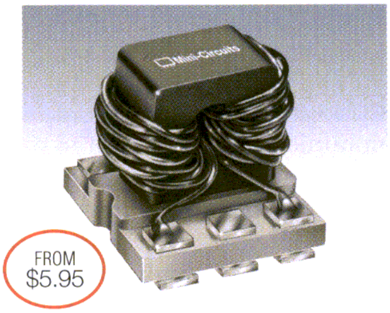
AMPLIFIERS

PA driver amplifier

RF Micro Devices has announced the release of the RF2367, a low noise CDMA/TDMA/GSM PA driver amplifier with a very high dynamic range. The frequency range of 150 to 2500 MHz is optimized for operation in the DCS and PCS bands for applications where low transmit noise power is of concern. Operating from a single 3-volt power supply, the RF2367 features adjustable bias current, a high intercept point and power down control. Manufactured using advanced gallium arsenide HBT process technology, the RF2367 is offered in a small SOT23-6 pack-

NEW PRODUCTS NO. 75

RF/IF MICROWAVE COMPONENTS



"DO-IT-YOURSELF" 10dB COUPLER LOWERS COSTS

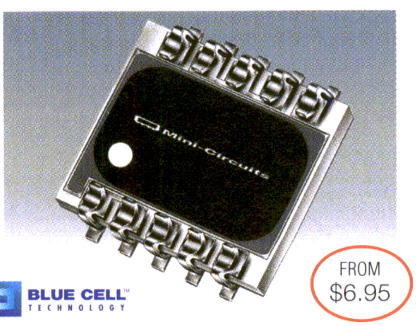
Mini-Circuits TCD-10-1W-75 needs only a commercially available external chip resistor, and a complete 10 to 750MHz 75 ohm directional coupler is realized. Designed for automated manufacturing to lower costs, this rugged "do-it-yourself" SM coupler provides 10.5dB ±0.5dB nominal coupling (±0.7dB max. flatness) with 1.4dB mainline loss and 18dB directivity typical midband. The 50/75 ohm TCD family contains 9 units with 9 to 20dB coupling for 5 to 1000MHz.

5W TYPE-N ATTENUATORS FOR DC TO 18GHz

Mini-Circuits family of broadband DC to 18GHz precision fixed attenuators contain 15 models with nominal attenuation values from 1 to 10dB plus 12, 15, 20, 30, and 40dB. Built tough to handle 5W average with 125W peak power, these small 1.90" units exhibit high temperature stability, outstanding phase linearity, and excellent VSWR. Equipped with stainless steel Type-N Male/Female connectors. Model number is BW-NXW5 substituting X with desired value.

612 TO 1200MHz VCO HAS LINEAR TUNING

Mini-Circuits ROS-1200W voltage controlled oscillator performs in the 612 to 1200MHz band targeting cellular and test equipment applications with low -139dBc/Hz SSB phase noise typical at 1MHz offset, wide 3dB modulation bandwidth typical at 20MHz, and 26-68MHz/V (typ) tuning sensitivity. Housed in a miniature 0.5"x0.5"x0.18" industry standard surface mount package, typical power output is 10dBm. Harmonics is -28dBc typical (specified to the fourth).



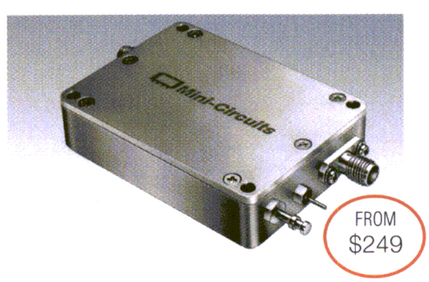
BLUE CELL™ TECHNOLOGY

ULTRA-THIN 0.07" MIXER PERFORMS 1600 TO 3200MHz

The MBA-18L level 4 (LO) frequency mixer, part of Mini-Circuits patented family of Blue Cell™ mixers, operates over 1600MHz to 3200MHz with 24dB L-R, 16dB L-I isolation and low 6.5dB midband conversion loss (all typ). The Blue Cell™ mixer series delivers a unique combination of performance repeatability, low conversion loss, superb temperature stability, thin profile, and very low cost. This model is ideal for satellite, GPS, WLAN, and PCMCIA applications.

CELLULAR BAND AMPLIFIER IS LOW NOISE SOLUTION

Mini-Circuits announces a new 25dBm (typ. output at 1dB comp.) medium high power 50 ohm amplifier ideal for cellular applications in the 800 to 900MHz band. The ZQL-900MLNW typically displays ultra-low 1.2dB noise figure and high +41dBm IP3 to suppress intermodulation products. Gain is 27dB (±1.8dB flatness) and VSWR is 1.3:1 in/1.4:1 out (all typ). This tough built coaxial amplifier is equipped with SMA-Female connectors and operates within -40°C to +70°C (max.).



BROADBAND 2WAY SPLITTER SPANS 20 TO 3000MHz

Designed to split a signal 2ways 0°, the 1 watt (max. input as splitter) ZAPD-30 power splitter from Mini-Circuits covers the broad 20 to 3000MHz frequency band. This SMA-Female coaxial unit displays 0.1dB amplitude and 1 degree phase unbalance (typ), plus low 1.1dB insertion loss (above 3.0dB) and 16dB isolation typical at midband. Housed in a rugged metal case, applications include UHF TV/DTV, aircraft radio navigation, and PCS/Cellular/GSM. Value priced.

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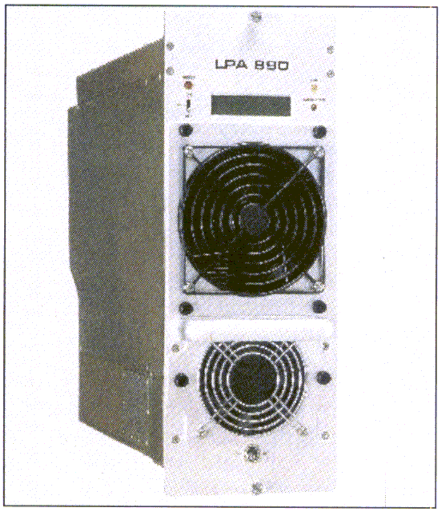
age, and sells for $0.50 in orders exceeding 10,000 units.
RF Micro Devices
Circle #197

Multi-carrier amplifiers

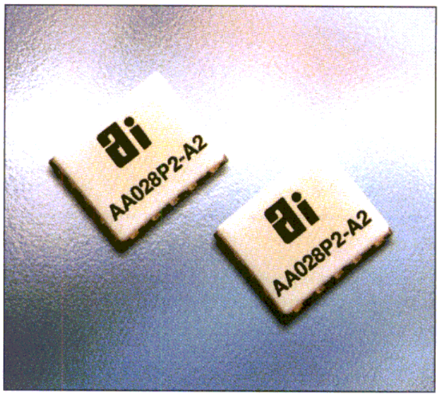
HC Electronics has released a new selection of high output 90-watt and 60-watt multi-carrier linear power amplifiers operating in the 800 MHz band. Two models are offered: the 90-watt LPA890, which functions in the 824 to 894 MHz or 869 to 894 MHz bands; and the LPA860, which operates in the 869 to 894 MHz band. Both modules can achieve higher output through the use of intelligent rack combiners. The LPA890 model features 60 dB gain with intermodulation distortion rated better than –60 dBc. The LPA860 model achieves 58 dB gain with intermodulation distortion levels better than –60 dBc.



HC Electronics
Circle #198

Surface mount amplifier

Alpha Industries introduces its latest 25 to 31 GHz surface mount amplifier, offering 13 dB gain. The amplifier is designed for use in communication and sensor systems as a gain stage in the receiver, transmitter, or LO chain when high gain and linearity are required. It is ideally suited for high volume millimeter wave applications, such as point-to-point and point-to-multipoint wireless communications systems. The unit is priced at $12.60 each in quantities of 10,000.



Alpha Industries
Circle #199

FABRICATION

GALLIUM ARSENIDE

Linear and EM Simulation Cuts Design Time of a 3.8 GHz Amplifier

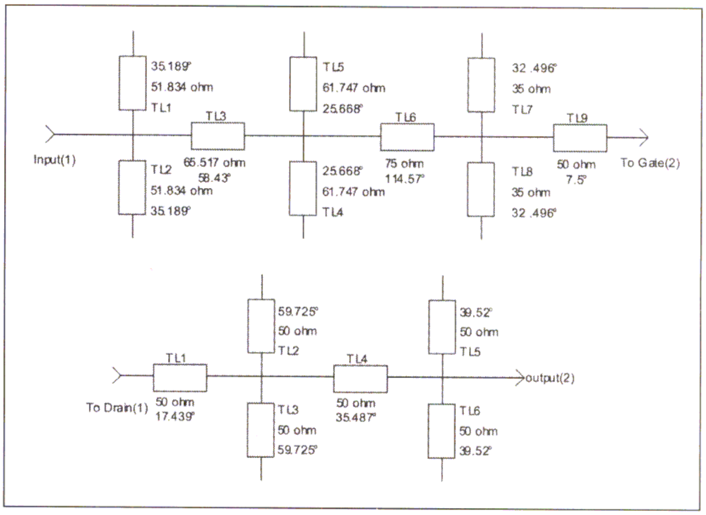
CAD design tools are valuable aids in circuit design and validation

By Sunny Lo
SignalCraft Technologies Inc.

As the operating frequency goes beyond 3 GHz, the feasibility of lumped component matching networks diminishes due to component parasitics as well as tolerance in the component values. An alternative is to use distributed matching networks. However, this leads to another obstacle: discontinuities in these networks are not always modeled with sufficient accuracy in linear simulators.

To avoid the "garbage in, garbage out" pitfall, planar electromagnetic (EM) simulations of these networks are necessary. While most designers appreciate the utility of such a tool, many have limited or no access due to cost of the software. With the introduction of Sonnet Lite, a free planar EM simulator from Sonnet Software, any RF designer can now benefit from the accuracy of EM simulations. While there are limitations to this free program, there are also "tricks" to make it useful in "real" designs [1].

Although accurate, EM simulations consume considerably more CPU time. Random changes to the circuit and re-simulation to optimize a circuit, as often used in linear circuit simulators, would be inefficient if applied to EM simulations. Here is where the speed and flexibility of a linear simulator such as the one included in the GENESYS 7 software from Eagleware can help out. By combining the capabilities of both simulators, the circuit has a much better chance of being right the first time. Using a 3.4 to 4.2 GHz amplifier design, this article will demonstrate how the two programs can be used together to produce such a circuit.

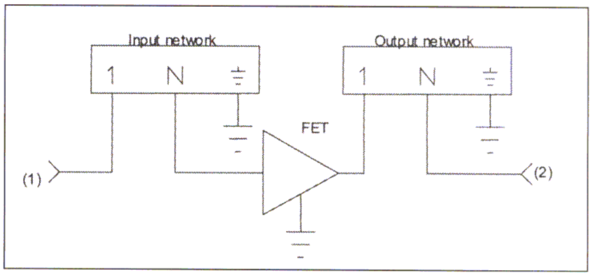


▲ **Figure 1. Synthesized matching networks from =MATCH=, with ideal transmission lines.**

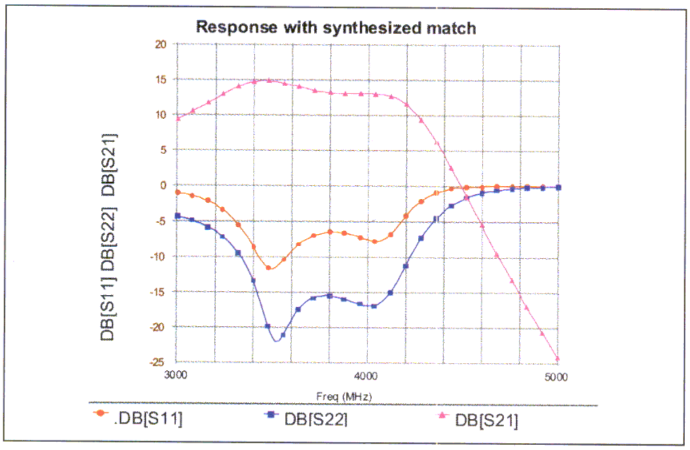
The basic performance requirements for the amplifier are:

- Frequency range: 3.4 to 4.2 GHz
- Gain: >12 dB
- Output Return Loss: < –15 dB

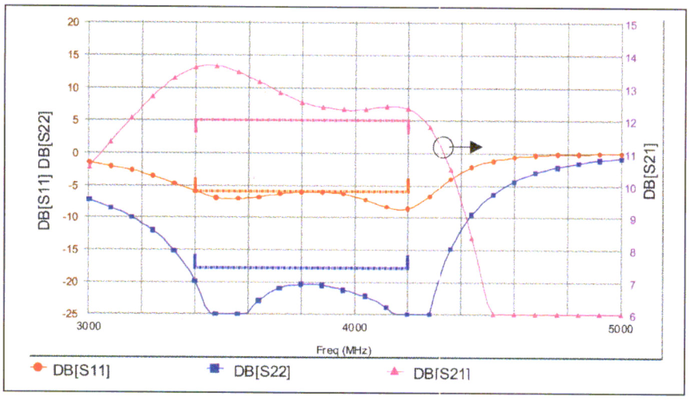
Once the active device has been selected, the



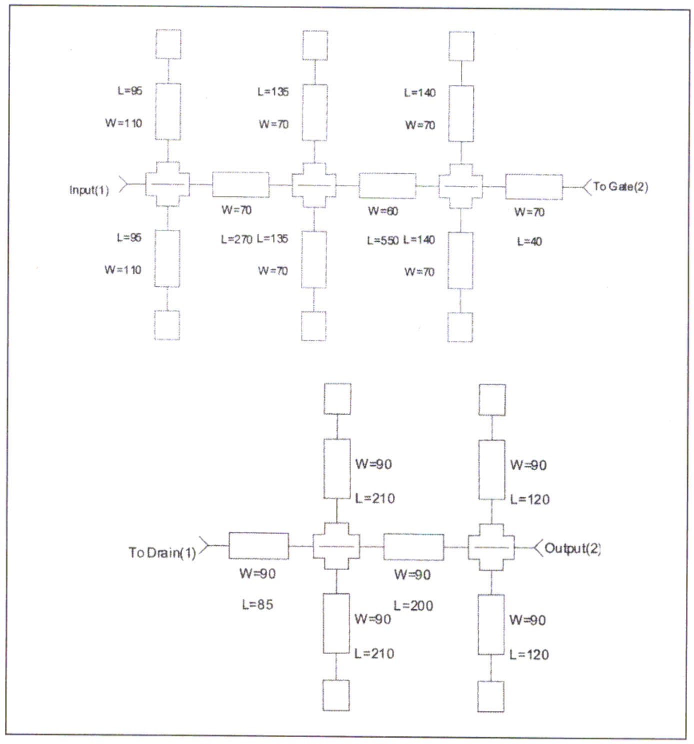
▲ **Figure 2. PA block diagram.**



▲ **Figure 3. PA response with synthesized matching networks from =MATCH=.**



▲ **Figure 5. PA response with microstrip elements as optimized in GENESYS 7.**



▲ **Figure 4. Matching networks as optimized by GENESYS 7 to include microstrip discontinuities.**

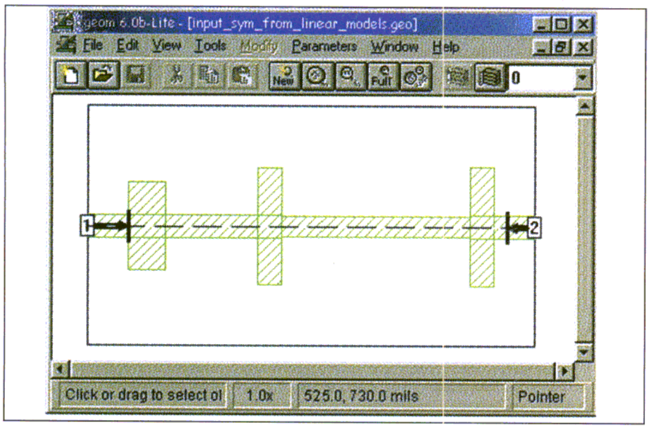
S-parameters will be used to generate the termination impedance for a "simultaneous conjugate match." This, in turn, is used in a synthesis program such as =MATCH= to generate a matching network.

Obtaining accurate *S*-parameters for the active device before starting the matching synthesis is very important. This means measuring the device even if there is published data. This will save time and frustration.

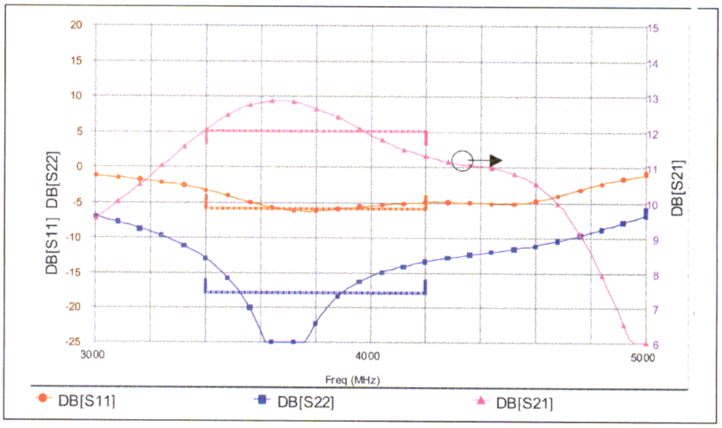
To move one step closer to a "real" circuit, the synthesized networks are transferred to GENESYS 7.0's linear simulator, where microstrip discontinuities are included in the simulations. To conserve memory usage in Sonnet Lite, the matching networks are made symmetric about the x-axis [1], as shown in Figure 1. The circuit parameters are then optimized to meet the design goals.

Necessity of EM simulations

The matching networks from GENESYS 7 are



▲ **Figure 6. Input network in Sonnet Lite using dimensions generated by GENESYS 7.**



▲ **Figure 7. PA response with EM simulations of the matching networks optimized by GENESYS 7.**

	Starting Value (mils)	After optimization to match Sonnet results (mils)	Error (mils)	Corrected value for Sonnet (mils) Starting Value + Error
L1	140	98	42	182
L2	135	90	45	180
L3	95	53.5	41.5	137
Lout1	210	147	63	273
Lout2	120	70	50	170

▲ **Table 1. Summary of matching network stub lengths.**

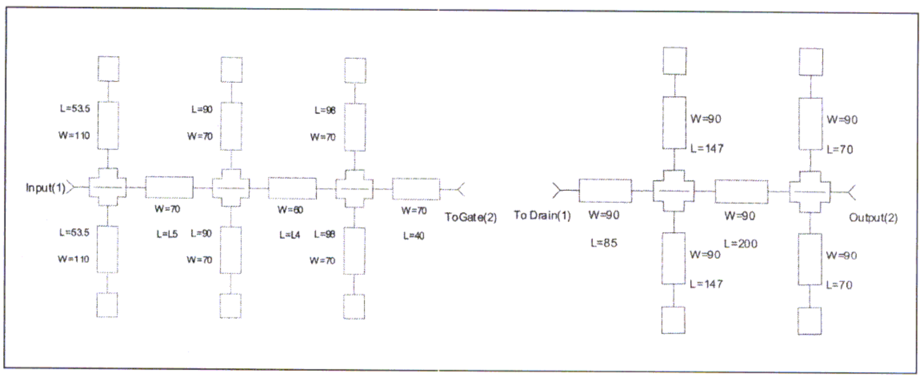
reproduced in Sonnet Lite, and the EM simulation results from Sonnet Lite are used for comparison. A screen shot of Sonnet Lite showing the input matching network is shown in Figure 6. The assumption here is that the EM results will be representative of the "real" results if the circuits were built. The simulation results are shown in Figure 7. The responses are significantly different from the desired response of Figure 5.

Circuit optimization

Figure 7 indicates that the matching networks need to be modified. To determine which element to change and by how much, we will first use the optimization feature of the linear circuit simulator (GENESYS 7, in this example) to modify the original networks to match the results of the first set of EM simulations. The differences in the dimensions are the "errors" and will be applied to the original circuit (generated by GENESYS 7). These new dimensions will then be incorporated into the Sonnet Lite simulations to check their effectiveness.

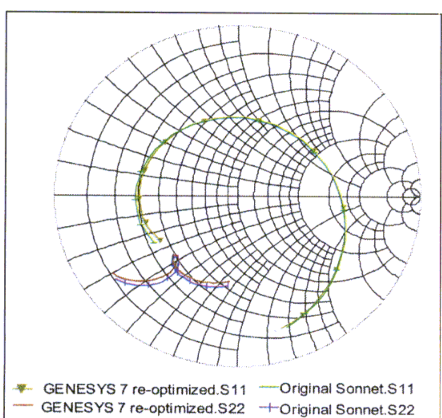
To keep things simple, only the lengths of the stubs will be varied in the optimization. If the results are not satisfactory, more parameters, such as line width and through-line length, can be made available to the optimizer. Figures 8 through 11 show the optimized dimensions and the resulting S-parameters. A summary of the stub lengths before and after "optimization" is given in Table 1. L1, L2, and L3 refer to the input network. Lout1 and Lout2 refer to the output network.

Even with relatively large correction factors (26 ~ 46%), the corrected dimensions from GENESYS 7 produced excellent results when re-simulated in Sonnet Lite. Using a 5-mil grid in Sonnet Lite, the dimensions in Table 1 are rounded off to the nearest increment.

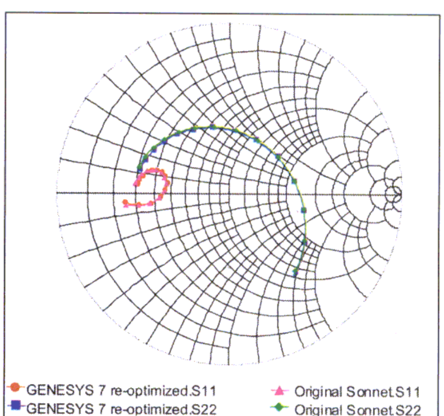


▲ **Figure 8. Matching networks "optimized" to match results from Sonnet Lite.**

As shown in Figures 12 through 14, the corrected dimensions produced EM results that are quite close to the intended response. One final modification to the input network is needed before the design can be fabricated: the stubs closest to the FET gate need to be altered to clear the solder pad for the Source connection of the FET (SOT89). This is another situation where EM simulation is valuable, enabling arbitrary conductor shapes to be used when needed. The resulting layout is shown in Figure 16.



▲ **Figure 9. *S*-parameters of the input network.**



▲ **Figure 10. *S*-parameters of the output network.**

Measurements vs. simulations

The final test of any simulation is, of course, actual results from a real circuit. Figures 17 through 19 compare the simulation results with the network analyzer measurements.

The measured results correlate well with the simulations. The frequency shift in the passband is related to the slight differences in the test jig that was used to measure the S-

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ERA-2	DC-6000	15.6	12.8	4.7	26.0	40	1.95
ERA-2SM	DC-6000	15.2	12.4	4.6	26.0	40	2.00
ERA-3	DC-3000	20.8	12.1	3.8	23.0	35	2.10
ERA-3SM	DC-3000	20.2	11.5	3.8	23.0	35	2.15
ERA-4	DC-4000	13.5	▲17.0	5.5	▲32.5	65	4.15
ERA-4SM	DC-4000	13.5	▲16.8	5.2	▲33.0	65	4.20
ERA-5	DC-4000	18.8	▲18.4	4.5	▲33.0	65	4.15
ERA-5SM	DC-4000	18.5	▲18.4	4.3	▲32.5	65	4.20
ERA-6	DC-4000	11.3	▲18.5	8.4	▲36.5	70	4.15
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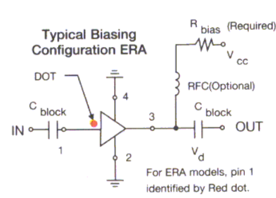

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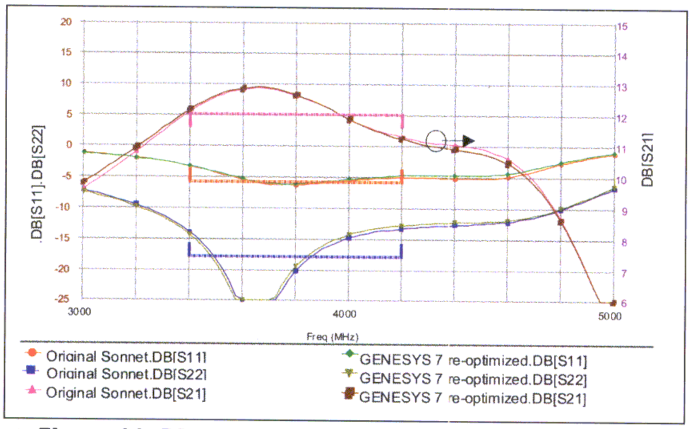
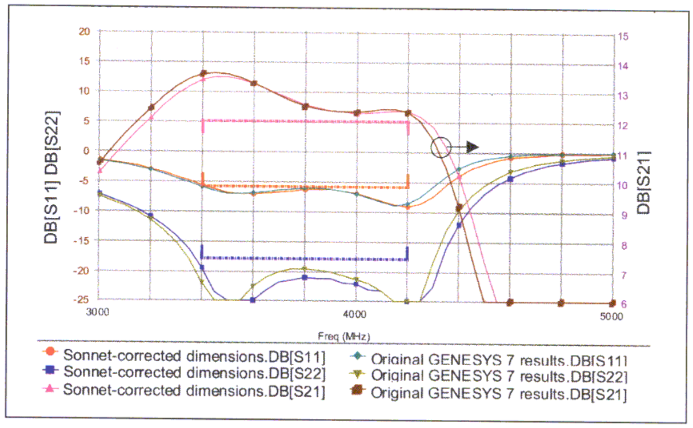
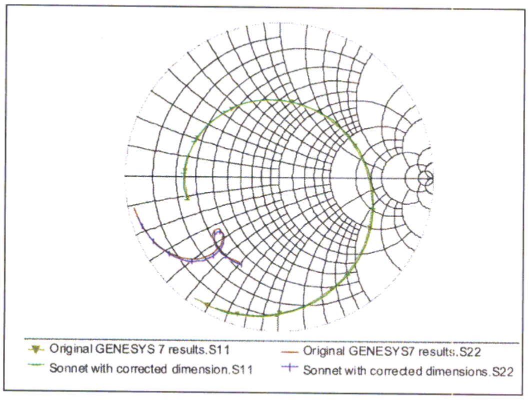


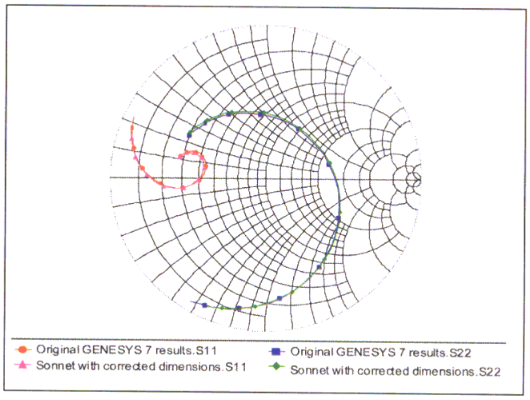
▲ **Figure 11. PA response with microstrip networks tuned to match Sonnet Lite results.**



▲ **Figure 14. PA response with Sonnet results using "corrected" dimensions.**



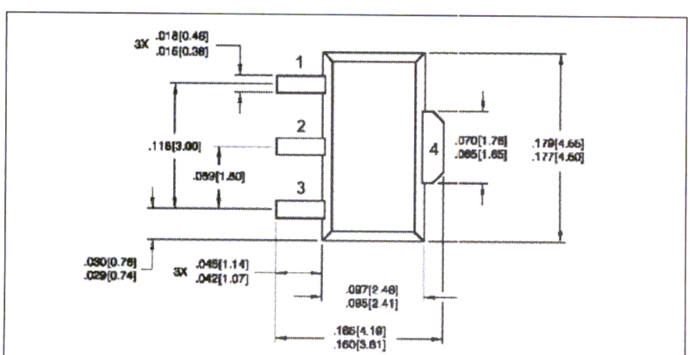
▲ **Figure 12. Sonnet results for input network with corrected dimensions.**



▲ **Figure 13. Sonnet results for output network with corrected dimensions.**

parameters of the FET and the actual layout of the PA. When the PCB for the PA was made, a new test jig was added that would mimic the grounding scheme of the actual layout (see Figure 16). With the new test jig, the S-parameters were re-measured. The results are shown in Figure 20.

Besides showing improved correlation between simulation and measurements, the results in Figure 20 also illustrate the importance of accurate device S-parameters.



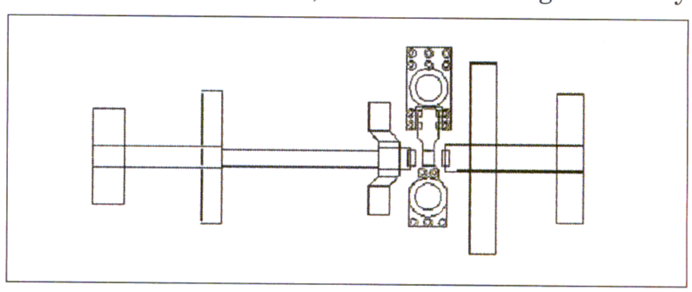
▲ **Figure 15. SOT89 package outline.**

Practical design considerations

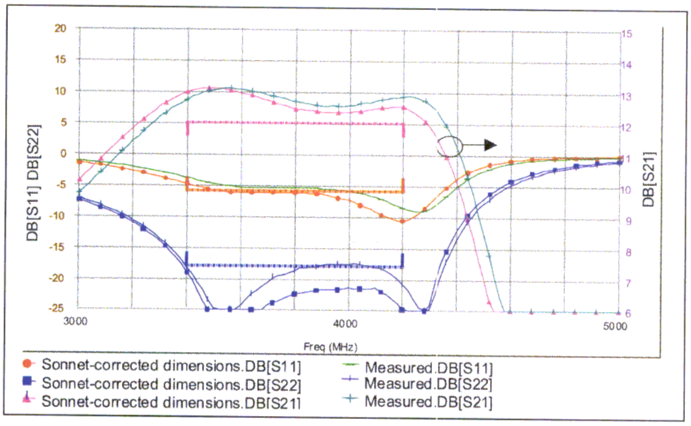
This article has thus far bypassed the biasing networks that would be needed to complete the design (the above results were obtained using the bias tees of the network analyzer). Now that the main concepts and results have been presented, some comments on the topic are appropriate.

Due to the moderate bandwidth requirement (~20%), $^{\lambda}/_{4}$ high impedance transmission lines or self-resonant chip inductors can be used to provide a DC bias path. The RF "short" at the end of the transmission line or inductor can be a chip capacitor that is at or near its series self-resonance. The effect of the bias networks on the overall circuit can be simulated by attaching it to the existing input and output ports of the PA.

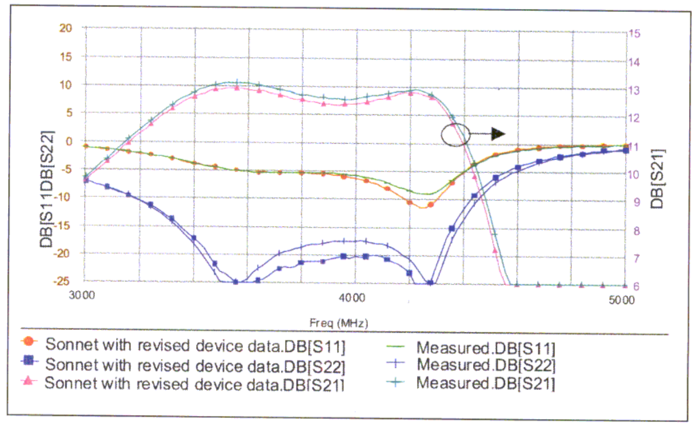
Another practical issue that arose for the above example is the length of the input matching network. To conserve space, the long section of transmission line can be folded into a "U," as shown in Figure 21. By



▲ **Figure 16. Layout for PA with input network modified to clear the ground tab of the FET.**

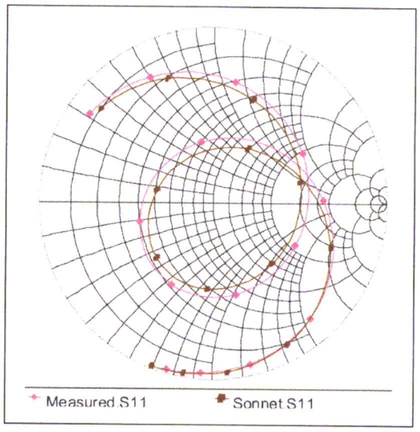


▲ **Figure 17. Measured results vs. simulation.**

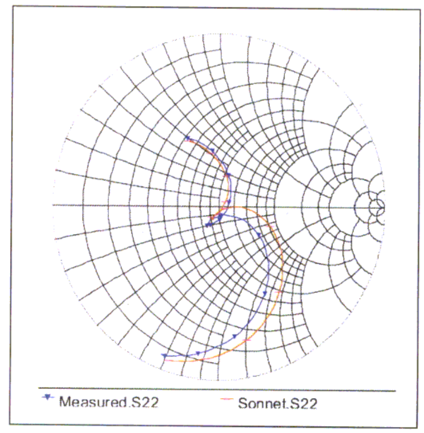


▲ **Figure 20. Measured results vs. simulation with revised *S*-parameters for the FET.**

introducing four bends, the dimensions of that section of transmission line need to be modified. Once the dimensions of the "U" have been optimized to match the results of the straight piece, the new network can be re-simulated. Instead of simulating the entire network, which would exceed the memory limit of Sonnet Lite, the network will need to be cut into three sections and their results recombined to get the overall response. A more detailed example of this partitioning trick is presented in [1].



▲ **Figure 18. Measured results vs. simulation: PA S_{11}.**



▲ **Figure 19. Measured results vs. simulation: PA S_{22}.**

PCB variations

PCB parameters such as dielectric constant, board thickness and etching variations can have a significant effect on circuit performance. Through Monte Carlo simulations, the speed of the linear circuit simulator can provide some assurance that the design will be repeatable in production. Small (less than grid size) changes in conductor dimensions are not easily handled in Sonnet Lite, and even then it would be difficult to determine the worst-case scenario for the different parameters.

Conclusions

Fast and accurate simulations need not always be expensive. By combining the efficiency of the optimizer in GENESYS 7 with the enhanced accuracy of Sonnet Lite, a wide band 3.8 GHz amplifier was able to meet design objectives in a single pass with minimal number of EM simulations. ■

Acknowledgements

The data presented in this article was collected while I was on subcontract at Microlynx Systems of Calgary, Alberta, Canada. I would like to thank Bill Durtler of Microlynx for permitting me to publish the work, and my colleagues for their comments on the article.

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1. James C. Rautio, "Tips and Tricks for Using Sonnet Lite — Free EM software will radically change the way you do high frequency design," *Microwave Product Digest,* November 1999, pp. 30–34, 67–70.

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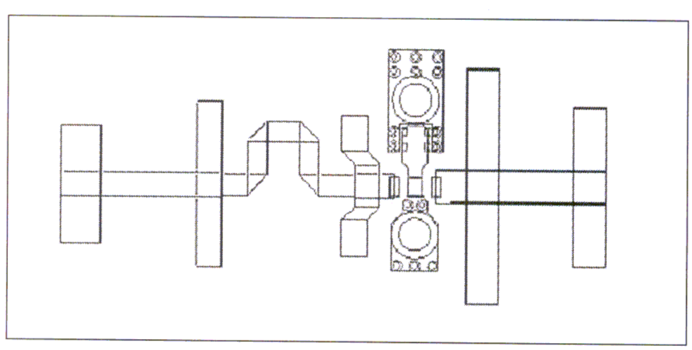

▲ **Figure 21. Input matching network with folded section.**

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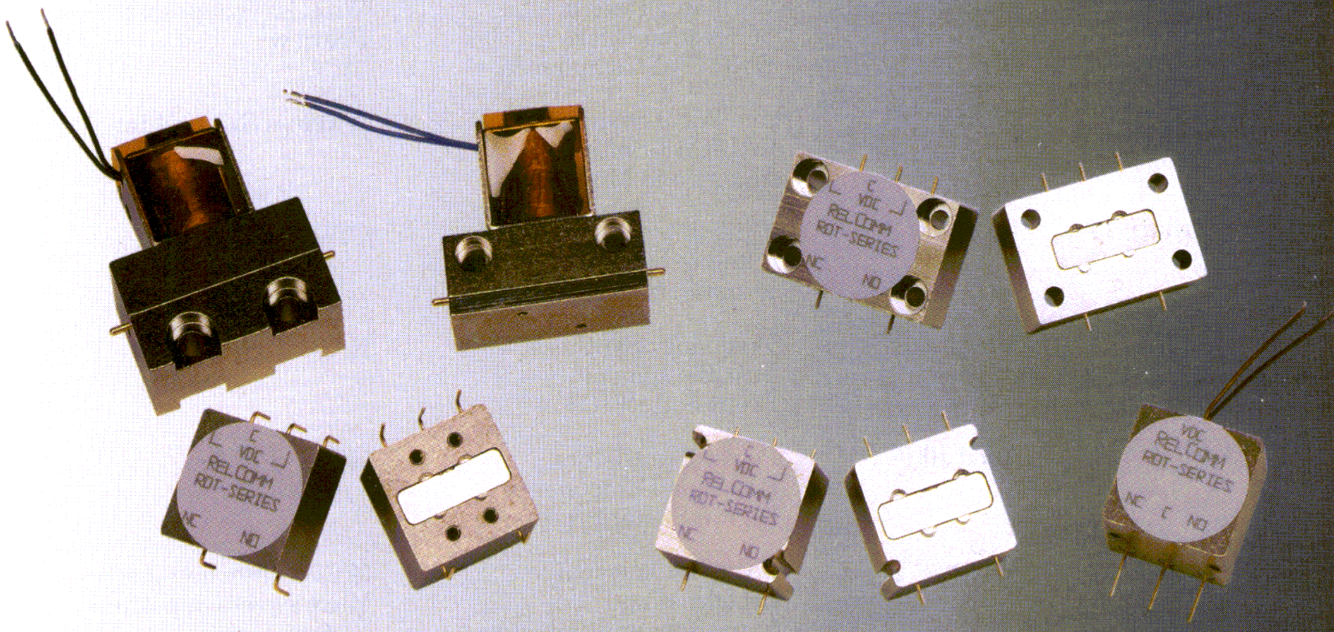

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Dielectric Materials at Microwave Frequencies

The effects of microwave energy on materials are important in industry, biology, medicine and your microwave oven

Kurt Fenske and Devendra Misra
University of Wisconsin-Milwaukee

Radio frequency and microwave signals have numerous scientific and industrial applications in modern technology. Among others, these include wireless communications, telemetry, biomedical engineering, food science, material processing and process controls in the industry. These applications require the electrical characteristics of media with which the signal interacts or through which the signal propagates. This article briefly reviews the characteristic parameters associated with dielectric materials and their characterization techniques.

Microwave energy is frequently employed in industrial processing of materials not only because of its efficiency but also because it yields a superior product. For example, the conventional process to dry the paper-ink using hot air in a printing press reduces the moisture content of the paper as well. This, in turn, reduces the life span of the product. On the other hand, microwaves leave the paper almost unaffected. Since microwave energy can pass through non-conducting (dielectric) materials, it can be used to cure composite structures efficiently. It also allows constructing internal images of the dielectric objects.

In biomedical engineering, microwaves are used to produce controlled heating for hyperthermia. Wireless communication signals interact with the medium they propagate through. Therefore, knowledge of the medium characteristics can assist in designing an efficient communication network.

In brief, there are many engineering systems that involve interaction of electromagnetic energy with the material medium. The design of these systems requires a good understanding of the phenomenon that is directly related to electrical characteristic parameters of the medium. This paper summarizes the characteristic parameters associated with non-magnetic dielectric materials and their characterization techniques.

Dielectric materials

In dielectric materials, most of the charge carriers are bound and cannot participate in electrical conduction. However, if an external electric field is applied to the material, these bound charges may be displaced. This displacement of charge creates a dipole field that opposes the applied field and the material is polarized.

In a linear and isotropic medium, the volume density of polarization is directly related to the applied electric field intensity. This is given by

$$\vec{P} = \varepsilon_0 \chi_e \vec{E} \tag{1}$$

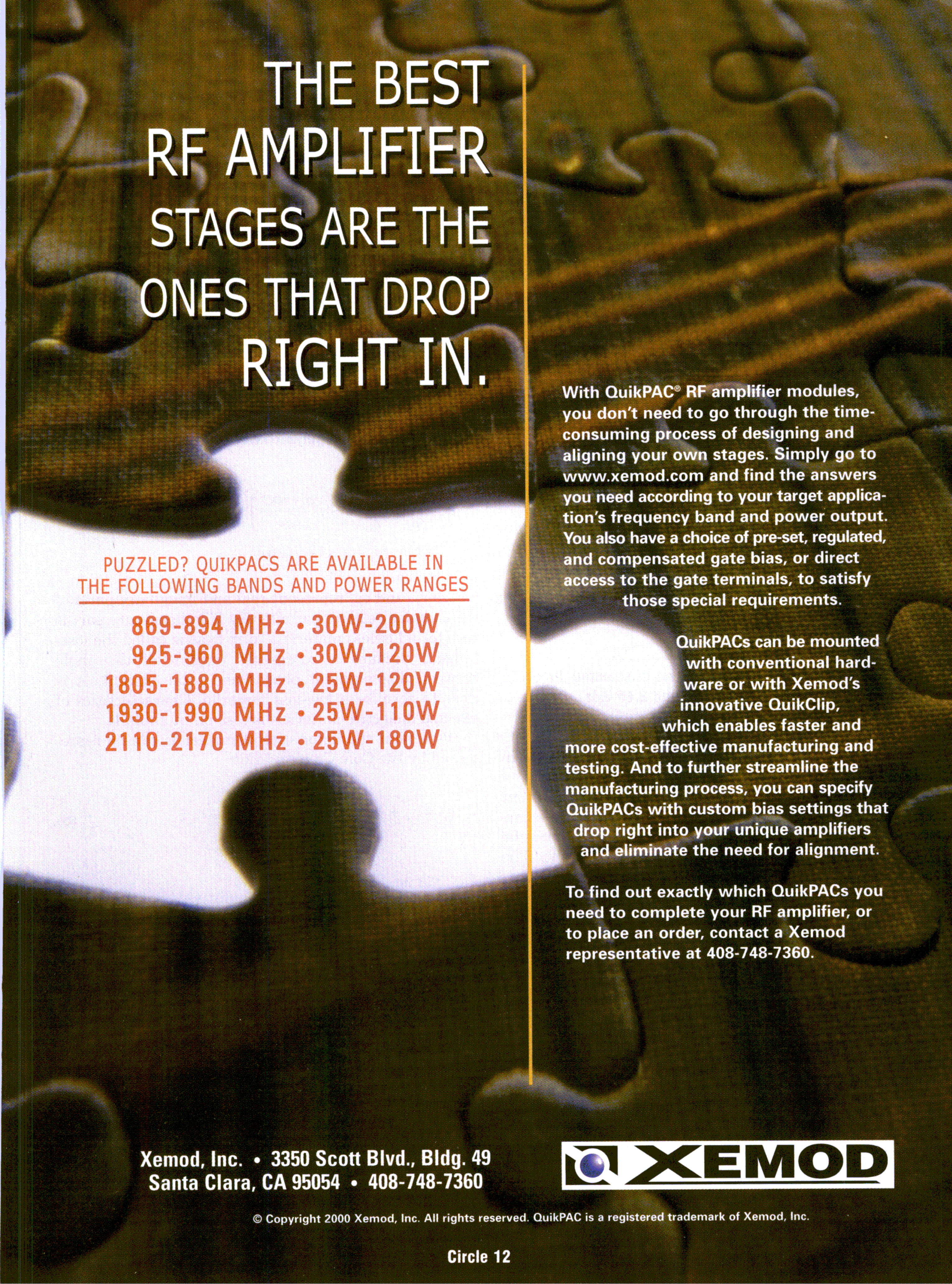
where ε_o is the permittivity of free-space and χ_e is electrical susceptibility of the material.

Electric flux density or displacement, $\vec{D}$, is related to the polarization density and the electric field intensity as follows:

$$\vec{D} = \varepsilon_0 \vec{E} + \vec{P} = \varepsilon \vec{E} = \varepsilon_0 \varepsilon_r \vec{E} \tag{2}$$

where ε is called the permittivity of the material. Since it is a very small number, the relative permittivity, er, of a material is generally specified for convenience. It is also known as the dielectric constant of that material.

Equation (2) is valid only in the frequency domain unless permittivity is independent of the frequency. Most of materials are dispersive, that is, their permittivity is frequency-dependent. If this is the case, then the right-hand-side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Material	Dielectric constant	Loss-tangent
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Bacon (smoked)	2.50	0.05
Beef (frozen)	4.4	0.12
Beef (raw)	52.4	0.3302
Blood*	58	0.27
Butter (salted)	4.6	0.1304
Butter (unsalted)	2.9	0.1552
Borosilicate glass	4.3 0	0047
Concrete (dry)	4.5	0.0111
Corn oil	2.6	0.0077
Cottonseed oil	2.64	0.0682
Sandy soil (dry)	2.55	0.0062
Egg white	35.0	0.5
Fused quartz	4.0	0.0001
Fat*	5.5	0.21
Glass ceramic	6.0	0.0050
Lard	2.5	0.0360
Lung*	32	0.3
Muscle*	49	0.33
Nylon	2.4	0.0083
Olive oil	2.46	0.0610
Paper	3-4	0.0125–0.0333
Soda lime glass	6.0	0.02
Teflon	2.1	0.0003
Thermoset polyester	4.0	0.0050
Wood	1.2–5	0.0040–0.4167
*At 37°C		

▲ **Table 1. Characteristic parameters of selected dielectric materials at room temperature and 2.45 GHz.**

of Equation (2) transforms to a convolution integral in the time-domain.

A time-varying electric field induces two different kinds of currents in a material medium. Conduction current is produced by a net flow of free charges, and the bound charges generate a displacement current. The former is related to the electric field intensity by Ohm's law as follows:

$$\vec{J}_C = \sigma\vec{E} \tag{3}$$

where J_C is the conduction current density that is expressed in Ampere per meter, and σ is the conductivity of material in Siemens per meter.

Displacement current density J_D is related to electric flux density by

$$\vec{J}_D = j\omega\vec{D} \tag{4}$$

Total current density J_T is the sum of conduction and displacement current densities. Hence,

$$\vec{J}_T = \sigma\vec{E} + j\omega\varepsilon\vec{E} \tag{5}$$

Conduction current represents the loss of power. There is another source of loss in dielectric materials. When a time-harmonic electric field is applied, the polarization dipoles flip constantly back and forth. Since the charge carriers have a finite mass, field must do work to move them and the response may not be instantaneous. Hence, the polarization vector may lag behind the applied electric field, which is noticeable especially at higher frequencies. In order to include this phenomenon, equation (5) is modified as follows:

$$\begin{aligned}\vec{J}_T &= \sigma\vec{E} + \omega\kappa''\vec{E} + j\omega\varepsilon\vec{E} \\ &= j\omega\left(\varepsilon - j\frac{\sigma + \omega\kappa''}{\omega}\right)\vec{E} = j\omega\varepsilon^*\vec{E}\end{aligned} \tag{6}$$

where ε* is called the complex permittivity of material.

Complex relative permittivity of a material is defined as follows:

$$\begin{aligned}\varepsilon_r^* &= \frac{\varepsilon^*}{\varepsilon_o} = \frac{1}{\varepsilon_o}\left(\varepsilon - j\frac{\sigma + \omega\kappa''}{\omega}\right) \\ &= \varepsilon_r' - j\varepsilon_r'' = \varepsilon_r(1 - j\tan\delta)\end{aligned} \tag{7}$$

where ε_r' and ε_r'' represent real and imaginary parts of the complex relative permittivity. The imaginary part is zero for a lossless material. Term tanδ is called the loss-tangent because it represents tangent of the angle between displacement phasor and total current. It is close to zero for a low-loss material. Characteristics of selected dielectric materials are given in Table 1.

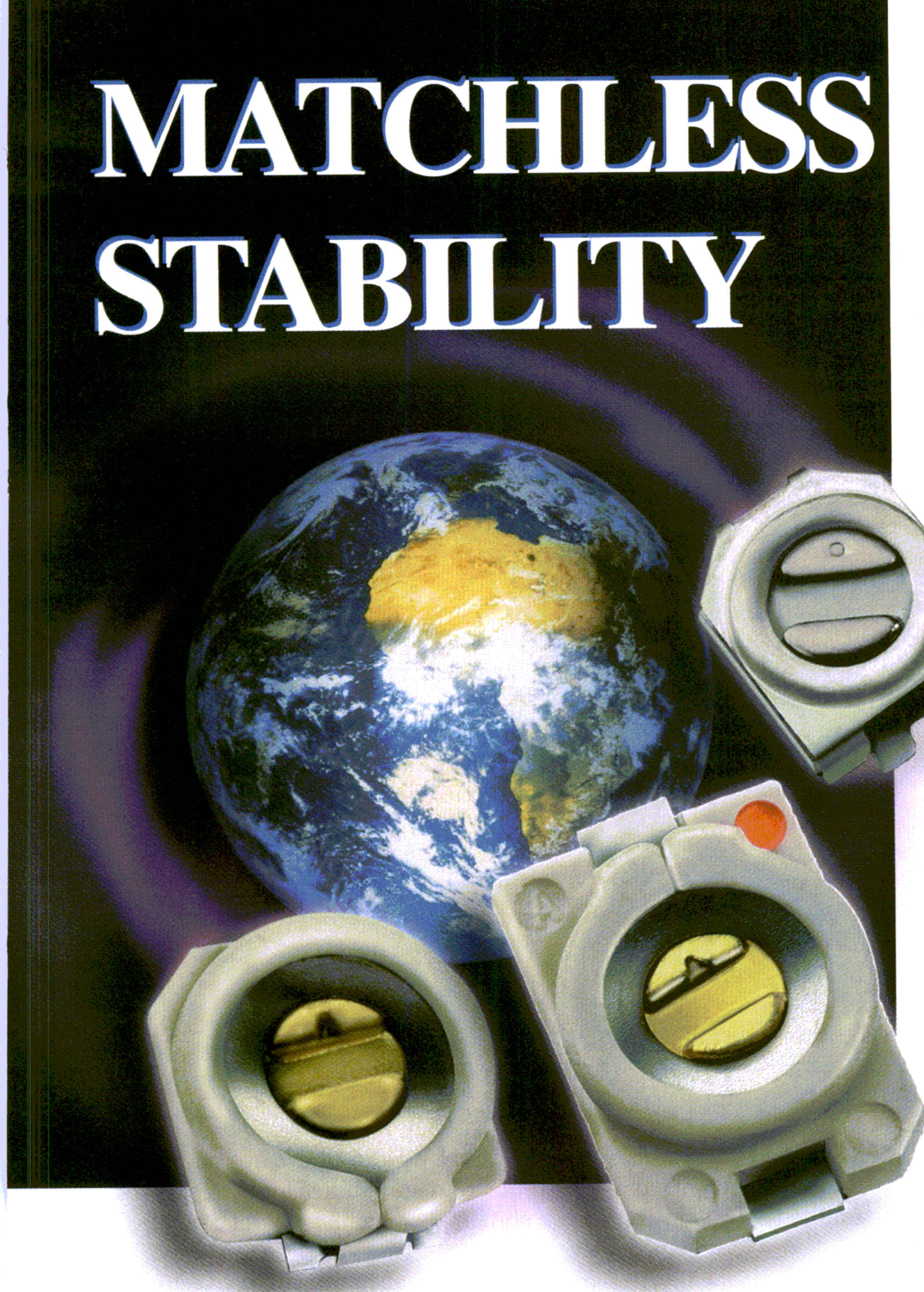
Dispersion characteristics of materials can be represented by the Cole-Cole equation

$$\varepsilon_r^* = \varepsilon_\infty + \frac{\varepsilon_s - \varepsilon_\infty}{1 + \left(j\frac{f}{f_r}\right)^{1-\alpha}} \tag{8}$$

where ε_∞ and ε_s are relative permittivities of the material at infinite and zero frequencies, respectively. Frequencies f and f_r (in Hertz) represent the signal frequency and the characteristic relaxation frequency of the material, respectively. If α is zero, then Equation 8 can be reduced to the Debye equation. Dispersion parameters of selected liquids are given in Table 2.

Experimental methods

At low frequencies, complex permittivity of the material is generally determined using a capacitive fixture. Capacitance and dissipation factor of the lumped capacitor are measured using a bridge or a resonant circuit [1]. Complex permittivity of the material is then calculated from this data. At microwave frequencies, the sample may be placed inside a transmission line or a reso-



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Butanol	2.95	17.1	0.08	0.33
Chlorobenzene	2.35	5.63	0.04	15.5
Distilled Water	5	78	0	19.7
Ethanol	4.2	24	0	1.24
Ethylene glycol	3	37	0.23	2.0
Methanol	5.7	33.1	0	3.0
Propanol	3.2	19	0	0.54

▲ **Table 2. Dispersion parameters for some liquids at room temperature.**

nant cavity. The resulting characteristics are then measured to compute the dielectric parameters. Since propagation characteristics of electromagnetic waves are influenced by complex permittivity of the medium it propagates through, the material can be characterized by monitoring the reflected and transmitted waves as well. Some of these high frequency techniques are summarized below [1–5].

Resonant cavity method—Measurement techniques employing lumped capacitive sample-holders are useful only up to the lower end of VHF band. At microwave frequencies, a number of techniques have been developed on the basis of distributed networks. Reflection and transmission characteristics of waves inside a transmission line or in free-space are used in this case. Resonant cavities are also used to determine the complex permittivity of materials at discrete frequencies.

Since resonance characteristics depend on the material loaded in a cavity, its quality factor and resonance frequency can be monitored to determine the dielectric parameters. If a cavity can be filled completely with the sample, then its dielectric properties can be determined by the following method:

- Measure the resonant frequency f_1 and the quality factor Q_1 of an empty cavity,
- Repeat the experiment after filling the cavity completely with the sample material.

If the cavity's new resonant frequency is f_2 and the quality factor is Q_2, then the dielectric parameters of the sample are

$$\varepsilon_r = \left(\frac{f_1}{f_2}\right)^2 \tag{9}$$

and

$$\tan\delta = \frac{1}{Q_2} - \frac{1}{Q_1}\sqrt{\frac{f_1}{f_2}} \tag{10}$$

For smaller samples, a cavity perturbation technique may be used. A circular cylindrical sample is placed in the region of maximum electric field inside the rectangular cavity that operates in its TE_{101} mode. Resonant frequency and quality factor of this cavity are measured with and without sample. Complex permittivity of the sample is then calculated as follows:

$$\varepsilon_r' = 1 + \frac{1}{2}\left(\frac{f_1 - f_2}{f_2}\right)\frac{V}{v} \tag{11}$$

and

$$\varepsilon_r'' = \frac{V}{4v}\left(\frac{Q_2 - Q_1}{Q_1 Q_2}\right) \tag{12}$$

where V and v are cavity and sample volumes, respectively.

Similarly, for a small spherical sample of radius r that is placed in a uniform field at the center of the rectangular cavity, the dielectric parameters are found by:

$$\varepsilon_r' = \frac{ab\ell}{8\pi r^3}\left(\frac{f_1 - f_2}{f_2}\right) \tag{13}$$

and

$$\varepsilon_r'' = \frac{ab\ell}{16\pi r^3}\left(\frac{Q_2 - Q_1}{Q_1 Q_2}\right) \tag{14}$$

As illustrated in Figure 1, a, b and l are the width, height, and length of the rectangular cavity, respectively. The frequency shift $(f_1 - f_2)$ must be very small for better accuracy.

Modified infinite sample method—This technique can be used for the liquid or powder samples. In this method, a waveguide termination is completely filled with the sample as shown in Figure 2. Since a tapered termination is embedded inside, it absorbs the incident wave completely. Therefore, it looks as if the sample is extending to infinity. Impedance at its input port depends on electrical properties of the sample filling. Following the slotted-line method of impedance measurement, its VSWR S and location of first minimum d from the load-plane are measured. Complex permittivity of the sample is then calculated by [4]:

$$\varepsilon_r' = 1 - \left(\frac{\beta}{\beta_o}\right)^2 + \frac{\left[\frac{\beta}{\beta_o}\right]^2 \times \left[S^2\sec^4(\beta d) - \left(1 - S^2\right)^2\tan^2(\beta d)\right]}{\left[1 + S^2\tan^2(\beta d)\right]^2} \tag{15}$$

and

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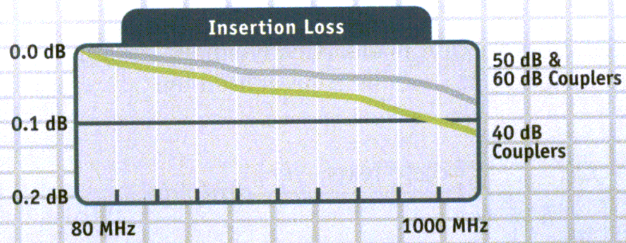

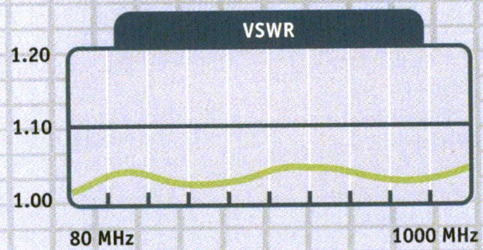

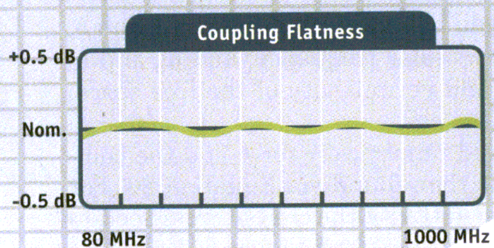

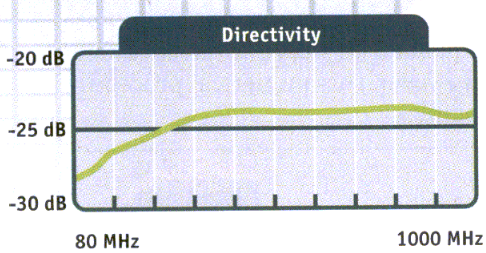

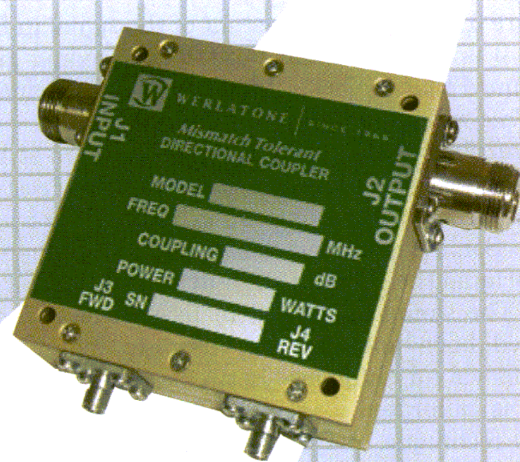

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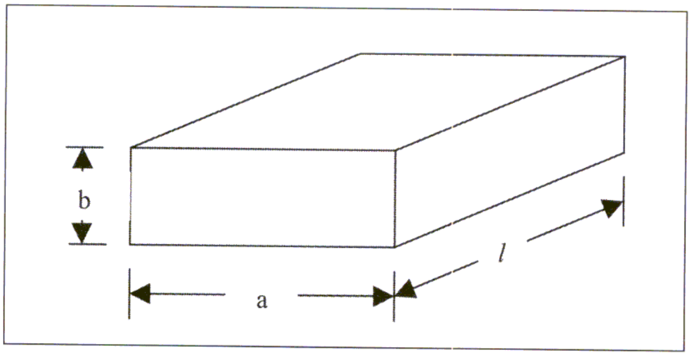
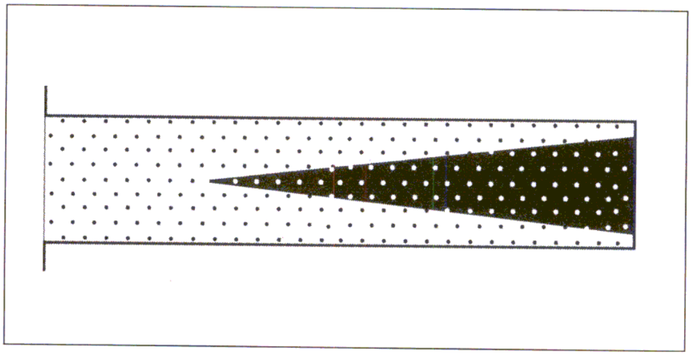
Circle 41



▲ **Figure 1. Geometry of a rectangular cavity.**



▲ **Figure 2. A waveguide termination filled with the sample.**

$$\varepsilon_r'' = \frac{\left[\frac{\beta}{\beta_o}\right]^2 \times \left[2S\left(1-S^2\right)^2 \sec^4(\beta d)\tan(\beta d)\right]}{\left[1+S^2\tan^2(\beta d)\right]^2} \qquad (16)$$

where β_o is the phase constant of the signal in free-space, and β is the phase constant in the feeding guide. It is assumed that the waveguide supports TE_{10} mode only.

Free-space method for the measurement of complex permittivity—Reflection and transmission of an electromagnetic wave at the interface of two dielectric materials depend on the contrast in their dielectric parameters. Some researchers have used these phenomena to determine the complex permittivity of dielectric materials. The sample is placed in free-space and phase-corrected horn antennas are used to monitor different waves via an automatic network analyzer [5]. The system is calibrated using the TRL (through, reflect, and line) technique. A time domain gating may be used to minimize the error due to multiple reflections. The sample of thickness t is placed in front of a conducting plane and its reflection coefficient Γ_1 is measured. Complex permittivity of the material is then determined by:

$$\Gamma_1 = \frac{j\tan\left(k_o t\sqrt{\varepsilon_r^*}\right) - \sqrt{\varepsilon_r^*}}{j\tan\left(k_o t\sqrt{\varepsilon_r^*}\right) + \sqrt{\varepsilon_r^*}} \qquad (17)$$

where k_o is the free-space wave-number of electromagnetic signal.

Open-ended coaxial line method—Sometimes it may not be possible to cut out the sample of a material for the measurement. This is especially important in the case of biological specimens to perform in-vivo measurements because the material characteristics may change otherwise, in which case the following technique may be used:

- An open-ended coaxial line is placed in close contact with the sample, as shown in Figure 3;
- Its input reflection coefficient (or admittance) is measured using an automatic network analyzer [2].

If the coaxial cross-section is electrically small, then its aperture admittance, Y_L, can be represented by:

$$Y_L = \frac{j2k^2}{\omega\mu_o\left[\ln\left(\frac{b}{a}\right)\right]^2}\int_a^b\int_a^b\int_0^\pi \frac{\cos(\phi')\exp(-jkr)}{r}\,d\rho\, d\rho'\, d\phi' \qquad (18)$$

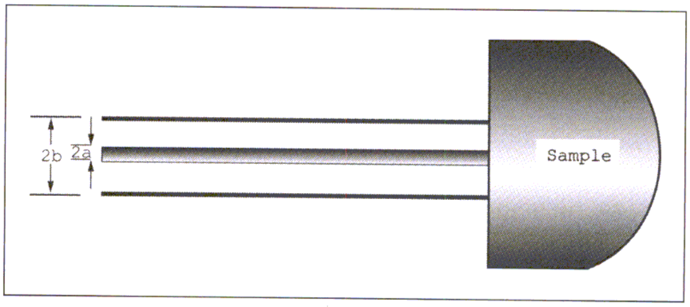
where

$$r = \sqrt{\rho^2 + \rho'^2 - 2\rho\rho'\cos(\phi')} \qquad (19)$$

$$k = \omega\sqrt{\mu_0\varepsilon_0\varepsilon_r^*} \qquad (20)$$

Inner and outer radii of the coaxial line are a and b, respectively; ω is the angular frequency; and μ_o and ε_o are the permeability and permittivity of the free space, respectively.

Equation 18 is solved numerically for k and the complex permittivity ε_r^* is then found via Equation 20. For materials with high permittivity, the coaxial cross-section may not be small enough; therefore, higher order modes of the field over the aperture need to be taken into account. This is accomplished through an electric field integral equation [2]. However, this numerical procedure becomes much more complex.



▲ **Figure 3. Coaxial line geometry with terminating sample.**

Using the manufacturer's recommended technique, the network analyzer is calibrated initially using three standards: an open-circuit, a short-circuit, and a matched load. The reference plane is then moved to the sample end of the coaxial line using a short circuit. Alternately, three standard materials (materials with known ε_r^*) can be used in conjunction with (18) to calibrate the system directly at the sample plane.

Conclusions

An electrical characterization of materials is desired in many scientific and industrial applications. This paper has summarized the characteristic parameters associated with dielectrics along with the properties of a variety of materials. Selected measurement techniques have also been briefly described; these can help the practicing engineers in selecting one for their specialized application. Relevant references, though by no means complete, are included that can provide further details. ■

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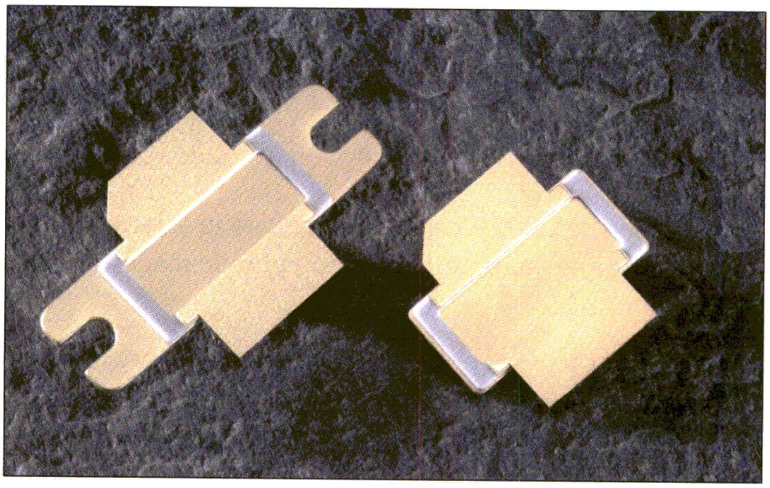
Cu/W Packages Combine High Thermal Conductivity and Low Expansion

Brush-Wellman recently introduced a major development in copper/tungsten (Cu/W) heatsinks with a new functionally graded material (FGM) for power semiconductor packaging. The new Cu/W material achieves the high thermal of copper with lower thermal expansion for higher reliability and improved compatibility with insulating materials and substrates.



▲ **Copper/Tungsten heatsink material offers high thermal performance with low thermal expansion.**

Using the new Cu/W metal matrix composites, the company offers FGM heatsinks for microelectronic packaging and as mounts or submounts for semiconductor laser die. The new FGM flanges consist of a high thermal conductivity core surrounded and constrained by a lower thermal conductivity, lower thermal expansion body. The combination permits the designer to tailor thermal conductivity and expansion separately for maximum heat transfer with lower mechanical stress on the high power die.

Depending on the grade of Cu/W material used as the surrounding body, an effective thermal conductivity for the die area of 320 W/mK is obtained with a coefficient of thermal expansion of 5.5 to 7.3 ppm/°C. Independent testing has shown a reduction of 18.6 percent in thermal resistance compared to a standard 150 W/mK flange. This value is close to the maximum attainable 23.1 percent reduction in thermal resistance obtained when using an OFHC copper flange.

An example of packaging available using the new material is the RF705 package. Aimed at high power dissipation requirements of 2-1/2G and 3G systems, the package is suitable for devices using Si, SiC, GaAs, GaN and other semiconductor materials.

As pictured above, the RF705 provides power dissipation up to 150 watts or more in both flanged and flangeless versions. ■

For more information, contact:

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Radio Transceiver Supports Broadband Wireless at 23.5 to 43.5 GHz

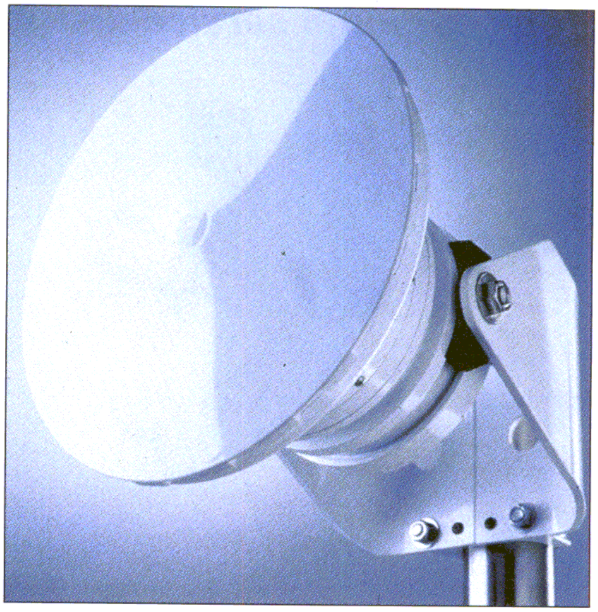
Galleon Wireless Broadband announces its Outdoor Unit (ODU), a high performance radio transceiver for Broadband Wireless Access (BWA). The ODU offers a new price/performance benchmark for systems operating at 24 GHz and higher. Formerly operating under the name BELSTAR, Galleon is developing a complete suite of BWA network products. The ODU is the result of development work and field trials with Galleon's principal technology subcontractor, Northrop Grumman Corp.

BWA provides telecommunications network operators a wireless method for providing customers with two-way voice, data and video transmission services at a 155 Mbps data rate, faster than DSL (50 Mbps) or cable modem (6 Mbps) services.

The unit's price/performance value is obtained through the use of MMIC technology, which dramatically lowers the number of components as well as associated manufacturing and maintenance costs. The Galleon ODU offers up to 2 GHz of bandwidth with a 16-gigabit bandwidth yield, assuming 64 QAM. Standard interfaces (DAVIC, DOCSIS, UHF and others) permit easy integration into a network.

Chipset and filter modules are available to cover RF bands ranging from 23.5 GHz to 43.5 GHz. This design flexibility permits network operators to adapt their equipment to changing customer needs.

The ODU maintains communication reliability and signal quality through advanced voltage-controlled oscillator (VCO) technology. The VCO delivers high stability and very low phase noise, and can be enhanced with optional circuitry to further reduce phase noise for critical high performance applications.



▲ **A new outdoor unit for Broadband Wireless Access has been introduced by Galleon Wireless Broadband.**

By the second quarter of 2001, Galleon expects to offer the ODU outdoor unit at a price of $1,000 each.

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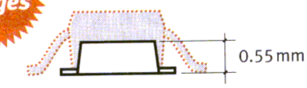

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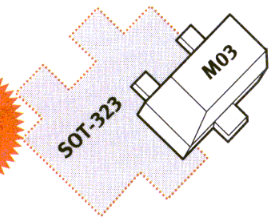

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NE662M04	23 GHz f_T LNA	1.1 dB	20 dB	2 GHz	M04

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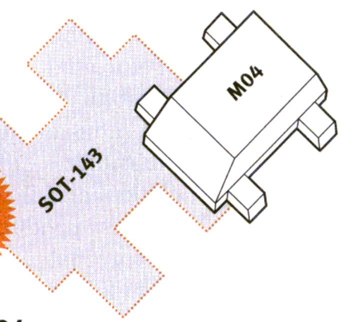

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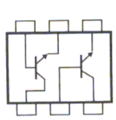

Part Number	Description	Q1 Spec	Q2 Spec
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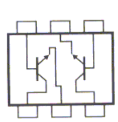
UPA814TC	Matched Die/Cascade LNA	NE688	NE688



Part Number	Description	Q1 Spec	Q2 Spec
UPA826TC	Matched Die/Osc-Buffer Amp	NE685	NE685
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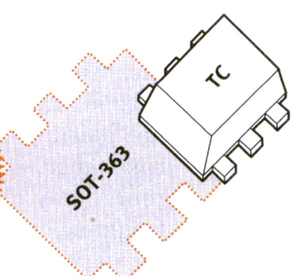

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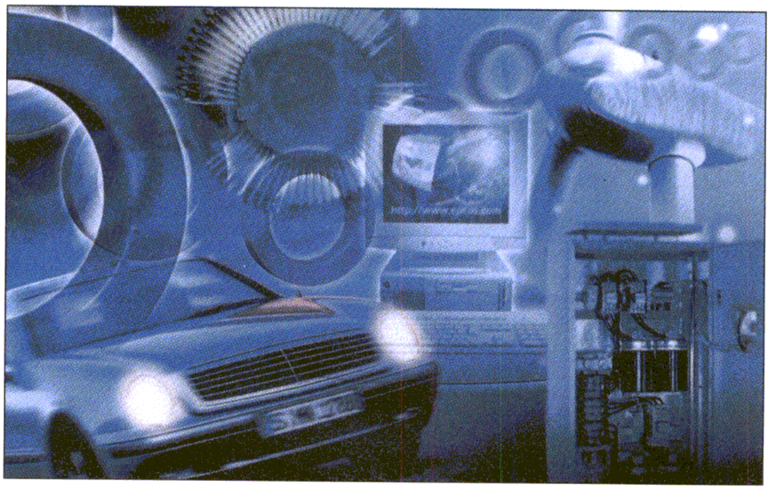

▲ **EPCOS offers a wide range of ferrite toroids for LAN and EMI reduction applications.**

New products include a series of toroidal cores with diameters from 2.5 to 4 mm. Also, T55 material that can handle high bias current and T38 material with μ_i=10,000 have been added to a broad line of cores using MnZn and NiZn materials.

MnZn materials have high permeability and sufficiently high Curie temperature for many applications, but NiZn materials are gaining popularity. They have high ohmic resistance, minimizing the effects of eddy current. They also have high impedance values at frequencies beyond 100 MHz. The limiting factor to these materials is that Curie temperature rises with permeability (μ_i).

For LAN signal transformers, the T55 material offers up to 8 mA DC current, which can result from asymmetry in multi-level coded data signals. T55 has an initial permeability of 4000 and a high saturation flux density of 470 mT at 25°C. When combined with EPCOS K10 material in the common-mode choke, LAN coupling needs are fully addressed.

EMI applications require suppression of primarily common-mode radiation or reception of interference. To meet safety requirements (staying within leakage current limits), chokes with high asymmetrically effective inductance material must be used. Current-compensated chokes are well-suited for this application, using equal windings that cancel magnetic flux in the core.

EPCOS materials meeting these requirements are available with nine values of initial permeability (μ_i), from 700 (K10 material) to 20,000 (T56 material). Toroid sizes range from 2.5 mm diameter to 202 mm. ■

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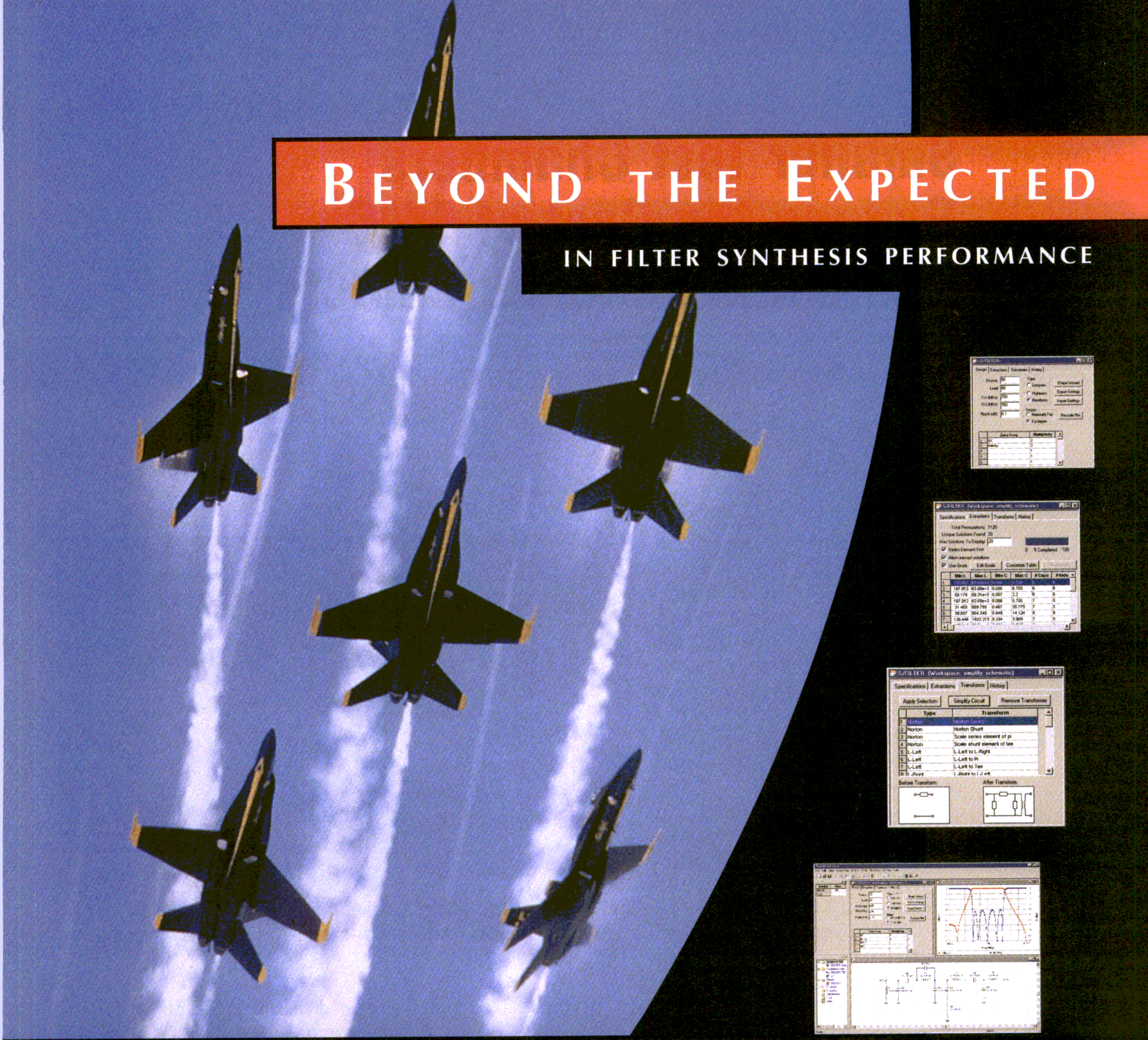

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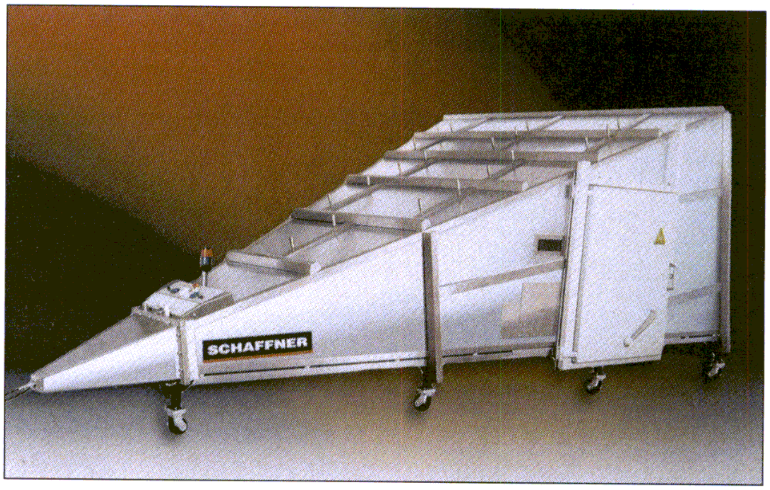

▲ **Schaffner EMC offers smaller, lower cost models in its line of GTEM test cells, enabling low-cost EMC compliance testing.**

Conventional chambers can cost as much as $400,000 to several millions of dollars, while Commercial test laboratory rental charges are $1,500 to $3,000 per day. GTEM installations are much lower in cost, ranging from $30,000 to $150,000.

Schaffner's GTEM systems can perform radiated (emissions) and immunity testing in a single, fixed, shielded environment that covers test frequencies from DC to 18 GHz. The test cells exhibit excellent field uniformity and reproducibility over a given test volume. They are well suited for field calibration applications.

The company's line includes both standard versions and a no-frills GTEM Lite version. The standard model includes a list of features that are usually options on other systems, such as mains filtering, lighting and safety interlocks. The GTEM Lite has a narrower test frequency range (DC to 5 GHz), but is also less costly, slightly smaller in size and requires less cabling than an equivalent standard system. Both GTEM cell types offer low setup and operating costs compared to conventional chambers and open sites. Industry standard software (IEEE 488 bus) is provided that can run fully automated testing. For example, the need to move the object in the x-y-z axes can be eliminated, significantly reducing test time.

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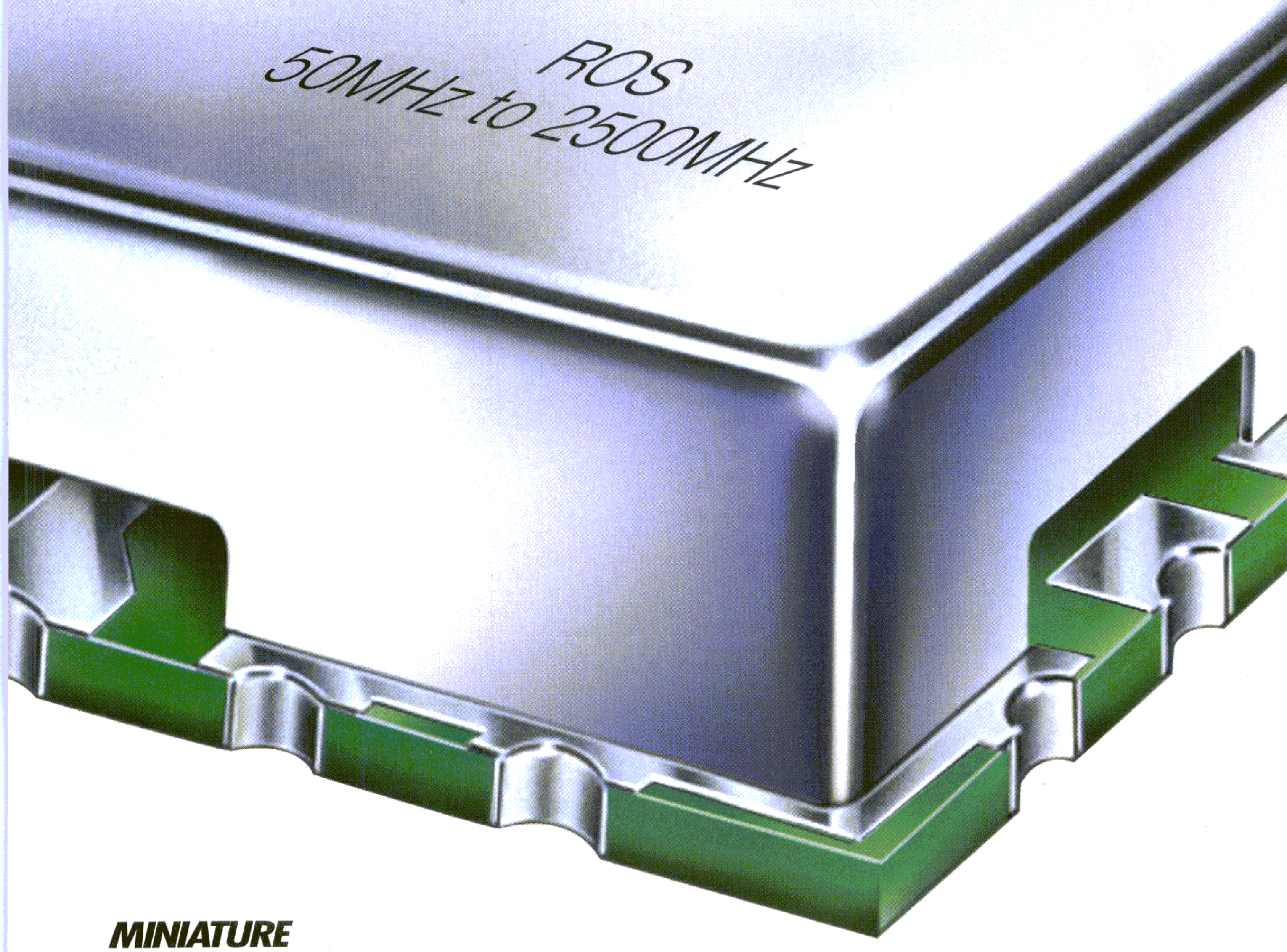

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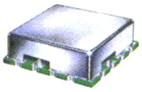
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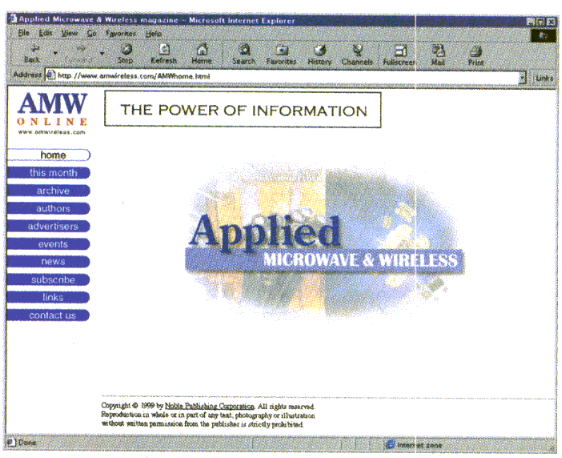

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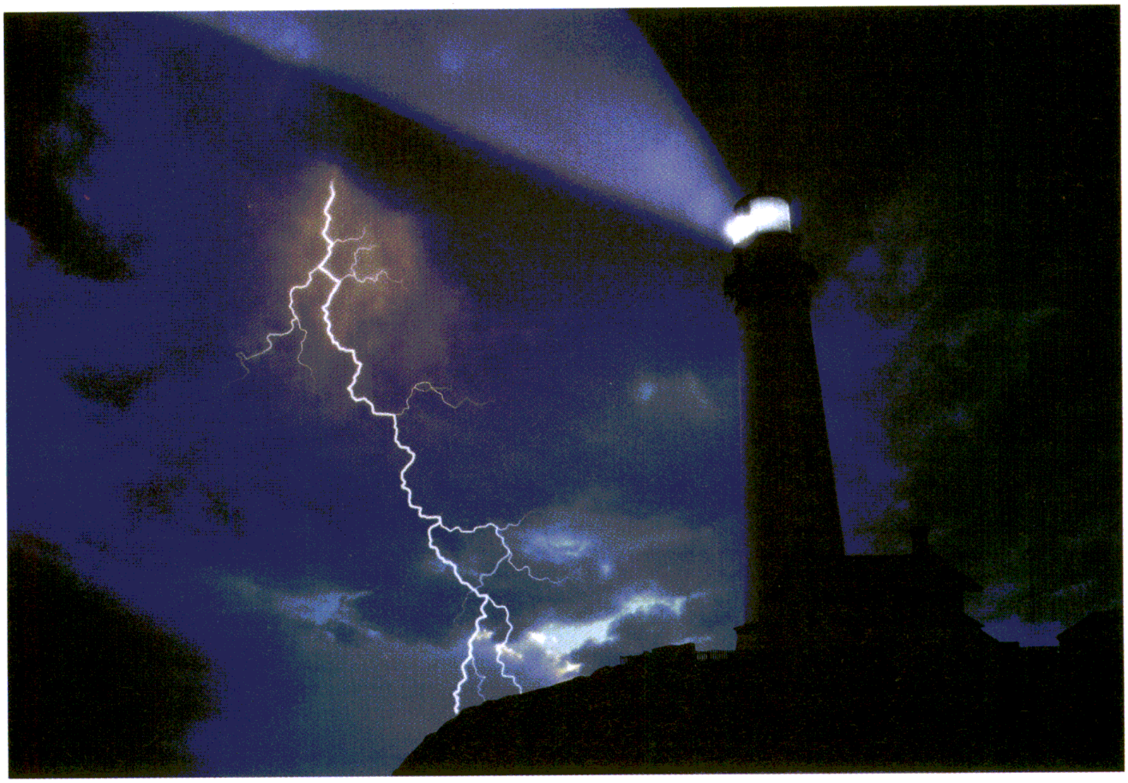

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	SGA-6286 SGA-6289	SGA-6386 SGA-6389	SGA-6486 SGA-6489
Frequency (GHz)	DC-3.5	DC -3.0	DC-1.8
Gain (dB)	13.8	15.4	19.7
TOIP (dBm)	34.0	36.0	34.0
P1dB (dBm)	20.0	20.0	20.0
N.F. (dB)	3.9	3.8	2.9
Supply Voltage (Vdc)	4.2	5.0	5.2
Supply Current (mA)	75	80	75

All data measured at 1GHz and is typical. MTTF @ 150C T_j = 1 million hrs. (R_{TH} = 97C/W typ)

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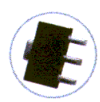

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alloys (Al-Cu-Si, for instance) can be made very resistant to electro-migration and are considered useful for high-power RF devices. These two systems can be made functionally acceptable for electro-migration characteristics.

Where gold shines (pun intended) as most superior is in the non-traditional reliability considerations for transistors—*beyond* electro-migration. Consider the table of characteristics at right.

Gold	Aluminum
Superior electrical resistivity 2.1 E-6 Ohm-cm	Electrical resistivity 2.6 E-6 Ohm-cm
Superior thermal conductivity 300 W/mK	Thermal conductivity 220 W/mK
Chemically relatively inert	Chemically active—oxidizes and corrodes easily
Melting point ~ 1,063 C	Melting point ~ 660 C
High Atomic weight—197	Low Atomic weight—27
Excellent step coverage	Sometimes poor step coverage and etch uniformity problems
Excellent flexural fatigue resistance—especially important in wires	Poor flexural fatigue resistance—especially important in wires
Superior CTE—14 ppm/C	CTE—24 ppm/C
All gold metal system possible—package, wires and chip metal—no intermetallics	Gold, aluminum and copper metals—package, wires and chip metal are different (potential intermetallics like "purple plague")

▲ Table 1. Non-traditional reliability considerations for the use of gold vs. aluminum in transistors.

A good example of a reliability problem beyond electro-migration occurred in the mid 1970s. The PAVE PAWS program (Precision Acquisition of Vehicle Entry-Phased Array Warning System) demanded high-peak RF power pulsed transistors. The company I worked for built a transistor (that I designed) using an excellent aluminum-copper-silicon metal system for devices that performed admirably—for a while. Unfortunately, after 1 million to 2 million pulses, the aluminum bond wires began to break and open up. Performance degraded slowly, then dramatically. The failures were caused by aluminum bond wires being temperature cycled by high currents, which made them expand, flex slightly and begin to work-harden. Eventually, the aluminum wires would break, and the transistor performance would degrade. The problem took months to discover in systems depending on maximum operating conditions. In this case, we switched to a gold metal system and gold wires and completely resolved the failure mechanism. But it was too late for the application, and many millions of dollars were lost.

Aluminum susceptibility to environmental acids, bases, salts and oxidation is also well known. Corrosion can readily damage any RF power transistor, especially under voltage bias conditions. Gold is much more inert to chemical assault.

Intermetallics (the old "purple plague" failure mechanisms) are a potential problem for aluminum-based systems because packages are gold metallized. Package gold plating reduces oxidation, corrosion and soldering wetting problems and is recognized universally as ideal for electrical contacting. A gold transistor metal system uses only one metal (gold) for contact between package, wires and chips (dice). Therefore, no intermetallics such as purple plague can occur. An all-gold system is inherently more reliable.

Finally, the methods for applying gold to the chip surface tend to be very conformal. As a consequence, gold processing-induced uniformity problems are minimal when compared to the aluminum etch and step-coverage difficulties often encountered. Consistency of final product minimum cross-section dimensions are therefore more easily achieved with gold. The result can be that a supposed aluminum cost advantage is negated or even tilted toward a yield increase with gold metallizations.

With this whole picture in mind, I would ask the reader to go back and study the physical property table one more time, then tell me: should the need ever arise, which metal system you would prefer in your pacemaker? ■

The Gold vs. Aluminum War Revisited

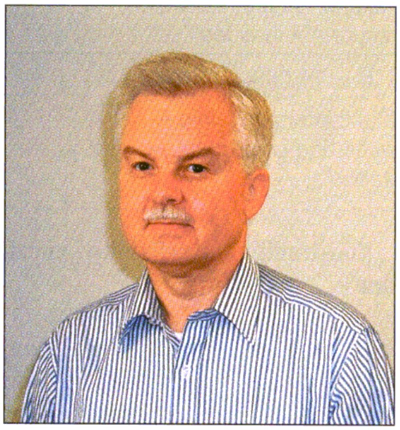
By Howard Bartlow
HdB Engineering

Isn't controversy an amazing thing? Claims and counter-claims are lobbed back and forth. Technical facts and opinions spin incessantly, though not inerrantly — especially when sales and marketing (read money) are involved. Well, the gold versus aluminum debate has resurfaced after 25 years, and discovering the truth seems nearly as difficult now as it was then.

In the 1970s, RF power transistor manufacturers fought each other tooth-and-nail for years to gain a customer acceptance advantage for their chosen transistor metallization systems. Remember the advertisements stating that aluminum is good for "ladders and lawn chairs" but *not* for RF power transistors? This was the catchy phrase trumpeted by PHI in one of its ad campaigns. I thought the campaign was rather creative, especially since I worked for a company on the aluminum alloy side of the fence. Electro-migration was *the* issue at the time, and 1 GHz to 3 GHz high-power aluminum metallized transistors could degrade in a matter of weeks or months. Evidence of electro-migration of aluminum and aluminum-silicon conductors was real and available and not pretty.

Remember "hillocks," "whiskers" and "void growth," "step coverage thinning," etched "mouse holes," metal grain size, passivation effects and Jim Black's electro-migration equation? The theory and practice of improving electro-migration test results were quite well researched and understood. Reducing current densities by increasing cross-sectional areas (thickening metal layers), as well as employing aluminum-copper-silicon alloys, allowed a manufacturer to build acceptable aluminum RF power transistors that had electro-migration lifetimes of at least 50 to 150 years. We built hundreds of thousands, maybe even millions, of transistors with that system.

Nonetheless, the damage to aluminum (and aluminum-based alloys) had been done. Gold was superior in direct "apples-to-apples" electro-migration comparisons, so every aluminum-based system began losing out in the marketplace. Nearly all RF power transistor manufacturers eventually switched to gold-based metallization systems.



Howard Bartlow has been designing, fab processing, assembling, testing and trouble-shooting RF power transistors for more than 30 years. He has served in engineering and management positions at Monsanto Research, TRW Semiconductor, Communications Transistor Corp., Acrian, Inc., Microwave Modules and Devices and Spectrian Inc. He has a bachelor's degree in Engineering Physics. He serves as president of HdB Engineering, working on BJT, LDMOS, GaAs MESFET, SiC MESFET, RF power module and packaging technologies for both commercial wireless and military applications. He holds 11 patents plus two pending, and lives in a suburb of Portland, OR. The RF power industry continues to challenge him today with the development of newer materials and technologies that demand creative engineering and problem solving.

So why has the debate re-surfaced now? Why not keep using the overall superior gold-based metallization systems? In a word — LDMOS. New, power LDMOS transistors look a lot like some transistors built in CMOS fabs for integrated circuits. The thousands of lessons learned in process controls, techniques and engineering developments for CMOS integrated circuits over the years can apply directly to LDMOS devices. And LDMOS power transistors work particularly well for many linear power amplifier applications.

However, CMOS fabs are highly reluctant to use any gold processing for fear of cross-contamination of gold atoms into CMOS wafers. The fear is great enough that no gold is even allowed near many CMOS wafer fabrication areas. Since CMOS integrated circuits easily enjoy a dominant business level versus RF, the CMOS line dictates what fab processes can be used. Unless the fab is completely separate, gold metallized LDMOS metallizations sometimes are simply ruled out. Thus, for some companies, the need arose to make aluminum-based metallizations work in RF power again. We now hear about "gold-free" metallizations and hot metal processes as if gold metallizations were somehow "bad" or outmoded. Not so.

Compared head-to-head with equal stripes, currents, temperatures and conditions, gold will always be superior to aluminum and its alloys. Physical properties still matter. At the same time, I will say that some aluminum

continued on page 115

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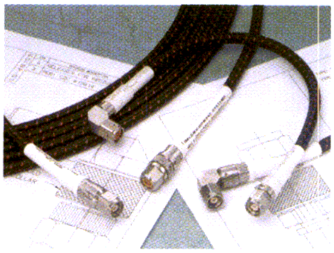

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