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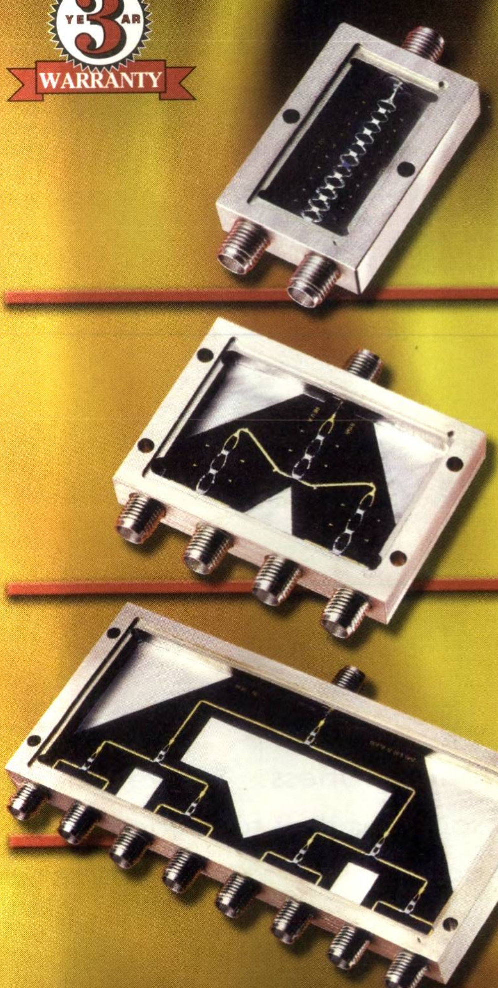

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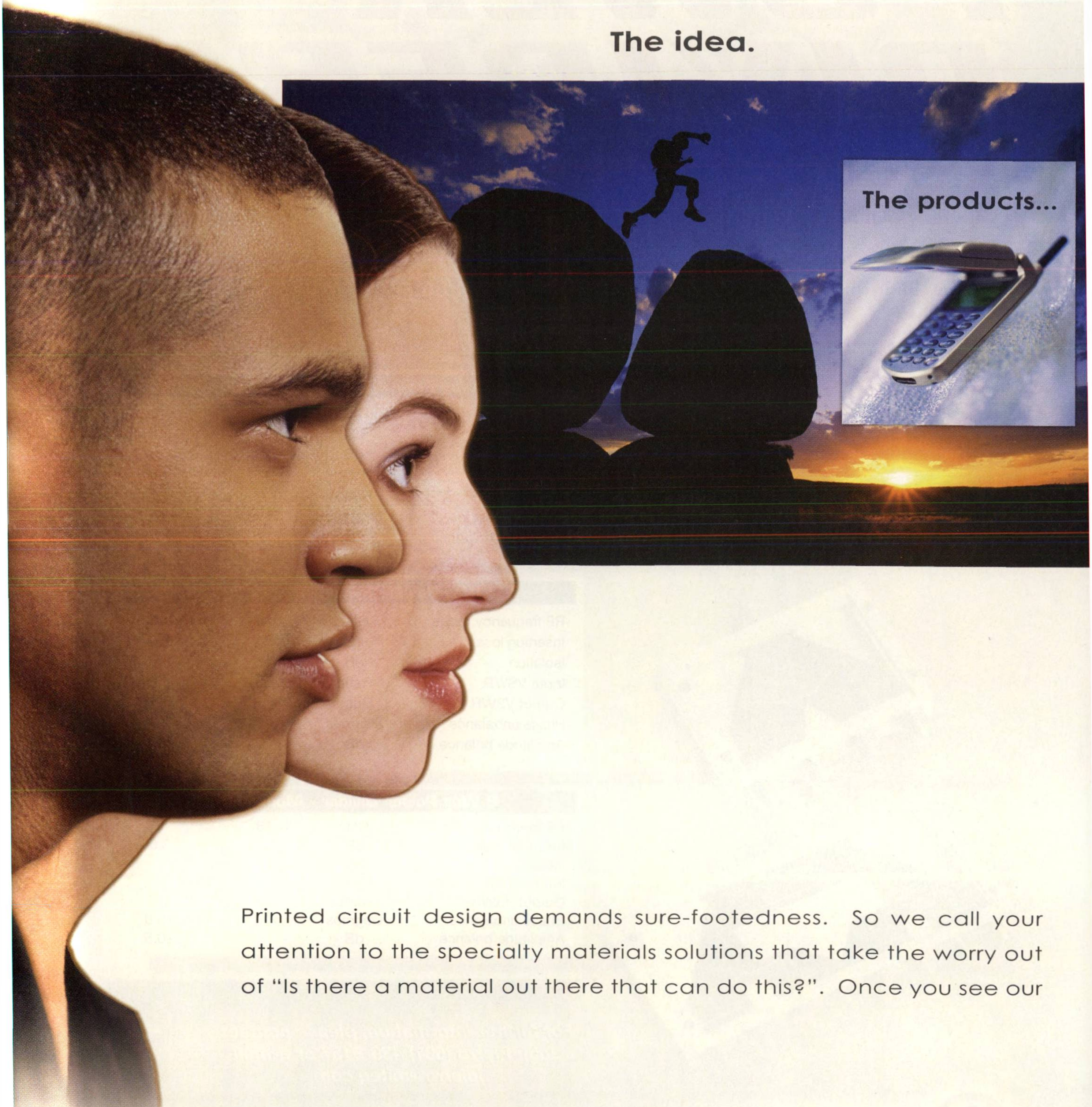

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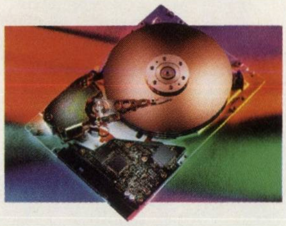

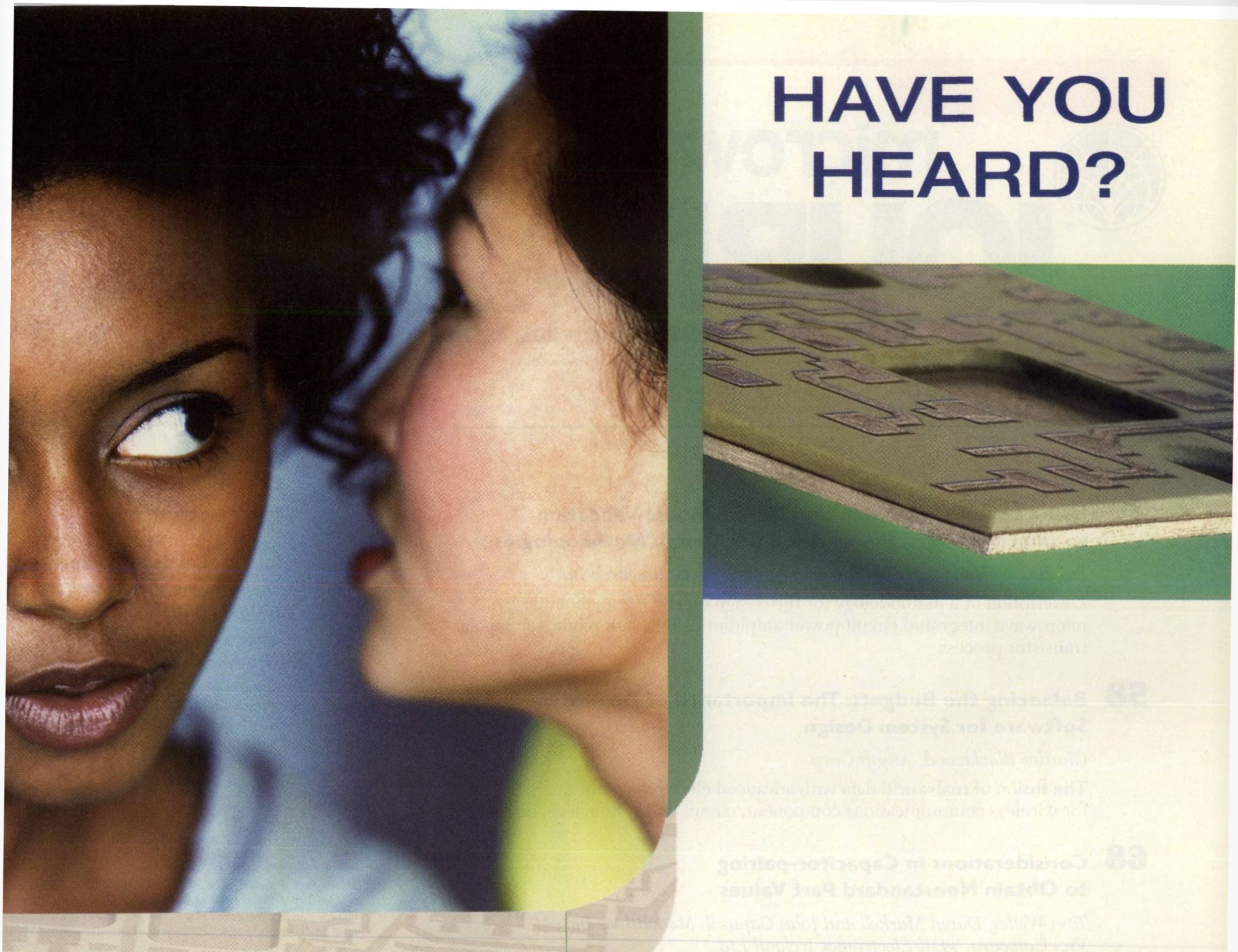

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ON THE COVER

A new family of active frequency multipliers and dividers is featured on this month's cover

Cover art courtesy of Hittite Microwave Corp.

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60th Microwave Measurement Conference
December 5–6, 2002
Washington, DC

Sponsor: Automatic RF Techniques Group (ARFTG). This conference will explore cutting edge microwave measurements for the industrial, academic, and government-related engineering communities. Original papers demonstrating approaches to the characterization of microwave devices, circuits and systems will be presented. Topics: test challenges for high power on-wafer devices, measurements for emerging technologies, nonlinear measurement, new developments in measurement instrumentation and other areas of automated RF measurements. The conference will also feature an exhibition, for information contact: Leonard Hayden, exhibits chair, Cascade Microtech (503) 601-1580, fax (503) 601-1601 or e-mail: leonard@cmicro.com. Contact: J. Gregory Burns, conference chair, Northrop Grumman (410) 765-7331, fax (410) 981-4874 or e-mail: john_g_burns@mail.northgrum.com.

IEEE International Electron Devices Meeting (IEDM)
December 8–11, 2002
San Francisco, CA

IEDM is a forum for reporting breakthrough in the technology, design and manufacturing of semiconductors and other electron devices. Additional information is available on the meeting's Web site at www.ieee.org/conference/iedm, or contact: Phyllis Mahoney, conference manager, Widerkehr & Associates, 16220 S. Frederick Ave., Suite 312, Gaithersburg, MD 20877 (301) 527-0900 or e-mail: phyllism@widerkehr.com.

Wireless Communications Association Technical Symposium
January 13–15, 2003
San Jose, CA

This ninth annual event will convene more than 1000 broadband wireless experts from 20 nations, collocated for the first time with the plenary meeting of the IEEE 802.16 Working Group on BWA. For more information, visit www.wcai.com, or contact: Colleen O'Reilly at colleen@wcai.com.

7th International Commercialization of Military and Space Electronics Conference and Exhibition (CMSE)
February 10–13, 2003
Los Angeles, CA

This conference is organized by Components Technology Institute Inc. in cooperation with EIA/ECA, IEEE/CPMT and IMAPS. It will specialize in COTS systems, subsystems, circuit boards and components while also emphasizing new technology, processes and design practices. Emphasis will be placed on practical solutions, new techniques and how to assess the risks of COTS and make cost-effective decisions that meet the mission requirements. Working groups and discussion sessions are planned on specific topics of concern to the industry. New issues identified by the delegates to the previous year's conference will also be addressed. Topics include: design practices, technology trends, applications, case studies/history, obsolescence management, radiation hardness, risk mitigation, selecting COTS and commercial suppliers, testing requirements and results, and constructive and destructive physical analysis. For further information, contact: Dale Stamps or Leon Hamiter at (256) 536-1304 or e-mail: dale@cti-us.com or lhamiter@cti.us.com. Information can also be accessed online at www.cti.us.com.

International Conference on Subsurface and Surface Sensing and Imaging IV
March 2–6, 2003
San Diego, CA

This conference reports advances and progress in the research and development of subsurface and surface sensing and imaging techniques,

sensors and applications, and addresses the technical barriers encountered in multiple domains of subsurface and surface sensing and imaging. Topics include: surface and ground penetrating sensors, signal and data processing and propagation and modeling, material properties and characterizations, and cross-cutting commonalities across subsurface and surface sensing applications. Additional information is available at http://ee.tamu.edu/subsurface-sensing-conference. For further information, contact: Cam Nguyen, Department of Electrical Engineering, Texas A&M University, College Station, TX 77843 (979) 845-6259 or e-mail: cam@ee.tamu.edu.

International Wireless Communications Expo (IWCE)
March 9–15, 2003
Las Vegas, NV

This show plays host to more than 350 exhibiting companies and 10,000 attendees in the two-way mobile communications industry. Along with base station workshops, the IWCE conference program features the latest developments in business, regulatory, public safety and homeland security. For more information, visit www.iwceexpo.com, or e-mail: trade_shows@primediabusiness.com.

IEEE Sarnoff Symposium
March 12, 2003
Trenton, NJ

This symposium brings together professionals and industry experts to exchange information on the latest developments in the fields. The conference includes an exhibition of components, technologies, systems and services, and also features tutorials. Topics: broadband wireless systems, network security, satellite communications, signal processing for communications, microwave device technology, modeling and simulations, optical networking, ultrawideband systems, VoIP and QoS, military communications, 3G mobile systems and wireless LANs, smart antennas and phase arrays, microwave photonics, and software radio. For additional program and registration information, visit http://ewh.ieee.org/r1/princeton-centraljersey/Sarnoff_Symposium.htm or contact: Peter Zalud, symposium chair, Sarnoff Corp. at pzalud@sarnoff.com.

IEEE International Symposium on Electromagnetic Compatibility
May 11–16, 2003
Tel-Aviv, Israel

This symposium will provide opportunities for EMC researchers, scientists, engineers and vendors to present the latest research results, discuss problems of current and mutual interest and exchange views and experience related to new EMC components, materials and equipment. For further information, contact: ORTRA Ltd., 1 Nirim Street, PO Box 9352, 61092 Tel-Aviv, Israel +972-3-63844, fax +972-3-6384455 or e-mail: emc2003@ortra.co.il or visit www.ortra.com/emc2003.

IEEE MTT-S International Microwave Symposium and Exhibition
June 8–13, 2003
Philadelphia, PA

This symposium will serve as the centerpiece of Microwave Week 2003. Topics: research, development and application of RF and microwave theory and techniques. In addition to IMS2003, a microwave exhibition, a historical exhibit, the RFIC symposium and the ARFTG conference will be held during Microwave Week 2003. The technical sessions will run Tuesday through Thursday of the Microwave Week. Workshops will be held Sunday through Tuesday, and the ARFTG Microwave Measurements Conference will be held on Thursday and Friday. For further information, contact: Richard V. Snyder, general chair, RS Microwave Co. Inc., (973) 492-1207, e-mail: r.snyder@ieee.org. For exhibition information, contact: Kristen Dednah, Horizon House Publications, 685 Canton St., Norwood, MA 02062 (781) 769-9750 or e-mail: kdednah@mwjournal.com.

WORKSHOPS & COURSES

NIST/ARFTG Microwave Measurements Short Course

■ **Topics:** Microwave measurement fundamentals, practical issues such as cables, fixtures, probes and on-wafer measurements, as well as more advanced measurement topics and an overview of wireless test instrumentation and measurement.
■ **Site:** Washington, DC
■ **Dates:** December 3–4, 2002
■ **Contact:** Dave Walker, NIST, e-mail: dwalker@boulder.nist.gov.

Advanced Wireless and Microwave Techniques

■ **Topics:** Antennas and filters are covered briefly, followed by a detailed discussion of figures of merit. Mixers and oscillator designs are evaluated, as well as defining, classifying and improving the efficiency and linearity of power amplifiers.
■ **Site:** Sunnyvale, CA
■ **Dates:** December 9–13, 2002
■ **Contact:** Besser Associates, 201 San Antonio Circle, Building E, Suite 280, Mountain View, CA 94040 (650) 949-3300.

RF Wireless System Design Fundamentals

■ **Topics:** Digital wireless communication systems concepts and performance limitations, system degradation due to RF components, wireless communication system budget profiles, propagation losses and link budgets, and cost vs. performance issues.
■ **Site:** Los Angeles, CA
■ **Dates:** December 11–13, 2002
■ **Contact:** Besser Associates, 201 San Antonio Circle, Building E, Suite 280, Mountain View, CA 94040 (650) 949-3300.

RF Wireless Engineering

■ **Topics:** Designed to give electrical engineers the specialized training they need to achieve competence in RF and wireless engineering. Students learn practical skills, such as component selection and impedance matching network design.
■ **Site:** Atlanta, GA
■ **Dates:** January 27–31, 2003
■ **Contact:** Georgia Institue of Technology, 613 Cherry Street, Swann Bldg., 3rd Floor, Atlanta, GA 30332 or visit: www.gtconted.org.

RF Wireless System Design Fundamentals

■ **Topics:** Combines theory with real-life examples to provide a complete foundation in digital communication techniques and their effects on RF circuit parameters.
■ **Site:** Phoenix, AZ
■ **Dates:** March 10–12, 2003
■ **Contact:** Besser Associates, 201 San Antonio Circle, Building E, Suite 280, Mountain View, CA 94040 (650) 949-3300.

Phased Array Antennas for Communications and Radar

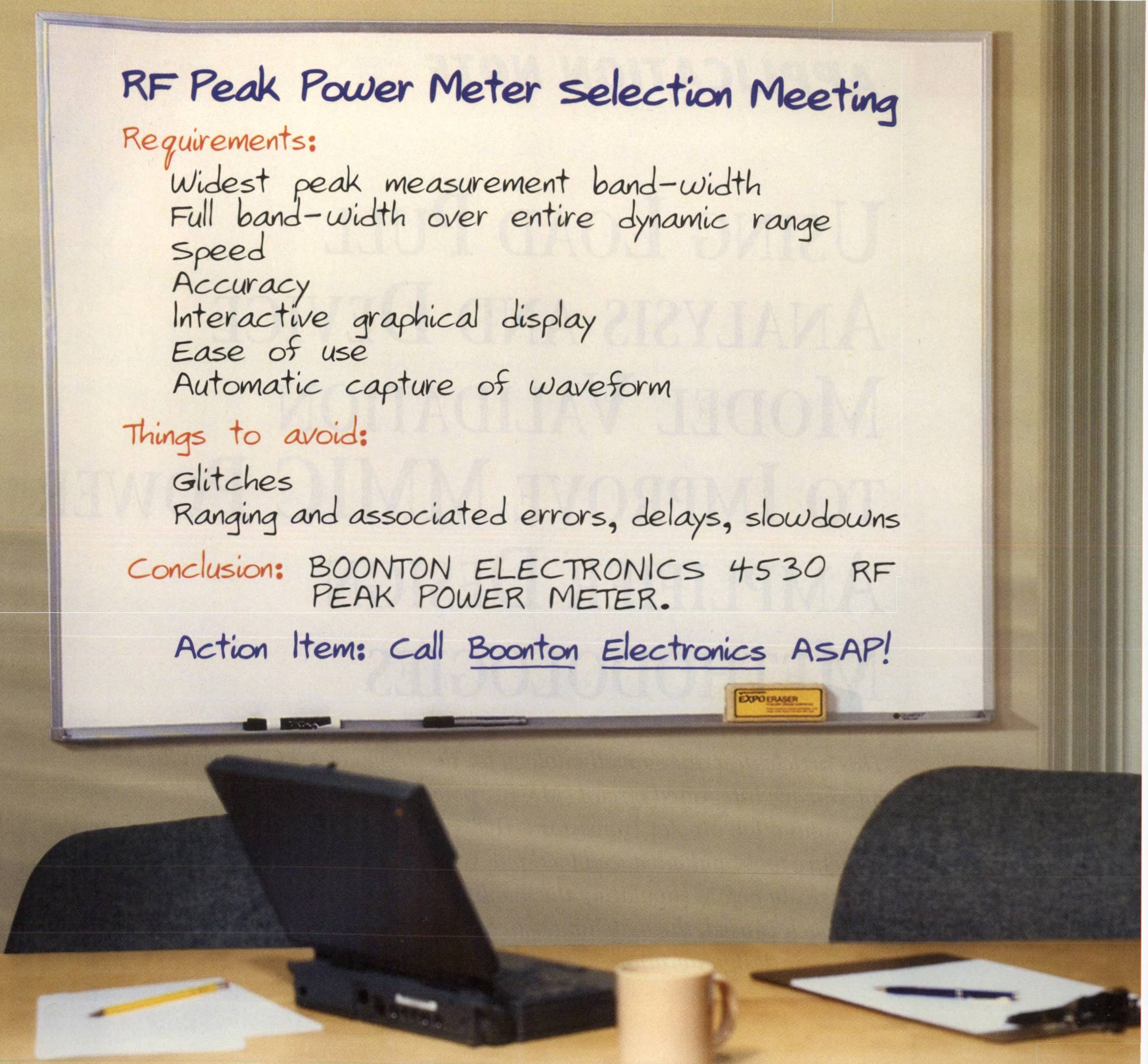
■ **Topics:** The necessary theory, basic principles of operation components, and important design parameters of phased array antennas. The gain, efficiency, polarization, pattern, bandwidth and input impedance characteristics are presented and matched to communication and radar system requirements.
■ **Site:** Davos, Switzerland
■ **Dates:** March 17–21, 2003
■ **Contact:** CEI-Europe AB, PO Box 910, S-612 25 Finspong, Sweden +46-122-175 70, fax +46-122-143 47.

Antenna Engineering

■ **Topics:** Theory and practice of antenna engineering covering the range of antenna properties and types from basic to state-of-the-art. The antennas presented in the course cover a wide spectrum of frequency.
■ **Site:** Atlanta, GA
■ **Dates:** April 28-May 2, 2003
■ **Contact:** Georgia Institute of Technology, Continuing Education, PO Box 93686, Atlanta, GA 30377 (404) 385-3501.

RF/Microwave Circuit Design: Linear/Nonlinear Theory and Applications

■ **Topics:** Enhance the design capability of the engineer by introducing modern linear and nonlinear design techniques. Seeks to combine the theory and practice of modern, computer-aided high frequency circuit design with greater intuition and increased insight.
■ **Site:** Cambridge, UK
■ **Dates:** May 12–16, 2003
■ **Contact:** CEI-Europe AB, PO Box 910, S-612 25 Finspong, Sweden +46-122-175 70, fax +46-122-143 47.


RF Peak Power Meter Selection Meeting
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Full band-width over entire dynamic range
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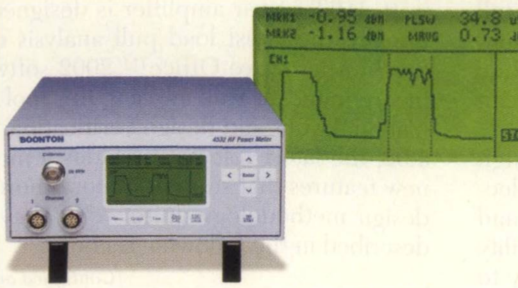
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APPLICATION NOTE

Using Load Pull Analysis and Device Model Validation to Improve MMIC Power Amplifier Design Methodologies

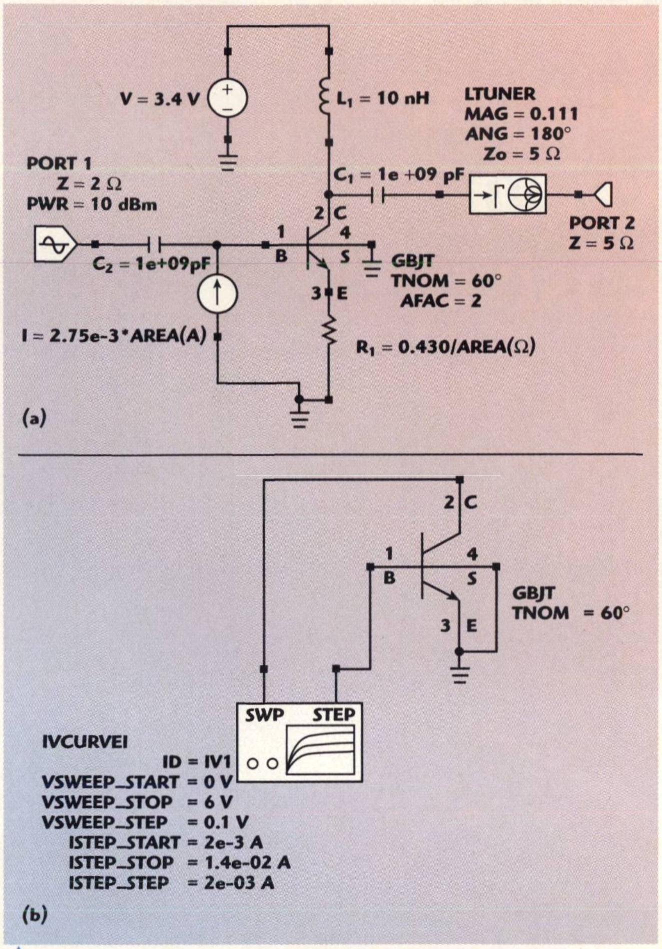
This article describes a methodology for the design of a 2 GHz monolithic microwave integrated circuit (MMIC) power amplifier (PA) using a heterojunction bipolar transistor (HBT) process. It is based on the use of modern software tools and advanced load pull simulations to validate the device model extensively before simulating the amplifier circuit as a whole. When this approach is used, the matching networks and final design require virtually no tuning or design iterations.

There is certainly no shortage of information about designing power amplifiers these days, and several books address the subject quite admirably.[1-6] However, as well as these books cover the theory and underlying logic of such designs, they do not tell the reader precisely how to do it. As a result, most designers develop their own ad-hoc methodologies for such designs, which can vary significantly in effectiveness.

This article describes a thorough design methodology that leverages advances in electronic design automation (EDA) and test and measurement data to improve the probability of first pass success. It is not the only way to design a power amplifier, but experience has proven it to be quite effective. It is based on the use of simple analyses and measurements to obtain all the necessary information before the design commences; the design then becomes a relatively simple process of creating a few simple building blocks, and connecting them together. The most important part of the design is a methodical evaluation of the power device. As an example, a single-stage, half-watt, HBT power amplifier is designed. Additionally, the latest load pull analysis capabilities in Microwave Office™ 2002 software are incorporated, as well as using this tool for simulation, layout and design verification. Version 5.02, the latest release, provides a number of new features that support a more thorough PA design methodology quite well. They will be described in the following sections.

[Continued on page 22]

Dr. Stephen A. Maas
and Ted Miracco
Applied Wave Research Inc.
El Segundo, CA



▲ *Fig. 1 The RF evaluation circuit (a) and DC evaluation circuit (b).*

THE AMPLIFIER

The goal is to design a single-stage heterojunction bipolar transistor (HBT) amplifier integrated circuit. The amplifier must cover 1.6 to 2.1 GHz, have at least 22 dB gain at full output and operate at a power-supply voltage of 3.4 V. To save chip area and minimize output loss, the output matching circuit is off-chip. An HBT process from Global Communication Semiconductors Inc. (GCS) will be used. The GCS foundry offers indium gallium phosphide (InGaP) HBT technology, having an f_{max} of approximately 50 GHz. The DC bias regulator will also be off-chip, using a CMOS IC. Thus, the output matching network and DC bias need not be part of the MMIC design.

The design of the amplifier starts at the output and proceeds toward the input. The process is as follows: evaluate the device using nonlinear analysis and load pull simulations; determine the device size, bias point and optimum load; determine the device input impedance; synthesize an input matching network; connect the input network to the device and make sure the combination works properly; and design the output matching network. Each step is addressed in that order.

Evaluate the Device

Before beginning the design, it is essential to perform a "sanity check" on the device models that the foundry typically provides. Many foundries use the SPICE Gummel-Poon (G-P) model to characterize their devices due to the fact that the model is uniformly supported across all circuit simulators. G-P is an older model and has many limitations for HBT design. More advanced models, such as the UCSD HBT model[7] and Anholt HBT model, are implemented in Microwave Office 2002 design software, along with the MEXTRAM and VBIC models. Nevertheless, such advanced models are not uniformly supported, so many foundries only extract G-P models. Thus, users will probably be using G-P for HBT modeling well into the foreseeable future.

It is particularly dangerous to design RF and microwave circuits without understanding the models used in the design. There is certainly adequate information available on SPICE models, especially G-P.[8] Often, parameters just "don't look right," and questionable parameters should be resolved before the design begins. It is a simple matter to evaluate the device. Two simple test circuits are used, as shown in ***Figure 1***. The same RF test circuit can be used to calculate both S-parameters and power performance; no special set-up is necessary other than to create graphs for the desired quantities. The device's small-signal current gain and maximum available gain are also calculated to make sure they are close to the expected f_t and f_{max} advertised for the device. ***Figure 2*** shows the current gain, H_{21}, and the maximum available gain G_{max}, indicating f_t and f_{max} of 45 GHz, in good agreement with the expected 50 GHz. To make certain that any potential instability problems do not exist, stability circles are computed. Finally, the HBT's I/V characteristics are swept to make certain that the DC part of the model is reasonable, and to determine the base bias current that provides the proper collector bias current. The I/V characteristics are shown in ***Figure 3***.

Determine the Device Size, Bias Point and Optimum Load

Beginning with the load impedance, for a class-A amplifier, the resistive part is given by the well known relation,

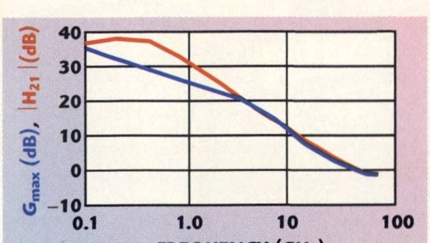
$$R_L = \frac{V_{cc} - V_{min}}{I_{cc} - I_{min}} \tag{1}$$

and the output power, P_{out}, is

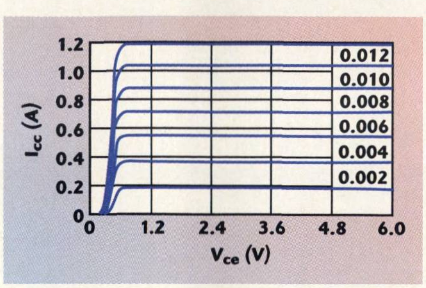
$$P_{out} = 0.5\,(V_{cc} - V_{min})(I_{cc} - I_{min}) \tag{2}$$

where

V_{cc} = bias voltage
I_{cc} = bias current

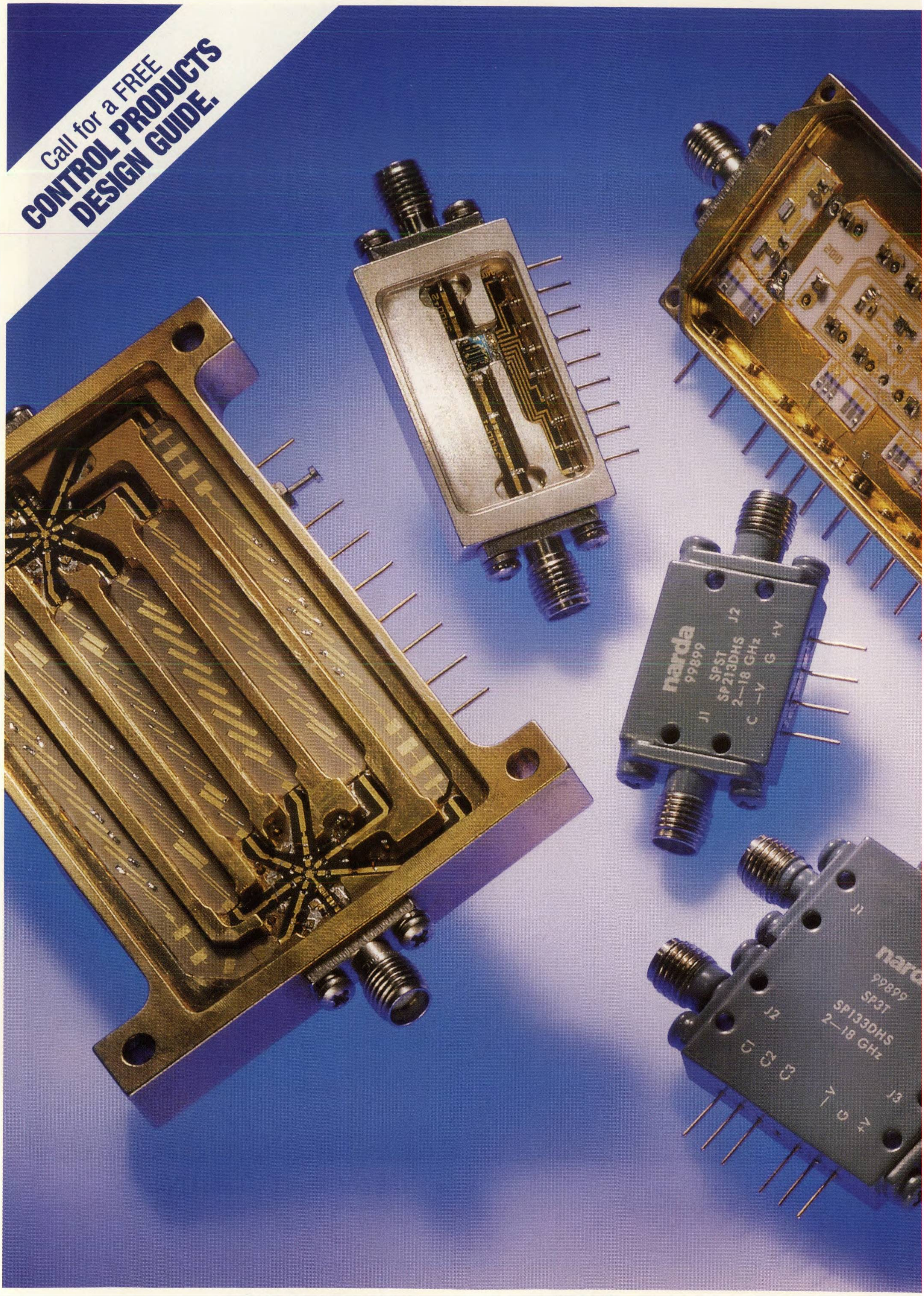


▲ *Fig. 2 G_{max} and current gain (H_{21}) of a single cell device calculated from the RF evaluation circuit.*



▲ *Fig. 3 I/V characteristics calculated from the DC evaluation circuit.*

[Continued on page 26]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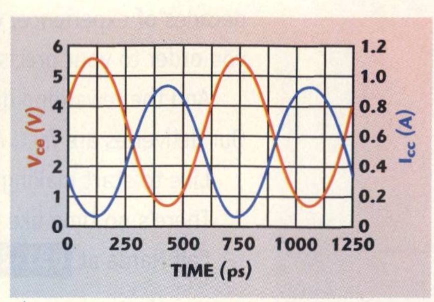
V_{min} = minimum collector-to-emitter voltage
I_{min} = minimum collector-to-emitter current

The device's I/V curves show that V_{min} is approximately 0.5 V, and the estimated I_{min} is approximately 0.05 A. Noting that V_{cc} = 3.4 V, and experimenting a little, I_{cc} = 0.5 A. This results in P_{out} = 0.65 W and R_L = 6.4 Ω. These are starting values, and may have to be modified somewhat. According to the foundry, the current density in the devices must be limited, under bias conditions, to 25 kA/cm^2. The individual cells have areas of 50 μm^2, so a device having 40 cells is needed. The foundry offers a 20 cell device, so two of these can be used in parallel.

In most power amplifier designs, a shunt inductance must be provided to resonate the device's output capacitance. However, from S-parameters, it is found that, at this relatively low frequency, the output capacitance is negligible, so no reactive tuning is needed. The load is purely resistive.

Now the harmonic-balance simulator is used to optimize the bias and load impedance using the RF evaluation circuit shown previously. No attempt has been made to match the input at this time; the excitation is simply increased until maximum output power is achieved. The load impedance is adjusted while monitoring the collector waveforms and adjusting the power. It is a simple matter to do this with the "tune" mode; numerical optimization, although available, is not necessary. The optimum condition is achieved when both the voltage and current minima are near zero, but not clipping. If the resulting output power is not right, the bias current and load resistance are adjusted until the correct power is achieved. Note that an extra fraction of one dB in output power is allowed to compensate for losses in the output matching network. The final collector current is 0.47 A and load resistance is 5.0 Ω.

Figure 4 shows the collector voltage and current waveforms. These are *internal* quantities; that is, they are the current in, and voltage across, the collector-to-emitter controlled source. It is essential to calculate the voltage and current at this point, not at the terminals. The terminal quantities include voltage drop across the collector and emitter resistances, and current in any output capacitance that might exist. Thus, it is impossible, by viewing external quantities, to know whether the load is optimized. In this case, the internal voltage and current exhibit a precise 180° phase



▲ *Fig. 4 Collector current and voltage waveforms at 1.8 GHz calculated from the RF evaluation circuit.*

[Continued on page 28]

difference, showing that the output reactance is negligible (or if it were not negligible, proper output tuning) and no saturation or clipping. ***Figure 5*** shows the dynamic load line. Its lack of ellipticity also indicates that there is no significant reactance in the output.

As a final test, a load pull simulation of the device is performed using the "load pull wizard" in the simulator. (A *wizard* in Microwave Office 2002 software is a separate software program that controls the simulator through an interface based on Microsoft's *component object model* (COM)). COM is a software technology that allows versatile interaction between separate software modules. ***Figure 6*** shows load pull contours of output power. These indicate that the load impedance is indeed optimized, and that the circuit is not unduly sensitive to the load.

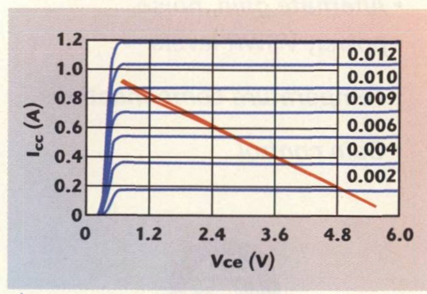
Determine the Input Impedance

The Microwave Office 2002 design suite enables the operator to calculate a large-signal input impedance, $Z_{in}(\omega)$. This is defined as

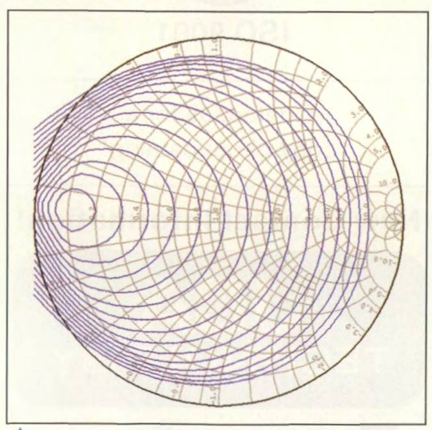
$$Z_{in}(\omega) = \frac{V_{in}(\omega)}{I_{in}(\omega)} \qquad (3)$$

where $V_{in}(\omega)$ and $I_{in}(\omega)$ are the input voltage and current Fourier components at the excitation frequency. This is the device input impedance that should be used for designing a matching circuit.

To design a matching circuit, it is helpful to have a lumped-element model of the HBT's input impedance. The dominant input elements in the HBT model are the base resistance and collector-to-emitter capacitance, so it is no surprise that a series RC network models the input impedance quite well. By plotting the input impedance of the HBT and the model on the same graph, the model can be easily adjusted to fit the input impedance. Again, optimization could be used for this task, but it is a simple,



▲ *Fig. 5 Dynamic load line.*



▲ *Fig. 6 Load pull contours showing output power in 1 dB increments. The curve closest to the center of the chart represents 26 dBm.*

[Continued on page 30]

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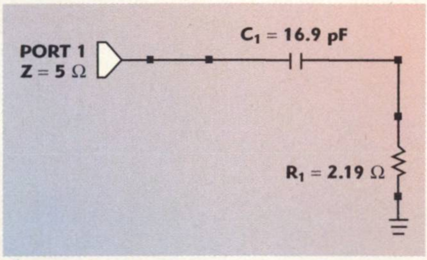
two-minute job with the tuner. The input model, consisting of 16.9 pF and 2.2 Ω, is shown in ***Figure 7***, and a comparison of the measured and modeled impedances is shown in ***Figure 8***.

Synthesize the Input Matching Network

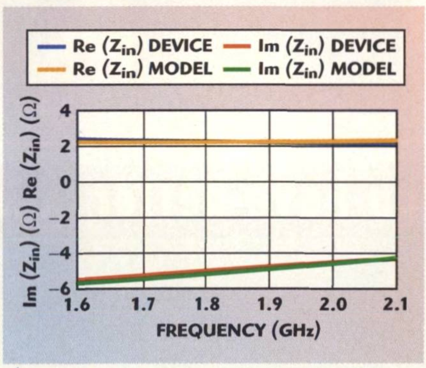
Several considerations drive the design of the input network. To eliminate low frequency gain, it should have a high pass structure, and should allow for easy biasing and DC blocking. Because of the high Q of the load, and the need to transform from a very low impedance to 50 Ω, the design of the network is not simple.

To meet these requirements, a series-L, shunt-C design is used. A "constant-Q" approach is employed, in which elements are selected by moving along a contour of constant Q on the Smith chart. This is an entirely graphical process which can be performed with Microwave Office 2002 software using the *"real-time tuning"* mode. Additionally, resistive loading is used to optimize the input match over the relatively wide bandwidth of 1.6 to 2.1 GHz. The loading introduces loss, of course, but the gain of modern HBTs is so great that it is acceptable. It also reduces the sensitivity of input return loss to uncertainties in the device model.

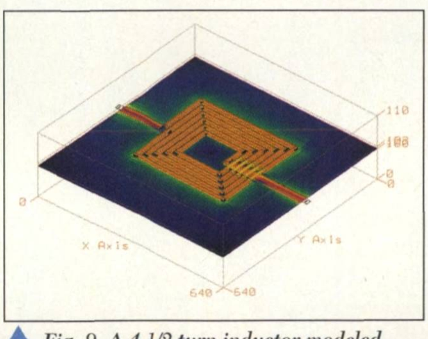
Once the element values are found, the inductances are replaced, one at a time, with square-spiral inductors, re-optimizing the circuit after each inductor is replaced. For best accuracy, the inductors should be analyzed by a planar electromagnetic (EM) simulator. Conventional lumped-element models of foundry devices are often less accurate and do not allow the designer to customize and optimize his own inductors or capacitors. For this task, the integral EMSight™ electromagnetic simulator in the Microwave Office design suite is used. The EMSight simulator will generate and export a SPICE equivalent circuit of the rectangular spiral inductor, so it is easy to determine whether the inductor has the desired value. ***Figure 9*** shows the current and electric field analysis for a four and a half turn spiral inductor. ***Figure 10*** shows the final matching circuit which resulted in an input re-



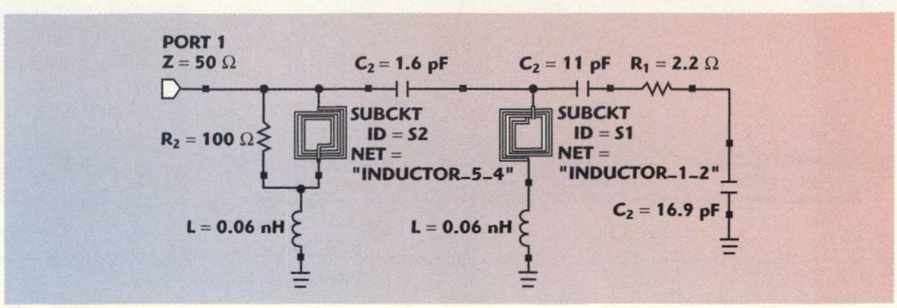
▲ *Fig. 7 The input model of the device.*



▲ *Fig. 8 A comparison of the input impedance of the device and the model.*



▲ *Fig. 9 A 4 1/2 turn inductor modeled with the electromagnetic simulator.*



▲ *Fig. 10 Input matching circuit.*

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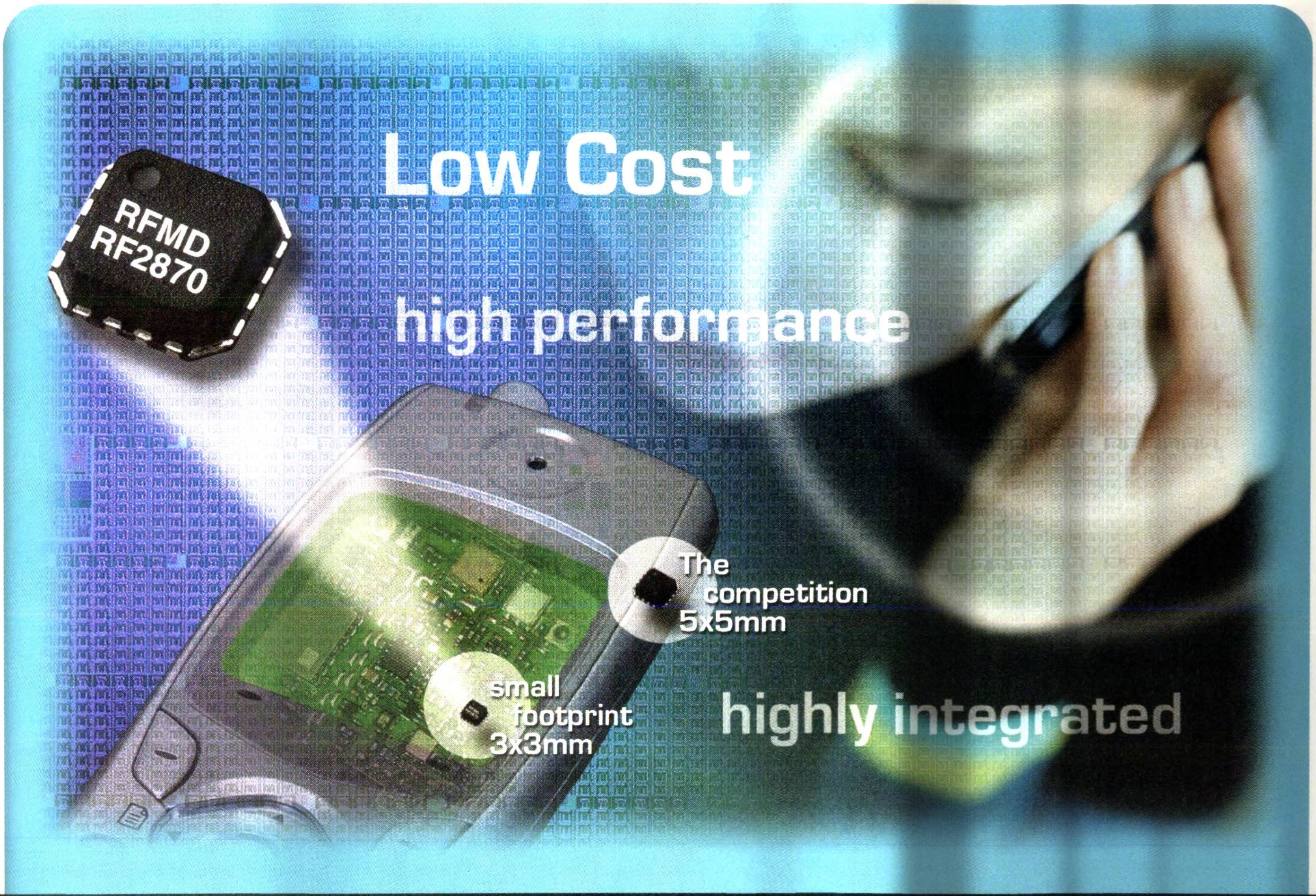

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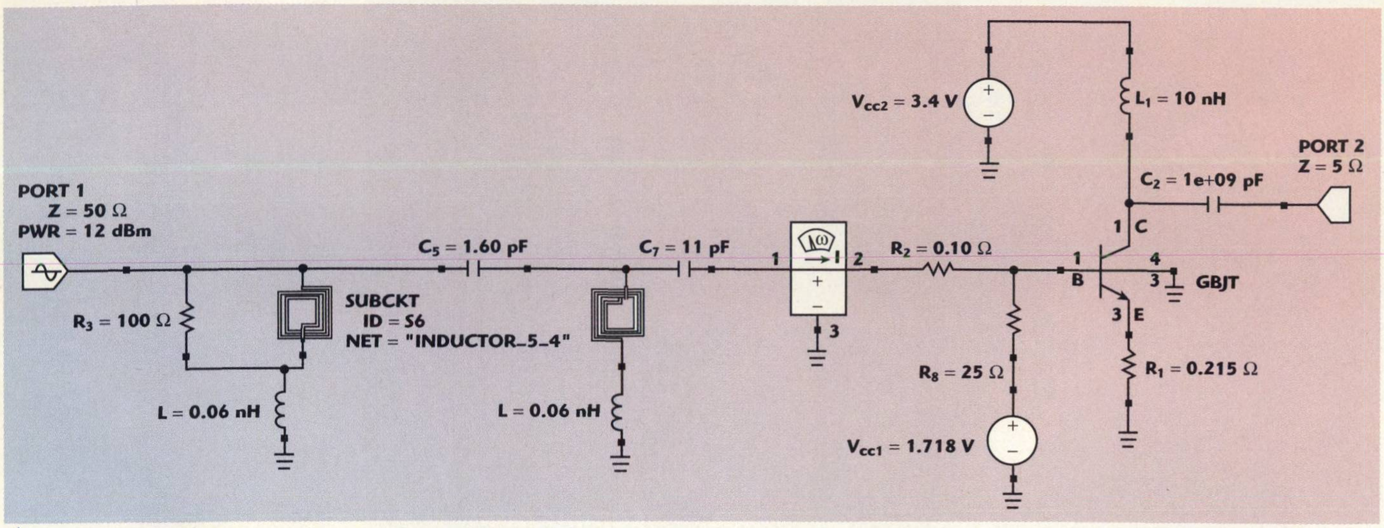
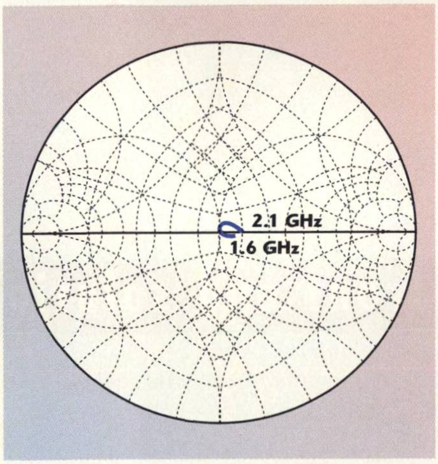
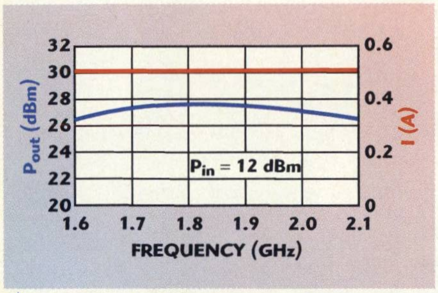


▲ *Fig. 11 The complete amplifier circuit.*



▲ *Fig. 12 Input reflection coefficient of the entire amplifier.*

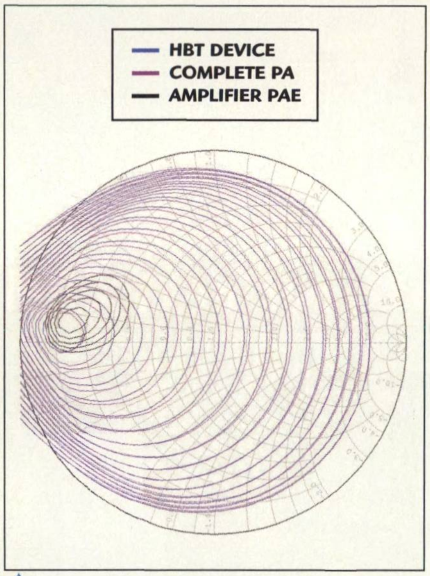


▲ *Fig. 13 DC bias current and output power of the amplifier.*

turn loss of better than 20 dB across the 1.6 to 2.1 GHz band.

Connect the Matching Circuit to the HBT

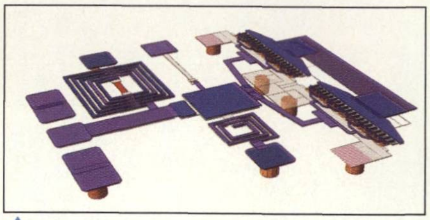
Now the matching circuit is connected to the HBT, and the combination, shown in ***Figure 11***, is analyzed. ***Figure 12*** shows the input reflection coefficient, and ***Figure 13*** shows the output power and bias current *without any further tuning or optimization of the circuit*. In ***Figure 14***, a set of load pull contours for the complete amplifier is shown, which is very similar to the one for the device alone. Finally, ***Figure 15*** shows the amplifier layout, which was also produced using the layout capabilities in the Microwave Office software. The 40 cell HBTs were realized as two 20 cell devices in parallel, and the inductors and capacitors are readily identifiable.



▲ *Fig. 14 Load pull contours for the complete amplifier compared to the HBT alone.*

Design the Output Matching Circuit

The output matching circuit is not part of the chip design, but an amplifier will not work very well without one. Because of the low load impedance required by the amplifier, an output matching circuit is unavoidably lossy. Most of the loss is generated where the currents are greatest, in the elements closest to the chip. Ideally, these should use capacitive microstrip stubs, but size limitations may dictate the use of chip capacitors instead. In this case, the main problem is a trade-off between capacitor cost and Q. A second problem is the large impedance transformation between 5 Ω at the chip and the invariably 50 Ω outside world, which creates a direct trade-off between bandwidth and loss.



▲ *Fig. 15 Layout view of the complete amplifier.*

It was found that a matching circuit consisting of series transmission lines and shunt capacitors represents a good trade-off between loss and size. High quality RF ceramic chip capacitors are used. The chip must also be designed to allow the use of multiple bond wires, as even bondwire loss can be significant.

CONCLUSION

In designing power amplifiers, the combination of a methodical approach and the use of modern CAD tools allows a designer to produce a valid design with minimal effort. The MMIC design methodology becomes

[Continued on page 34]

extremely efficient when the designer has access to a single CAD environment that provides EM simulation, harmonic balance, load pull analysis and integrated foundry libraries for laying out the chip. The key to the process is to determine as much as possible about the power device to be used in the amplifier before the design itself begins. The old saying, "knowledge is power," applies here more than you might imagine.

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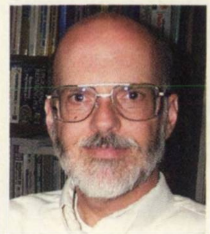


Dr. Stephen A. Maas *received his BSEE and MSEE degrees in electrical engineering from the University of Pennsylvania in 1971 and 1972, respectively, and his PhD in electrical engineering from UCLA in 1984. Since then, he has been involved in the research, design and development of low noise and nonlinear microwave circuits and systems at the National Radio Astronomy Observatory, Hughes Aircraft Co., TRW and the Aerospace Corp. He is now chief scientist at Applied Wave Research Inc. (AWR), where he specializes in nonlinear circuit simulation technology. For several years he was a member of the electrical engineering faculty at UCLA, where he still periodically teaches regular and extension courses.*

Ted Miracco *received his bachelor of science degree in electrical and computer engineering from Carnegie-Mellon University in 1987. He is a co-founder and the executive vice president of Applied Wave Research Inc. (AWR), an El Segundo, California-based company that was founded in 1994. He has worked in microwave product development for over a decade, specializing early on in microwave filter design. Before joining AWR, he was a senior account executive at Cadence and, prior to that, was responsible for business development and worldwide corporate accounts at EEsof. He can be reached via e-mail at miracco@mwoffice.com.*

Raytheon Adds Reliability Engineering Consulting

Raytheon Co. is expanding its Reliability Analysis Laboratory (RAL) technical offerings with the incorporation of reliability engineering consulting services. Reliability is a core feature that can make or break the long term success of both products and companies. Design for Reliability (DFR) engineering tasks are just as important as development engineering tasks in keeping repairs and warranty expenses low and thus profits, product quality and customer satisfaction high.

Raytheon's experience in reliability engineering is extensive in the areas of problem solving, program planning, product and system modeling, testing failure analysis, process improvement and field data analysis for large and small programs, both in the government and commercial sectors. "Planning for product reliability is just as important as planning for a product's design and manufacturing. The amount of product reliability must be in proportion to a product's usage and warranty goals. Too much reliability and the product will be too expensive. Too little reliability and the warranty and repairs costs will be high," said Bill Tice, RAL's department manager.

Joe Dzekevich, the RAL's reliability engineering lead added, "Reliability is especially important to start-up companies, which have to go to market quickly with a reliable product or go out of business. Up-front DFR methodologies are a good way to get there ahead of your competition or to maintain a lead."

The added reliability engineering services will encompass areas such as reliability program planning; reliability and maintainability predictions; system redundancy and availability modeling; failure mode prediction and analysis and accelerated stress test planning and analysis. "Add to these reliability engineering services RAL's already extensive failure analysis capabilities and one now has a premier one-stop shopping solution for reliability engineering needs," Tice said.

For more information on Raytheon's complete set of reliability engineering services, visit the RAL's Web site located at: http://www.reliabilityanalysislab.com.

Lockheed Martin Receives $131 M for Leadership of Missile Defense Team

Lockheed Martin received a $131 M funding increment to continue its leadership of the Battle Management/Command and Control/Communication (BM/C2/C) program for the Missile Defense Agency's (MDA) Missile Defense National Team.

The contract runs through December of 2003 and funds the team's efforts to devise and field an operational structure that seamlessly and effectively links the numerous sensors, weapons and command and control systems currently utilized in the individual US missile defense programs or elements. Other BM/C2/C team members include TRW, Boeing, Northrop Grumman, General Dynamics and Raytheon. In addition to leading the National Team's BM/C2/C program, Lockheed Martin serves as a member of the Systems Engineering and Integration team.

Missile Defense Agency Director Lt. Gen. Ronald Kadish created the Missile Defense National team in January 2002 to bring together the best engineers and scientists in America. The team is focused on enabling the US military services to field missile defense elements of the overall Ballistic Missile Defense System as soon as practical; developing and testing technologies; and improving the effectiveness of deployed capabilities by inserting new technologies as they become available. "This is a bold, new approach to missile defense. Lockheed Martin is committed to delivering MDA's vision of an effective, integrated, layered missile defense, capable of defeating all ranges of threat," Kier said.

Since receiving its initial contract and funding of $23 M in February 2002, the BM/C2/C team has built a staff of approximately 130 individuals, drawn from the six member companies, and has identified capabilities to be developed and deployed for an initial block deployment in 2004. "We are fortunate to have assembled the best and brightest minds from American industry to work as members of a team on this vitally important national project," Kier said.

US Air Force Accepts First Article ASR-11 DASR Systems

Raytheon Co.'s first article ASR-11 digital airport surveillance radar (DASR) systems located at Eglin Air Force Base, FL, and Stockton, CA, have been accepted by the US Air Force. This contract milestone has been achieved through confirmation of compliant performance through testing.

Acceptance of these first article systems clears the way for acceptance of low rate initial production (LRIP) systems, which are being installed at multiple locations throughout the US. Raytheon is currently manufacturing 38 LRIP systems for the Department of Defense (DoD) and the Federal Aviation Administration (FAA).

The all solid-state ASR-11 radar provides a modern, highly reliable upgrade to replace the existing aging ASR-7 and 8 radars. The ASR-11 provides airport terminal area primary surveillance coverage to 60 nautical miles and secondary surveillance coverage to 120 nautical miles. The ASR-11 digital processing provides greatly improved target and weather processing to support the improved performance associated with deployment of the Standard Terminal Automation Replacement System (STARS). The ASR-11 DSAR is a joint procurement for the DoD and

FAA. Raytheon is under contract to develop and install up to 93 ASR-11 systems for the DoD ad 122 for the FAA over the next several years.

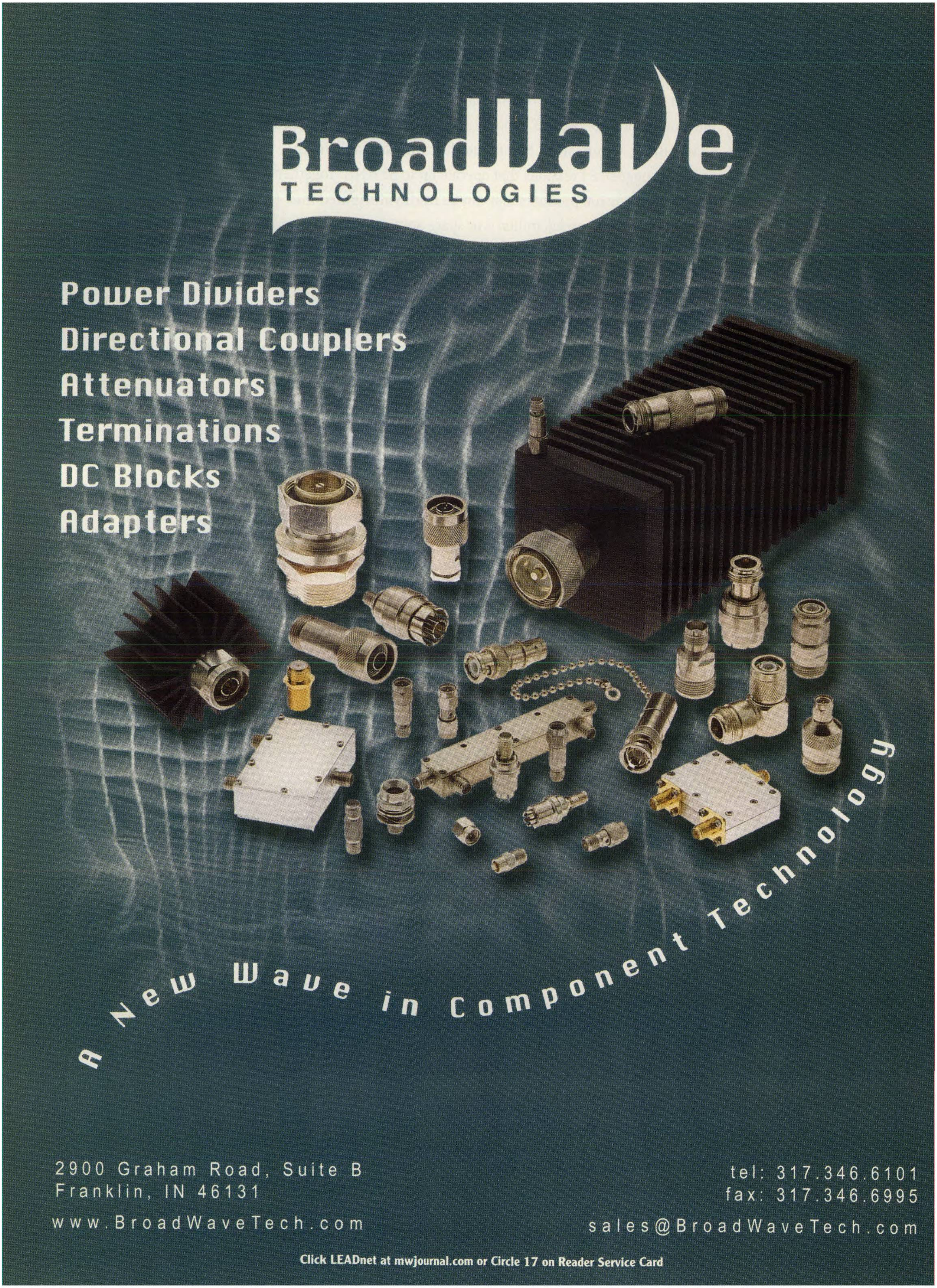
Northrop Grumman Wins AF Mission Planning Contract for Precision Guided Munitions

Northrop Grumman Corp.'s Integrated System sector has been awarded a $49 M, nine-year contract by the US Air Force to provide Precision Guided Munitions Planning software (PGMPS). This program will give the Air Force a common, integrated personal computer-based mission planning application for all current and future USAF PGMs and their variants. The Air Armament Center at Eglin Air Force Base, FL, selected Northrop Grumman PRB Systems, a unit of the Integrated Systems' Airborne Early Warning and Electronic Warfare (AEW & EW) Systems business area, to develop the PGMPS user interface to provide a common, consistent weapon planning process for all USAF PGMs.

PGMPS will be a unique planning component of the Joint Mission Planning System and will replace the current weapon planning modules for the AGM-130/GBU-15, Joint Direct Attack Munitions (JDAM), Joint Standoff Weapon (JSOW), Wind Corrected Munitions Dispenser (WCMD), Small Diameter Bomb (SDB), Joint Air-to-Surface Standoff Missile (JASSM) and other future PGMs.

"This program clearly establishes Northrop Grumman as the leader for PGM planning in the Air Force and presents a strategic opportunity for Integrated Systems to be a major contributor to streamlining the joint targeting process," said Larry Schadegg, president of PRB Systems. "With our teammates, we are positioned to develop a robust and cost-effective solution to give our war fighters the edge."

Northrop Grumman's team consists of the company's Information Technology sector, San Pedro, CA; Zeltech, Hampton, VA; and 21st Century Business Consultants Inc., Waldorf, MD. The first phase of the contract has a value of $4.1 M over 18 months and includes work for WCMD, AGM-130, JDAM and JSOW. Over the life of the contract, Northrop Grumman will analyze requirements, design, develop, document, test and qualify the initial releases of PGMPS and updates; provide training to support the testing and fielding of the software; and support training and integration to achieve operational use and fielding. The work will be performed at AEW & EW Systems' site in Hollywood, MD, and Niceville, FL. ■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INTERNATIONAL REPORT

Martin Streetly, International Correspondent

Surrey Aids Nanotechnology Research with UHV STM/SEM

The University of Surrey's (Guildford, Surrey, England) acquisition of an Ultra High Vacuum Scanning Tunnelling Microscope/Scanning Electron Microscope (UHV STM/SEM) is being billed as a central figure to a £1.2 million future nanotechnology research project that is headed up by the institution's Professor of Solid-State Electronics, Ravi Silva. Built to specification, the Surrey UHV STM/SEM is housed in the University's Advanced Technology Institute (ATI) and is claimed to be one of only six such instruments worldwide. Professor Silva's team will use the device (which is described as representing a 'completely new dimension [in] the fabrication of [nano-scale] devices') to research carbon nanotubes (the Buckyball derivative of Carbon 60) as a precursor to a new generation of devices that will be smaller and faster than current silicon technology. The Silva programme is being channelled into four areas to facilitate rapid exploitation in all, with the key issue being small device structures. Areas being explored include field emission-based electron energy sources, quantum dots as light emitters and the development of ultra fast terrahertz electron beam and carbon nanotube energy storage devices.

Surrey's carbon nanotube research is initially funded for a three-year period and it is hoped that other projects will be able to exploit the new UHV STM/SEM facility well into the next decade. For its part, Surrey's ATI is described as bringing together six research groups from three Schools within the University to stimulate cross-disciplinary research. Alongside the described carbon nanotube effort, overall research areas will include microwave subsystems, lasers and optoelectronics, large area electronics, ion beam applications and biosensors. To maximise utility, Surrey further notes that the ATI is a multi-purpose facility that has been designed for the rapid re-direction of use when required.

Philips Opens New Semiconductor Design Centre

On 8 October 2002, Netherlands contractor Royal Philips Electronics announced the opening of the Philips Delft Design Centre (PDDC) at Delft, Holland. A cooperative venture with Delft University of Technology (DUT), Philips cites the new facility as a symbol of its commitment to investing in the future of standard analogue semiconductor technology and to providing the best possible learning environment for electronic engineering students. Initially, the PDDC will have three permanent Philips employees (dedicated to design work in support of standard analogue Integrated Circuit (IC) developments) and three professors, two PhD students and two MSc students from DUT. The PDDC project builds on a 17-year relationship between Philips and DUT and is designed to capitalise on DUT's prowess in the analogue field. The emphasis on analogue design is said to reflect the importance of analogue devices (such as interface and power management ICs) to the development of digital applications. Looking to the future, Philips engineers from France, the Netherlands and the USA will be able to rotate through the facility to build competence and there is the expectation that the PDDC will have grown to 30 full-time employees by 2005-2007.

Sweden Starts Automobile Radar Interference Tests

Sweden's National Testing and Research Institute has launched a programme to provide a capability whereby it can test the impact of radar-generated electromagnetic interference on processors used to control automobile functions such as throttle control, gear (shift) changing and braking. While not yet as common as potential risks from sources such as mobile phone base stations, the Swedes feel that the potential threat from radars such as air traffic control sensors at airports is sufficient as to warrant an interference test programme. Accordingly, UK contractor TMD Technologies Ltd. has developed and supplied a dedicated L-band (1 to 2 GHz) amplifier for use in the effort. Designated as the PTC6468, the new equipment operates from 1.1 to 1.5 GHz at 20 kW (min), with 24 kW (typ) at the band edges and 35 kW at mid-band. Created over a six-month period, PTC6468 incorporates a high power L-band travelling wave tube source, a switch mode power supply and control circuitry. The unit is liquid cooled (water with anti-freeze and anti-corrosion additives) and control is via a digital front panel or remotely using an IEEE standard interface. Of the various components, the amplifiers switch mode power supply is a re-engineered variant of TMD's standard 50 kW radar power supply and the company sees PTC6468 as the starting point for an interesting niche market in such test equipment.

Portugal Selects Italian Radar

Portugal has selected Italian contractor Galileo Avionica's APS-717 search and navigation radar for installation aboard the 12 EH.101 utility/search and rescue helicopters it is procuring from the Anglo-Italian AgustaWestland joint venture. Operating in the I-band (8 to 10 GHz), APS-717 is described as being suitable for both fixed- and rotary-wing applications, and as offering search and rescue, surveillance, navigation and target designation capabilities. The equipment can be integrated with its host platform's navigation system and a forward-looking infra-red (FLIR) sensor. A 32 target track-while-scan facility is a system option. Usually reliable sources suggest that APS-717 variants have been installed aboard Italian Air Force AS-61A-4/HH-3F search and rescue helicopters and G222 transport aircraft together with AB 412HP helicopters of the Italian coast guard service. As currently scheduled, Galileo Avionica will deliver the Portuguese radars during the summer of 2003.

BAE Systems Wins Kuwaiti HIDAS Order

UK-based BAE Systems Avionics has announced that Kuwait has selected the company's Helicopter Integrated Defensive Aids System (HIDAS) for installation aboard its AH-64D Longbow Apache battlefield attack helicopters. Alongside the equipment itself, the deal involves a comprehensive training and support package that includes integration of the system into a Longbow Apache crew trainer that is being supplied to the Kuwaiti Air Force. The overall package is understood to form part of the US Foreign Military Sales effort under which Boeing is supplying Kuwait with the AH-64.

The HIDAS configuration being supplied to Kuwait is the same as that being fitted to the British Army's Apache AH.1 attack helicopter. As such, the suite is described as being an integrated modular equipment that incorporates the BAE Systems Avionics' Sky Guardian 2000 radar warning receiver (RWR) and Series 1223 laser warning receiver (LWR), the BAE Systems North America AN/AAR-57(V) Common Missile Warning System (CMWS) and a Thales Optronics (Vinten) Vicon 78 Series 455 CounterMeasures Dispensing System (CMDS). Of these, the Sky Guardian 2000 RWR covers the 0.5 to 40.0 GHz frequency range and houses (in its electronics unit) the suite's controller. The HIDAS LWR's processors are housed in the system's Data Transfer Unit (DTU), as are those for its CMDS. For its part, the CMDS features dedicated chaff (64 shots each magazine) and infra-red decoy flare (32 each magazine) dispensing modules, while the heads for the suite's LWR, CMWS and RWR are located in such a way as to provide 360° coverage in azimuth. The previously cited DTU is the medium by which the system receives necessary pre-flight messages, with data being loaded using a standard 40 MB PCMCIA card. ■


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THE COMMERCIAL MARKET

MEMS Moves Stealthily into the Wireless Sector

Driven by innovation, pushed by legislation, and pulled by market need, wireless MEMS will experience strong growth in the next couple of years. In-Stat/ MDR reports that worldwide revenues for wireless MEMS (in the form of RF MEMS components and wireless MEMS sensors) will experience a 32.2 percent compound annual growth rate between 2001 and 2006, with the cell phone, automobile, and industrial processing/condition monitoring markets benefiting most from what the technology has to offer.

"Development of the MEMS piece of the wireless pie has required significant innovation at the engineering level, while the electronics that provide the wireless portion, by and large, are already well established," says Marlene Bourne, a senior analyst with In-Stat/MDR. "It is the pairing of the two that is opening the doors to markets and applications that have been waiting for such a breakthrough."

In-Stat/MDR also found that:

- While packaging hurdles appear to have been overcome for RF MEMS, the next real issue being faced is getting the devices into the market — no one has yet proven volume manufacturability.
- The use of wireless MEMS sensor networks has been limited to date by a lack of interface standardization and the basic question of which wireless standard to choose — both of which are quickly being resolved.
- Sales of RF MEMS (switches/relays, filters, inductors, etc.) are forecast to reach the $200 M mark by 2006, which is down considerably from earlier forecasts. This is the result of better pricing and volume guidance from companies in this brand new market segment.
- Sales of wireless MEMS sensors are forecast to reach about $700 M in 2006, largely based on the use of sensors in tire pressure monitoring systems. However this application is not yet a done deal — a pending government decision could either completely eliminate this segment beyond 2006, or provide for exponentially higher revenues.

New Ultra Wideband Report

Ultra Wideband (UWB) is a wireless technology that is trying to fill a void for extremely high data rate wireless connectivity, beyond what 802.11a/b and Bluetooth are capable of. Until February 14, 2002, when the Federal Communications Commission (FCC) approved commercial use of the technology, the military and the government were the only users of this technology. UWB has a data rate potential of greater than 100 times that of Bluetooth solutions and over twice as fast as 802.11a, while using 80 percent less power than current Wi-Fi implementations. UWB will allow for the transmission of video files and streams between digital TVs, projectors, camcorders, PCs, and digital set-top boxes. We can expect to see UWB-enabled electronics and chipsets devices to start appearing in 2003.

Allied Business Intelligence (ABI) forecasts the total global shipments for UWB-enabled electronics and chipsets under the moderate scenario could reach 45.1 million units by 2007, with resulting revenues of $1.3 B. These projections include shipments into market segments including communications, imaging, vehicles, locators and military and government use.

The new ABI report, "Ultra Wideband (UWB) Wireless — An Evaluation of Technology Prospects and Potential Market Applications," investigates the world markets of UWB and examines if this will be the next generation in wireless communications such as wireless LANs, personal area networks, imaging, vehicles, locators and radar. The report employs a scenario analysis approach (moderate, aggressive and pessimistic) in providing forecast for 20 different market segments through 2007.

Smart Appliances: Bringing the Digital Home Closer to Reality

Smart appliances cover a diverse range of products by both small and large companies. According to a report from In-Stat/ MDR, most segments have been affected by the economy and many smaller companies have been forced out of business or have been acquired by larger ones. However, there are still many small companies interested in smart appliances.

Smart appliances include two main categories: Internet-enabled consumer electronics and white-goods. Smart Internet-enabled consumer electronics include audio products, picture frames, analog Internet TV, and e-mail devices, while white-good segment is divided into major and portable appliances. Consumers remain confused about the value of smart appliances and whether they are worth the premium prices being demanded, and manufacturers remain concerned about applications, upgrade ability, standards, and subscription fee requirements. Internet and home networking penetration rates and price also affect the market for smart appliances.

There have been considerable product introductions in the Internet audio segment, which is expected to continue as manufacturers and consumers decide which applications are most desirable and prices decline. Products such as e-mail devices and Internet picture frames will be spurred on by continuing interest in complementary technology products and growing consumer awareness. These products have been primarily targeted to the North American market. However, this is beginning to change.

CDMA Leads the Pack in the Race to 2.5G and 3G Networks

The race to deploy more advanced wireless networks is in high gear in North America, with next generation CDMA networks currently leading, says Allied Business Intelligence (ABI). While both TDMA/ GSM and CDMA operators have picked up the pace in recent months, it is the CDMA operators who currently dominate this closely watched race. Extensive 2.5G coverage now exists in both the US and Canada for more than 55 million CDMA subscribers in the region.

While CDMA2000 1X technology allows operators to roughly double voice capacity and provide a theoretical data rate of 144 kbps, the GSM operators should not be discounted just yet. Despite the arduous transition from TDMA to GSM/GPRS, the next step for many North American GSM operators is EDGE technology, a much less complex transition. EDGE offers data rates more than double that of CDMA2000 1X. Of course, the CDMA camp has their answer to EDGE technology with higher evolutions of CDMA2000. This constant jockeying for dominance has heightened the awareness of both investors and consumers alike as they have become increasingly interested in observing the evolution of wireless technology.

"In North America, CDMA is clearly ahead, benefiting from its easier upgrade path from 2G. However, we cannot discount the enormous presence of GSM/GPRS networks in the rest of the world. In the end, spending on upgrading these networks to WCDMA will dominate total 3G investments over the next five years. It's important to differentiate between first deployed and total deployed," declares ABI senior analyst Edward Rerisi. He adds, "That said, we are still closely examining the situation in Europe for signs of a further delay in WCDMA."

In the end, the consumer will likely win as 3G coverage begins to proliferate worldwide. Despite the current lead of CDMA networks in the race to 3G, those based upon GSM technology will likely become more ubiquitous. In a new study by ABI, findings indicate that spending on GSM-based 3G networks, or WCDMA, will far exceed that of CDMA-based networks. The report, entitled "Wireless Base Stations: Global Deployment & Revenues for 2G, 2.5G and 3G Systems," reveals that WCDMA spending will represent a staggering 78 percent of total 3G investments over the next five years. Note that ABI considers CDMA2000 1X a 2.5G technology. Updated information on 3G developments will be available from ABI in first quarter of 2003 as part of the Wireless Base Station study. ■

INDUSTRY NEWS

■ The **Connecticut Microwave Corp.** acquired the combiner, duplexer and filter product lines of **Davicom Technologies Inc.** The acquisition included all engineering, tooling, manufacturing process data and work in progress. These products are primarily used in dedicated systems including business, safety and municipal.

■ **Littelfuse Inc.** has purchased **Semitron**. The all cash FAB and product purchase completes Littelfuse's current product line of varistors, low power silicon diodes, polymer-based, high speed ESD suppressors and a range of resettable and single-use fusing approaches.

■ **Laird Technologies** has acquired **Boldt Metronics International (BMI)** for $28.7 M in cash and the assumption of $1 M of long term capital equipment leases. The companies will combine their technical, geographic and human resources to become the premier supplier to the electronics industry globally.

■ **EMS Technologies Inc.** has formed a new division, **EMS Satellite Networks**. The new division includes all of the satellite broadband products formerly managed under the company's Space & Technology Group/Montreal, including all DVB-RCS systems, hubs and terminal products.

■ **The Tru-Connector Corp**. has changed its official name to **TRU Corp**. The new name is intended to reflect the broader nature of the business while maintaining the company's basic identity.

■ **TriPoint Global Communications** announced that the microwave antenna products formerly produced by the company's **Mark Antenna** business will be manufactured under the **Prodelin** brand. The company will discontinue the use of the Mark Antenna name and brand.

■ **Vitelec Electronics** is re-branding as part of the **Emerson Network Power–Connectivity Division**. An important element of Emerson's corporate brand strategy, the name change marks a new phase in the development of Vitelec and lays the foundations for the introduction of a new range of advanced products and enhanced levels of customer service for the European marketplace.

■ **Sirenza Microdevices** has established a new customer support office in Shenzhen, China. This office will provide critical real-time applications engineering support to key OEM accounts throughout China. Sirenza's new office is located at the Pavilion Hotel Business Building, Room 2308, 4002 Huaqiang Road North, Fu Tian District, Shenzhen, China +86 755 82074046 or e-mail: China-apps@sirenza.com.

■ **Rohde & Schwarz Inc. (R&S)** has opened an expanded test and measurement systems center located in the metro Dallas, TX, area. The facility will support all test and measurement systems, as well as EMI receiver sales. In related news, the company has announced that all EMI test receivers and software packages formerly offered by **Tektronix** are now available directly through R&S. For all other Rohde & Schwarz general purpose test and measurement equipment, Tektronix remains the exclusive distributor in North America.

■ **Agilent Technologies** has recently inaugurated the new Agilent Technologies Millimetre-wave Laboratory at the University of Manchester (UK) Institute of Science and Technology. The new laboratory has facilities for measurements up to 140 GHz, includes a 110 GHz single sweep vector network analyzer, a 8510C vector network analyzer, a 50 GHz portable spectrum analyzer and a modular DC source, semiconductor parametric analyzer.

■ **Ansoft Corp.** has refocused operations on its core software business and will discontinue its efforts to apply its IP technology core to hardware design through **Altra Broadband**'s Irvine Technology Center. The center closed operations in September.

■ **SV Microwave Inc.** has entered into a manufacturing agreement with **Andrew Corp.** under which Andrew is licensed to manufacture certain SV connector products including the Blind Mate and Base Transceiver Station connector product lines. The new agreement gives SV immediate capacity in China where the designated connector products will be manufactured at Andrew's Suzhou location.

■ **Teravicta Technologies Inc.** and **Read-Rite Corp.** have entered into an agreement to utilize Read Rite's manufacturing capabilities for the high volume production of Teravicta's line of low loss, high linearity RF MEMS switches. Teravicta currently produces its RF MEMS switches on its own low volume pilot production line, but this alliance will transition it to high volume production at Read-Rite.

■ As part of a recently concluded high volume production agreement, the RF Components division of **Raytheon Co.** will supply **Hyundai/Curitel Inc.** with the model RMPA0951A-102 GaAs HBT power amplifier for the handset manufacturer's DD-500 PCS platform. Hyundai/Curitel has earmarked the DD-500 for the Korean market.

■ **Spirent plc** and **UbiNetics** have entered into a strategic partnership to expand their test solutions for the 3G market. Under the terms of the agreement, Spirent has acquired UbiNetics' CS100 product line and rights to associated intellectual property for £6.5 M cash. In addition, under the agreement Spirent and UbiNetics will partner for the further development of the CS100 product line and future test platforms.

[Continued on page 50]


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■ **Avnet RF & Microwave** has signed a distribution agreement with **Directed Energy Inc. (DEI)**. With this agreement, Avnet will now distribute DEI's Fast Power MOSFET transistors, a class of RF transistors designed for high power RF applications.

■ **Anritsu Corp.** has entered a strategic partnership with **Flextronics Test** whereby the two companies will co-develop integrated test solutions utilizing Anritsu's leading-edge test instrumentation and Flextronic's test software and application programs. The agreement enables Anritsu to provide global customers with complete ATE solutions for R&D, manufacturing and maintenance applications.

■ **RF Micro Devices Inc.** and **Jazz Semiconductor** have agreed to enter into a strategic relationship for silicon manufacturing and development. Under the agreement, RF Micro Devices will obtain a guaranteed lower cost source of supply for wafers fabricated utilizing Jazz's manufacturing processes, including SiGe, BiCMOS and RF CMOS. In addition, the companies will collaborate on the development of wireless technology roadmaps.

■ **Applied Wave Research Inc. (AWR)** has agreed to provide process design kit library support for the GaAs foundry process at **OMMIC**, formerly **Philips Microwave Limeil**. MMIC designers will use the libraries, within Microwave Office™ 2002, to speed the design of high end wireless and fiber optic infrastructures while helping to ensure electrical specifications are met.

■ **United Monolithic Semiconductors (UMS)** and **IMST** have agreed on a closer cooperation. Under this co-operation, IMST will act as a UMS certified design house, having a preferred access to UMS process information, models and design tools. IMST will design and develop products for its own or common customers, based on the UMS processes.

■ **Phihong** has selected **Allied Electronics** as a new catalog distributor for its standard model power supplies and adapters. Allied brings additional customer support for power source design-ins as well as additional distribution support in Canada.

■ **E2V Technologies** became one of the first companies in the UK to achieve certification to the new ISO 9001:2000 international quality standard.

FINANCIAL NEWS

■ **REMEC Inc.** reports sales of $53.5 M for the second quarter ended August 3, 2002, compared to $60.4 M for the same period in 2001. Net loss was $16 M (35¢/diluted share), compared to a net loss of $15.4 M (34¢/diluted share) for the second quarter of last year.

■ **SEMX Corp.** reports sales of $8.2 M for the second quarter ended June 30, 2002, compared to $11.2 M for the same period in 2001. Net loss was $1.5 M (77¢/common share), compared to $684 K (14¢/common share) for the second quarter of last year.

■ **Ansoft Corp.** reports sales of $9.3 M for the first quarter ended July 31, 2002, compared to $11.2 M for the same period in 2001. Net loss was $3.3 M (28¢/diluted share), compared to $498 K (4¢/diluted share) for the first quarter of last year. In related news, the company's board of directors voted to amend its existing common stock repurchase program to permit the company to acquire an additional million shares of its common stock. Under the original program approved in 1998, the company had purchased 937,000 of the one million shares authorized for repurchase.

CONTRACTS

■ **PC Dynamics**, a subsidiary of **Integrated Performance Systems**, has signed a $440 K contract with **Lockheed Martin NE&SS–Surface Systems** to supply microwave circuit components for its Aegis program. If all options are exercised and PC Dynamics meets all performance criteria, the contract's value would be approximately $3.1 M over the next five years.

■ **Reynolds Industries** has completed negotiations for another release of a continuing master agreement with Northrop Grumman Corp.'s Integrated Systems sector. The contract is for high voltage and fiber optic cable assemblies for RF countermeasures on the F/A-18E/F Super Hornet strike fighter program.

[Continued on page 52]

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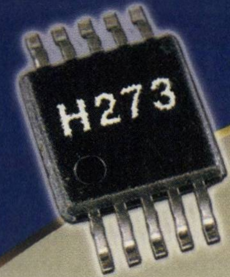

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NEW!	HMC425LP3	6 Bit Digital	2.4 - 8.0	0.5 to 31.5	LP3	$8.22
	HMC273MS10G	5 Bit Digital	0.7 - 3.7	1 to 31	MSOP10	$1.70
	HMC306MS10	5 Bit Digital	0.7 - 4.0	0.5 to 15.5	MSOP10	$2.49
	HMC307QS16G	5 Bit Digital	DC - 4.0	1 to 31	QSOP16	$2.49
	HMC230MS8	3 Bit Digital	0.75 - 2.0	4 to 28	MSOP8	$1.25
	HMC288MS8	3 Bit Digital	0.7 - 3.7	2 to 14	MSOP8	$1.35
	HMC290	2 Bit Digital	0.7 - 4.0	2 to 6	SOT26	$1.05
	HMC291	2 Bit Digital	0.7 - 4.0	4 to12	SOT26	$1.05
	HMC173MS8	VVA	0.8 - 2.0	0 to 30	MSOP8	$2.24
	HMC210MS8	VVA	1.5 - 2.3	0 to 40	MSOP8	$2.24
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AROUND THE CIRCUIT

Enon Microwave Inc., a subsidiary of **Micronetics Wireless Inc.**, has won a three-year contract with a US government defense agency. The company will provide microwave switches for surveillance radar to be used around the world. The award is the company's second three-year contract for high powered microwave switches in less than a month.

Radio Waves, a **Smiths Interconnect** company, has received a substantial order from Aperto Networks for antenna solutions to be installed in its PacketWave Systems. The orders include integrated antennas for Aperto Networks' 5.8 GHz subscriber units, as well as base station antennas for Aperto Networks' 3.5 and 5.8 GHz systems. These orders will fulfill Aperto's forecasted antenna needs for the next 12 months.

PERSONNEL

MCE Technologies Inc. has appointed **Craig Lindberg** president of MCE/Inmet Corp. Inmet is located adjacent to MCE Technologies' headquarters in Ann Arbor, MI, and is one of seven MCE Technologies' business units located throughout North America, the UK and China. Prior to being named president, Lindberg was the VP of business development.

Peregrine Semiconductor has appointed **Jim Cable** to the role of president and CEO. He replaces Stavro E. Prodromou, who will continue to advise the company as director and chairman of the board. With over 20 years experience in semiconductor research, development and production, Cable is the former chief operating officer of Peregrine.


▲ Elliott R. Brown

MEMGen Corp. is hoping to boost its RF expertise with the appointment of **Elliott R. Brown** to its advisory board. Brown is currently a professor of electrical engineering at the University of California, Los Angeles (UCLA) and has over 25 years of experience in RF research and development, including RF micro-devices.

Elgar Electronics Corp. has appointed **John P. Mei** as vice president finance and CFO. Most recently, Mei was group controller for Acterna's multimedia group and previously was finance manager for GE's industrial systems group.

RF Micro Devices Inc. announced that **Daniel A. DiLeo** has been named a member of the board of directors. DiLeo was nominated and confirmed by the board as a new director at an August board meeting. He joins the board with experience in market development, engineering, manufacturing and sales and marketing for high growth, global semiconductor businesses.

CAP Wireless Inc. has named **Mark E. Lampenfield** executive VP. In this position, Lampenfield will oversee the operations sales and marketing functions of the company. He brings over 30 years of sales, marketing and operations experience to the company.


▲ Douglas J. Sober


▲ Derrick Neo

GIL Technologies Inc. has hired **Douglas J. Sober** as its new VP, quality assurance. Sober is expected to play an important role in furthering the company's reputation for high manufacturing standards. In related news, GIL has named **Derrick Neo** business director of Asia. Neo will manage the company's new Asian sales office in Shenzhen, China, and will direct all of the company's sales operations throughout the Asian rim.


▲ Michael St. Lawrence

Rogers Corp. has named **Michael St. Lawrence** to the newly created post of director of strategic marketing. St. Lawrence has been with the company for over 15 years and has held a variety of positions, most recently as research and development group manager for the high density products group. He will be responsible for helping chart Rogers' future course.

[Continued on page 54]

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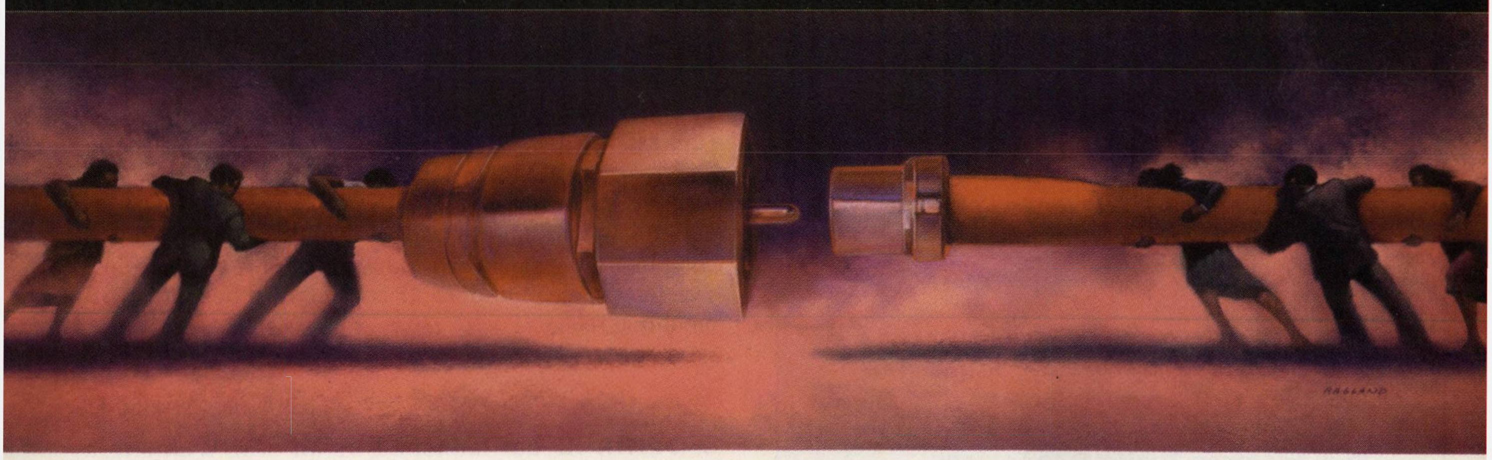

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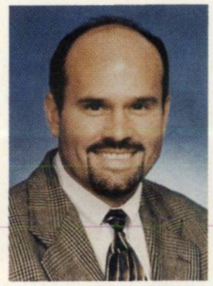
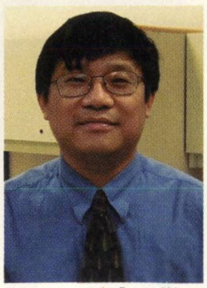
■ StratEdge has hired **Casey Krawiec** as a senior account manager. Krawiec spent seven years at Kyocera America where he held the positions of offshore sales manager and sales engineer. Prior to that he was a design engineer and engineering project manager for the Naval Surface Warfare Center.


▲ Casey Krawiec

■ Olympus Partnership Development Group (PDG) has selected **Peter Wang** to serve as optical product marketing manager. Wang has more than 12 years engineering experience in laser system design, fiber optical passive/active components, optical system design, and semiconductor related industries, and more than five years experience in project management.


▲ Peter Wang

■ Filtel Microwave Inc. has appointed **Perry Pawliuk** quality assurance manager. Pawliuk comes to the company with 27 years of experience having worked for companies such as Develcon Electronics, Advantech, Wavesat and Trackcom Systems. He will be responsible for establishing and maintaining the company's ISO 9001 quality management system.

■ Barry Industries has named **Rob Sinclair** sales and marketing manager. Sinclair will be responsible for launching the company's new LTCC product line and capabilities and will also manage the existing product lines in the New England region.

■ SyQor has added **Charlie Landino** as the process owner of technical services. Landino will be responsible for customer support for both pre-sales and post-sales activities at the company. Landino brings with him over 30 years of experience in the field of power electronics, having held senior hardware design engineering positions for a number of telecommunications firms.

REP APPOINTMENTS

■ **AMCOM Communications Inc.** has announced the appointment of **Northwest Rep-Tek**, with offices in Calgary and Alberta, Canada, to be the exclusive representatives for British Columbia, Alberta, Sasakatchewan and Manitoba, including Winnipeg.

■ **Filtel Microwave Inc.** has made a series of representative appointments. **Jose Delgado** of **CP Technologies** was named as the company's exclusive representative in Southern California and **RF Electronic Sales** has been appointed as the exclusive representative for New York state and northern New Jersey. Internationally, **Lancer Communication** has been appointed as the exclusive representative for Taiwan, **EG Components** for Scandinavia, and **SM Engineering** for Korea.

■ **Matsushita Electric Works UK Ltd.** has appointed **Link Microtek** as UK distributor for the company's range of microwave switching devices. Primarily designed for applications in measurement instruments, automatic test equipment, mobile phone base stations and broadcast equipment, the range of products includes a comprehensive lineup of 50 and 75 Ω coaxial switches and PCB-mounting relays.

WEB SITE

■ **OPHIR RF** has launched its all new Web site at www.ophirrf.com. It now includes product search capability, a simple to use 'request for quotation' facility and a local company representative identifier. Designed for ease of use, the site allows visitors to request literature, ask for help from an applications engineer and receive new product alerts.

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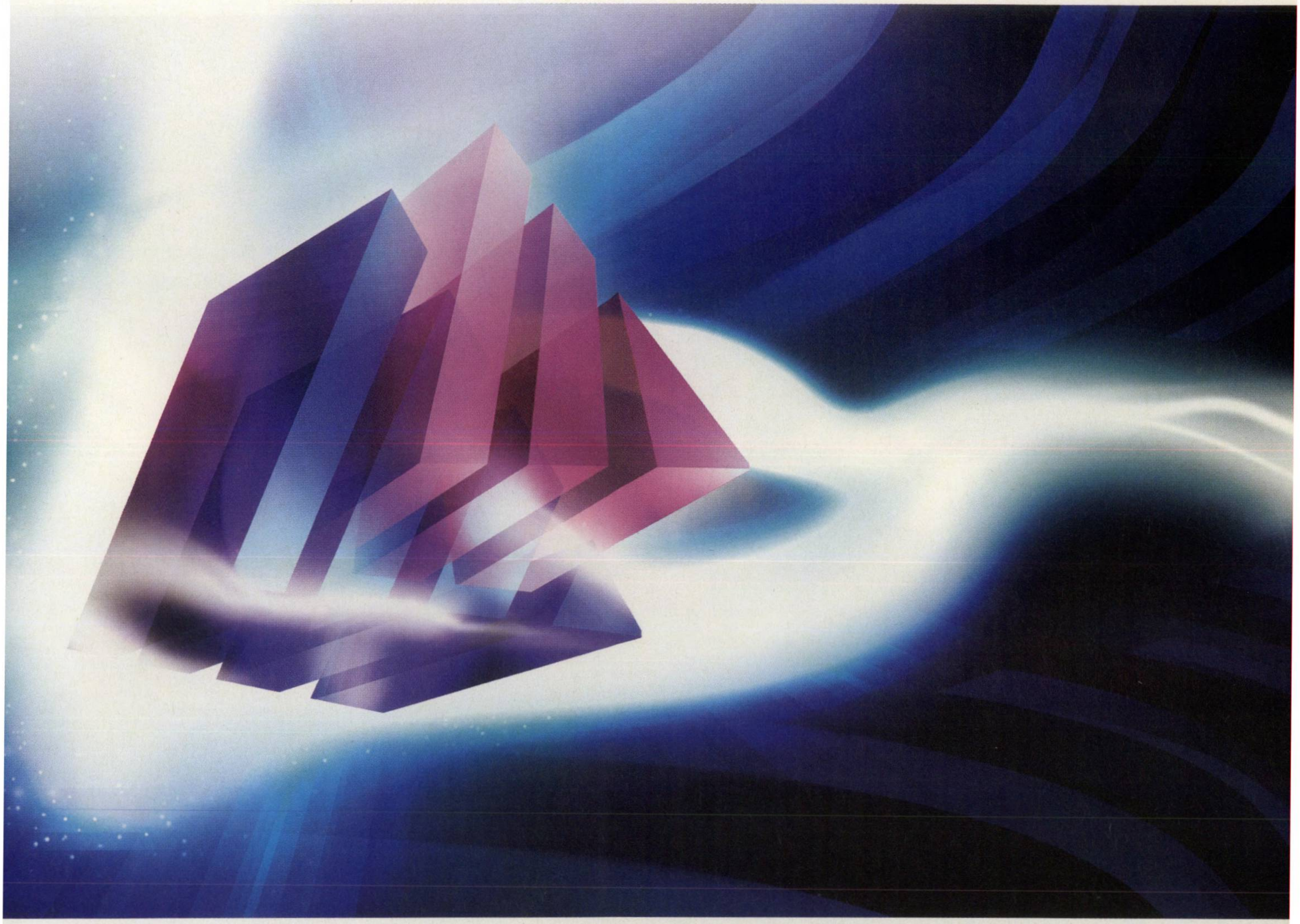

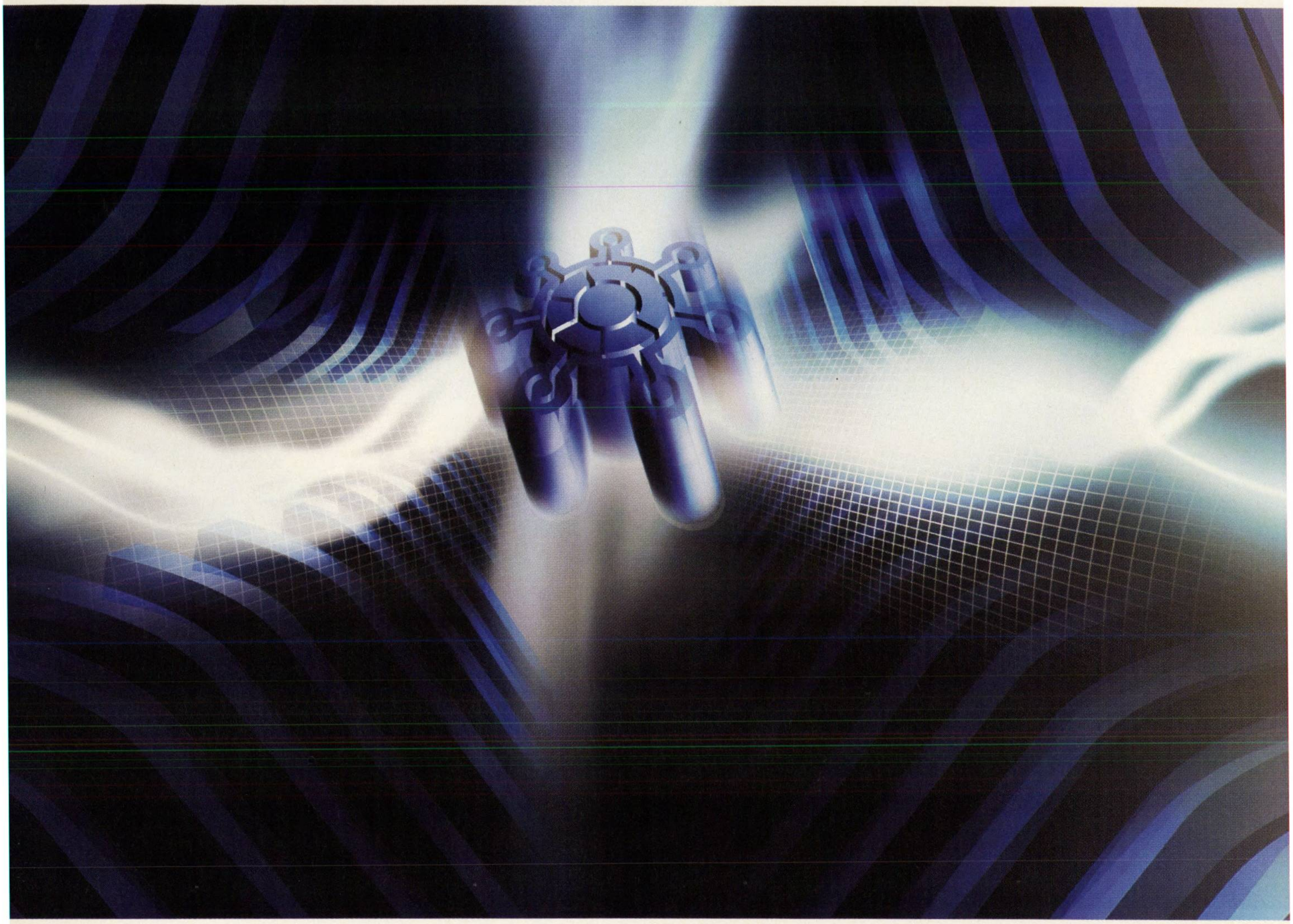

BALANCING THE BUDGETS: THE IMPORTANCE OF DEDICATED SOFTWARE FOR SYSTEM DESIGN

The problem of exporting system level design models to the component engineering disciplines, and vice versa, is addressed. For instance, common office spreadsheet programs are often used by system engineers to model power budgets, filter losses and signal-to-noise ratios. To offer a new solution, an example is given of a Tx/Rx channel in a WCDMA base station. Also, a worked example of a feed-forward amplifier is given along with sampled data.

The International Telecommunication Union (ITU) standard Wideband Code-Division Multiple Access (WCDMA) was derived from code-division multiple access (CDMA), and is officially known today as IMT-2000 direct spread. It is a third-generation (3G) mobile wireless technology offering much higher data speeds to mobile and portable wireless devices than commonly offered in today's market. WCDMA can support mobile/portable voice, images, data and video communications at up to 2 Mbps (local area access) or 384 kbps (wide area access). The input signals are digitized and transmitted in coded, spread-spectrum mode over a broad range of frequencies.

When 3G networks are fully deployed across Europe, mobile Internet access will provide users with many types of new services and content over a wide range of different networks and user devices. These services, many of which have yet to be developed, will be capable of accommodating user's personal preferences, location and circumstances at a particular moment in time.

However, before the 3G dreams can be fully realized, the industry must wait for the public to decide whether they can afford the extra services the bandwidth will provide. Clearly, that will be dependent upon the content available, and simply browsing the Web faster will not be sufficient to generate public interest. While the content debate continues, the task of installing a viable WCDMA infrastructure that will enable 3G connectivity has to be addressed.

To achieve this, the engineering challenges are formidable and, over the past decade, the role of the RF systems engineer has made a notable change. Up until the 1990s most system engineers dealt with requirements and functions by using a black box approach to design. Today, however, the demands being placed on the designers to reduce power, while at the same time improving perfor-

[Continued on page 60]

CHARLES BLACKWOOD
Ansoft Corp.
Twickenham, UK


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mance, means that there is also a need to know what goes on in every black box. Only then can the various parameters be tuned and optimized to allow the overall system requirements to be met.

Design software has come a long way, and there are software tools available that allow the system engineer to go into far more detail, have access to accurate simulations, and handle far larger communications chains than ever before. Unfortunately, there is a price to pay, and that price only becomes apparent when a team of engineers tries to pass the various aspects of a system design between groups of design specialists.

POWER BUDGETS

Computer modelling plays a vital role in all system design but, surprisingly, one of the most widely utilized tools today is the Microsoft spreadsheet program — Excel. It is an excellent tool for keeping track of parameter budgets, but the fundamental drawback is that it is not an engineering tool.

For instance, take the example of a transmit/receive (Tx/Rx) channel in a WCDMA base station, where the system designer needs to define the RF power, filter characteristics, signal-to-noise ratios, local oscillator stability and response of a phase-locked loop. A spreadsheet allows the designer to build up a simple description of the Tx/Rx path, with each module, such as filter, amplifier and modulator, being described by one spreadsheet cell, allowing access to a basic budget analysis, highlighting points of high noise levels, or highlighting a need for amplification.

Many small RF design teams of less than 10 to 15 engineers use the spreadsheet approach. Undoubtedly, this method does the job, provided the system engineers using this approach have an in-depth knowledge of circuit design, in addition to their expertise in systems engineering. More commonly, however, the number one reason for using Excel is because it is perceived as being free, as it comes bundled with the office PC in many cases. The cost of ownership of this type of software tool is very low, but for complex system work it often requires a disproportionate amount of time and effort to make it productive.

While Office spreadsheets are fine for tracking discrete values, they are poor, or even useless, for handling S-parameters or the type of equations used by RF systems engineers. So, if spreadsheets are not really suitable, what is? Some commercially available RF systems software packages have a built-in library of modulation schemes, RF parts such as filters and mixers, as well as functional parts, including digital logic gates, probes and switches.

These packages allow the designer to have greater control of the characteristics of each part, and to make them as realistic as possible. By following the analysis reports available from such tools, bit error rate (BER) plots, eye diagrams and spectral plots can ensure that a system will be designed for optimum performance.

[Continued on page 62]

Some pure mathematical tools currently available will allow an RF engineer to model functional blocks in the transmit/receive channel of a WCDMA base station, for example, but they can also allow the creation of impossible circuits. For instance, zero impedance transmission lines, or unfeasibly large capacitors, can be required. Therefore, if meaningful results are to be generated, the user must be both a component designer and a system designer.

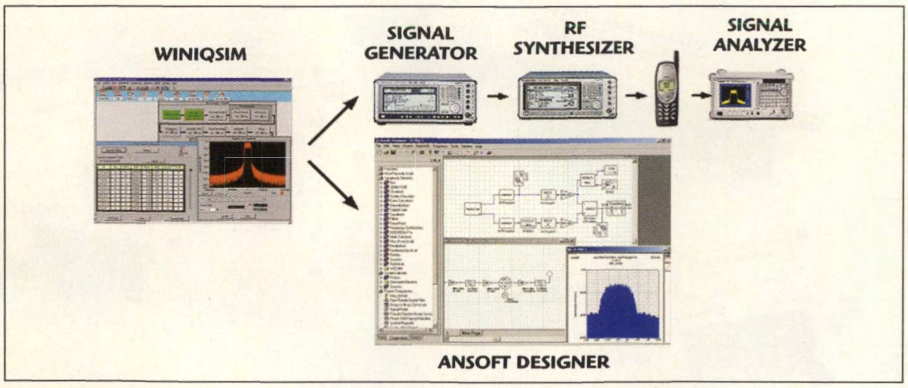
Plugging tools together from multiple vendors always sounds easy, but in practice this task often proves to be very difficult and time consuming. Multivendors are often chosen because companies believe they can save money by "shopping around" and purchasing what they need. It is only once the tools have been delivered that the real meaning of seamless integration becomes apparent.

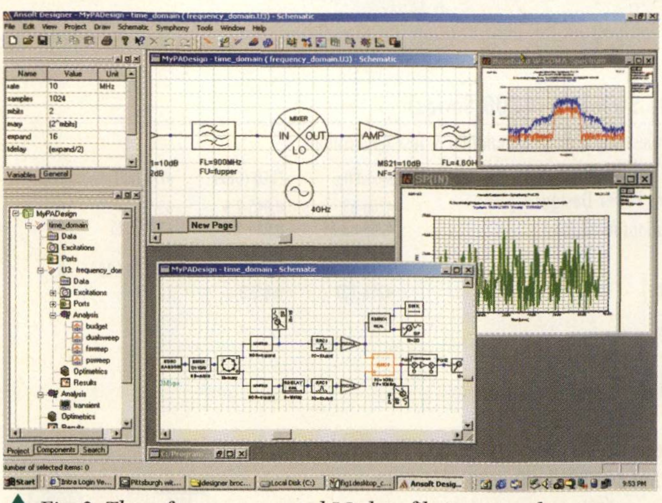
MIXED-MODE SIMULATION

Seamless integration — that well-used marketing phrase — usually means: "Spend only a few weeks getting your software engineers to write conversion scripts in order to make all the tools work together!" The problems usually arise when RF engineers, who use and work with S-parameters, need to export their system level designs to digital component engineers. The component engineers do not generally have a problem with S-parameters, but tend to use SPICE models in their design environment.

The challenge for the tool vendors has been to develop the required interfaces that will enable the export of S-parameters into SPICE libraries, and vice versa, to enable designs to be passed between system level and component design teams.

Fig. 1 By using Ansoft Designer and WinIQSim, the data capture flow uses the same waveform as the actual hardware. ▼





▲ *Fig. 2 The software encrypted IQ data file is accessed as a source component in the mixed-mode simulation.*

To illustrate how this can be achieved consider the Ansoft Designer interface in which the mixed-mode simulation environment utilizes I and Q baseband signals that are generated with the Rohde & Schwarz WinIQSim software package. The capability offered enables engineers to apply RF signals from many of today's communication standards, including 3GPP WCDMA, TD-SCDMA, Bluetooth, Hiperlan/2, IEEE802.11b and CDMA2000 directly to their circuit and system simulations. Having the simulation engine within the environment's interface allows engineers to use real-world parameters within the function block library, and therefore, unlike a pure mathematical tool, any attempt to use a circuit value outside of a usable range will be restricted.

Users can generate arbitrary waveforms by modifying the value of parameters and impairments affecting the waveform. ***Figure 1*** shows a typical data capture flow. These user-specified waveforms may then be stored as signal source library components for use in driving the mixed-mode simulation environment (see ***Figure 2***).

In this way RF designers and system architects can simulate communication systems under the same conditions used in hardware testing and product development. Through access to the latest communication standards, designers have the advantage of qualifying component and system specifications under actual operating conditions much earlier in the design process.

The software permits the calculation of I and Q baseband signals, which are in turn used with signal generators to excite communication transceiver radios. Having access to a library of built-in RF system level components is a major advantage. Amplifiers can be included that model both the linear region and intermodulation distortion (IMD) created by the nonlinear transfer characteristic, voltage-controlled oscillators that include phase noise characteristics, mixer models that incorporate spurious products, modulators and demodulators including IQ, AM, FM and PM, black boxes that support importing of S, Y and Z parameters, noise and nonlinear measured or software generated data can be included.

In mixed-mode simulation, component specification can be as simple or as complex as the user demands, with components initially being specified with a few key parameters. This enables engineers to quickly build a system model for "what if" analyses. As the design progresses and more data become available, it can be included to refine the simulation. For example, an amplifier can be modelled several ways — gain, P1dB and IP3, frequency, temperature and/or power dependent S-parameters, and AM-AM and AM-PM data. In addition, imported data can be used to model inputs, arbitrary modulation methods and noise sources.

FEED-FORWARD EXAMPLE

In the example given below, a feed-forward amplifier with a WCDMA input is discussed. First, the software imple-

[Continued on page 64]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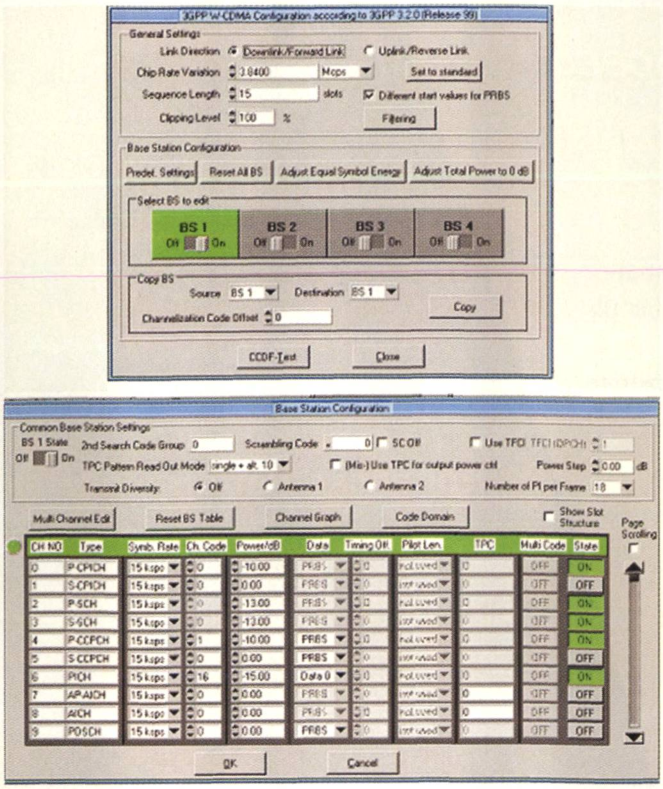

▲ *Fig. 3 Up to four mobile stations, each operating in a different mode, can be simulated.*

ments the physical layer of WCDMA frequency division duplexing (FDD) with all physical channels defined by 3GPP. Parameters such as symbol rate, channelization code, channel power, channel data, channel state, and transmit power control bits and scrambling codes are independently settable for each code channel base station or mobile station.

A 3GPP mobile station can operate in three different modes:

- Physical Random Access Channel (PRACH) only, where the mobile station generates a single PRACH, which is used when a call is set up from the mobile to the base station.
- Physical Common Packet Channel (PCPCH) only, where the mobile station generates a single PCPCH. This channel is used for the transmission of packet-oriented services (such as SMS).
- Dedicated physical control and data channels (DPCCH + DPDCH) is the standard mode for speech and data transmission.

Then the mobile station generates a control channel (DPCCH) and up to six data channels (DPDCH), depending on the required data rate. Taking the example of WinIQSim and Ansoft Designer, the signal and response of up to four mobile stations can be simulated, each operating in one of the modes — PRACH only, PCPCH only, DPCCH and DPDCH (see ***Figure 3***).

[Continued on page 66]

While base station configurations are more complex than mobile station modes, not all downlink channels are essential for simulation or testing. Using typical configurations will provide adequate simulation conditions for understanding device or system behavior. A basic base station signal contains the ever-present control channels required for synchronization and a varying number of dedicated physical channels (DPCH). The number of DPCHs and their symbol rates depend on the number of connections and the required data rates. If high data rates are required, several DPCHs can be linked together and transmitted to one and the same mobile (multicode state). For base station simulations, the software generates signals of up to 512 data channels distributed to a maximum of four base stations with 128 code channels each.

STATISTICAL PROPERTIES OF 3GPP SIGNALS

The superposition of many code channels may lead to very high crest factors in the WCDMA sum signal. Because crest factor influences the linearity requirements of system components, particularly the power amplifier, it is important to simulate designs with statistically correct signals. The signal's statistical properties — expressed in the complementary cumulative distribution function (CCDF) — generally depend on four parameters: the number of code channels; the selection of channelization codes; the correlation of the user data; and the timing offsets between code channels.

The mixed-mode simulation method allows users to fully investigate intermodulation distortion or spectral regrowth due to the nonlinearity of a power amplifier at the circuit level or behavioral (system) level. Also, circuit envelope analysis provides the capability to look at spectral waveforms, eye diagrams, constellation plots or adjacent channel power ratio (ACPR) as a function of any swept parameter, including power, bias and impedance tuning. Signal-processing probes allow direct plotting of crest factor, error vector magnitude (EVM) or CCDF.

MULTICARRIER WCDMA AND MULTISTANDARD SIGNALS

The capability of the software to support generating and simulating multicarrier and multistandard signals is critical for tests on amplifiers and other components with multicarrier signals. The single WCDMA carriers are generated separately with the 3GPP WCDMA system and then mixed in the multicarrier mixed signal system.

This function is not limited to multicarrier WCDMA. Mixed systems (that is WCDMA carriers combined with GSM or other systems) are also possible. These features can be used to simulate transmitters of multisystem base stations. Such units will become more important as they provide integration of new WCDMA technologies on existing networks.

CONCLUSION

If RF engineers and system architects are to achieve first-pass designs in an evolving, time critical market such as WCDMA, then it is no longer practical to rely on "home brewed" tool chains from several software vendors. This is because the time needed to integrate all the tools together, and gain sufficient levels of confidence will be too long. Only a design environment capable of fully modelling high frequency component behavior offers the greatest likelihood of success.

The partnership between an EDA vendor with its designer software environment, and a test and measurement company with its software combines accurate signal generation with RF modelling and simulation. It provides a fusion of real-world data with advanced electromagnetic analysis for designing at component, circuit and system levels.

While it is common knowledge that a screwdriver can be used as a lever, or a chisel, it's not the right tool for the job. Similarly, it is rather like using a spreadsheet to design 3G communications. Given enough time and effort a spreadsheet will serve a purpose, but if your competitor has a cost-effective tool that is right for the job, and your team does not, do not expect to win the time to market race. ■

Charles Blackwood *earned his BSc degree in communication engineering from the University of Kent at Canterbury. He is currently the Northern European sales manager for Ansoft Corp. Europe and is based in the UK. Before working for Ansoft, he worked for several telecommunication companies, including Siemens and Matra.*

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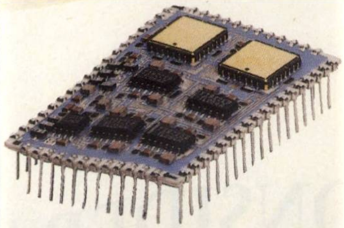

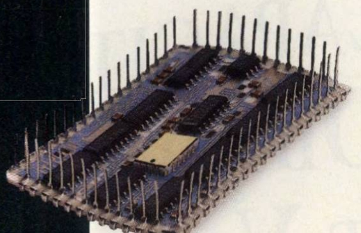

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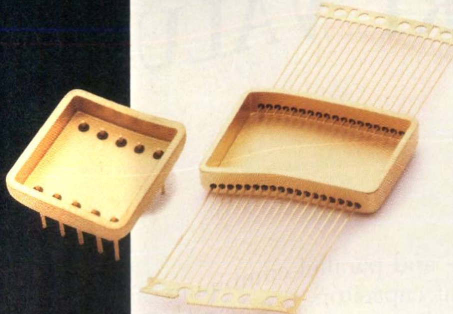

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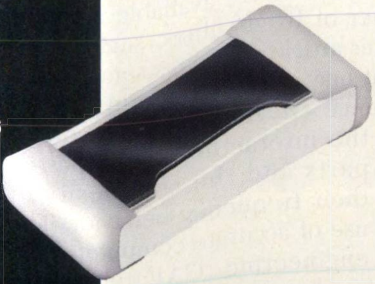

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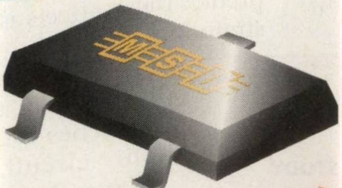

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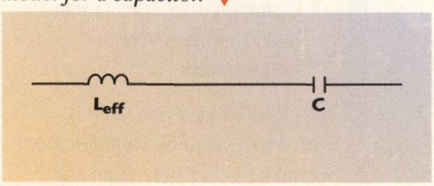
CONSIDERATIONS IN CAPACITOR-PAIRING TO OBTAIN NONSTANDARD PART VALUES

Series- and parallel-combining of surface-mount capacitors is a standard approach to realize semi-arbitrary capacitance when designing a printed circuit board layout. This pairing of components is often necessary as surface-mount parts are available in only a limited number of fixed values, such as 0.5, 1.0, 2.0 pF. This article addresses some potential pitfalls that may arise in the form of unexpected resonance effects, brought about by the interaction of the paired parts and the complexity of their frequency behavior. The use of accurate computer-aided engineering (CAE) models to predict these effects is demonstrated.

Fig. 1 Simple LC circuit model for a capacitor.



For the purposes of this analysis, a simplified series L-C equivalent circuit model (***Figure 1***) was used to extract an effective series inductance for each part, based on the measured first resonance:

$$\omega_0 = \frac{1}{\sqrt{LC}} \qquad (1)$$

The characteristics of the capacitors used in the study are listed in ***Table 1***, as measured on 14 mil-thick FR4 using series, two-port, microstrip, test fixtures. None of the capacitors exhibited secondary, higher order resonances below 15 GHz. The series L-C model is often used to represent capacitor performance, but it will become evident in the examples presented that models with greater physical representation are required for accurate performance prediction, even in simple, two-capacitor arrangements.

TABLE I

CAPACITORS USED IN THIS STUDY (0402 BODY STYLE)

Capacitance (pF)	1st Series Resonance (GHz)	L_{eff} (nH)
0.5	14.5	0.24
0.7	11.2	0.29
1.2	8.7	0.27
1.5	8.0	0.26
3.3	5.1	0.30
6.8	3.2	0.36

The three multiple-capacitor configurations considered here are the series-parallel (***Figure 2***), shunt-parallel

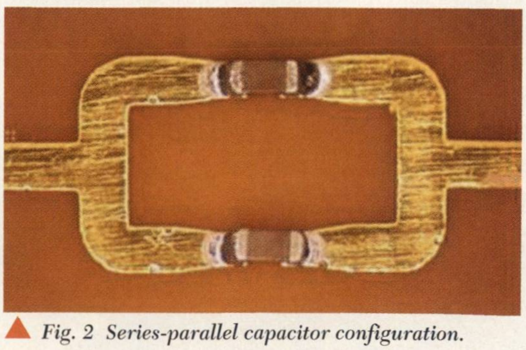


Fig. 2 Series-parallel capacitor configuration.

[Continued on page 70]

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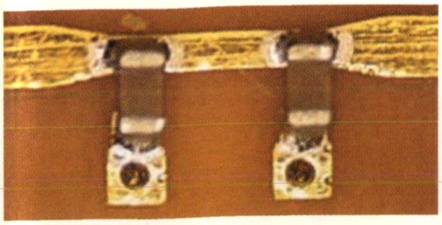
(***Figure 3***) and shunt-series arrangements (***Figure 4***). Measured

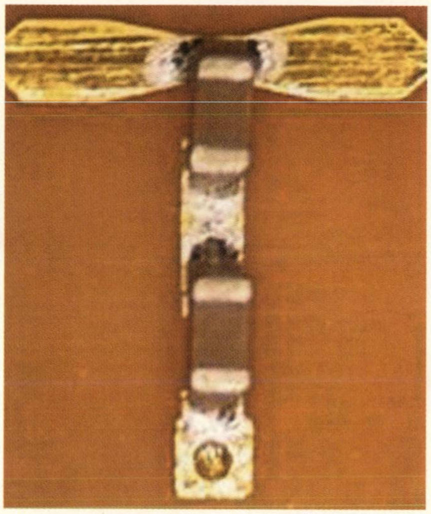


▲ *Fig. 3 Shunt-parallel capacitor configuration.*

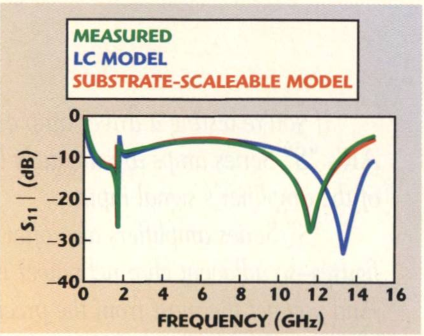
results for these configurations are analyzed below. In each case, comparisons are made with accurate, substrate-scaleable capacitor models.[1] The schematics used to generate the simulated results included a complete representation of the interconnect discontinuities.

SERIES-PARALLEL RESULTS

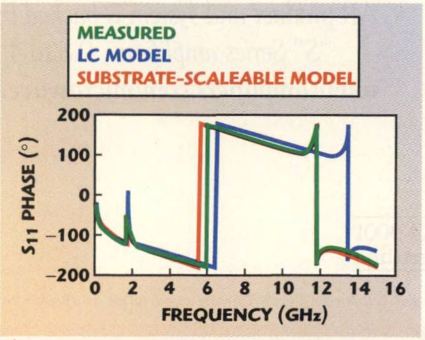
A comparison of the S_{11} magnitude and phase for a series-parallel combination of a 3.3 and 6.8 pF capacitor is shown in ***Figures 5*** and ***6***, respectively. In these graphs, the measured data is compared against the simulated results obtained using accurate, substrate-scaleable equivalent circuit models, and those from the simple L-C models. It is apparent that when paired together the series resonance of neither individual capacitor — which would manifest as a low S_{11} — is observed (5.1 and 3.2 GHz for the 3.3 and 6.8 pF capaci-



▲ *Fig. 4 Shunt-series capacitor configuration.*



▲ *Fig. 5 S_{11} magnitude for a 3.3 and 6.8 pF capacitor in the series-parallel configuration.*



▲ *Fig. 6 S_{11} phase for a 3.3 and 6.8 pF capacitor in the series-parallel configuration.*

[Continued on page 72]

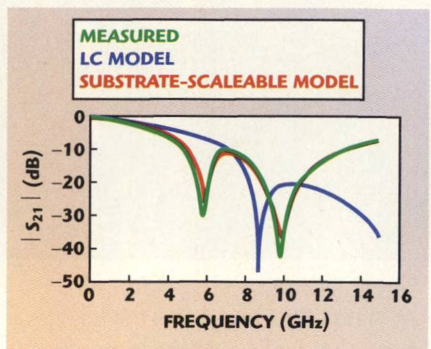
tors, respectively, as listed in the table). Furthermore, if one applies the simple L-C circuit model for each capacitor and combines the two in parallel, an effective capacitance of 10.1 pF is obtained with a net inductance of 0.17 nH. From these values, a parallel resonance (minimum return loss or large S_{11}) at 3.84 GHz is predicted, which is not evident in the data. On the contrary, a sharp series-parallel resonant frequency pair appears near 1.8 GHz, along with a strong series resonance at 11.8 GHz.

As noted above, the circuit schematics for the substrate-scaleable and simple L-C models both accounted for all discontinuity effects in the layout. The low frequency correlation between the simulated and the measurement data argues the importance of proper interconnect modeling, as the resonant effects that would be expected from even the simple L-C model analysis are not observed via simulation. The high frequency behavior points out the limitations of the simple L-C model, however, even when embedded within proper interconnect elements.

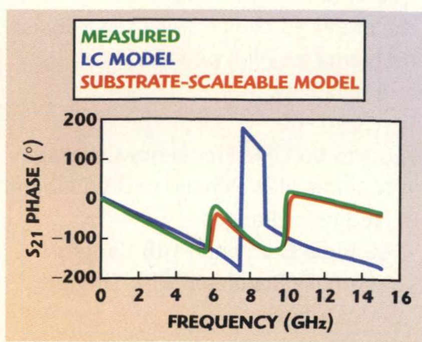
SHUNT-PARALLEL RESULTS

Measured data and simulation results of S_{21} for a shunt-parallel combination of a 0.5 and 1.5 pF capacitor are given in ***Figures 7*** and ***8***. In the shunt configuration, the fundamental (series) resonance of the capacitors results in low impedance to ground and thus high insertion loss. The measured data show such effects at approximately 5.8 and 9.8 GHz, as do the simulated results using the substrate-scaleable model.

As an approximate analysis of this layout, each capacitor can be considered independently. Assuming the nominal capacitance value for each part, and attributing the low and high resonances individually to the 1.5 and 0.5 pF capacitor, respectively, an effective inductance of approximately 0.5 nH is extracted. Using the data in the table, it may be determined that each ground-via contributes an additional 0.25 nH inductance to the circuit.



▲ *Fig. 7 S_{21} magnitude for a 0.5 and 1.5 pF capacitor in the shunt-parallel configuration.*



▲ *Fig. 8 S_{21} phase for a 0.5 and 1.5 pF capacitor in the shunt-parallel configuration.*

[Continued on page 74]

Bipolar Amplifiers

Standard Options:

- *Connector Options: SMA, BNC, Type N*
- *Integrated Limiters*
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OPERATING FREQUENCY (MHz)	MODEL NUMBER	GAIN (dB, Min.)	GAIN FLATNESS (±dB, Max.)	VSWR IN/OUT	NOISE FIGURE (dB, Max.) NF_1	NF_0	NF_2	OUTPUT POWER (dBm, Min.)
0.001 – 500	AU-1534	30	0.5	2.0:1	1.3	1.4	1.5	+8
0.01 – 200	AU-1442	35	0.5	2.0:1	1.2	1.2	1.2	+5
0.01 – 200	AU-1447	56	0.5	2.0:1	1.2	1.2	1.2	+12
0.01 – 250	AU-1559	11	0.5	2.0:1	4.2	4.2	4.2	+16
0.01 – 400	AU-1565	54	0.75	2.0:1	1.2	1.2	1.3	+14
0.01 – 500	AU-1310	30	0.5	2.0:1	1.3	1.4	1.5	+8
0.01 – 1000	AU-1402	18	1.0	2.0:1	6.0	5.0	5.0	+16
0.01 – 1000	AM-1300	27	0.75	2.0:1	1.4	1.6	1.8	+8
0.01 – 1000	AM-1431	35	0.75	2.0:1	1.4	1.6	1.8	+8
0.1 – 2000	AM-1364	9	1.5	2.0:1	6.0	6.0	6.0	+10
1 – 200	AU-1464	35	0.5	2.0:1	1.2	1.2	1.2	+6
1 – 400	AU-1421	24	0.5	2.0:1	2.4	2.4	3.1	+17
1 – 500	AU-2A-0150	30	0.5	2.0:1	1.3	1.4	1.5	+8
1 – 500	AU-3A-0150	44	0.5	2.0:1	1.3	1.4	1.5	+10
1 – 500	AU-4A-0150	60	0.75	2.0:1	1.3	1.4	1.5	+10
1 – 1000	AM-2A-000110	26	0.75	2.0:1	1.4	1.6	1.8	+6
1 – 1000	AM-3A-000110	35	0.75	2.0:1	1.4	1.6	1.8	+8
5 – 200	AUP-1568	26	0.75	2.0:1	5.0	4.5	4.5	+28
5 – 300	AUP-1495	11	0.75	2.0:1	15	9.0	9.0	+28
5 – 300	AUP-1496	23	0.75	2.0:1	8.0	7.0	7.0	+28
5 – 300	AU-1021	24	0.5	2.0:1	2.7	2.8	2.9	+20
5 – 300	AUP-1479	36	1.0	2.0:1	2.5	2.7	2.9	+28
5 – 1000	AM-1475	36	0.75	2.0:1	1.4	1.6	1.8	+15
5 – 2000	AM-1573	18	1.5	2.0:1	4.0	4.0	4.0	+21
5 – 2000	AM-1590	36	2.5	2.0:1	3.8	3.8	3.8	+20
5 – 2000	AM-1591	48	2.5	2.0:1	3.8	3.8	3.8	+20
100 – 1000	AM-1412	35	0.75	2.0:1	1.4	1.6	1.8	+14
100 – 2500	AM-1585	26	2.0	2.0:1	3.6	3.6	3.6	+20
200 – 2000	AM-1569	20	1.5	2.2:1	4.2	4.3	4.6	+14
1000 – 2000	AM-1477	37	1.0	2.0:1	1.8	2.1	2.4	+15

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Thus, there is some consistency in treating the capacitors separately, since the analysis on each resonance yields approximately the same via inductance.

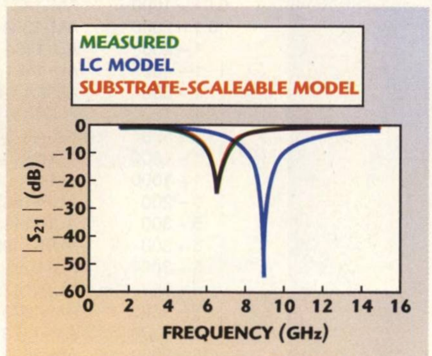
From a different perspective, one might consider first each capacitor in isolation (with its nominal capacitance and an overall inductance of 0.5 nH, after accounting for the via) and then evaluate the parallel combination of the two. This calculation leads to a net capacitance of 2.0 pF and an effective series inductance of 0.25 nH. The corresponding resonant frequency is approximately 7.1 GHz, an effect that would lead to high insertion loss and one not noticeably present in the graphs. The simple L-C representation does, however, falsely predict a single resonance behavior (near 8.8 GHz).

SHUNT-SERIES RESULTS

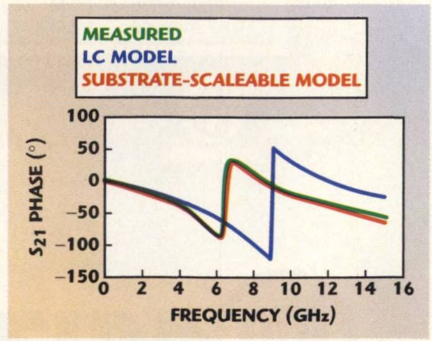
Finally, consider the shunt-series combination of a 0.7 and a 1.2 pF capacitor. Measured data and simulation results of S_{21} for this case are shown in ***Figures 9*** and ***10***. As before, a preliminary analysis based on the simple series L-C model is used for each capacitor.

From the data in the table, along with the 0.25 nH via inductance determined above, the total inductance of the combined circuit is found to be 0.81 nH. The effective capacitance of the series combination is 0.44 pF. This analysis leads to the prediction of a resonance at 8.4 GHz, whereas the measured results show it to occur at 6.5 GHz. It might be argued that the interconnect-line between the two capacitors introduces an additional inductance, tending to lower the resonant frequency. While this is true to some extent, the overall inductance would have to be greater than 1.3 nH in order for the frequency to drop to 6.5 GHz (assuming the net capacitance remained at 0.44 pF). It is unreasonable to assume that the short interconnect-line, which is less than 1 mm in length, adds more than 0.5 nH to the circuit.

As was found with the shunt-parallel circuit, simulated results using the simple L-C model, along with com-



▲ *Fig. 9 S_{21} magnitude for a 0.7 and 1.2 pF capacitor in the shunt-series configuration.*



▲ *Fig. 10 S_{21} phase for a 0.7 and 1.2 PF capacitor in the shunt-series configuration.*

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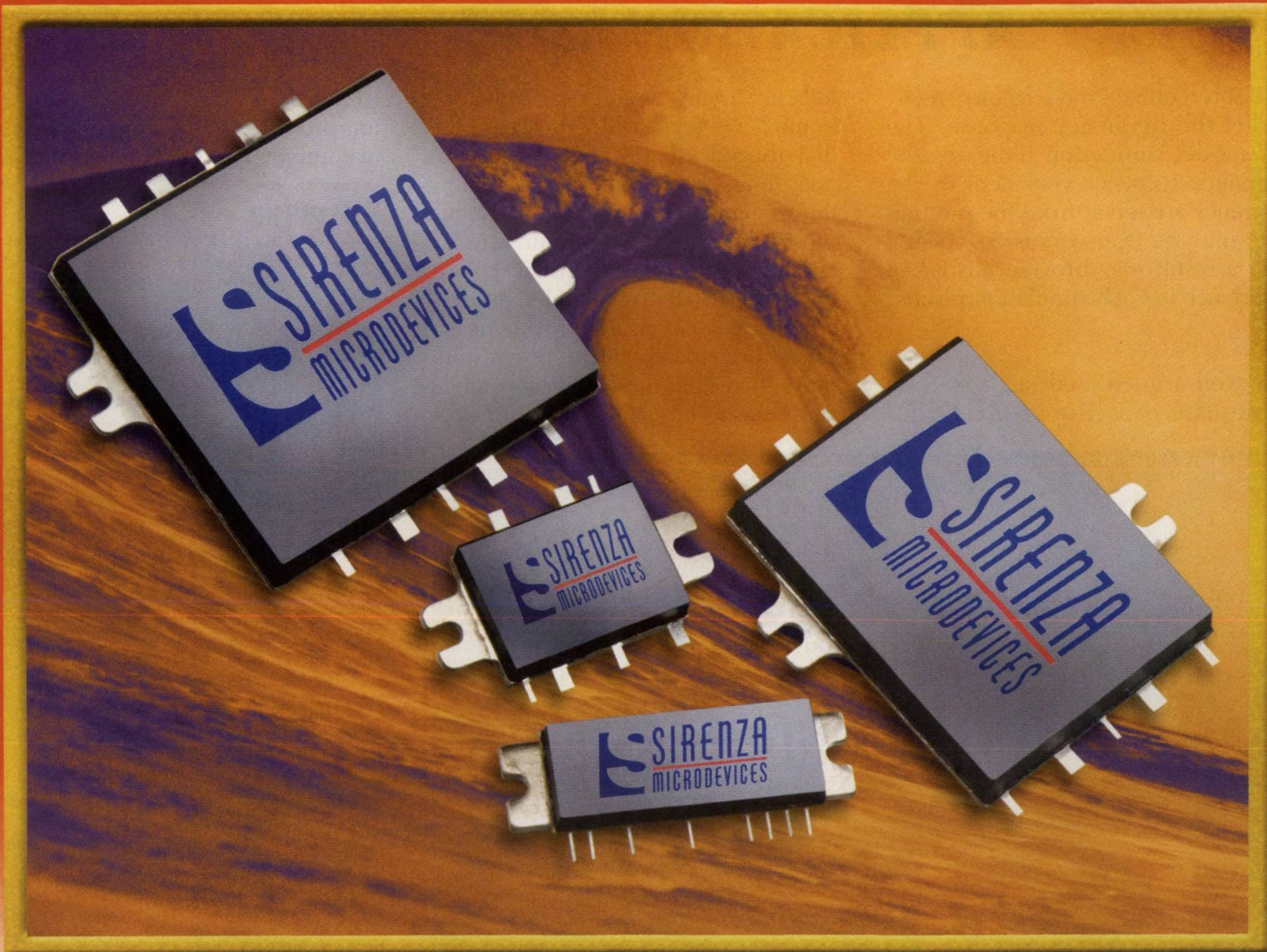

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For more information, visit us at www.sirenza.com • 800.764.6642 plete interconnect models, incorrectly predict the frequency response. The L-C model simulation indicates the resonance to occur very near to the frequency arrived at from the preliminary analysis (8.9 versus 8.4 GHz). Such a result may provide misplaced confidence in CAE simulation results.

CONCLUSION

A well-argued analysis, based on inadequate capacitor models, thus fails in varying degrees for the three examples described. In each case the distributed effects of the interconnect-lines play a role in determining the frequency response. However, the most significant factor required in accurately predicting the circuit behavior is the capacitor model. The substrate-scaleable models provide nearly exact comparisons to the test results. Simplified L-C models, which might closely emulate measurement data from a series two-port test fixture, often prove to be inaccurate when used in common circuit configurations. ■

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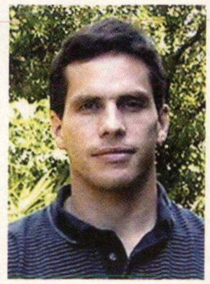
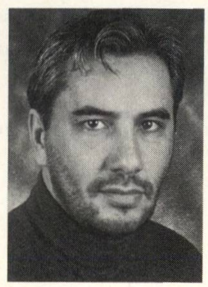


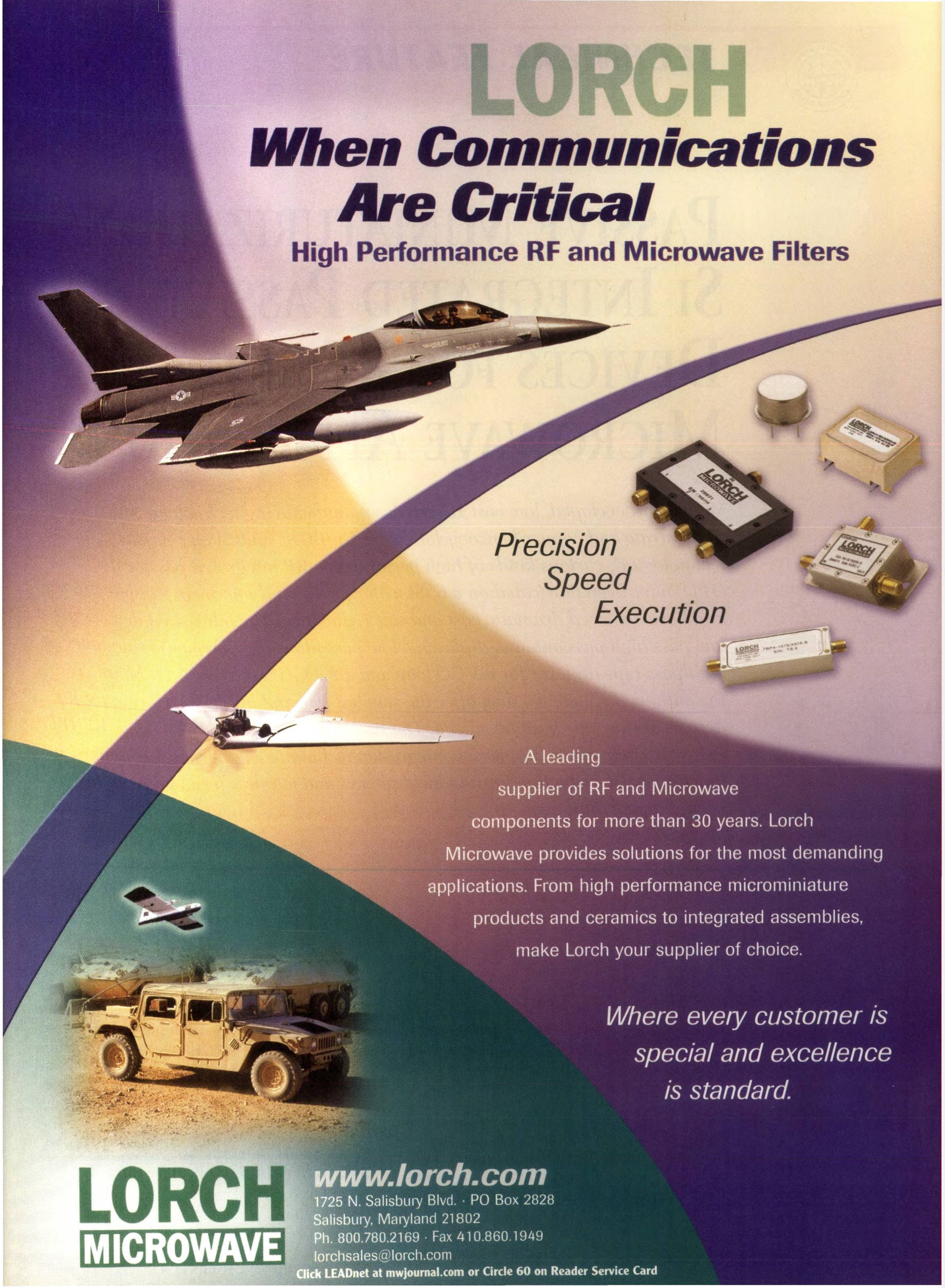
Tom Weller *received his BS, MS and PhD degrees in electrical engineering from the University of Michigan in 1988, 1991 and 1995, respectively. He was a member of the technical staff at Hughes Space & Communications Group from 1988 to 1990. He is currently an associate professor in the department of electrical engineering at the University of South Florida (USF), and a member of the Wireless and Microwave (WAMI) Center. He co-founded Modelithics Inc. with Larry Dunleavy in 2001.*

David Markell *received his BS degree in electrical engineering from the University of South Florida (USF) in 1998, where he is currently pursuing his MSEE degree. He has worked at USF as a lead engineer of a research project designed to generate microwave substrate-scaleable SMT chip capacitor, resistor and inductor models for the completion of a CAE library for a major manufacturer. He is currently a member of the technical staff at Modelithics Inc.*

John Capwell *received his BS degree in electrical engineering from the University of South Florida (USF) in 1999, where he is currently pursuing his MS degree in electrical engineering. He is currently a member of the technical staff at Modelithics Inc.*



Vicentiu (Vivi) Cojocaru *received his MSc degree from Politehnica University in Bucharest, Romania, and his PhD degree from University College Dublin, Ireland. Between 1996 and 2000, he worked as a research scientist and as a lecturer in the department of electrical and electronic engineering at University College Dublin. He currently works as an engineering manager for TDK Electronics Ireland Ltd.*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PASSIVE MINIATURIZATION: SI INTEGRATED PASSIVE DEVICES FOR RF AND MICROWAVE APPLICATIONS

With well-developed, low cost RF passive manufacturing technology (thick oxide Si substrate and copper/benzocyclobutene (Cu/BCB) multi-layer passive process technologies), various kinds of high performance RF integrated passive devices (IPD) have been fabricated on a 6" Si wafer for RF and microwave applications, and have achieved dramatic cost and size reductions. The fabricated devices are a low pass/high pass antenna diplexer, a low pass filter with harmonic resonance, a bandpass type diplexer for a VCO loop and a 2.4 GHz wireless LAN balun. To the authors' knowledge, they offer the smallest size and highest performance for devices of this type built on silicon. The size of the wafer-level, packaged, RF IPD is 1 to 1.5 mm². This RF passive integration technique will permit the 40 percent functional size reduction for handheld phones and wireless terminals that has been pursued until now.

When considering mobile electronic products, smaller is usually better. Most portable devices often have particularly stringent miniaturization requirements in order to meet market place expectations. The average passive device count in cellular phones has not dramatically dropped in recent years[1] and current handset designs are far from eliminating passive components. The size improvements, to date, have come from both smaller passive components and more efficient packaging of the components on the boards. The market demands, such as ultra-miniaturization and pricing pressure, make the integration of passive components on wafer to be the solution for the next generation of mobile terminals. There has been considerable research focused on low temperature and low cost co-fired ceramic (LTCC) technology,[2] but the research on thin film multi-chip module with deposited substrate (MCM-D) technology is not active. It is mainly focused on glass carrier substrates,[3] not on Si substrates because of many limiting factors such as substrate conduction.

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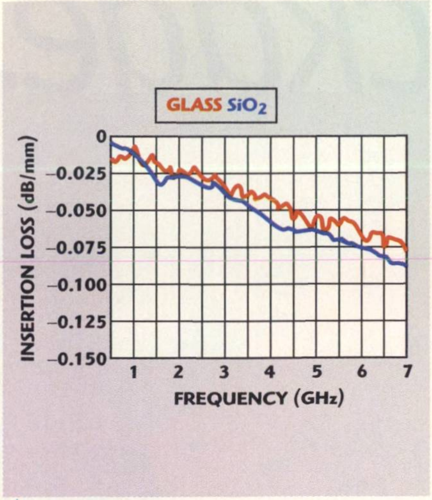
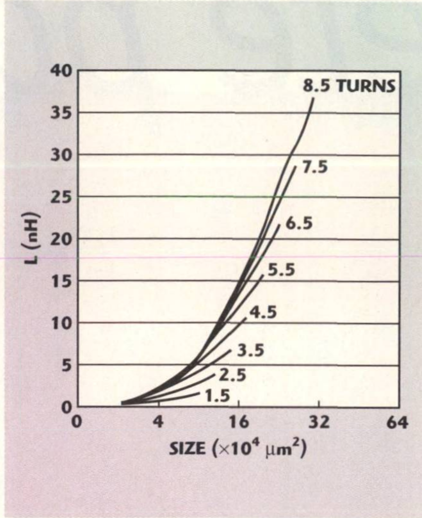


▲ *Fig. 1 Insertion loss of coplanar transmission lines on 25 μm thick oxide Si substrate and high quality glass substrate.*



▲ *Fig. 2 Inductance as a function of size and number of turns.*

A thick oxide technology has been developed to transform the Si wafer in a new RF substrate and synergistically combine it with a high quality passive Cu/BCB process.[4,5] Device examples and measured results for various RF IPDs are provided to demonstrate their superior performance and attractive size reduction. To reduce the size of the passive component to the maximum, PbSn eutectic solder balls are used to attach them directly on the board.

PASSIVE INTEGRATION ON SI SUBSTRATE

Silicon is the most stable and reliable semiconductor and has been used in many electronic applications. Unlike other semi-insulating substrates, silicon has had to rely on the properties of silicon dioxide for isolation. But a thin silicon dioxide layer cannot effectively isolate passive devices on Si substrates because of capacitive effects. The signal loss is also high because of the finite conductivity of the silicon substrate. These effects make it difficult to use Si technology for RF and microwave applications. In addition to the substrate loss, the conductor loss in the metals, due to the low conductivity of materials such as aluminum, limits the application of silicon to less than a few hundreds of megahertz.

A thick silicon dioxide layer of 25 μm on Si reduces the transmission losses by confining most of the electromagnetic field in the low loss dielectric layer beneath the conductors and not in the conducting silicon region. Also, the use of an 11 μm thick Cu metal layer and low dielectric constant BCB material makes it possible to implement low loss transmission lines over a broad frequency range. For example, a 50 Ω coplanar transmission line of 50 μm width and 15 μm gap on Si showed a total insertion loss of only 0.04 to 0.07 dB/mm at 5 GHz. This RF loss is very comparable to that of a high quality glass substrate in the low GHz region, though the latter shows slightly higher performance in the frequency region above 10 GHz. However, the glass substrate is very fragile and can break easily during the process. ***Figure 1*** shows a comparison of the insertion losses of coplanar waveguide (CPW) lines built on a thick oxide Si substrate and a high quality glass substrate. The results show that the specialized Si substrate and Cu process can be the optimal choice for low cost and high performance RF passive integration. For small-size passive devices, the area occupied by an inductor should be reduced, while maintaining satisfactory RF performance. To achieve this goal, a line width of 10 μm and a line spacing of 10 μm are mostly used for functional passive devices on Si. ***Figure 2*** shows the inductance value as a function of the number of turns and inductor size. Most of the planar spiral inductors

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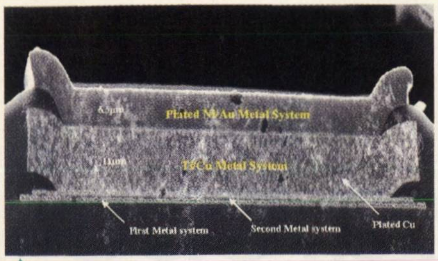
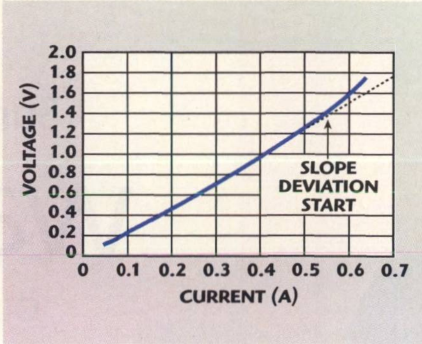


▲ *Fig. 3 Cross-sectional view of a bonding pad.*



▲ *Fig. 4 DC current rating test of the spiral inductor.*

occupy less than 0.5 × 0.5 mm area and are mostly concentrated in the 0.09 to 0.16 mm^2 region. They are much smaller than any other inductors implemented on PCB, LTCC or glass substrates.

SMALL-SIZE AND HIGH POWER RF INTEGRATED PASSIVE DEVICES

The on-chip inductor, a major technological challenge, is fabricated on a thick oxide Si substrate using a Cu metal system with 11 µm plated metal thickness and BCB interlayer material. The fabricated inductors have a spacing of 10 µm between turns and several different metal widths (10, 20, 30, 40, 50 µm) and show the maximum quality factor Q for the range of 40 to 60 µm, depending on the geometrical parameters and for inductances from 0.3 to 35.0 nH. ***Figure 3*** shows a cross-sectional view of the bonding pad structure composed of Ti/Cu – Ni/Au and BCB passivation. The Ni (6 µm) and Au (0.5 µm) layers are electroplated and used for the bonding of Au wire. They are not needed when the chip is bonded with solder balls during packaging. The spiral inductor with thick Cu metal has much more DC current and RF power capabilities than conventional Al- or Au-based inductors, even though the underlying interconnection between inner and outer sides of the inductor is relatively thin. The DC current test results for a spiral inductor with a 10 µm line width and line spacing is shown in ***Figure 4***. Every current step was sustained for 15 minutes and the current was increased to 640 mA. According to the data, some deviation was observed around 550 mA and the underlying metal started to burn at more than 600 mA. No physical or performance change was noticed after the inductor was submitted to a DC current of 500 mA for 168 hours. RF components for wireless applications are typically required to have an RF CW power handling capability of 3 W. The input and output matching circuits fabricated using the Cu process endured 5 W of RF power for 168 hours without change, which means the developed small-size Si RF IPD can be considered to have greater power handling capability than required by commercial specifications for handheld terminals.

In the following paragraphs, high quality Si integrated passive devices with high volumetric efficiency will be introduced and their measured performance provided. All the passive devices are wafer-level bonded to the boards, using eutectic solder balls for direct attachment, or are finished with Ni/Au plated pads for wire bond attachment on multi-chip modules. Typically, wire bondable passive devices are 20 to 30 percent smaller than solder ball bonded devices.

LUMPED L-C TYPE LOW PASS FILTER FOR PAM OR FEM APPLICATIONS

In modern communications systems, low pass filters are used to pass the wanted signal and eliminate or attenuate harmonics, mainly at the output stage of the power amplifier. With conventional low pass filter circuits, it is not easy to meet the insertion loss in the pass-band and the attenuation levels at the second- and third-harmonic frequencies required by strict commercial specifications. While achieving low insertion losses, enough harmonic attenuation was obtained by modifying the Chebyshev filter and the elliptic filter topologies to resonate at harmonic frequencies.

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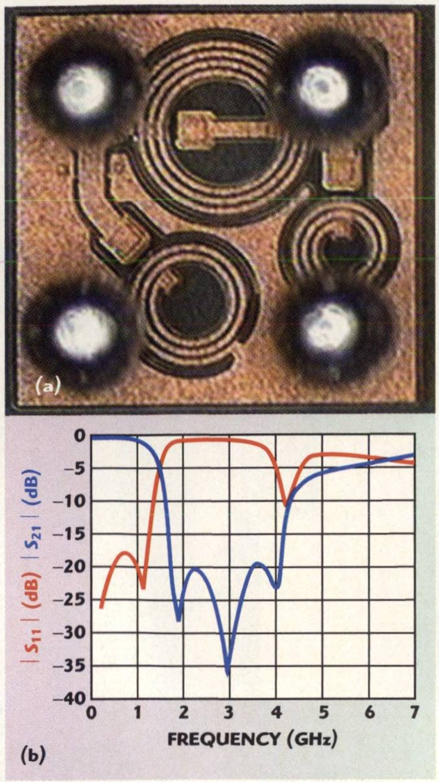
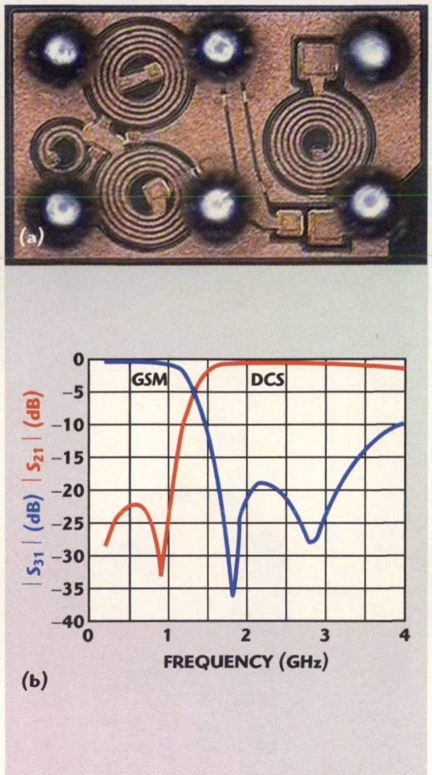


▲ *Fig. 5 Wafer-level packaged low pass filter (a) and its measured results (b).*



▲ *Fig. 6 GSM/DCS diplexer (a) and its measured results (b).*

Figure 5 shows a 900 MHz low pass filter for GSM applications and its measured results. As shown, the insertion loss is typically 0.45 dB and, in the case of a wafer-level packaged device, the second- and third-harmonic attenuation levels are greater than 25 dBc. The size of the filter is 1.0 mm^2 and it can be directly attached on the PC board by a conventional solder-reflow process. This ultra-miniaturization is sure to be an important step for a 40 percent size reduction in the functional integration of mobile phones.

LOW PASS/HIGH PASS LUMPED L-C DIPLEXER FOR ASM OR FEM APPLICATIONS

The diplexer handles two different carrier frequencies on the same signal path. It consists of a combination of low pass and high pass filters or two bandpass filters. The bandpass diplexer shows good attenuation characteristics but is very lossy, so it cannot be used in the front end of a mobile phone that does not allow for any excessive loss. ***Figure 6*** shows a fabricated GSM/DCS diplexer and its measured performance. To achieve the high band-rejection level and harmonic attenuation level, four inductors are used to obtain series and parallel resonances. The insertion losses of the wafer-level packaged diplexer are 0.45 dB at 900 MHz and 0.6 dB at 1800 MHz. The band-rejection levels are greater than 25 dBc at 900 MHz and 30 dBc at 1800 MHz. The third-harmonic attenuation is also more than 25 dBc. The size of the wafer diplexer is 1.5 mm^2, corresponding to a 60 percent reduction over a conventional device size.

BANDPASS TYPE LUMPED L-C DIPLEXER FOR VCO LOOP APPLICATIONS

This bandpass type diplexer, providing bandpass filtering and impedance matching, is designed to operate as an EGSM/PCS/DCS transmitter VCO sampling diplexer. It can be used for a two-port, low power transmitter VCO feedback to the synthesizer in GSM handsets. The device circuit consists of bandpass filter channels where suppression of second- and third-harmonic signals is required. Two parallel resonators are used in se-

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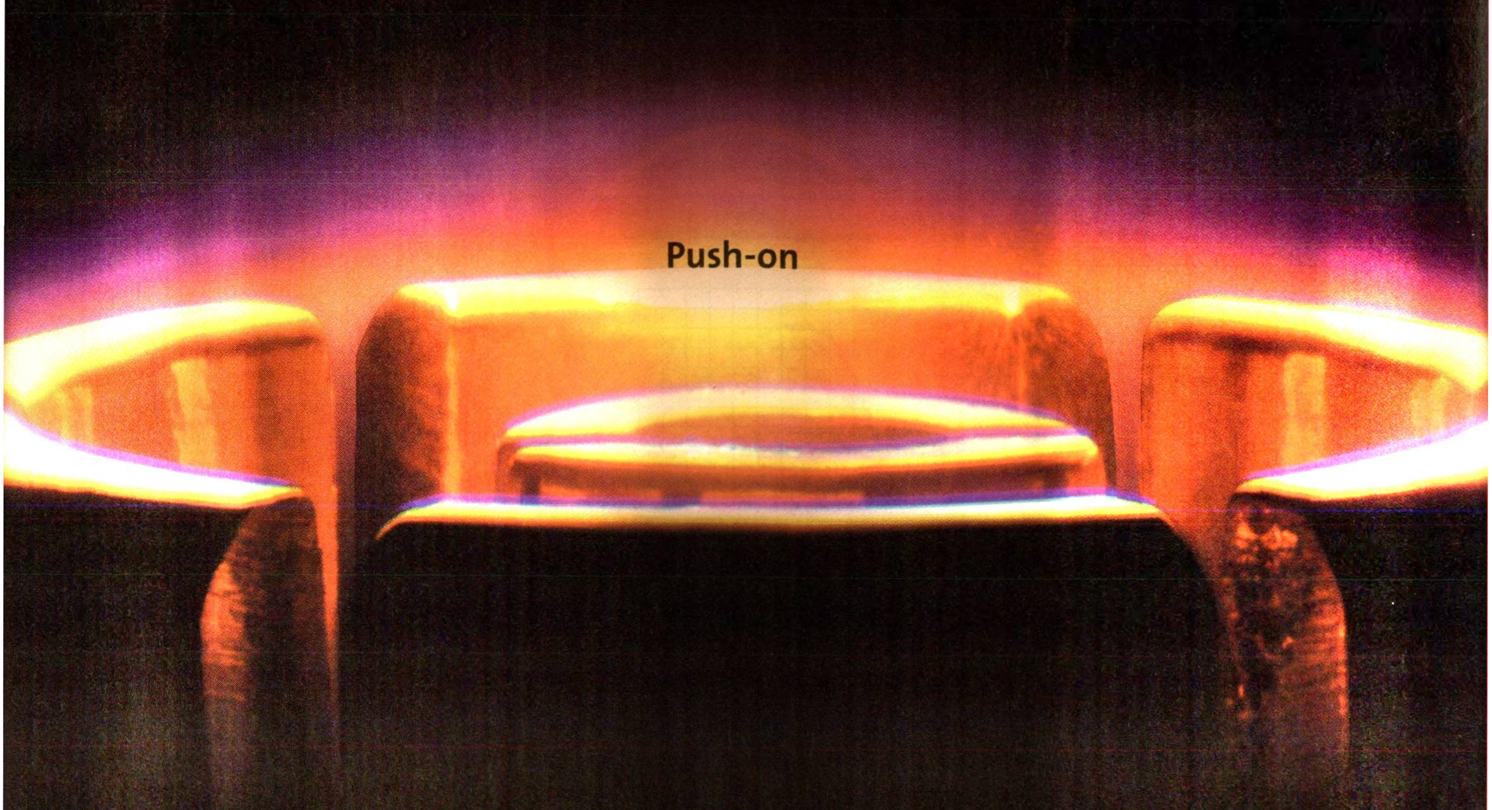

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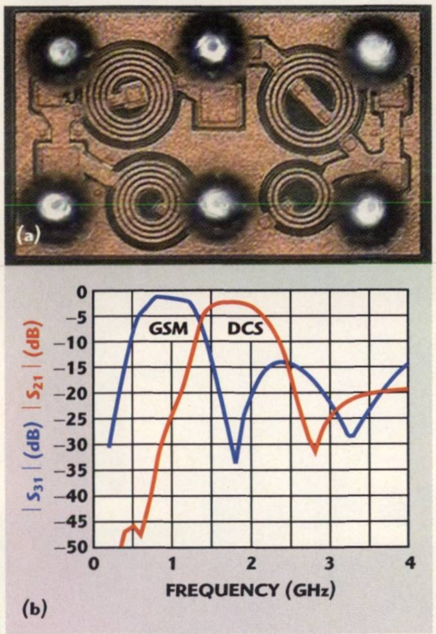
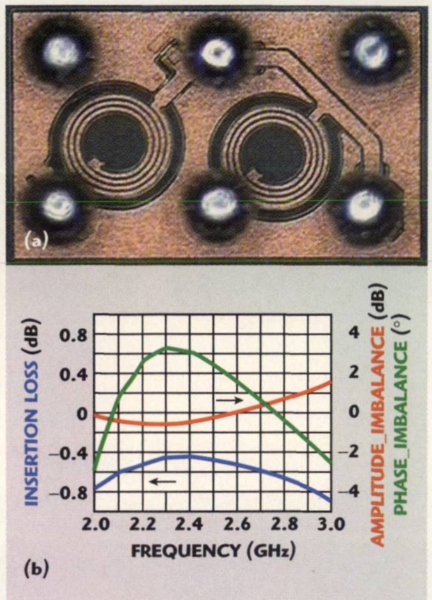


▲ *Fig. 7 Bandpass type EGSM/PCS/DCS diplexer (a) and its measured results (b).*



▲ *Fig. 8 Balun for 2.4 GHz wireless LAN applications (a) and its measured results (b).*

ries connection and shunt connection in each channel. ***Figure 7*** shows the fabricated chip and its test results. The nominal values of the pass-band insertion loss are 1.2 dB at 880 to 915 MHz and 2.2 dB at 1710 to 1910 MHz. The attenuation levels of the stop-band are more than 30 dBc at 1800 MHz and 20 dBc at 3600 MHz. The return loss is greater than 15 dB. The packaged device size is 1.5 mm^2.

LUMPED L-C BALUN FOR WIRELESS LAN APPLICATIONS

A balun converts a balanced signal to an unbalanced one or vice-versa. The photograph and the measured performance of a fabricated 2.4 GHz balun are shown in ***Figure 8***. The insertion loss of 0.5 dB is obtained from back-to-back measurement of the baluns. The phase imbalance is measured to be less than 3.5° and the amplitude imbalance is less than 0.5 dB. These results are superior to those from their ceramic counterparts, which show typical phase imbalance of 10° and amplitude imbalance of 2.0 dB maximum. The smaller size, 1.5 mm^2, corresponding to a reduction of 60 percent over the conventional device, is another advantage.

DISCUSSION

To make RF integrated passive devices small and inexpensive, no packaging was used, except eutectic solder balls for wafer-level bonding. The device with eutectic solder balls can be directly attached to the board, using conventional surface-mount technology (SMT) techniques and a solder-reflow process. Solder balls with a diameter of 130 or 300 μm can be used, depending on the requirements. The former is obtained during the plating process and the latter is typically achieved with the ball placement process. The fabricated passive devices can be RF-tested using a membrane probe card. According to the experiments performed, the solder balls are not damaged too much and the RF measurement results are accurate and reproducible if the probe contact is kept to less than 20 times for the same device. When less than 20 contacts have been made, the measurement error is less than 0.05 dB and the standard deviation is 0.016 dB.

CONCLUSION

Small-size RF integrated passive devices are fabricated on 25 μm thick oxide Si substrates, using a thick Cu and low dielectric constant BCB interlayer processes. The devices show high RF power handling capability (more than 3 W) and low loss. Solder bumps for passive device bonding are used and to the authors' knowledge, the fabricated wafer-level packaging circuits shown here are the smallest ever re-

[Continued on page 88]

ported, compared to conventional ceramic passive devices. The low pass filter shows an insertion loss of 0.45 dB and the harmonic attenuation level is greater than 25 dBc. The packaged device size is 1.0 mm^2. The antenna diplexer for GSM/DCS applications shows insertion losses of 0.45 and 0.60 dB at 900 MHz and 1800 MHz, respectively. The band-rejection levels are more than 25 dBc at 900 MHz and 30 dBc at 1800 MHz. The insertion losses of the diplexer for VCO applications are 1.2 dB at 900 MHz and 2.2 dB at 1800 MHz, in spite of the bandpass topology. A balun is also fabricated for 2.4 GHz wireless LAN applications and has a loss of 0.5 dB, a phase imbalance less than 3.5° and an amplitude imbalance less than 0.5 dB. The size of the above three passive device circuits is 1.5 mm^2 and shows a dramatic reduction compared to conventional approaches. The "cheap and small" passive devices are achieved by using passive integration technology on Si substrates and wafer-level solder bump bonding, while maintaining their high power handling capability and RF performance. This RF passive integration on Si will provide a step forward for the next generation of mobile technology and will be cost-effective. It is an optimal solution for handheld wireless applications that require stringent cost and size reductions. ■

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Dong-Wook Kim *received his BS degree in electronics engineering from HanYang University, Seoul, South Korea, in 1990, and his MS and PhD degrees in electrical engineering from the Korea Advanced Institute of Science and Technology (KAIST), Taejeon, South Korea, in 1992 and 1996, respectively. From 1991 to 2000, he worked as a member of the technical staff at the LG Electronics Institute of Technology, Seoul, South Korea, where he was involved in developing microwave and millimeter-wave circuits and modules. Since 2000, he has been a principal engineer and general manager of Telephus Inc., where he is currently leading the product development division to develop Si integrated passive devices and multi-chip modules for microwave and millimeter-wave applications. His areas of interest are low cost passive integration techonology, RF and millimeter-wave integrated circuit design, multi-chip module based on low cost passive integration process and system-in-package. Dr. Kim is a member of both IEEE and IMAPS (International Microelectronics and Packaging Society).*

In-Ho Jeong *received his MSEE and PhD degrees from the Korea Advanced Institute of Science and Technology (KAIST), Taejeon, South Korea, in 1997 and 2001, respectively. He currently works for Telephus Inc. as a senior R&D engineer. His research interests include RF MEMS technology and integrated passive devices for wireless communications.*

Jong-Soo Lee *received his BSEE, MSEE and PhD degrees in electrical engineering from the Korea Advanced Institute of Science and Technology (KAIST) in 1994, 1996 and 2001, respectively. In 2000, he was a research engineer at the DaimlerChrysler Research Center, Ulm, Germany, working with GaN power amplifiers. He joined Telephus Inc. that same year. His work includes passive device modeling for MCM-D technology, design of passive circuits, and the design of RF modules integrated with passive circuits, SAW-based filters and RFICs. His major interests are in active devices such as Si BJT or BiCMOS and GaAs (InP) MESFET, HEMT or HBT to design RF subsystems or modules such as transceiver modules including diplexers, LNAs, filters and mixers.*

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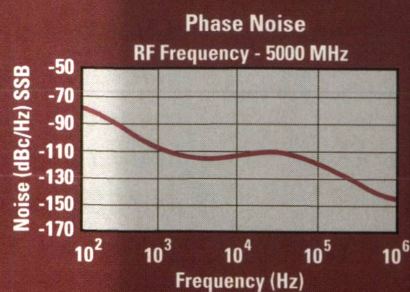

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Effects of AM/AM and AM/PM Distortion on Spectral Regrowth in 3GPP W-CDMA BS Power Amplification

The effects of amplitude and phase nonlinearities in a power amplifier, which lead to spectral regrowth, are examined in the context of a 3GPP W-CDMA forward link (base station) environment. Behavioral modeling and simulation techniques are applied first to independently examine the relationship between AM/AM and AM/PM conversion and spectral regrowth. This analysis suggests that compressing the peak envelope excursions by several decibels of gain compression still results in a tolerable level of spectral regrowth. Conversely, limited phase distortion appears as a key element in developing linear power amplifiers with linearity margin to the 3GPP specification. Experimental results, obtained by digitally predistorting a 3GPP W-CDMA compliant signal to correct only for the amplifier's AM/PM conversion, provide supporting evidence.

The transmitter unit, contained in base stations for third generation cellular systems, requires a linear power amplifier to preserve signal integrity when amplifying composite signals associated with multiple channels. While the power amplifier must meet a number of requirements to be system level compliant, the design process generally emphasizes and places particular attention to optimizing amplifier linearity and efficiency for a given output power level. Developing such a power amplifier requires exploring the design space (bias point, load line, device sizing, harmonic terminations, etc.) in great detail using either empirical means such as load pull measurements, analytical techniques based on applying large signal model representations of the active device, or a combination of the two. While these techniques can be very useful, effective and powerful for design purposes, qualitative insight into those design constraints affecting amplifier linearity is often less obvious. The intent of this article is to provide some qualitative insight into spectral regrowth due to the power amplifier nonlinearities for a 3GPP W-CDMA compliant base station (BS) signal.[1] The amplifier nonlineari-

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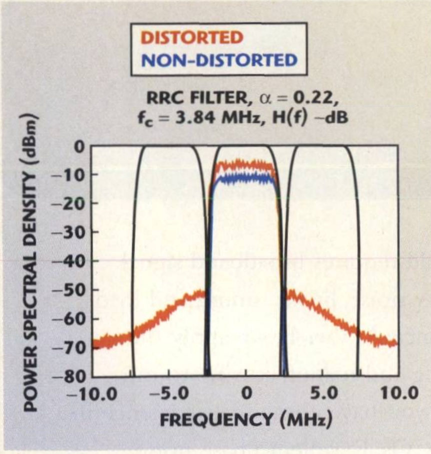
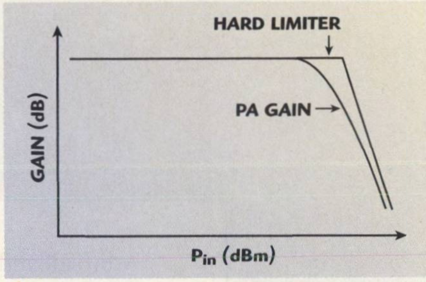
J. Staudinger
Motorola, Semiconductor Products Sector
Tempe, AZ



▲ *Fig. 1 PSD for both a distorted and non-distorted 3GPP W-CDMA Test Model 1 signal (30 kHz measurement BW).*

ties are represented in terms of gain compression (AM/AM conversion) and phase distortion (AM/PM conversion). Behavioral modeling and simulation algorithms are applied to independently examine and quantify the effects of each on spectral regrowth. Simulation results illustrate that while both mechanisms contribute to an increase in spectral content, limited AM/PM conversion is a key element to achieving high amplifier linearity. Phase distortion resulting from AM/PM conversion is experimentally examined by correcting an amplifier's AM/PM response by digitally predistorting a W-CDMA compliant signal to the amplifier's inverse static phase characteristics.



▲ *Fig. 2 Gain compression characteristics typical of class A/B power amplifiers.*

SPECTRAL REGROWTH IN 3GPP W-CDMA

A traditional measure of power amplifier linearity, applied in 2G and 3G cellular systems, considers spectral regrowth or the increase in spectral content of the amplified signal resulting from nonlinearities in the power amplifier. This concept is shown in ***Figure 1***, which graphically illustrates the power spectral density (PSD) for both a non-distorted as well as an amplified and distorted W-CDMA signal.[1] The increased spectral content due to the amplifier nonlinearity is clearly evident. A figure of merit known as adjacent channel leakage ratio (ACLR) quantifies spectral regrowth. In the 3GPP standard, it is defined as the ratio of average power in channels offset in frequency by 5 and 10 MHz to power in the main transmit channel. Note that all power measurements are made after passing the signal through a receive filter which is specified as a root-raised-cosine (RRC) filter with a roll off factor (α) of 0.22 and cut-off frequency of 3.84 MHz. The response of the RRC filter is illustrated where it is overlaid on the PSD plots for both main and adjacent channels.

Since a BS signal is the composite or sum of many individual traffic channels as well as control and paging channels, it becomes important to precisely define the properties of each individual constituent channel. This is especially important in evaluating amplifiers as the characteristics of the signal have a major influence on their performance. Factors such as phase alignment, power level, code selection and timing offsets of the individual channels result in widely differing composite signals.[2] Hence, it is critical to have a well-defined test signal for purposes of consistency as well as for the test signal to represent a realistic base station traffic scenario. The 3GPP standard[1] addresses this issue by including specific details regarding the generation of a compliant signal and specifies that Test Model 1 be used for ACLR measurements. A compliant Test Model 1 signal was developed consisting of 64 traffic channels with spreading factors, timing offsets, code and level settings per the 3GPP specification.

AMPLIFIER AM/AM AND AM/PM NONLINEARITIES

A relatively simple method of describing power amplifier nonlinearities considers AM/AM and AM/PM conversions in a behavioral model.[3,4] Assuming a narrow band signal nonlinearity and first zonal components only, a gain transfer function is developed which relates instantaneous complex input and output envelope voltages based on AM/AM and AM/PM data. In practice, power amplifiers generally exhibit near-constant gain with increasing input power until the onset of gain compression, as shown in ***Figure 2***. At this point, further increases in input power result in slight increases in output power until the device/amplifier saturates. To a first order, and neglecting the soft character in the compression response, it can be represented by a piecewise linear approximation where beyond the onset of compression, output power does not increase. While this representation is at best a loose description of actual PA gain characteristics, it nonetheless is useful in approximating the effect of compressing peak envelope excursions on spectral regrowth.

Estimating the amount of headroom or power back-off needed is an important parameter in designing a linear power amplifier. As greater amounts of back-off are required to meet linearity performance, the amplifier efficiency decreases while the thermal power dissipation increases.

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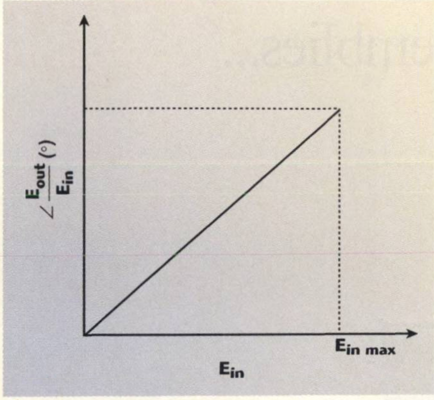
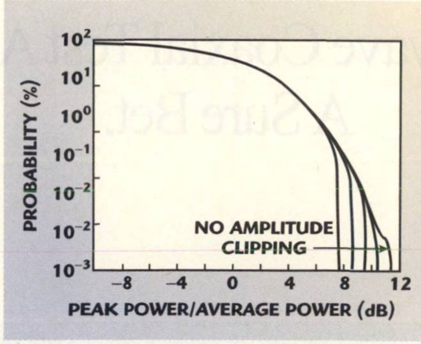
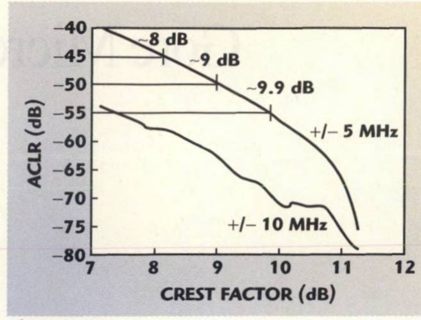


▲ *Fig. 3 AM/PM conversion modeled as a linear variation with input envelope voltage.*



▲ *Fig. 4 CCDFs for Test Model 1 with 64 traffic channels, with and without amplitude compression.*



▲ *Fig. 5 Simulated results showing ACLR as a function of crest factor.*

In addition, the size and/or saturated power capability of the amplifier must increase accordingly. Hence, larger active devices are necessary; load line impedances decrease thereby making it more difficult to implement amplifier matching structures, all of which add cost and size.

In addition to gain compression, variations in phase with input power (AM/PM conversion) must be considered. This characteristic is much more difficult to generalize. In practice, power amplifiers exhibit an AM/PM response, which may be monotonically increasing, monotonically decreasing, or non-monotonic. In general, AM/PM conversion becomes more pronounced at higher input drive levels, approaching gain compression. Here and for the purpose of simplicity, a representation is adopted, where AM/PM distortion is linearly proportional to the magnitude of the signal's envelope voltage, as shown in ***Figure 3***. That is, the phase is assumed to vary linearly with envelope voltage over the entire signal's dynamic range.

Estimating the amount of amplifier AM/PM conversion that can be tolerated is also an important design parameter. Allowances in the power amplifier design process can often be made to mitigate, at least to some extent, AM/PM conversion.

Based on the AM/AM and AM/PM representations shown, a behavioral model consisting of a complex gain transfer function is developed based on techniques illustrated in previous work.[3,4]

SIMULATION RESULTS

From the previous discussion a compliant Test Model 1 signal is developed, consisting of 64 traffic channels formed from data sequences occupying five time slots (3.3 ms in length, 5 slots × 666.667 Ωs/slot). This signal is used in all simulations. The complementary cumulative distribution function (CCDF) illustrating the statistical distribution for this signal is shown in ***Figure 4*** (curve labeled "No amplitude clipping"). The upper portions of the graph indicate that very large envelope excursions are possible as peak-to-average (PAR) power ratios in excess of 10 dB occur. These large excursions, however, occur relatively infrequently as indicated by their low CCDF probability.

Simulations were first conducted to calculate ACLR as a function of the amount of clipping applied to peak envelope voltages. Note that the AM/PM conversion component is set to zero. The effect of amplitude clipping on the CCDF is illustrated by the various traces. The resulting ACLR performance at frequency offsets of 5 and 10 MHz is shown in ***Figure 5***. Both are shown as a function of crest factor,

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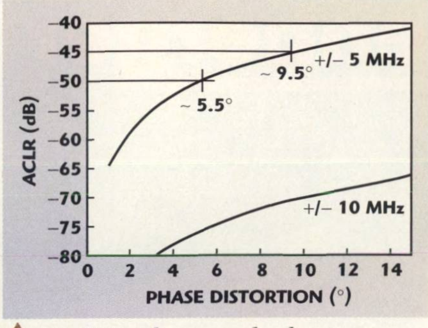
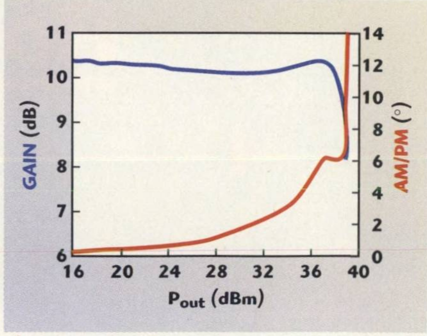


▲ *Fig. 6 Simulation results showing the relationship between ACLR and phase distortion.*

which is simply the maximum PAR of the waveform. Given system level ACLR specifications of –45 and –50 dB (at offset frequencies of 5 and 10 MHz, respectively), the W-CDMA signal can be compressed to a crest factor of near 8 dB and still meet stated system level ACLR performance. Further, lower ACLR levels of –50 and –55 dB (5 MHz offset frequency) have corresponding crest factors of approximately 9 and 10 dB, respectively. These simulations suggest that very high peak envelope voltage excursions, which occur with relatively low probability, can be compressed and still meet ACLR performance.



▲ *Fig. 7 Measured complex gain response for a 7 W power amplifier.*

A second set of simulations was performed to examine AM/PM conversion effects. In this case, AM/AM conversion is set to zero (that is no gain compression). AM/PM conversion is modeled as a linear phase variation with envelope voltage. Simulations were conducted with total phase variations ranging from 1° to 15°. Results are shown in ***Figure 6*** with ACLR values corresponding to 5 and 10 MHz plotted as a function of total signal phase distortion. The phase distortion shown is the total linear phase variation over the full dynamic range of the signal. These results suggest that a relatively modest level of approximately 10° AM/PM conversion results in an ACLR of –45 dB at a 5 MHz offset. ACLR at the 10 MHz offset frequency remains small. However, it must be noted that the –45 dB ACLR level is a system level specification and not necessarily that of the power amplifier. The power amplifier is often specified with significant margin to the system level requirements. Therefore, if ACLR levels of –50 dB and lower are considered, for example, the tolerable amount of AM/PM conversion becomes quite small — on the order of a few degrees at most. Hence, in developing power amplifiers with linearity margin to the 3GPP specification, AM/PM conversion effects may well be dominant over gain compression.

EXPERIMENTAL MEASUREMENTS OF AM/PM CONVERSION AND ACLR

The preceding analysis was based on modeling the power amplifier nonlinearity using relatively simple methods. In this section, AM/AM and AM/PM conversion effects are considered, via experimental measurements on a relatively high power (7 W) amplifier driven by a Test Model 1 W-CDMA 3GPP compliant stimulus.

The complex gain characteristics of the amplifier were first measured using a pulsed CW stimulus by driving the amplifier from linear operation into several decibels of gain compression. A pulsed CW stimulus was used to mitigate the influence of thermal and self-heating effects on the complex gain response (***Figure 7***). The amplifier exhibits a P_{1dB} of 38.5 dBm or 7 W (that is, an output power of 38.5 dBm at 1 dB gain compression). The gain response shows some variation with input power — the gain decreases, then expands slightly prior to the onset of compression. The insertion phase is almost monotonic and increases gradually with drive level except at high power levels where it rises very rapidly.

Given the amplifier's static complex gain response, pre-distortion techniques are applied next, which predistort the input W-CDMA signal to correct only for the amplifier's AM/PM conversion. This is done algo-

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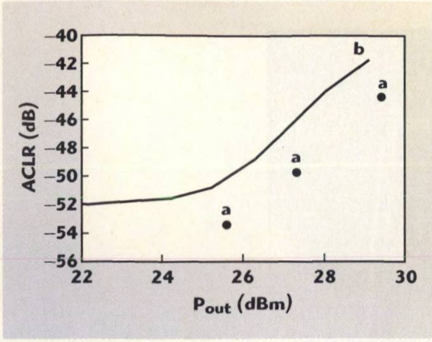

▲ *Fig. 8 Measured power amplifier ACLR response (at a 5 MHz offset frequency), (a) with and (b) without predistorting the input signal.*

rithmically by digitally predistorting the base-band I/Q components. Beginning with a W-CDMA Test Model 1 compliant base-band signal s(t), where

$$s(t) = I(t) + jQ(t) \qquad (1)$$

A pre-distorted signal taking the form

$$s_{pd}(t) = I_{pd}(t) + jQ_{pd}(t) \qquad (2)$$

is developed by applying a pre-distorter complex gain function $G_{pd}(v)$ to s(t)

$$s_{pd}(t) = G_{pd}\langle I(t) + jQ(t)\rangle \qquad (3)$$

The function $G_{pd}(v)$ is the inverse of the phase constituent of the amplifier's complex gain response only. That is

$$|G_{pd}(v)| = 1 \qquad (4)$$

and

$$\arg(G_{pd}(v)) = -\phi(v) \qquad (5)$$

where

$\phi(v)$ = amplifier's AM/PM conversion
v = magnitude of the input envelope voltage ($v = |I(t) + jQ(t)|$)

The function $\phi/(\nu)$ is developed from the measured amplifier's AM/AM data based on a pulsed CW stimulus. The function $G_{pd}-(\nu)$ is then derived from $\phi/(\nu)$ and is implemented as a polynomial series expansion. Note the magnitudes of s(t) and $s_{pd}(t)$ are identical for all points in time and only the phase of the two signals differ. The difference in phase is equal, but opposite in sign, to what the amplifier exhibits. Finally, the signal $s_{pd}(t)$ is modulated with a carrier and applied to the input of the power amplifier. Note that a unique predistorted input signal must be developed for each output power level.

To form a reference to compare against, the ACLR performance of the amplifier is first measured with a Test Model 1 signal over a broad input power range. Next, a predistorted (phase only) Test Model 1 signal is developed and applied to the amplifier. The results are shown in ***Figure 8***.

Consider the performance and the resulting ACLR at an output power level of 25.5 dBm with the amplifier driven by W-CDMA stimulus. This point represents the power amplifier operation at 13 dB output power back-off from the 1 dB gain compression point. Therefore, very limited amplitude distortions occur, mainly at the extreme peak envelope excursion. However, they occur rather infrequently as illustrated by the CCDF. Additionally, lower envelope voltages are distorted somewhat since the amplifier exhibits some gain variation (AM/AM) prior to the 1 dB compression point. Nonetheless, at this power level, it is expected that to a large extent, the amplifier's AM/AM response does not contribute significantly to ACLR. On the other hand, at this power level, the signal is distorted by a significant amount of AM/PM, approximately 8.5° for a PAR of 12 dB. Predistorting the signal to correct for the amplifier's AM/PM contributions significantly improves ACLR to a level near –54 dB.

Measured results are also shown for average output power levels of 27.5 and 29.5 dBm. These power levels represent output power back-offs of 11 and 9 dB, respectively. These signals are distorted by larger amounts of AM/PM conversion, well in excess of 15 for a signal PAR of 11 dB. Predistorting the input signal to account for the amplifier's AM/PM again improves ACLR. The higher ACLR values in these cases suggest that gain compression is a much more dominant effect. Note that these ACLR values are somewhat higher than those predicted by the hard limiter case. This is likely due, in large part, to limitations in the method used to implement the predistorter that become more pronounced at high power levels. Note that it becomes more difficult to account for

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the amplifier's AM/PM response when peak envelope excursions extend several decibels into compression. As illustrated previously, the AM/PM response for this amplifier rises very rapidly at power levels that drive the amplifier several decibels into gain compression. Therefore, it becomes particularly difficult to measure/model this region of operation accurately. In addition, it must be emphasized that in this work, the predistortion function $G_{pd}(v)$ is developed from static measurements and as such does not account for any memory effects which may be thermally or electrically produced. Nonetheless, these measurement results clearly confirm that to achieve ACLR values well below –45 dB, only limited AM/PM conversion can be tolerated.

CONCLUSION

The effects of AM/AM and AM/PM conversion on ACLR performance has been examined for an amplifier driven by a 3GPP W-CDMA BS signal. The results suggest that compressing the peak envelope excursions by several decibels can likely be tolerated. Alternatively, achieving a more linear amplifier, which exhibits ACLR values with margin to the system specification of –45 dB, requires the amplifier to exhibit very limited AM/PM conversion. Experimental results conducted on a high power amplifier provide strong supporting evidence. ■

ACKNOWLEDGMENT

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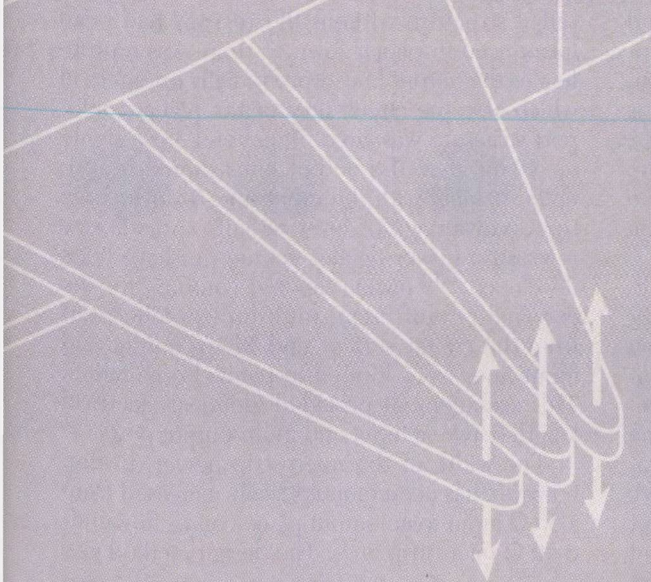

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PRACTICAL DESIGN CONSIDERATION OF A MODIFIED STRUCTURE FOR A PLANAR MULTIPORT POWER DIVIDER AT 2 GHz

This article proposes a modified structure at 2 GHz of IMT-2000 for the planar multiport power divider published in the literature.[8–10] Design parameters for the practical realization of the structure have also been investigated using HFSS (FEM method) simulation. Modification of the circuit structure has been accomplished by reducing the width of the output port plane. For 1:2 and 1:3 power dividers, the measured S-parameters in this study have been compared with simulation results obtained by using the method described by Kobeissi and Wu.[10] The comparison demonstrates that significant improvement of $|S_{11}|$*, broadening of the bandwidth and compactness of the circuit size have been achieved by the modification.*

A power divider is a basic microwave and millimeter-wave circuit dividing the input RF power by an arbitrary ratio. This circuit has a wide range of applications such as antenna systems, power amplifiers, power oscillators, balanced mixers and so on. Multiport power dividers/combiners have been used in power amplifiers to increase their output power.[1]

The Wilkinson type of microstrip power divider has been predominantly used as a multiport power divider. But λ/4 length transmission lines for impedance matching and parallel connection of resistors between output lines are required.[2,3] If the resistor is eliminated in the power divider, a more efficient one-step fabrication process of the circuit can be accomplished. Therefore, in order to develop a power divider without a resistor, a circular microstrip resonant structure had been suggested for the multiport circuit.[4,5] Since this circuit had to have a coaxial feeding structure at the input, it was not easy to realize an all planar type circuit as required for MMIC or HMIC topology. Next, even though a radial structure without the resistor had been introduced in planar form,[6,7] there was a problem as the output lines were radially oriented, in which case the circuit area of the planar multiport structure was large. In general, for a smaller circuit area, the output lines should be oriented parallel with the input line. To overcome the disadvantage of lines radially spread, new microstrip power divider geometries have been developed,[8–10] having tapered contours for the external circuit edge, multiholes etched out from the circuit surface and where output and input lines were located in parallel orientation. This arrangement provided identical electrical length between input and multi-output ports.

The width of the microstrip power divider, however, has been monotonically increased from a 50 Ω input to an output plane connecting multi-50 Ω microstrip lines. The design method giv-

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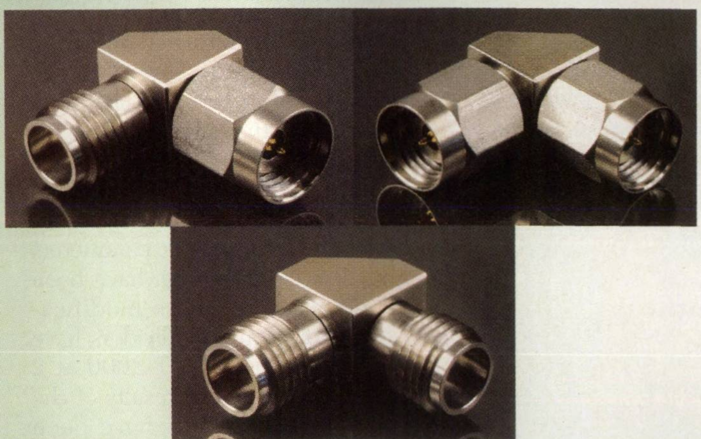
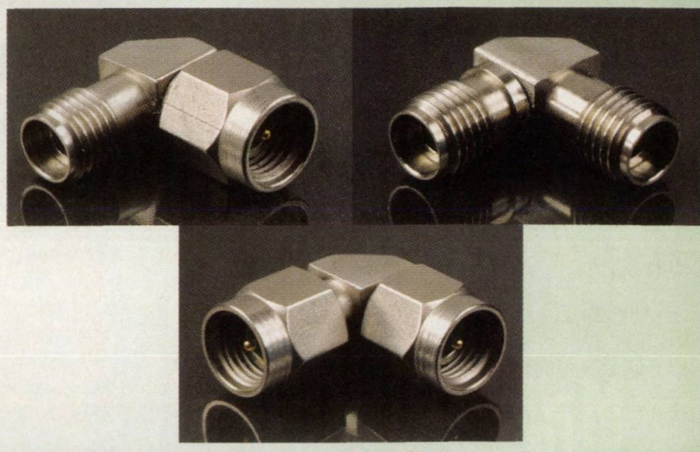
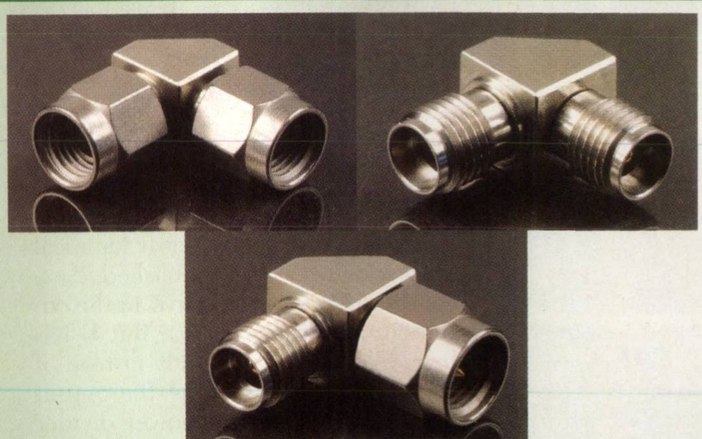
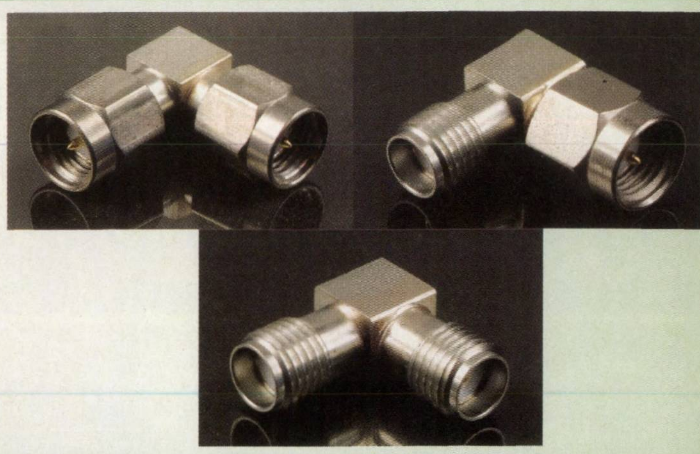
YONGIN HAN AND IHN S. KIM
KyungHee University
Kyungki-do, South Korea









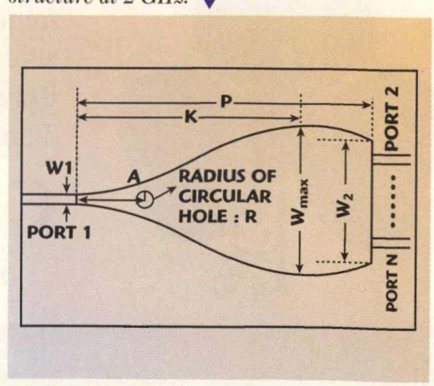
en in the references[8–10] has introduced an external contour geometry where the width is monotonically increased from input to output. For better input return loss, the physical width at the output plane should be narrower as the frequency is increased. For example, at frequencies higher than 10 GHz, the discontinuity between the final transverse plane and the frequency-independent multi-50 Ω microstrip lines becomes small. The width W_2, in reference 10, at the output transverse plane should be maintained at approximately 0.6λ. If the number of output ports is increased, a width of approximately 0.5 to 1.0λ must be normally adopted. However, if this type of circuit contour is considered below approximately 10 GHz, the width W_2 must be physically increased. At 2 GHz, since the physical width at the output plane becomes wider, the discontinuity between the output plane and the multi-50 Ω microstrip output lines can be increased. When 1:2 and 1:3 power dividers at 2 GHz were designed by the method described in Kobeissi and Wu,[10] 10 and 11 dB return losses were shown as the best HFSS simulation results obtainable even if optimum tuning was carried out.

In order to overcome this problem, the contour geometry for the previously described multiport power divider[8–10] has been modified. Modification analysis has been carried out based on HFSS simulation. The modified geometry has an external shape such that the width at the output plane is reduced again after passing through a maximum width, located in the middle of the circuit. The frequency responses as a function of the circuit parameters of a practical circuit design have been included. Based on these modifications, 1:2 and 1:3 power dividers have been developed for IMT-2000 at 2 GHz. The measurement results of the modified power dividers have been compared with the simulation results of the power divider designed by using the method of reference 10 at 2 GHz.

MODIFIED STRUCTURE AND CIRCUIT PARAMETERS

The modified structure of a planar multiport power divider is shown in ***Figure 1***. The external contour of the power divider geometry is determined by Equation 1 that is identical to the one previously published,[10] except for the denominator K in the cosine argument (K was P). But K has been adopted in this study instead of P, to reduce the width at the output end of the microstrip power divider in which W_{max} must be placed in the middle of the structure. Therefore,

Fig. 1 Modified contour of a multiport power divider structure at 2 GHz.



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$$W_2 = W_1\left(1 - q\cos\left(\frac{\pi x}{K}\right)\right) \quad (1)$$

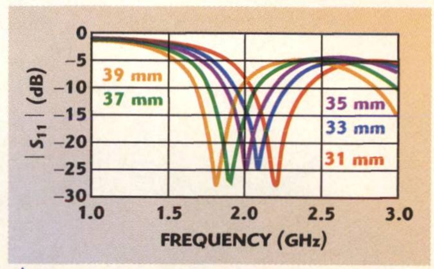
where $0 \le x \le P$ and K must have a value between P and P/2. The final width of the structure, W_2, is proportional to q. As the value of K becomes the center of the length of the structure, the output width, W_2, approaches zero.

The circuit parameters of Equation 1, indicated in the diagram, are important factors in determining the external contour of the tapered geometry. In the following section, information for a practical circuit design will be given. After the circuit parameters for the geometry of a 1:2 power divider were extracted for the frequency range from 1 to 3 GHz through HFSS optimization, a 1:3 power divider was constructed by using the design parameters.

The frequency responses as a function of the circuit parameters P, K, q, A and R have been studied in terms of S-parameters for a 1:2 microstrip power divider structure which has been fabricated on a substrate with a dielectric constant $\varepsilon_r = 10.2$ and a thickness of 1.27 mm (Rogers RO 6010).

Figure 2 shows the center frequency variation (depending on the length P) of the modified structure for $K = 0.4\lambda$ and $q = 7$. ***Table 1*** summarizes the relationship between the center frequency and P expressed in electrical length. It has been observed that the length P is a circuit parameter determining the center frequency. Optimum characteristics have been obtained when the length was approximately 0.6λ. K and q are the parameters for the location of W_{max} and the width of the structure, respectively. The location and radius of a circular hole etched out on the structure can be chosen for better electrical characteristics.

Figure 3 shows the center frequency variation and the input return loss



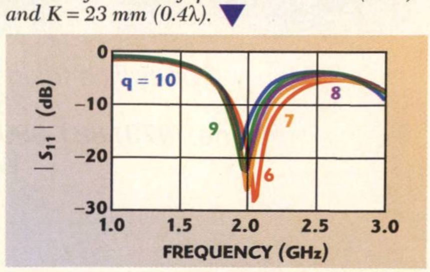
▲ *Fig. 2 Center frequency variation as a function of P with $K = 0.4\lambda$ and $q = 7$.*

TABLE I

THE 1:2 MODIFIED POWER DIVIDER PARAMETERS

Length P (mm)	Center Frequency (GHz)	Electrical Length
31	2.209	0.600λ
33	2.090	0.600λ
35	2.015	0.601λ
37	1.910	0.601λ
39	1.821	0.601λ

Fig. 3 Center frequency and input return loss as a function of q with P = 35 mm (0.6λ) and K = 23 mm (0.4λ). ▼



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1930-1990 MHz	MV192A-35TK	2.5	200°	+42/+40	$9.91
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2100-2200 MHz	VP212A-35NJ	2.5	360°		$10.30
2100-2200 MHz	VP212A-60BK	1.7	190°		$5.32
2100-2200 MHz	VP212A-60CN	1.0	100°		$5.03
1930-1990 MHz	VP192A-35NJ	2.5	360°		$10.60
1930-1990 MHz	VP192A-60BK	1.7	190°		$5.32
1930-1990 MHz	VP192B-60CN	1.0	100°		$5.26
SURFACE MOUNT VOLTAGE VARIABLE ATTENUATOR w/Attenuation 20 dB & Solder Reflow to 240° C					
2100-2200 MHz	VA212B-60TK	0.6		+45	$4.25
1930-1990 MHz	VA192B-60TK	0.6		+45	$4.25

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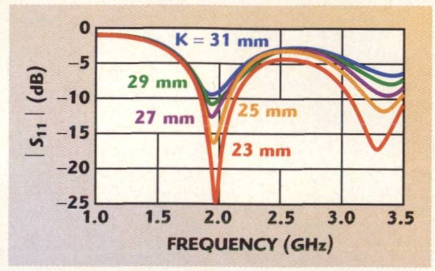
characteristics depending on q. W_{max} and W_2 increase with q. The discontinuity between the width at the output plane and the multi-50 Ω microstrip lines also increases with W_2. In order to be operating at a 2 GHz center frequency for the 1:2 power divider, K = 0.4λ (23 mm) has been chosen.

$|S_{11}|$ and the center frequency characteristics as a function of K are shown in ***Figure 4***. As the value of K approaches P/2 (longitudinal center of the structure), W_2 becomes narrower. The center frequency variation can hardly be distinguished, but is improved by reducing K. K should be determined by considering the number of output port and electrical characteristics.

Figure 5 illustrates that the center frequency and input return loss characteristics depend on the values of A (6, 7 and 8 mm) when P = 0.6λ (35 mm), K = 0.4λ (23 mm), q = 1 and R = 1 mm. The location (A) of the etched circular hole has a negligible effect on the center frequency, but the characteristic is affected. It has been found that as the radius R of the circular hole is increased, the center frequency moves higher and the characteristic is not improved, as shown in ***Figure 6***.

Based on this above analysis, the circuit parameters for the 1:2 power divider with the modified geometry have been determined at the center frequency of 2 GHz to be P = 0.6λ (35 mm), K = 0.4λ (23 mm), W_{max} = 0.31λ (17.86 mm), W_1 = 1.12 mm, W_2 = 0.16λ (8.98 mm), A = 0.12λ (7 mm), R = 1 mm and the width between output ports is 0.11λ (6.4 mm). A 1:3 power divider with the modified geometry has also been designed and optimized at 2 GHz. In this case, the circuit parameters are P = 0.6λ (35 mm), K = 0.4λ (23



▲ *Fig. 4 Center frequency and input return loss as a function of K with P = 35 mm (0.6λ) and q = 7.*

Fig. 5 Center frequency and input return loss as a function of A with P = 35 mm (0.6λ), K = 23 mm (0.4λ), q = 7 and R = 1 mm. ▼

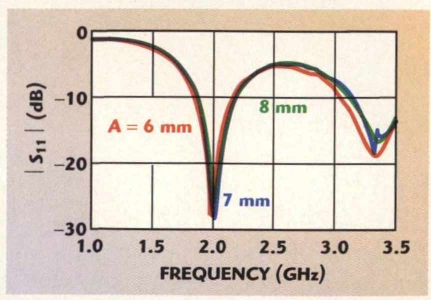
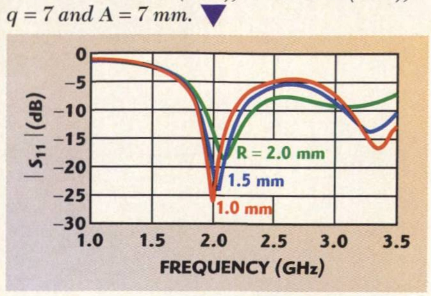


Fig. 6 Center frequency and input return loss as a function of hole radius R with P = 35 mm (0.6λ), K = 23 mm (0.4λ), q = 7 and A = 7 mm. ▼



[Continued on page 110]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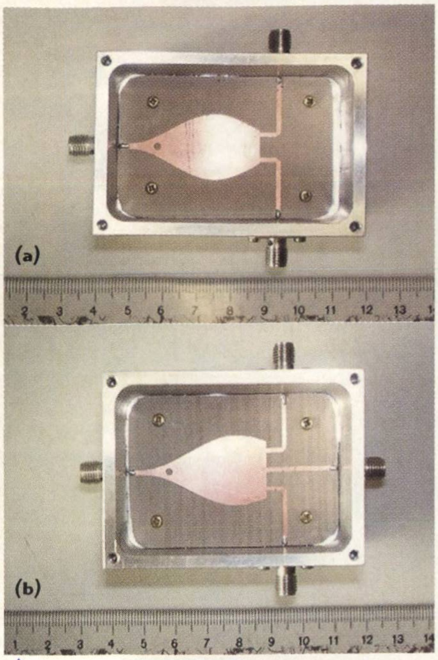
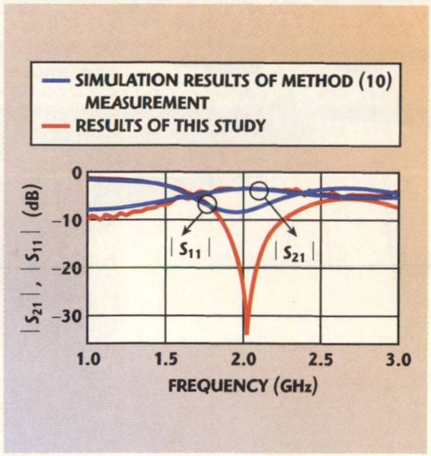


▲ *Fig. 7 Photographs of (a) 1:2 power divider and (b) 1:3 power divider.*

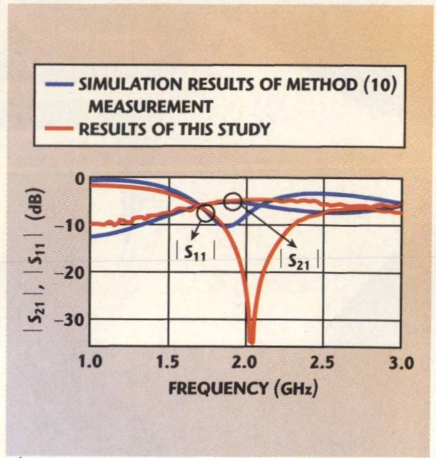
mm), W_{max} = 0.32λ (18 mm), W_1 = 1.12 mm, W_2 = 0.25λ (14.6 mm), A = 0.13λ (8 mm), R = 1 mm and the width between output ports is 0.08λ (5 mm).



▲ *Fig. 8 S-parameter comparison between simulation (method of ref. 10) and measurements (this study) for a 1:2 power divider.*

RESULT COMPARISON FOR THE MODIFIED STRUCTURE

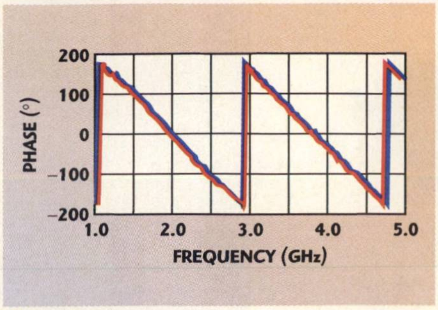
The S-parameters of the 1:2 and 1:3 multiport microstrip power dividers have been simulated and measured with the ε_r = 10.2 and 1.27 mm thickness substrate as described before. The simulation for the previously described structure of reference 10 has been accomplished using the HFSS and measurements were carried out for the modified structure of this study. The two results have been compared in the 1 to 3 GHz frequency range.



▲ *Fig. 9 S-parameter comparison between simulation (method of ref. 10) and measurements (this study) for a 1:3 power divider.*

Figure 7 shows photographs of the modified 1:2 and 1:3 power dividers. ***Figures 8*** and ***9*** illustrate the

[Continued on page 112]



▲ *Fig. 10 Measured phase differences between output ports for the modified 1:3 power divider.*

differences in S-parameters for the 1:2 and 1:3 power dividers developed by using the methods of Kobeissi and Wu[10] and this study. For the 1:2 power divider, the difference in S_{21} between simulated and measured results is not easy to distinguish. However, the approach of this study provides a significant improvement of $|S_{11}|$, from 9 to 33 dB. A similar improvement in $|S_{11}|$ has been obtained for the 1:3 power divider, from 10.5 to 37 dB. The phase characteristic of the 1:3 power divider with the modified structure is shown in ***Figure 10***, where a phase difference less than 1° between output ports has been obtained.

For a bandwidth characteristic based on a 20 dB return loss, this study has provided approximately 100 MHz for the 1:2 power divider and 120 MHz for the 1:3 circuit, but the structure simulated by the Kabeissi and Wu method[10] could not be directly compared since $|S_{11}|$ characteristics better than 15 dB could not be achieved.

CONCLUSION

The external geometrical contour of the planar multiport power divider having parallel orientation of input and output lines[8–10] has been modified for IMT-2000 application at 2 GHz. The circuit parameters for a practical design of the structure have been provided by HFSS analysis. Based on this analysis, 1:2 and 1:3 power dividers with the modified shape have been realized in microstrip form. S-parameter characteristics of the modified power divider have been measured and compared with simulation results of the 1:2 and 1:3 power dividers designed and optimized by the Kobessi and Wu method.[10] This comparison shows that the modified power divider provides a significant improvement of the $|S_{11}|$ characteristic from 9 to 33 dB for the 1:2 power divider and from 10.5 to 37.0 dB for the 1:3 divider. For a bandwidth characteristic based on a 20 dB return loss, this approach has provided approximately 100 MHz for the 1:2 power divider and 120 MHz for the 1:3 circuit, but the structure simulated by the previously published method[10] could not be directly compared since an $|S_{11}|$ characteristic better than 15 dB could not be achieved. By using the modified geometry, the physical circuit size has been reduced more than the one given by the previously published method.[10] ■

ACKNOWLEDGEMENT

This work was supported in part by Grant No. R02-2001-00862 from the basic research program of the Korea Science and Engineering Foundation and by KyungHee University.

[Continued on page 114]


Let us take your specs for a spin.
Frequency: 14.0 – 14.5 GHZ
10.75 – 12.75 GHZ
VSWR = 1.20 MAX
Insertion Loss = .20 Db MAX

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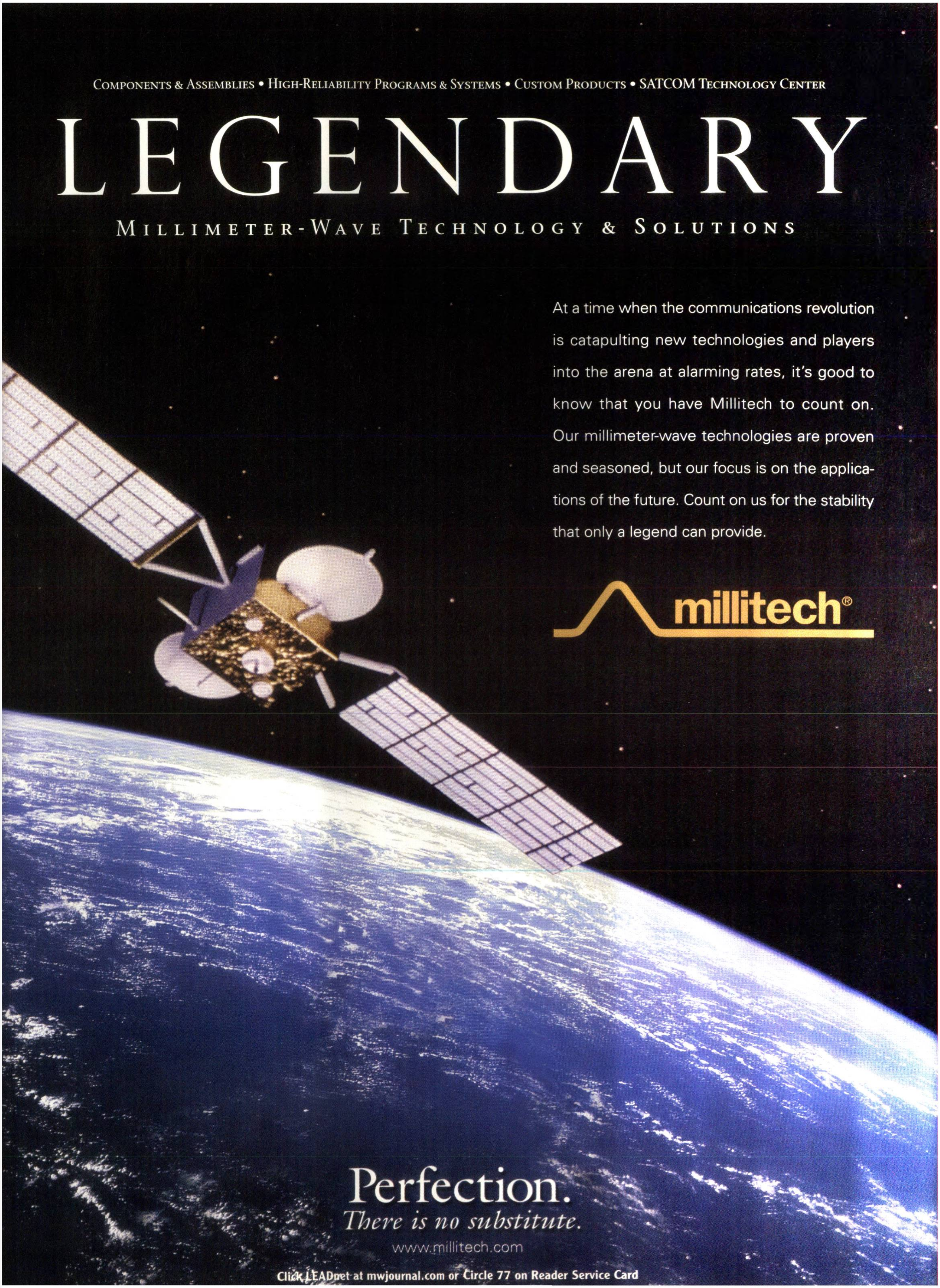
Yongin Han *received his BE and MAS degrees in electrical engineering from KyungHee University, Suwon, South Korea, in 2000 and 2002, respectively. He is currently working for Sewon Telecom, Seoul, as a research engineer. His research interests are in the areas of power dividers/combiners, microwave oscillators and RF part design in CDMA PCS phone.*

Ihn S. Kim *received his BE degree in electrical engineering from KyungHee University, Suwon, South Korea, in 1974, and his MAS and PhD degrees, both in electrical engineering, from the University of Ottawa in 1983 and 1991, respectively. He has worked for the Canadian Space Agency, Com Dev Ltd., General Instrument of Canada and the Korean Broadcasting System. From February 1999 to February 2000, he was on sabbatical at ETRI (South Korea), ETH (Switzerland) and Motorola Florida Research Lab (US). He is currently a professor teaching microwave engineering in the school of electronics and information at KyungHee University. His research involves commercial application of radar technology, modeling of microwave structures by numerical methods (FEM, FDTD and TLM) and their application to filters and power divider/combiners, and nonlinear microwave circuit development such as oscillators and mixers. He can be reached via e-mail at ihnkim@khu.ac.kr or by phone at +82-31-201-2587.*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Active Multipliers and Dividers to Simplify Synthesizers

Modern frequency synthesis uses a combination of frequency multiplication and frequency division to generate the required data and carrier signals for microwave radio and radar applications. Frequency multiplication increases the carrier frequency while increasing the phase noise of the carrier by 20log(N), where N is the multiplication factor. Conversely, frequency division decreases the carrier frequency and changes the phase noise of the carrier by 20log(1/M), where M is the division ratio.

For the most part, frequency dividers are used in a phase-locked loop (PLL) to divide the voltage-controlled oscillator (VCO) output frequency down to the reference frequency in order to achieve phase/frequency locking. These devices are traditionally digital circuits and for high divide ratios (greater than 16) operate at input frequencies less than ~2 GHz. At higher frequencies, digital frequency dividers are available in binary division ratios (/2, /4, /8...) up to approximately 14 GHz. A new class of GaAs/InGaP heterojunction bipolar transistor (HBT) active frequency dividers offering odd division ratios (/3 and /5), high input frequency (up to 8 GHz) and wide input power range (–15 to +10 dBm) are now available in low cost plastic packages providing the designer with more flexibility in synthesizer architecture.

Frequency multipliers, on the other hand, are most commonly used after the VCO to increase the carrier frequency to the radio or radar transmit/receive frequency. Various devices are utilized to achieve frequency multiplication, including Schottky diodes, step recovery diodes, balanced rectifiers (frequency doublers) and saturated amplifiers. The advantages and drawbacks of each approach are well known. A new class of GaAs/InGaP HBT active multipliers offering conversion gain, wide input power range (–15 to +10 dBm), high output frequency (up to 16 GHz) and high multiplication factors (× 4, × 8 and × 16) are now available in low cost plastic packages that will dramatically simplify the design and implementation of X- and Ku-band synthesizers.

[Continued on page 120]

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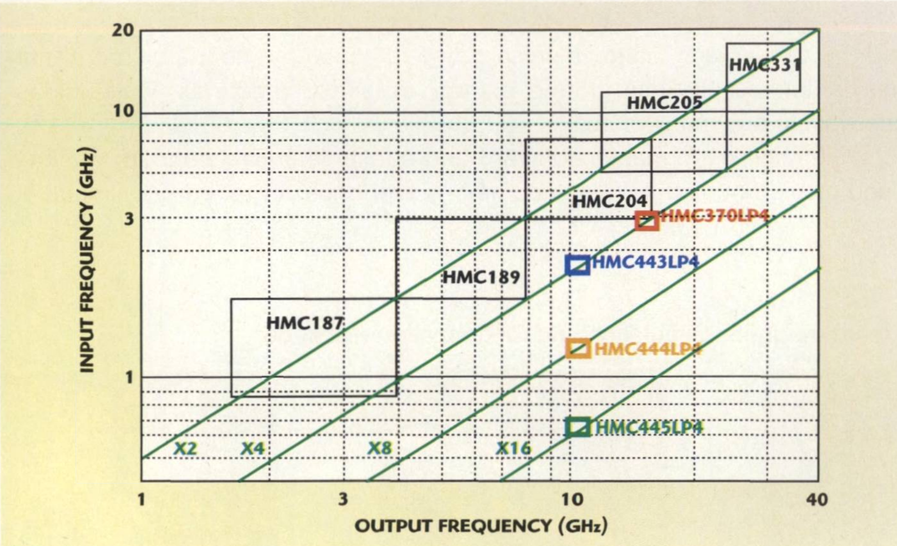
STANDARD FREQUENCY MULTIPLIER PRODUCTS

Four new active frequency multipliers have been added to the company's family of surface-mount frequency doublers. ***Figure 1*** depicts the frequency coverage of the passive and active frequency multiplier line. The models HMC370LP4 and HMC443LP4 are active ×4 frequency multipliers with output frequencies at X- and Ku-band, respectively. The models HMC444LP4 and HMC445LP4 have output frequencies in the 10 to 11 GHz frequency range. Inspection of data reveals that the new active multipliers are suitable for the OC-192 fiber optic market, military radar and the microwave radio markets.



▲ *Fig. 1 Frequency multiplier operating range.*

The key performance parameters of the active multipliers are summarized in ***Table 1***. The HMC445LP4 multiplier will convert an OC-12 clock frequency directly to OC-192 with up to +7 dBm output power operating from +5 V and consuming only 78 mA of current. Similarly, the HMC444LP4 and HMC443LP4 devices will convert OC-24 and OC-48 clock frequencies to OC-192 while using only 68 mA and 52 mA of cur-

TABLE I

ACTIVE MULTIPLIER PERFORMANCE PARAMETERS

Part Number	Function	Input Frequency (MHz)	Output Frequency (GHz)	Input Power (dBm)	Output Power (dBm)	SSB Phase Noise at 100 kHz (dBc/Hz)
HMC445LP4	X16 active	618.75 to 687.50	9.9 to 11.2	–15 to +5	+4 to +7	–130
HMC444LP4	X8 active	1237.5 to 1400.0	9.9 to 11.2	–15 to +5	+3 to +6	–136
HMC443LP4	X4 active	2450 to 2800	9.8 to 11.2	–15 to +5	+1 to +4	–142
HMC370LP4	X4 active	3600 to 4100	14.4 to 16.4	–15 to +5	–4 to 0	–140

[Continued on page 122]

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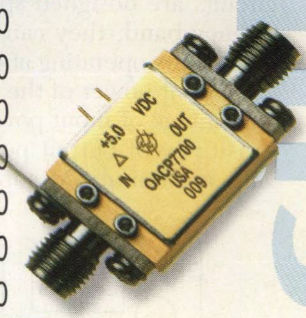
Oscillator with internal MMIC amplifier available in SMTO-8 or CougarPak™.								
OAS5100	4600-5100	0-15	10/2.0	-70/-100	30	-25	5.0	100
OAS6100	4700-6100	0-15	10/2.0	-70/-100	100	-25	5.0	100
OAS6500	5100-6500	0-15	10/2.0	-68/-98	100	-25	5.0	100
OAS6700	5400-6700	0-15	10/2.0	-68/-98	90	-25	5.0	100
OAS7700	5700-7700	0-15	10/2.0	-68/-98	150	-25	5.0	100
OAS8600	7000-8600	0-15	10/2.0	-65/-95	200	-25	5.0	100
OAS8900	7300-8900	0-15	10/2.0	-65/-95	200	-25	5.0	100
Straight Oscillator available in SMTO-8 or CougarPak™.								
OS5100	4600-5100	0-15	0/3.0	-70/-100	30	-5	5.0	30
OS6100	4700-6100	0-15	0/3.0	-70/-100	100	-5	5.0	30
OS6500	5100-6500	0-15	0/3.0	-68/-98	100	-5	5.0	30
OS6700	5400-6700	0-15	0/3.0	-68/-98	90	-5	5.0	30
OS7700	5700-7700	0-15	0/3.0	-68/-98	150	-5	5.0	30
OS8600	7000-8600	0-15	0/3.0	-65/-95	200	-5	5.0	30
OS8900	7300-8900	0-15	0/3.0	-65/-95	200	-5	5.0	30

Specifications are typical.



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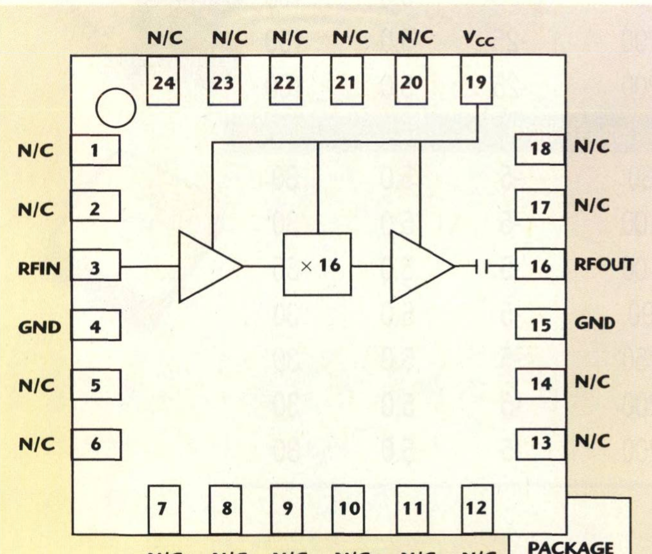
rent, respectively. All four active multipliers are supplied in 4.0 × 4.0 × 1.0 mm leadless surface- mount QFN plastic packages compatible with automated assembly and reflow soldering processes.

HBT ACTIVE MULTIPLIER ADVANTAGES

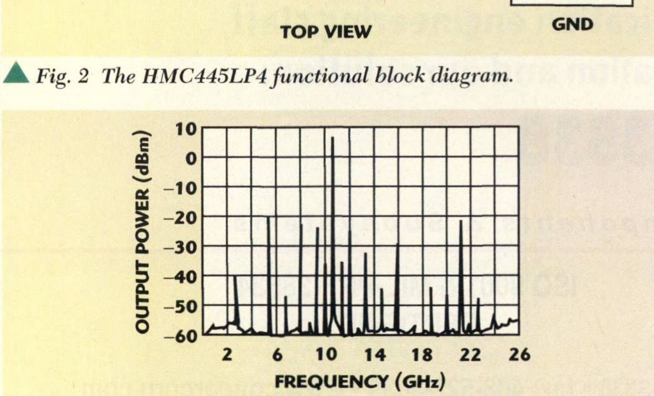
Active multipliers designed in GaAs/InGaP HBT technology operate on single positive (+5 V) bias, with low current (< 80 mA) and low residual phase noise (–130 to –140 dBc/Hz at 100 kHz) at frequencies > 16 GHz. The input and output of the HMC445LP4 multiplier is single-ended, as shown in ***Figure 2***. The only required external component is a bypass capacitor on the V_{cc} pin.

The ×16 architecture is implemented as a series of ×2 active multiplication stages. Each stage has integrated filtering for optimum subharmonic performance. This filtering circuit results in the narrow-bandwidth nature of the active multiplier line and is responsible for the excellent harmonic and subharmonic rejection of > 20 dB, as shown in the output spectrum of ***Figure 3***. Since these filtering circuits are designed specifically based on the input frequency band, they can be easily scaled to create active multipliers operating at other frequencies.

Output power of the HMC445LP4 device is flat over a wide range of input power, as shown in ***Figure 4***. The insensitivity of output power to input power variation results in natural AM suppression and coupled with excellent power flatness vs. temperature further simplifies synthesizer architecture by minimizing the need for limiting amplifiers or gain control.



▲ *Fig. 2 The HMC445LP4 functional block diagram.*



▲ *Fig. 3 The HMC445LP4 multiplier's output spectrum.*

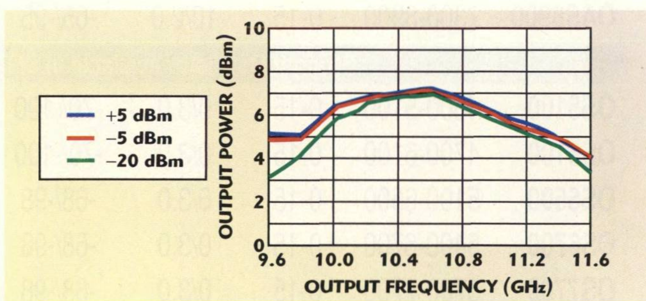
AN EXAMPLE 16 GHz APPLICATION

A typical application of the HMC370LP4 active ×4 multiplier as an LO generator for a microwave radio is shown in ***Figure 5***. LO chains for microwave radios in the 28 to 32 GHz band typically begin with an L-band source that is frequency multiplied to the transmit band through successive stages of amplification and frequency doubling. With the availability of low cost subharmonically pumped mixers for the 28 to 32 GHz band, the designer is only required to generate a 14 to 16 GHz LO signal at –5 dBm. A passive doubler solution to this problem would require three stages of amplification/doubling with each successive stage increasing in cost due to the drive level and filtering requirements. With the wide input drive range of the HMC370LP4 multiplier, the entire chain can be replaced with one passive doubler and one active ×4 multiplier with no additional amplification or filtering.

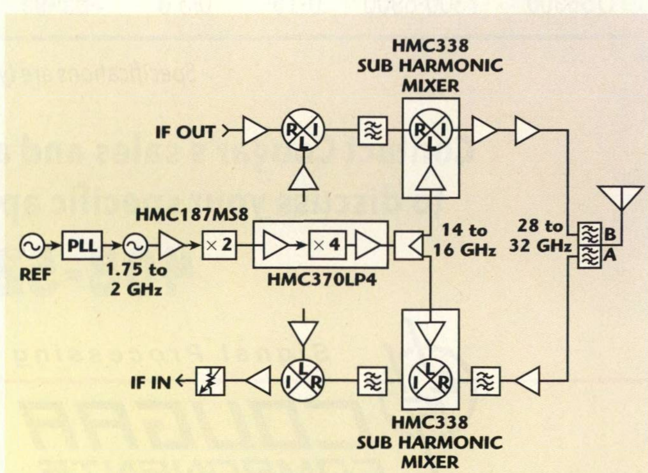
STANDARD ACTIVE DIVIDER PRODUCTS

The company has expanded its line of digital frequency dividers with the addition of the models HMC437MS8G (/3) and HMC438MS8G (/5) that operate over a DC to 7 GHz frequency range.

The key performance parameters of the company's family of GaAs/InGaP HBT frequency dividers are summarized



▲ *Fig. 4 The HMC445LP4 multiplier's output power vs. input drive level.*



▲ *Fig. 5 Typical 16 GHz multiplier application.*

[Continued on page 124]

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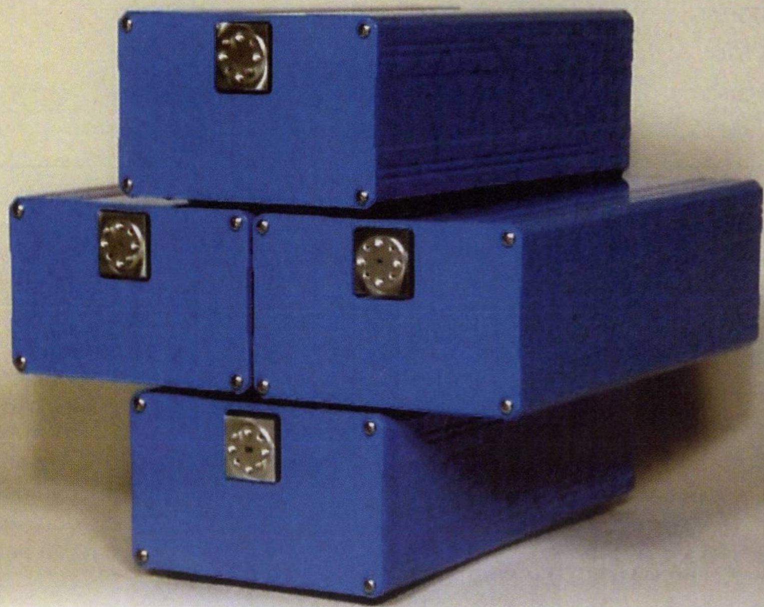

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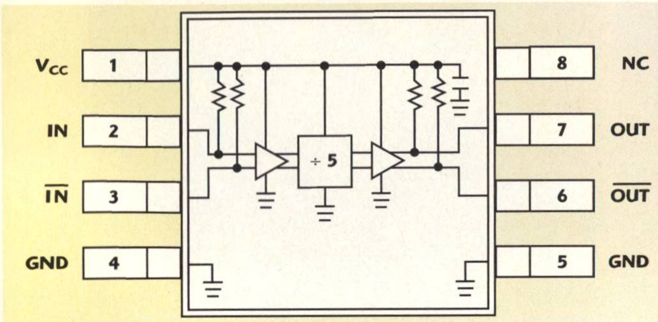
TABLE II

FREQUENCY DIVIDER PERFORMANCE PARAMETERS

Part Number	Function	Input Frequency (MHz)	Output Frequency (GHz)	Input Power (dBm)	Output Power (dBm)	SSB Phase Noise at 100 kHz (dBc/Hz)
HMC364S8G	÷2	DC to 12.5	DC to 6.25	–10 to +5	–8 to +5	–145
HMC437MS8G	÷3	DC to 7.0	DC to 2.60	–10 to +10	–4 to –1	–153
HMC365S8G	÷4	DC to 13.0	DC to 3.25	–15 to +10	+2 to +5	–151
HMC438MS8G	÷5	DC to 7.0	DC to 1.60	–12 to +12	–4 to –1	–153
HMC363S8G	÷8	DC to 12.0	DC to 1.50	–15 to +10	–9 to –6	–153



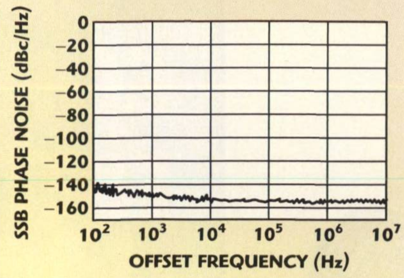
▲ *Fig. 6 The HMC438MS8G divider's functional block diagram.*

in ***Table 2***. The HMC437MS8G and HMC438MS8G operate from a single +5 V supply, have differential inputs/outputs and consume 69 and 80 mA of DC current, respectively. Both new dividers are supplied in 3.0 × 4.9 × 1.0 mm surface-mount plastic packages compatible with automated assembly and reflow soldering processes.

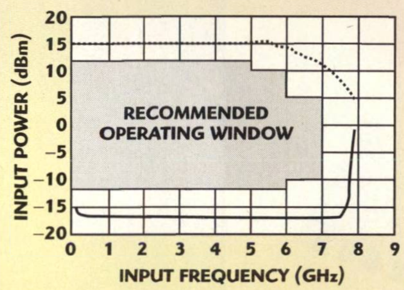
HBT DIGITAL DIVIDER ADVANTAGES

Digital dividers designed in GaAs/InGaP HBT technology operate on single positive (+5 V) bias, with low current (< 80 mA) and low residual phase noise (–145 to –153 dBc/Hz at 100 kHz) at frequencies up to 13 GHz. The input and output of the HMC437MS8G and HMC438MS8G dividers feature differential inputs, as shown in ***Figure 6***, but can be operated single-ended as well.

[Continued on page 126]

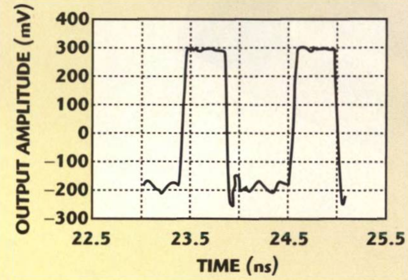


▲ *Fig. 7 The HMC438MS8G divider's SSB residual phase noise at P_{in} = 0 dBm, F_{in} = 6 GHz and T = 25°C.*

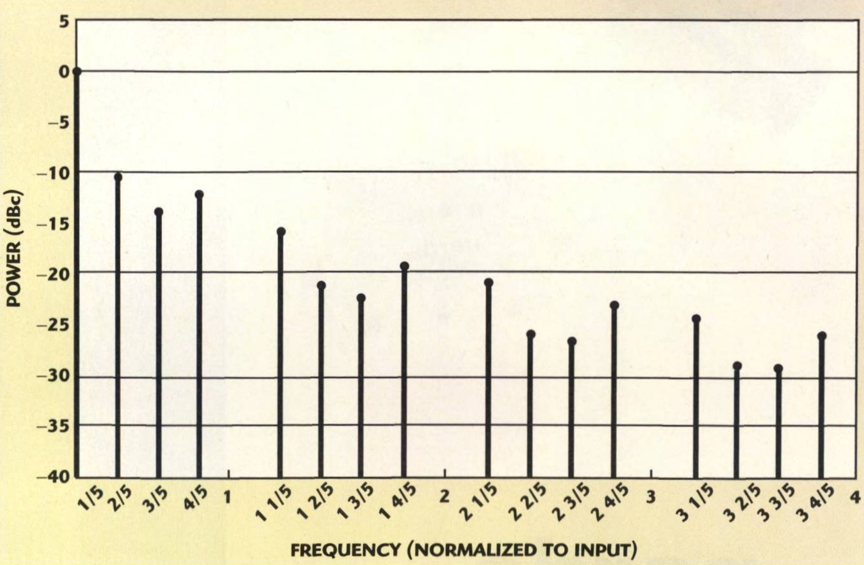


▲ *Fig. 8 The HMC438MS8G divider's input sensitivity.*

The low residual output phase noise and wide input power operating range are shown in ***Figures 7*** and ***8***. The output waveform of the HMC438MS8G divider is shown in ***Figure 9***. The 40 percent duty cycle results in an output frequency spectrum that has low (< −10 dBc) harmonics (***Figure 10***). In sharp contrast, a digital counter set to divide by five typically maintains the input carrier pulse width and hence has approximately a 10 percent duty cycle resulting in the rich harmonics shown



▲ *Fig. 9 Output voltage waveform of the HMC438MS8G /5 divider at P_{in} = 0 dBm, f_{out} = 882 MHz and T = 25°C.*



▲ *Fig. 10 Harmonic content of an ideal /5 divider with 40 percent duty cycle.*

[Continued on page 128]

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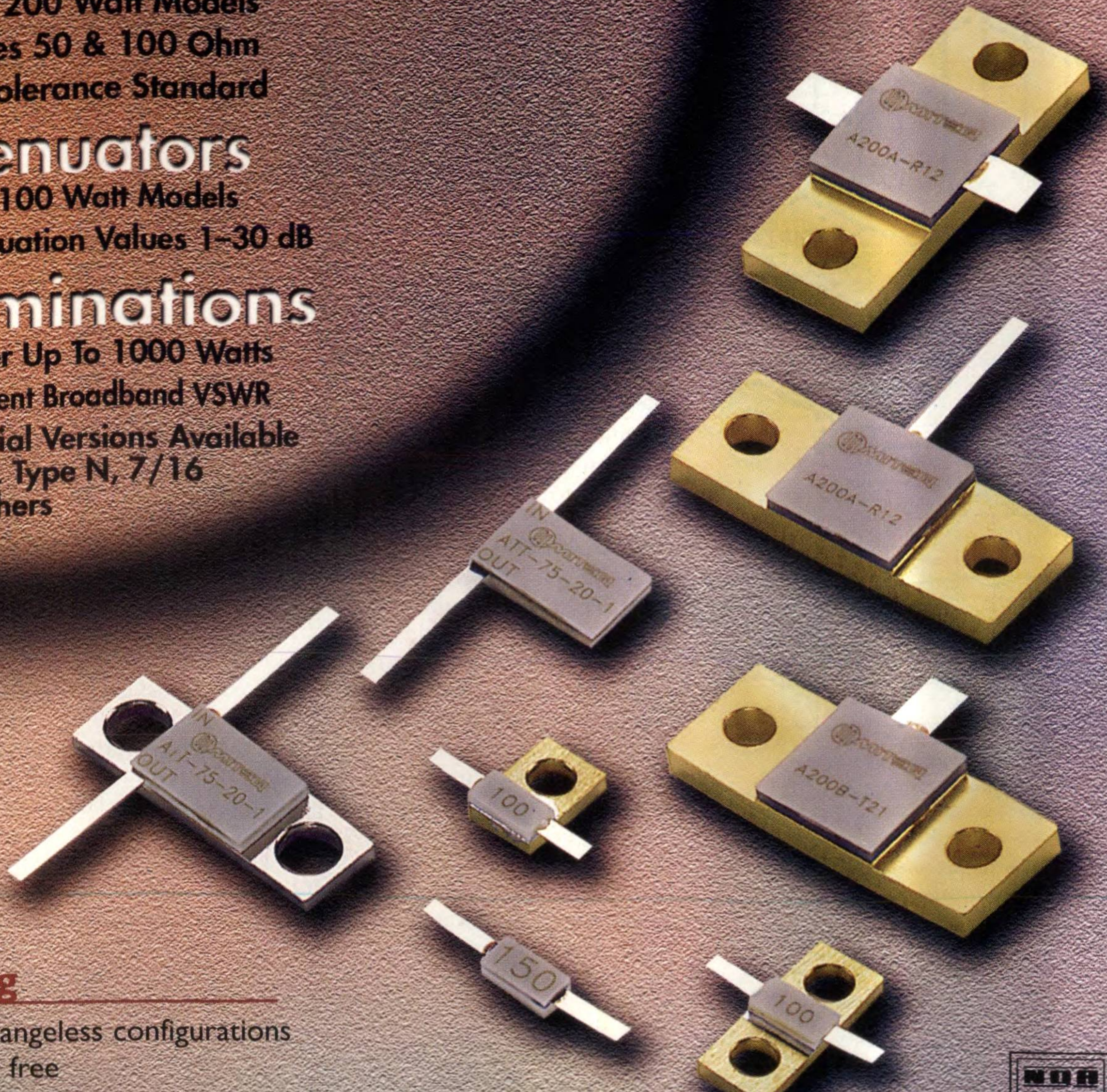

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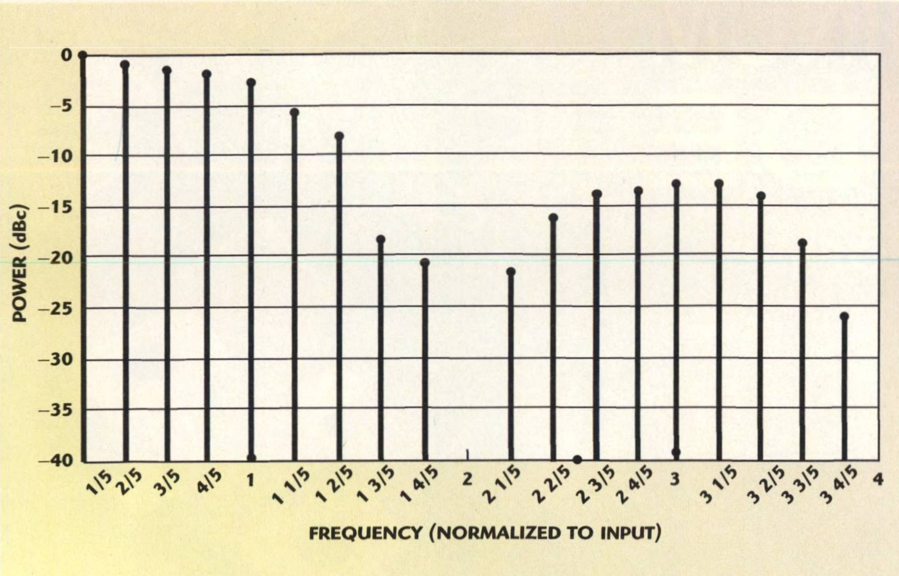
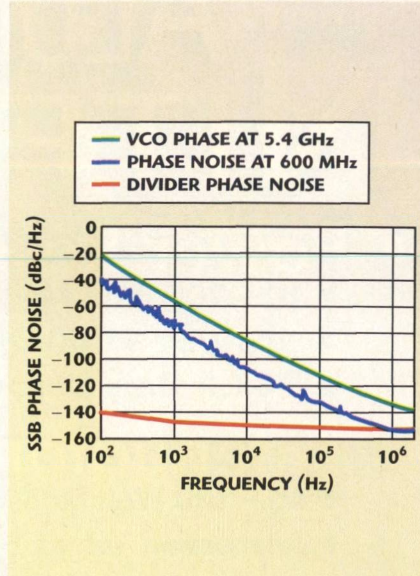
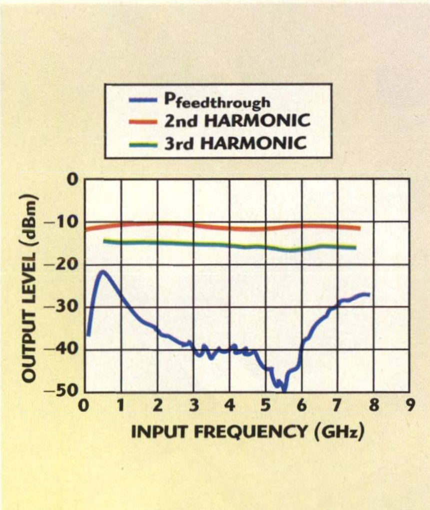


▲ *Fig. 11 Harmonic content of an ideal /5 counter with 10 percent duty cycle.*



▲ *Fig. 14 Base station LO phase noise.*



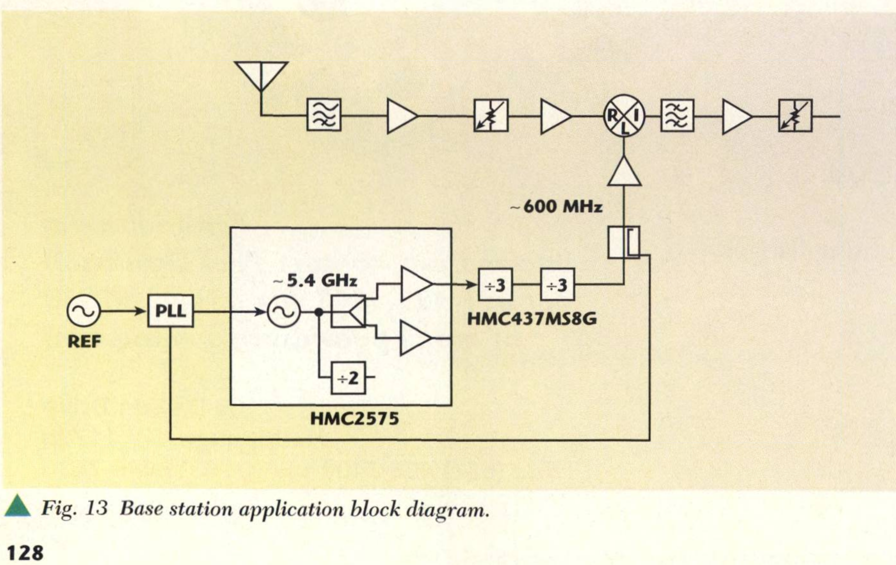
▲ *Fig. 12 Output harmonics and feedthrough suppression of the HMC438MS8G /5 divider at P_{in} = 0 dBm and T = 25°C.*

in ***Figure 11***. Since the fifth harmonic of the output is also equal to the input frequency, the HMC438MS8G device has excellent feedthru suppression, as shown in ***Figure 12***.

AN EXAMPLE 5 GHz APPLICATION

A typical application of the HMC437MS8G /3 divider as an LO generator for a mobile phone base station is shown in ***Figure 13***. LO chains for base station applications typically begin with a UHF VCO locked to a crystal frequency source that is frequency multiplied and amplified to the transmit/receive frequency. With the availability of low cost, low noise C-band VCOs and high frequency dividers, the same LO signal can be generated by a C-Band VCO and frequency division. In the example shown, a 5.4 GHz VCO output directly drives two successive /3 dividers to create the required 600 MHz LO frequency. The entire frequency generation chain consists of three low cost plastic-packaged components with no intermediate amplification or filtering stages.

The 600 MHz LO phase noise is shown in ***Figure 14***. The VCO phase noise (top curve) is reduced through frequency division by 20log(1/9) = −19 dB. At carrier offsets greater than 1 MHz, the divider residual phase noise (bottom curve) limits the output phase noise to approximately −154 dBc/Hz.



▲ *Fig. 13 Base station application block diagram.*

CONCLUSION

A family of active frequency multipliers and frequency dividers has been introduced that allows the synthesizer designer to take new approaches to solving traditional microwave radio and radar application problems. These GaAs/InGaP HBT devices are available in low cost plastic packages operating from single +5 V supplies. Conversion gain and excellent harmonic/subharmonic suppression minimize external support circuitry. Product samples and connectorized evaluation boards are available on request directly from the factory or online at www.hittite.com.

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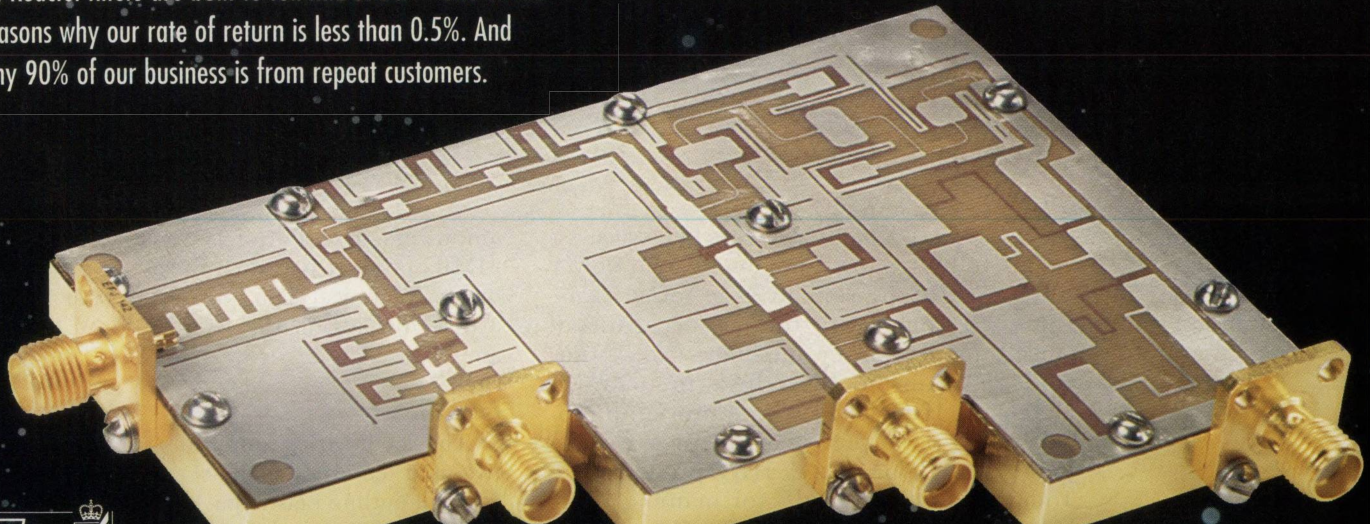

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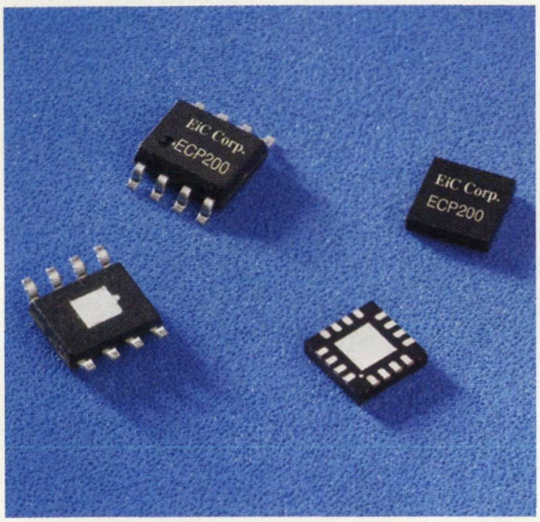


A COST-EFFECTIVE 2 W GAAS MMIC AMPLIFIER SOLUTION

GaAs HBT technology has been successfully deployed for low to mid-power MMIC RF amplifiers in recent years. EiC Corp. pioneered the use of an InGaP emitter for RF power amplifiers and first introduced gain block products in 1998. InGaP/GaAs HBT devices have proven to provide better performance and be more reliable than their AlGaAs/GaAs HBT predecessors. Today, other semiconductor manufacturers have followed suit, and InGaP HBT products are superceding AlGaAs.

Recently, InGaP HBT capability has been taken to the next level with the introduction of a series of intermediate power amplifiers. Using the same process applied to already-offered lower power gain blocks, one-half-, one- and two-watt MMIC RF amplifiers are now available. The ECP05x series are 1/2 W, the ECP10x are 1 W and the ECP20x are 2 W amplifiers. Exceeding 1 W capability for a MMIC product is truly a milestone in the industry. Technical data referred to later in this article focuses on the ECP200 2 W MMIC amplifier.

ECP DEVICE FEATURES

The ECP-series amplifiers are cost-effective, and feature a single-stage MMIC design in small outline plastic packaging with built-in ESD protection and the reliability benefits of InGaP HBT technology. The new amplifiers offer significant design flexibility by providing V_{de} up to 7 V, class A/AB selectivity, temperature compensation bias and power up/down sequencing. In addition, partial input match-

[Continued on page 132]

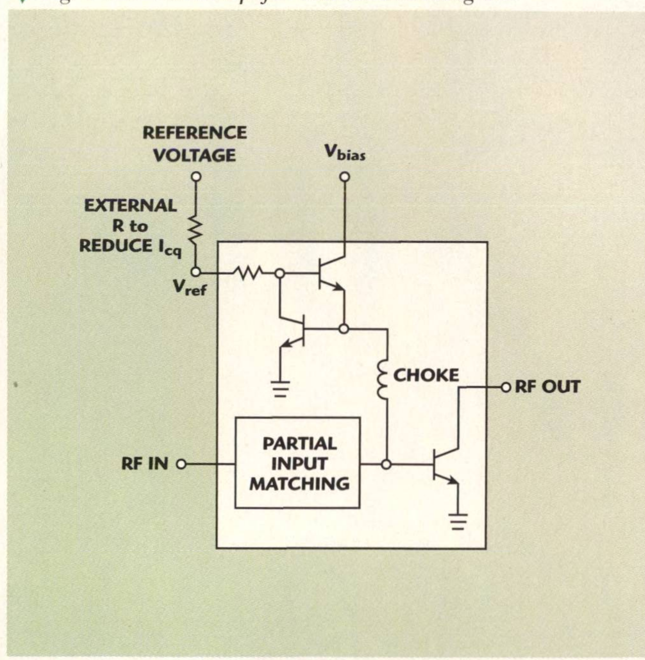
EIC CORP.
Fremont, CA

TABLE I
ECP-SERIES INTERMEDIATE POWER AMPLIFIER CHARACTERISTICS

Model	P1dB* (dBm)	Gain* (dB)	OIP3* (dBm)	Frequency (MHz)
ECP050	27	15	45	1800 to 2300
ECP052	27	15	45	800 to 1000
ECP053	28	17	45	2100 to 2700
ECP100	31	12	47	100 to 2300
ECP103	31	12	47	2100 to 2700
ECP200	33	10	49	100 to 2300
ECP203	33	11	49	2100 to 2700

*Typical values as stated on the data sheets

▼ *Fig. 1 ECP-series amplifier's circuit block diagram.*

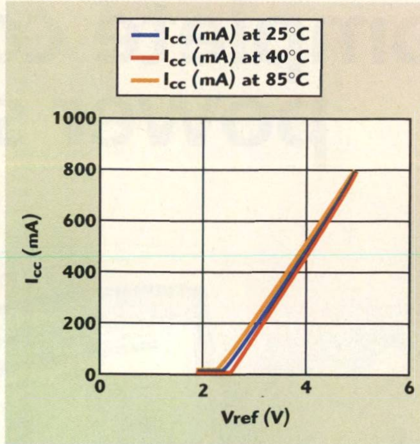


ing circuitry is built on-chip. ***Table 1*** lists the operating characteristics of the intermediate ECP-series.

TEMPERATURE COMPENSATED CURRENT MIRROR BIAS CIRCUIT

Figure 1 shows a block diagram of the amplifier bias circuit. The amplifier is a single-stage design. The output matching circuit is provided off-chip and the collector current is provided through an RF choke of the output matching circuit. The bias circuit is based on the current mirror principle and provides compensation over temperature for the bias current. V_{bias} requires a DC bias voltage, usually connected to V_{cc}, to provide the bias to the current mirror circuitry.

[Continued on page 134]



▲ *Fig. 2 ECP200 amplifier's I_{cc} vs. V_{ref} performance over temperature.*

Figure 2 shows measured results for the class A bias condition for the 2 W, ECP200 amplifier over temperature. The RFout and V_{bias} are at 5 V, and the I_{cc} is measured against V_{ref} as V_{ref} is swept to 5 V. The measurement was performed over a –40° to +85°C ambient temperature range. The result clearly shows the bias current changes little over temperature. The current mirror bias circuit also provides power down capability. As shown in the data plot, when V_{ref} is less than 2 V, the RF transistor is turned off. The power down capability allows the user greater flexibility in system design.

TABLE II

THE AMPLIFIER'S CLASS A AND AB CHARACTERISTICS

Class A	Class AB
Highest OIP3	Lower OIP3 than Class A
High power dissipation at back-off of RF power	Lower power dissipation (higher efficiency) at back-off of RF power than Class A
NF and P1dB are similar	NF and P1dB are similar

SINGLE-STAGE AMPLIFIER WITH ON-CHIP ESD PROTECTION

The ECP family features a single-stage design approach that provides flexibility for its users. ESD protection is included on-chip, protecting the circuit against 2000 V HBM (human body model). The ESD protection is not available with other discrete component solutions.

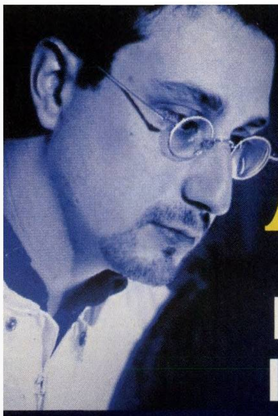
PARTIAL INPUT MATCHING CIRCUIT

As shown in the block diagram, the RF HBT is matched at the input, on-chip. For the ECP200 example, this input matching circuit is designed for the 2.14 GHz, W-CDMA base station transmitter frequency band. For lower frequency bands, the on-chip input matching circuit offers partial matching. Further matching made on the PCB will bring the input return loss to the desired level.

CLASS A OR CLASS AB BIAS

Table 2 compares the class A and class AB operation of the amplifier. The major trade-off is the output third-order intercept point (OIP3) with the power dissipation (efficiency) in a back-off state. For applications where OIP3 is not the top priority, class AB offers higher efficiency. Probably the most interesting feature of the ECP-series to the design-

[Continued on page 136]



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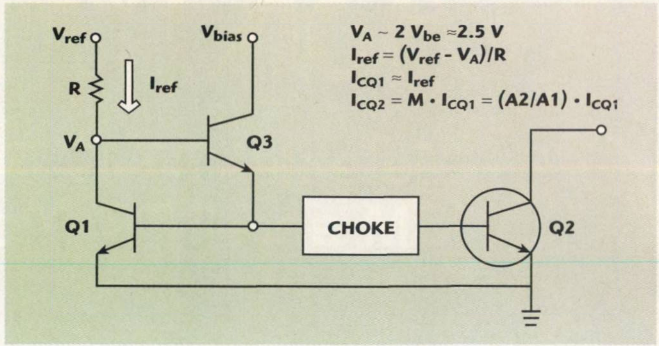


▲ *Fig. 3 Current mirror operation.*

er is the flexibility of bias selection that allows the user to operate the amplifier in either class A or class AB, or anywhere in between.

The bias current of an ECP amplifier can be easily set (selected) in two ways: (1) by setting the V_{ref} voltage or (2) by adjusting the value of the external resistor on the V_{ref} pin. Method 1 is clearly shown in the I_{cc}/V_{ref} data plot. I_{cc} versus V_{ref} is a linear relationship and the quiescent current can be set at any desired level, with little variation over temperature.

The user can change the external resistor value (Method 2) to modify the slope of the I_{cc} vs. V_{ref} response if needed for varying control capability. Method 2 can be understood by examining the diagram shown in ***Figure 3***. I_{cq2} of the RF transistor Q2 is proportional to I_{cq1} according to the ratio of transistor size, A2/A1. When I_{cq1} ($\sim I_{ref}$) is increased, I_{cq2} is increased accordingly. Since $I_{ref} \sim (V_{ref} - 2.5\ V)/R$, the quiescent current I_{cq2} can be adjusted by the resistor value of R, while keeping V_{ref} at a fixed level.

By adding an external resistor on the PCB, the total value of R is increased. The reference current I_{ref} and quiescent current I_{cq2} are reduced accordingly, changing the bias condition from class A to class AB.

Vcc UP TO 7 V

The HBT breakdown voltage is over 20 V for BV_{cbo} and over 10 V for BV_{ceo}. Therefore, the amplifier can operate at higher than 5 V collector voltage. However, the ESD protection circuitry limits the collector voltage to approximately 12 V, which means V_{cc} is limited to 7 V to avoid the kick-on of the ESD protection.

When the device voltage is raised, the quiescent current must be reduced accordingly so the total DC bias power is constant. For example, the ECP200 amplifier is biased at 5 V and 800 mA. At 7 V, the current must be reduced to 570 mA, so the total power dissipation is maintained at 4 W.

The input matching is only slightly affected by the change of the collector voltage. However, the load impedance should be adjusted according to

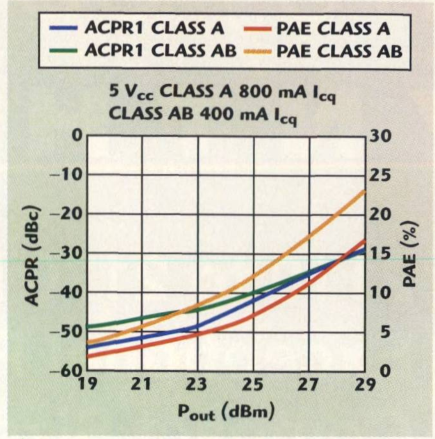
$$R_L = (V_{cc} - V_k)/I_{cc}$$

Since V_{cc} is increased and I_{cc} is reduced, the load resistance will increase accordingly.

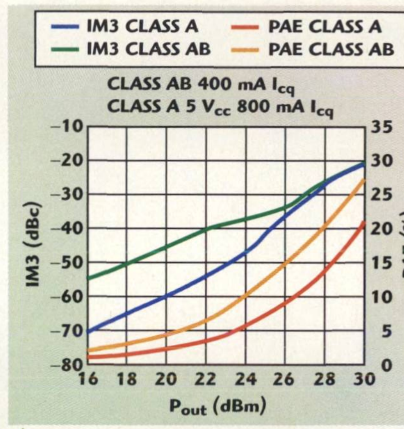
HIGH LINEARITY

Many wireless communication systems require high linearity as a result of complicated signal modulation and multi-carrier applications. As a result, the demand on OIP3 is

[Continued on page 138]



▲ *Fig. 4 ECP200 ACLR vs. P_{out} under class A and class AB operation at 2.14 GHz and 25°C.*



▲ *Fig. 5 ECP200 two-tone IM3 and efficiency vs. SCL output power at 2.14 GHz and 25°C.*

ever increasing. A high OIP3 can be achieved by using either a higher output power transistor or a more linear transistor. The higher output power transistor requires more DC power consumption, which is undesirable. Therefore, a more linear transistor is preferred.

The ECP-series is a family of high linearity amplifiers. The linear figure of merit (OIP3-P1 dB) for EiC's InGaP HBT process has historically been from 15 to 17 dB. This high figure of merit has been maintained for the higher power ECP-series devices and provides outstanding performance to satisfy many infrastructure applications.

Figure 4 shows the linearity results using a W-CDMA signal for a base station. Both class A and class AB (50 percent class A quiescent current) were tested. The adjacent channel leakage ratio (ACLR) for both bias conditions is similar at the same output power, but the efficiency is higher for the class AB mode.

Figure 5 summarizes the two-tone test results. The third-order intermodulation (IM3) in class A is better than class AB at lower power levels, until both reach 26 dBm single carrier level (SCL). Twenty-six dBm SCL corresponds to 32 dBm peak envelope power (PEP); therefore, above this power level, waveform clipping starts and the IM3 in both bias conditions becomes similar.

CONCLUSION

It can be seen from the reported test results that the ECP200 intermediate power amplifier, as well as its lower power bretheren, demonstrates outstanding performance. Now the RF designer has a truly cost-effective option for higher power amplifier requirements, a result of the company's leading edge InGaP/GaAs HBT technology. The devices can also be configured in parallel to meet even higher power out requirements, well under the cost of existing solutions.

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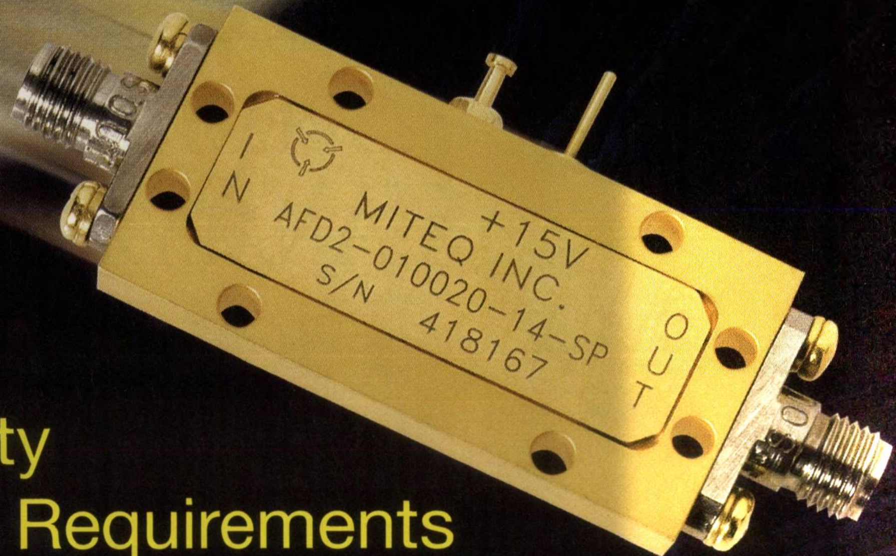

Amplifiers for a Variety of System Requirements

MODEL NUMBER	FREQ. (GHz)	GAIN (dB, Min.)	GAIN FLATNESS (±dB, Max.)	NOISE FIGURE (dB, Max.)	IN/OUT VSWR	POWER OUT (dBm, Min.)	CURRENT (mA, Typ.)
AFD2-010020-14-SP	1–2	20	1.50	1.4	2.0:1	+10	100
AFD3-010020-14-SP	1–2	34	1.25	1.4	2.0:1	+10	120
AFD3-022023-12-SP	2.2–2.3	30	0.50	1.2	1.5:1	+10	100
AFD3-023027-12-SP	2.3–2.7	30	0.50	1.2	1.5:1	+10	125
AFD3-027031-12-SP	2.7–3.1	30	0.50	1.2	1.5:1	+10	125
AFD3-031035-12-SP	3.1–3.5	30	0.50	1.2	1.5:1	+10	125
AFD3-037042-12-SP	3.7–4.2	30	0.50	1.2	1.5:1	+10	125
AFD3-040080-35-SP	4–8	24	1.25	3.5	2.0:1	+10	150
AFD3-020080-40-SP	2–8	23	1.50	4.0	2.0:1	+10	160
AFD3-040120-55-SP	4–12	18	1.50	5.5	2.0:1	+10	150
AFD3-080120-50-SP	8–12	18	1.25	5.0	2.0:1	+10	150
AFD1-010020-23P-SP	1–2	11	1.00	4.0	2.0:1	+23	200
AFD2-010020-23P-SP	1–2	25	1.50	3.5	2.0:1	+23	325
AFD3-020027-23P-SP	2.0–2.7	22	1.25	4.5	2.0:1	+23	350
AFD3-027031-23P-SP	2.7–3.1	22	1.25	4.5	2.0:1	+23	400
AFD3-031042-23P-SP	3.1–4.2	22	1.25	4.5	2.0:1	+23	350
AFD3-040080-23P-SP	4–8	20	1.25	5.5	2.0:1	+23	350
AFD3-020080-20P-SP	2–8	18	1.50	6.0	2.0:1	+20	350
AFD3-080120-20P-SP	8–12	15	1.25	5.0	2.0:1	+20	350
AFD3-040120-18P-SP	4–12	15	1.75	6.5	2.0:1	+18	350

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The FSH3 handheld spectrum analyzer is available in two versions, either as a pure 100 kHz to 3 GHz spectrum analyzer or as a spectrum analyzer with a tracking generator for scalar network analysis. Equipped with other options such as a power sensor and a VSWR bridge, it can be used as a power meter, a scalar network analyzer or an analyzer for distance-to-fault measurements on cables.

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At the heart of the instrument is an ASIC processor with a 32-bit RISC processor, a display controller and elements for controlling the periphery such as the RS-232 interface or the power sensor. The processor is used for operating the instrument, controlling the mea-

[Continued on page 142]

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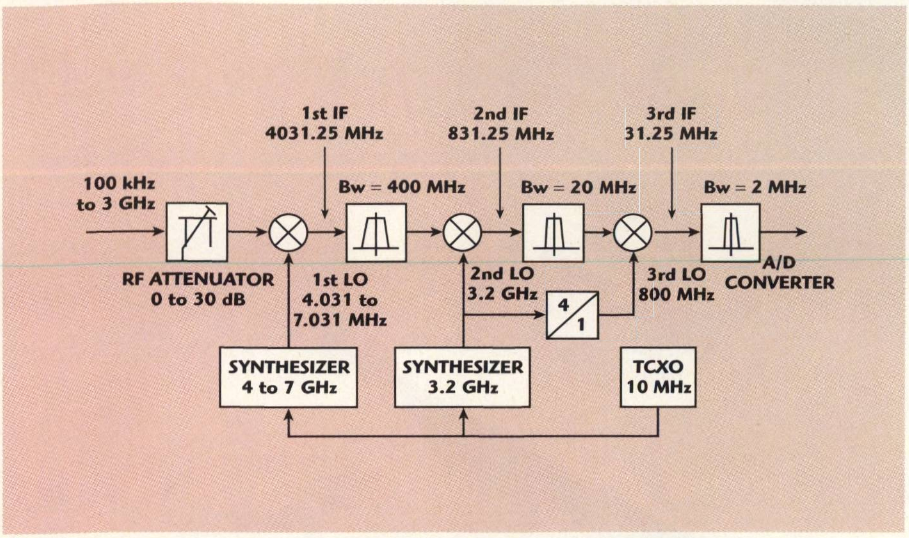
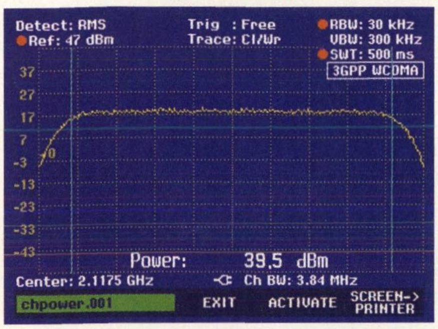


▲ *Fig. 1 The FSH3 spectrum analyzer's RF and IF section block diagram.*

Fig. 2 Power measurements on a 3GPP W-CDMA signal. ▼



surement sequences, and calculating and displaying the results. Management of the power supply and the nickel metal hydride battery is handled by a separate ASIC.

The RF receive path (shown in ***Figure 1***) is designed as a threefold converting superheterodyne receiver with a high first intermediate frequency. With damage prevention being a key consideration in such instruments, the RF input is particularly well protected by a combination of overvoltage arrester, PIN diodes and capacitive coupling so that the attenuator or the input mixer will not be damaged if high voltage is inadvertently applied or if electrostatic discharge occurs.

The instrument's mode of operation is such that the first local oscillator with a high frequency range of 4031.25 to 7031.25 MHz converts the input frequency to the first intermediate frequency (4031.25 MHz) and consists of three voltage-controlled oscillators (VCO), each covering a frequency range of 1 GHz. Two components specially developed for synthesizers — a high frequency predivider and a divider for fractional division ratios with integrated phase detector — generate a sweep that is synchronized to the 10 MHz temperature-compensated crystal oscillator (TCXO) at each frequency point. The FSH3 spectrum analyzer therefore represents receive signals at the correct frequency even with large display ranges of the order of 3 GHz.

Signal processing starting at the last intermediate frequency is purely digital. An A/D converter digitizes the 31.25 MHz IF signal and two integrated circuits perform all other processing steps such as IF filtering, envelope detection, logarithmation, video filtering and signal detection in real time. The main processor retrieves and displays the configured data.

Using this concept the spectrum analyzer offers the functions and characteristics provided by desktop units but implements them on the small footprint of a handheld unit. For instance, it offers resolution bandwidths from 1 kHz to 1 MHz in 1-3-10 sequence. These bandwidths are designed for optimum settling in the frequency sweep with Gaussian filter characteristics. The common method of using a four-pole analog filter with decoupled contacts also yields a virtually Gaussian filter characteristic in the passband but at the expense of selectivity. Here, the shape factor of the 60 dB bandwidth to the 3 dB bandwidth would typically be < 15. However, much higher selectivity can be obtained with digital implementation, so with the FSH3 instrument the shape factor of the 60 dB bandwidth to the 3 dB bandwidth is < 6. This makes it much easier to separate adjacent signals, especially if they differ greatly in amplitude.

Another feature is the practically error-free rectification and logarithmation of the filtered IF signal, whereby the IF signal envelope is determined mathematically. After envelope detection, the signal is logarithmized for logarithmic representation. Since the two operations are performed by calculation in the digital circuits they are practically error-free and the deviation of the display linearity from the ideal value depends almost exclusively on the linearity of the A/D converter.

Video bandwidths of 10 Hz to 1 MHz are also in 1-3-10 sequence. The attenuation characteristic of the video low pass filters is similar to that of RC low pass filters. This ensures that the smoothing of the trace is similar to that of analyzers with analog RC low pass filters when the video bandwidth is reduced.

The handheld spectrum analyzer also utilizes different detectors for signal weighting, namely auto-peak, peak, sample and the RMS detector that is applied to the power measurement of modulated signals. Since the detectors are digital, there are distinct advantages. The peak detector, for example, is free from effects often experienced by analog versions, such as charge time, discharging and storage effects.

There is also the advantage that an RMS detector can be implemented, which means that the power of signals can be measured without regard to the detector characteristic in the same way that a thermal power meter does. An RMS detector is particularly important in measuring the power of modulated signals such as W-CDMA signals, an example of which is shown in ***Figure 2***. With the channel power function the instrument measures the power accurately and reproducibly in a single sweep.

Due to the digital design concept the functions referred to are highly constant and reproducible. For example, the error caused by a bandwidth switchover is negligible. The display linearity depends entirely on the linearity of the A/D converter whose linearity error is also negligible in practice. The level measurement uncertainty is almost exclusively a factor of the absolute gain of the IF section and of the frequency response of the attenuator and the input mixer, with the FSH3 correcting the two parameters during measurement. The frequency response is stored in the instrument for all RF attenuator settings and during measurement the displayed level is corrected at each frequency. The instrument also monitors its internal temperature and corrects the total gain accordingly, with the result being a specified total level measurement uncertainty of 1.5 dB max.

Despite low power consumption and associated limited processor power, the digital design concept produces a relatively high measurement speed, due to the ASICs processing the IF data in real time. The minimum sweep time is only 100 ms if the entire frequency domain is displayed, while with time domain measurement (span = 0 Hz), the minimum sweep time is just 1 ms. The power available (7 W with tracking generator for an operating time of 3.5 hours with one battery charge) can thus be used more effectively for the RF dynamic range of the RF front-end. With a typical third-order intercept of 15 dB, at a noise figure of 32 dB (typical), the spectrum analyzer attains values comparable to those obtained with high end desktop analyzers.

EASY OPERATION

Because the spectrum analyzer is used for field maintenance and service, it must be easy to operate when working under difficult conditions as well as when accessing functions and reading results. There is direct access to all basic functions at a keystroke without soft key control and the keys are arranged so that if holding the instrument with both hands all keys and the roll key can easily be reached with the thumbs. The analyzer has only one submenu level, and markers and delta markers come in handy when processing results on the 14 cm color LC display. Up to 100 measurements, including all settings, can be stored in the internal memory at the press of a button.

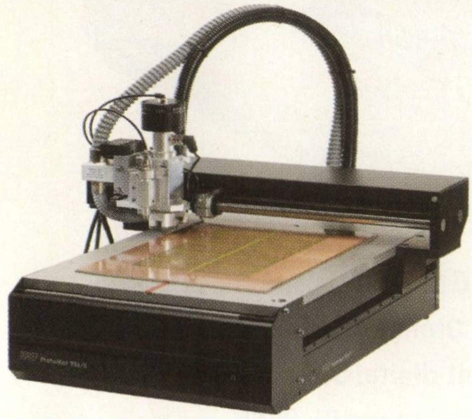
The instrument comes with a Windows™ software that takes the results of measurement from the analyzer and stores them in common graphics formats (Bitmap, Windows Metafile, PNG, PCX) as text files or in Excel™ format. The date and time of the measurements and all settings are provided to ensure transparent documentation, and additional comments can be added and stored in the software. Since a Word macro is available to generate a customer-specific report, the user can easily integrate the results into pre-defined forms.

CONCLUSION

The FSH3 handheld spectrum analyzer combines lightweight portability with sophisticated functionality and performance. It is user-friendly and easy to operate yet offers characteristics comparable with high end desktop analyzers.

Rohde & Schwarz GmbH & Co. KG,
Munich, Germany
+ 49 89 4129 11765, www.rohde-schwarz.com.

Circle No. 300



A Fast and Flexible Circuit Board Plotter

When employing circuit board plotters for product development the main requirement is for fast, efficient and flexible operation. Consequently, the high costs incurred by the long waiting periods inherent when externally procuring circuit board prototypes, particularly those with RF substrates, can be avoided. The ProtoMat 95s/II meets these requirements, and all without chemicals, as it utilizes pure mechanical techniques to produce a circuit board on a copper-coated base plate. The data is provided by a CAD development system, with data formats in Gerber, Excellon, DXF or HPGL all suitable for use.

In operation a cushion of air is created that enables the cutting head to glide over the material during processing to avoid direct contact with the surface. Permanent sensing of the surface of the circuit board material being processed guarantees precise compliance with the specified cutting depth.

Drives for the X and Y directions consist of five-phase stepper motors combined with precision recirculating ball screws and linear guides. This layout enables the plotter to structure circuit boards with a resolution of 5 μm and associated repetition accuracy <1 μm.

This process is supported by the cutting tools developed by LPKF. The RF-cutters,

The ProtoMat 95s/II at the Fraunhofer Institute

The Fraunhofer Institute for Integrated Circuits (IIS-A) in Erlangen, Germany, is one of the leading institutions in Europe for applied research and development in the electronics sector. Starting from concepts and taking them through to complete products, the institute provides services to industrial companies and public institutions.

This work often involves numerous steps along the development process and usually cannot be simulated using computer models alone. They actually have to be built, which has resulted in a successful working relationship with the LPKF Group.

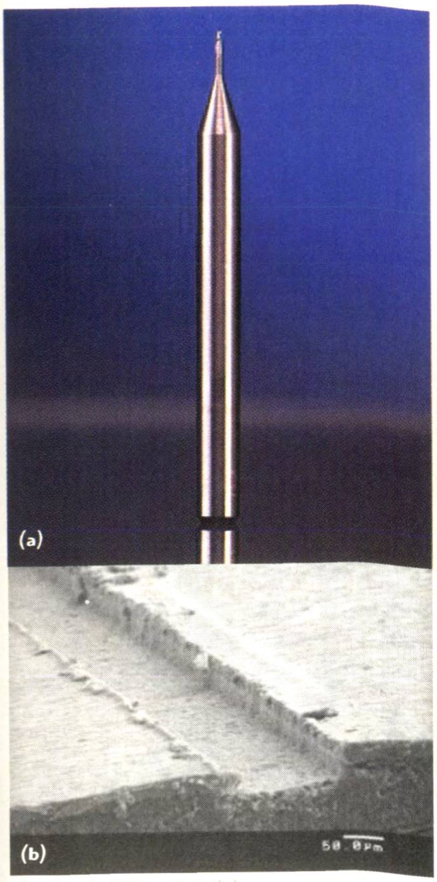
The ProtoMat 95s/II circuit board plotter is playing its part, as in the design and realization of adaptive multibeam or fractal multiband antennas. Another core area of activity is in RF measuring and system technology, involving the development and testing of RF modules for professional applications in digital radio repeaters and base stations.

LPKF Laser & Electronics
Garbsen, Germany

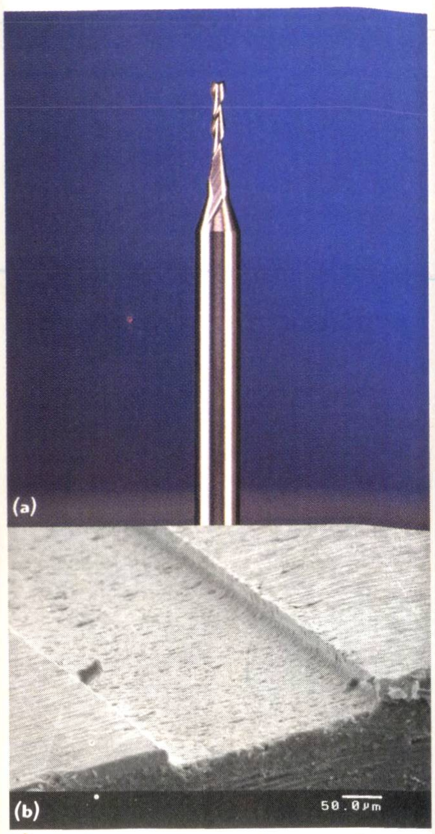
shown in ***Figure 1***, produce a right-angled cutting edge that ensures optimal compliance with the layout geometry and the best possible match with the mathematical specifications. They are available in diameters from 0.15 mm (6 mil) to 0.40 mm (16 mil). To cut wider insulation tracks, end-mills (shown in ***Figure 2***) are used, which also cut right-angled edges and are available in diameters ranging from 0.8 mm (31 mil) to 3.0 mm (118 mil). A key feature facilitating high speed, precision circuit board plotting is automatic tool changing that utilizes a magazine carrying 30 tools. This is particularly advantageous for the production of small series boards.

The plotter is suitable for all single-sided and double-sided circuit board types (including RF and microwave materials) in the highest

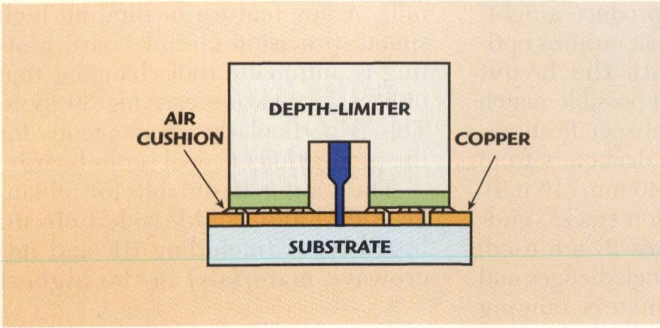
[Continued on page 146]



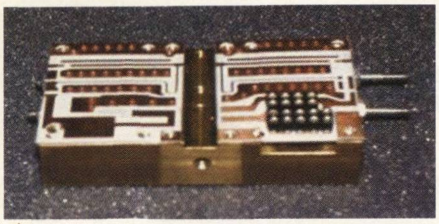
▲ *Fig. 1 RF-cutters (a) and the cut they produce (b).*



▲ *Fig. 2 Endmills (a) and an example of the cut they produce (b).*



▲ *Fig. 3 The contactless depth-limiter operation.*



▲ *Fig. 4 A high power transistor test adapter developed using the ProtoMat 95s/II.*

packing densities, and can be used for contour milling, front panel/label engraving and milling holes in aluminum front panels. In particular, the parameters of the machine, such as the rpm of the cutter or the advance and lowering speed (pneumatic lift), can be flexibly co-coordinated for the processing of standard RF and microwave substrates. A contactless depth-limiter (see ***Figure 3***) is ideal for thin material as well as flexible circuits, ensuring that the sensitive copper layer is not damaged during the processing of RF substrates.

Typical fields of application include antenna technology, RF measuring and system technology, the development of linear power amplifiers and the realization of laboratory prototypes for the matching of circuit elements, and the power supply of RF high power transistors, shown in ***Figure 4***. These are some of the areas of development undertaken at the Fraunhofer Institute (see sidebar).

CONCLUSION

The ProtoMat 95s/II is particularly suitable for cost-intensive development projects due to the wide range of materials it can process, the wide combination of boards and surface substrates offered, and its ability to adapt to differing production specifications. Speed and flexibility are key features and although the machining time is dependent on track density, simple boards take just minutes while even complex multilayer boards only take a few hours.

LPKF Laser & Electronics, Garbsen, Germany
Tel: + 49 (0)5131 7095-0;
Fax: + 49 (0)5131 7095-90;
e-mail: rp.sales@lpkf.de;
Web site: www.lpkf.com.

LPKF Laser & Electronics Inc., USA
Tel: 1-800-345-5753;
e-mail: sales@lpkfusa.com;
Web site: www.lpkfusa.com.

Circle No. 301

PRODUCT FEATURE

PRECISION E_B/N_O GENERATORS FOR BER TESTING TO 44 GHZ

With the continually increasing complexity of digital modulation techniques employed in wireless and satellite communications systems, instruments that accurately generate and maintain a set level of bit energy-to-noise density ratio (E_b/N_o) have become valued tools in optimizing bit-error rate (BER) performance. These instruments have been steadily enhanced over the years with greater accuracy, higher frequency ranges, wider power range and control, and many other features that help make characterizing digital receiver performance more straightforward. The Noise Com UFX-EbNo instrument is an E_bN_o generator that takes this performance a step further with comprehensive automation, output frequencies to 44 GHz and greater reliability.

The ratio of the actual E_bN_o to the required E_bN_o for a desired BER (system margin) is the broad determinant of how well a wireless communication system will perform under signal conditions that are affected by factors such as rain, hail or dust-induced attenuation. Every modulation scheme, such as BPSK, QPSK and QAM, has a different curve illustrating its theoretically-achievable BER for a given E_bN_o. The sharp E_bN_o curves of complex digital modulation schemes drive the need for extremely accurate instruments that generate a precise level of E_bN_o over a wide range of power levels and operating frequencies.

The UFX-EbNo generator is designed to satisfy these requirements. It is the successor to the UFX-Series, which was one of the industry's first solutions for E_bN_o generation, and was incorporated in test systems by manufacturers of subsystems and systems for wireless and satellite communications, as well as in various military test systems. The new instrument builds on the core capabilities of the company's precise noise generators, while focusing on the generation of accurate bit energy-to-noise density ratios rather than precise absolute noise power levels.

[Continued on page 148]

NOISE COM INC.
Parsippany, NJ

The basic application of the UFX-EbNo unit is to simplify and speed up the process of BER testing by automatically generating and setting carrier-to-noise ratio (C/N) and E_bN_o so that the performance of the system can be evaluated with confidence. It operates in five modes: C/N, E_bN_o, carrier-to-noise density (C/N_o), carrier-to-interferer (C/I) and as a true RMS power meter, while the focus is on E_bN_o and C/N ratios. It can also function as a precision generator of white Gaussian noise. All of the relevant input and output signal levels of the chosen operating mode can be displayed simultaneously, including carrier-to-noise ratios.

The various functions of the UFX-EbNo generator are highly automated, and the instrument compensates for bit rates, bandwidth and other factors. Most measurements can be performed by pressing a single button. For E_bN_o testing, the instrument automatically calculates noise density based on a user-specified bit rate, carrier output level and output E_bN_o ratio. To obtain a specific C/N_o ratio, the UFX-EbNo unit is simply set to E_bN_o mode and zero is entered for the bit rate. The instrument creates the required value of C/N_o.

The key to the accuracy and repeatability of this process is the instrument's ability to generate and maintain its specified characteristics over a wide range of power levels and operating frequencies. To achieve this, the UFX-EbNo generator is designed around precision components, including the calibrated noise sources, which have been the company's core product line since 1985. In addition, the instrument has an integral RMS digital power meter, which is customized for operation within the frequency range of the instrument, and provides a C/N accuracy of ±0.2 dB received signal strength (RSS) within its range of –55 to +5 dBm. Leveraging on the company's experience and the power meter technology of Boonton Electronics, a Wireless Telecom Group subsidiary, the instrument can measure both the carrier signal and noise with Gaussian noise crest factors up to 18 dB. The power meter uses various averaging methods to ensure accurate measurements, and all measurements are made through couplers that allow the signal to pass through to the output connector without degradation by the power meter circuitry.

The UFX-EbNo generator utilizes the substitution calibration method to set all of the desired carrier-to-noise ratios. This technique eliminates the nonlinearity effects that can result from the instrument's power meter by maintaining both signal and noise at the same level at the power meter's input. The noise power can then be offset by the desired ratio. While the attenuator that varies the instrument's noise output power is the greatest potential source of inaccuracy, its impact on overall performance is negligible because the attenuators are well-characterized, precision devices.

Secondary effects such as thermal drift are also negligible because the measurement samples of both noise and power are very short. Active components can also be contributors

[Continued on page 151]

TABLE I UFX-EbNo SPECIFICATIONS	
Operating modes	C/N, C/N_o, E_b/N_o, C/I, noise generator, power meter
Frequency ranges	5 MHz to 44 GHz, in standard or custom bands
Carrier path	
Input power range (dBm)	–55 to +5
Max. input power, no damage (dBm)	+21
Output power range (dBm)	–55 to +5
Gain resolution (dB)	0 to 60 in 0.1 dB steps
Gain flatness (dB)	±0.2 for 50 to 90 MHz
Noise path	
Output power range (dBm)	–55 to +5
Flatness (dB)	±0.2 dB/40 MHz
Attenuation range (dB)	60 in 0.25 dB steps (0.1 dB steps optional)
Ratio accuracy (dB)	0.2 RSS
Power meter	
Range (dBm)	–55 to +5
Averaging	10 to 65, 535 samples
Control	Local, IEEE-488-2 (RS-232C, RS-422, RS423, TCP/IP, 10/100 Ethernet optional)
Interferer input	–4 dBm ±2 dB; frequency range equals noise bandwidth
Power required	85 to 264 V AC, 2 A
Dimensions (in.)	17 × 5.25 × 17.5

to long term drift, but since they are common to the signal and noise paths, their variations do not affect the calibrated ratio. In addition, the signal path for both phase and amplitude is linear, which ensures that the desired signal passes through the instrument undistorted.

In every mode, data entry is a simple process, and front-panel indicators and a 4 × 20 line display provide clear indication of instrument settings. The UFX-EbNo instrument can be specified with remote control capabilities via IEEE-488-2, RS-232C, RS-422, RS-423, TCP/IP or Ethernet. A wide variety of options are available, such as a tracking feature that lets users with unstable input signals perform accurate long term testing at a given E_bN_o ratio by correcting input signal drift with 0.2 dB resolution to ensure a constant value of E_bN_o. Other options include 0.1 dB per step attenuation of output noise and C/N, dual channels, uninterrupted carrier during E_bN_o calibration, and 0.01 dB attenuation and E_bN_o resolution.

The number of standard frequency ranges available with the UFX-EbNo has been increased, and they now range (in bands) from 5 to 90 MHz, to 18 to 22 GHz, with other custom frequency ranges available between 5 MHz and 44 GHz. Each frequency range corresponds to a particular communications service, such as satellite communications frequencies, cable modems, 802.11a and 802.11b wireless LANs, and all current wireless communications bands. ***Table 1*** lists the generator's performance specificiations.

Noise Com Inc.,
Parsippany, NJ (201) 261-8797, www.noisecom.com.

Circle No. 304

NEW PRODUCTS

COMPONENTS

■ Ceramic Filter/Decoupling Capacitor

This ceramic filter/decoupling capacitor utilizes revolutionary X2Y® technology. The design incorporates common shielding electrodes that form a Faraday Cage around the standard capacitor. The addition of the common electrode creates two matched or balanced capacitors enabling simultaneous line-to-line and line-to-ground filtering. The matched capacitor design also eliminates temperature, voltage and aging differences. This component virtually eliminates parasitic inductance and provides superior filter and decoupling performance, allowing multiple component replacement and space savings on the PC board. The capacitor is available in chip sizes 0603 through 1812 and 1410. Price: 50¢ (1000). Delivery: stock to eight weeks.

Johanson Dielectrics Inc.,
Sylmar, CA (818) 364-9800.

Circle No. 218

■ WLAN Surface-mount Dielectric Resonator Filter

The model 3DL25-4950/X100 dielectric resonator filter is a three-pole bandpass Chebychev design used for WLAN applications. The unit's low profile, small size, low loss and surface-mount capability makes it suitable for this commercial application.

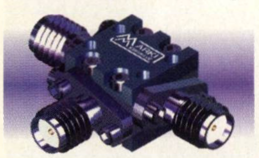
The filter has a low loss passband from 4900 to 5000 MHz, 2.0 SWR and 20 dB (min) rejection at 5250 to 5350 MHz. Dielectric resonator design provides for stable performance in the operating temperature range of –40° to +85°C. Size: 0.55" × 0.55" × 0.20".

K&L Microwave Inc.,
Salisbury, MD (410) 749-2424.

Circle No. 219

■ Mixer

The model M9-0950 mixer is a design unique to the company, covering 9 to 50 GHz on the RF and LO, with a 1 to 22 GHz IF. Created for wideband applications requiring a very high IF, it is a unique double-balanced mixer design, available in a 2.92 mm connectorized outline.



The local oscillator drive level is 9 to 13 dBm. Typical conversion loss is 10 dB.

Marki Microwave Inc.,
Morgan Hill, CA (408) 778-4200.

Circle No. 221

■ RF Relay

This miniature RF relay embodies new technology to provide low loss and high integrity switching of RF signals up to 2.5 GHz and 10 W. The relay is ideal for redundancy, filter and amplifier switching in mobile phone base station, wireless LAN, private mobile radio and associated test and measurement equipment. Engineered to minimize insertion loss and SWR, the relay offers good RF performance wherever high frequency isolation, termination and signal routing is required. The relay achieves HF characteristics of 60 dB isolation (min), and insertion loss of 0.2 dB (max) at 2.5 GHz (50 Ω). Size: 20.0 × 9.4 × 8.9 mm.

Omron Corp.,
London, UK +44 (0) 20 8450 4646.

Circle No. 223

■ Connector

This cut and fit solder connector is designed for annularly corrugated cables (7/8", 1-1/4", 1-5/8") and provides a simple and speedy soldered connection, resulting in a reduction of connector-sourced intermodulation harmonics.

The connector kit comprises a helical screw threaded inner conductor, a push-on slotted outer conductor assembly and a heatshrink sleeve. No solder or flux is required as the outer and inner parts of the connector are pre-loaded with enough solder in the optimal location to ensure an ideal electrical connection. The only additional tool required is a gas torch or an electrical heat gun.

Spinner GmbH,
Munich, Germany +49 89 12601-0.

Circle No. 225

■ Helical Filter

This helical filter has an input power of 2 to 4 W and is designed for D-amps, E-GSM, JCDMA, PDC, IMT2000 and PDC1.5 repeater systems. Suitable for all wireless systems and with good functionality, this filter exhibits a group delay variation of less than 20 ns.

Temwell Corp.,
Taipei, Taiwan 886-2-25652287.

Circle No. 228

■ Star Point Combiner

The star point combiner principle was chosen to implement a suitable filter combiner to achieve low attenuation loss with acceptable mechanical dimensions in combining the uplink and downlink ranges with the receiving and transmitting ranges in a common antenna field.

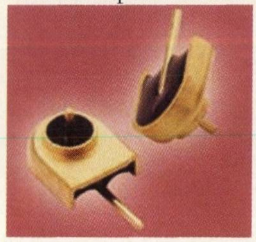
First the receiving and transmitting ranges of each of the two connections are separated in a two-way star point and routed through the filters. Then they are re-combined in a four-way star point and routed to the antenna. The combiner is made up of four temperature-compensated eight-cavity filters. They are combined in a star point network with strip line technology.

Spinner GmbH,
Munich, Germany +49 89 12601-0.

Circle No. 226

■ Surface-mount Feedthru

The Bell Pin™ surface-mount feedthru has been developed to work with a lightweight aluminum or brass housing, easily mounting to the bottom of the package creating true surface mountable parts with increased performance.



A SWR rating of less than 1.5 up to 15 GHz and offering true 50 Ω performance, this technique is an ideal solution to costly, heavy Kovar packaging problems. The Bell PIN SMT feedthru is made with a gold-plated Kovar base and center conductor fired in place with low loss corning glass. Advanced modeling techniques were used to achieve a 50 Ω frequency range compared to Kovar SMT housings. Hermetic reliability is exactly that of a round ferrule type feedthru.

Thunderline Z,
Hampstead, NH (603) 329-4050.

Circle No. 230

■ High Density Shielding Interconnects

These HDSI® assemblies are ideal for high speed digital or analog signal transmission in high end servers, telecommunication switches/routers and automatic test equipment. The low profile, high density assemblies offer optimal signal integrity, achieved through matched impedance, closely controlled line-to-line skew, reduced EMI and low crosstalk.

Flexible and low profile, the interconnect solutions are ribbonized for easy routing in tight spaces and board-to-board applications.

Tensolite,
St. Augustine, FL (877) 890-7483.

Circle No. 229

■ Chip Resistors

This line of miniature, high reliability chip resistors ranges in size from 0202 to 0705 with tolerances from 0.1 percent, power ratings from 20 to 200 mW, temperature coefficients of resistance as low as 25 ppm and voltage ratings from 15 to 50 V. The operating range for the resistors is from –55° to +125°C. A full range of termination styles and termination finishes are available. Reliability is assured by extensive in-house testing. The resistors are designed for use in a wide range of high reliability applications, including medical electronics, defense systems, microwave communications, aerospace electronics and satellite systems.

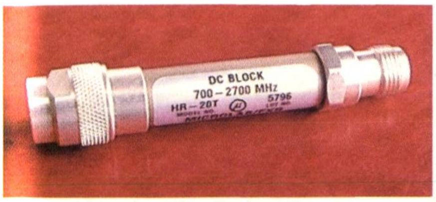
State of the Art Inc.,
State College, PA (814) 458-3401.

Circle No. 227

[Continued on page 154]

NEW PRODUCTS

DC Block



The model HR-20T DC block features a loss of less than 0.8 dB over the 700 to 2700 MHz cellular, PCS and UMTS frequency bands. It is used to prevent the flow of direct current and low frequency current surges along the inner conductor of a transmission line, while permitting the unimpeded flow of RF signals. The distributed series coupling capacitor gives the HR-20T a power capability of 250 W average, with a peak voltage rating of 2 kV. Applications include the DC isolation of wireless signal paths and to limit current surges at antenna sites that may occur during lightning storms. Special attention has been directed to maximize DC isolation and minimize passive intermodulation. Weight: 0.9 oz.

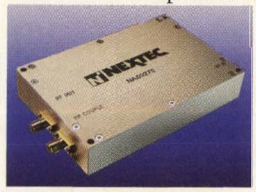
Microlab/FXR,
Livingston, NJ
(973) 992-7700.

Circle No. 222

AMPLIFIERS

Power Amplifier

The model NA00270 one-watt power amplifier for IMT-2000 operates from 2110 to 2170 GHz. It is a single bias (+12 V) medium power amplifier optimized for CDMA signals with 1 W average CDMA output power. It offers forward power detection and on/off control input signal ports, has a P1dB of 35 dBm and output third-order intercept of 51 dBm, with a gain of 30 dB.



Nextec Microwave & RF Inc.,
Santa Clara, CA (408) 727-1189.

Circle No. 234

Wideband Amplifier

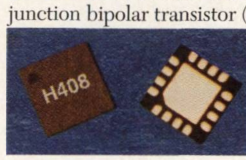
The model AML0016P2001 low noise highly linear wideband amplifier operates in the frequency range of 10 MHz to 6 GHz and delivers low orders of distortion with an IP3 of 36 dBm (typ). Gain is 21 dB (min) and the amplifier gives +23 dBm (min) output power at 1 dB gain compression and a noise figure below 3.2 dB from 200 MHz and above. Input and output SWR is 2.1. Operating at 15 V DC, this amplifier draws a nominal current of 480 mA. Internal DC regulator, reverse voltage protection and field removable SM connector shells are standard. Size: 1.54" × 0.75".

AML Communications Inc.,
Camarillo, CA (805) 388-1345.

Circle No. 231

HBT Power Amplifier

The model HMC408LP3 GaAs InGaP heterojunction bipolar transistor (HBT) MMIC power amplifier operates from 5.1 to 5.9 GHz with +30 dBm P1dB output power and +43 dBm output IP3. It is ideal for low cost UNII point-to-point/multi-point and 902.11a/HiperLAN WLAN access point applications. The power amplifier provides 20 dB of gain and +32.5 dBm of saturated power at 27 percent PAE from a single 5 V supply. A power down input provides RF output power and DC current control. The input of the amplifier is internally matched to 50 Ω and requires minimal external output matching components. Size: 3 × 3 mm.



Hittite Microwave Corp.,
Chelmsford, MA (978) 250-3343.

Circle No. 233

DEVICE

Intermediate Power Amplifiers

The model ECP052, ECP053, ECP103 and ECP203 RF MMICs are built with a highly reliable InGaP/GaAs HBT process. The 053, 103 and 203 are operable from 2100 to 2700 MHz, making them ideal for 802.11b and other applications at 2.4 GHz, while the ECP052 operates over the 800 to 1000 MHz range with a P1dB of 27 dBm, gain of 15 dB and OIP3 of 45 dBm. The ECP053 features a P1dB of 28 dBm, 17 dB of gain and the same OIP3 of 45 dBm. The ECP103 offers a P1dB of 31 dBm, gain of 12 dB and OIP3 of 47 dBm, while the ECP203 has a P1dB of 33 dBm, 11 dB of gain and an OIP3 of 49 dBm.

EiC Corp., Fremont, CA (510) 979-8938.

Circle No. 236

INTEGRATED CIRCUITS

HCI-level Bluetooth Module

This full host controller interface (HCI) level Bluetooth module, Blue Module,™ integrates

both active devices and passive components to create complete circuits in one chip, making it both a carrier of the IC and a functional device. The module, which has achieved logo certification for Bluetooth standard version 1.1, is ideal for wireless applications such as cell phones, laptop and desktop computers, headsets, telematics and personal digital assistants as well as advanced products including wireless printers, mouses and keyboards. The Blue Module enables engineers to bypass the process of designing and laying out additional RF circuits and is compatible with other Bluetooth enabled devices, allowing for greater connectivity.

Murata Electronics North America Inc., Smyrna, GA (770) 436-1300.

Circle No. 238

Single-chip Direct Conversion Transceiver

The model CX74063 single-chip direct conversion transceiver simplifies the design of multiband GSM handsets. The transceiver supports general packet radio service and enhanced data rates for global evolution standards and provides a well-defined roadmap to dual-mode, WCDMA/GSM UMTS handsets. These handsets will support advanced applications such as multimedia and high speed Web browsing. The device's advanced architecture reduces the number of external components required to build a mobile handset by more than one-third, significantly reducing the size, cost and power requirements of next-generation GSM handsets. The device integrates all of the circuitry associated with generating a 13 or 26 MHz system reference frequency, with the exception of the crystal. Size: 8 × 8 mm. Price: $5.50 (10,000).

Skyworks Solutions Inc., Woburn, MA (978) 241-7000.

Circle No. 239

MATERIAL

Conductive Silicones

The BISCO® EC-2000 series electrically conductive EMI/RFI shielding material is available in wide, continuous rolls. These products have excellent compression-set resistance, high shielding effectiveness and easy processing to increase yields. The nickel-graphite-tilled EC-2000 series silicones offer high design flexibility because they are softer and more compressible than similar products, meaning that a good mechnical seal can be obtained through compression of the gasket. In addition, these conductive elastomer sheets offer the combined advantages of high shielding performance, good flame resistance characteristics and corrosion resistance. This series of silicones can be easily die cut and delivered, offering additional design versatility and decreased time to market for OEM designers. The series also provides a range of thicknesses from 0.020" to 0.125".

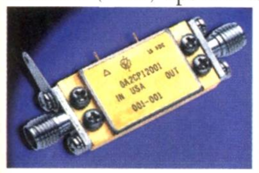
Rogers Corp., Rogers, CT (800) 237-2068.

Circle No. 240

SOURCES

High Performance VCO

The model OA2CP12001 voltage controlled oscillator (VCO) operates from 8000 to 12000



MHz and combines an oscillator circuit, voltage regulator, MMIC amplifier and bandpass filter in a two-stage CougarPak™ housing. The oscillator is designed for fast tuning speed and linear tuning. Typical performance includes post tuning drift

[Continued on page 156]

NEW PRODUCTS

of ±3.0 MHz at 1 μs relative to frequency at 1 s, 2:1 tuning linearity over frequency, and suppresses harmonics to better than –20 dBc. The output power is set to 10 dBm (min) over the –54° to +85°C temperature range, this VCO covers the frequency range in 0 to –12 V, and draws 250 mA from a 15 V DC supply. The new line of VCOs is suitable for challenging signal processing applications.

Cougar Components,
Sunnyvale, CA (408) 522-3838.

Circle No. 242

■ VCXO

The model 357 surface-mount voltage controlled crystal oscillator (VCXO) features a frequency signal output that will interface with HCMOS electronic circuitry. It is targeted to support SONET, DSL and cable modem applications where frequency servo is required. The model 357 is presently available in frequencies from 1.5 to 45.0 MHz. The tristate feature enables the output frequency to be electrically switched to enable or disable the frequency source. It also offers an average pull range of ±50 ppm or ±32 ppm to servo the frequency back to its nominal value as the product operates over its temperature range. Operating temperatures from 0° to 70°C and –40 to +85°C are offered.

CTS Corp.,
Elkhart, IN
(630) 924-3726.

Circle No. 243

■ VCO

The model MW500-1225 voltage-controlled oscillator (VCO) operates from 90 to 190

MHz and is designed for satellite and other wireless applications. It maintains a low phase noise level of –111 dBc/Hz at 10 kHz, –133 dBc/Hz at 100 kHz, and harmonic levels below –17 dBc over the 90 to 190 MHz band. The VCO is stable drifting –0.02 MHz/°C, pulling 0.2 MHz peak to peak, and pushing is 0.7 MHz/V. Tuning linearity is better than 1.6:1 across the bandwidth with power output of +3 dBm ±1 dB. Supply is +5 V at 15 mA and tuning voltage range is 0 to 17 V. Modulation bandwidth is 5.3 MHz. Size: 0.50" × 0.50" × 0.15". Price: $21.95 (1 to 10).

Micronetics VCO Products Division,
Hudson, NH (603) 579-0900.

Circle No. 246

■ L-band VCO

The model V626ME08 voltage-controlled oscillator (VCO) generates frequencies from

2200 to 2700 MHz and covers the entire frequency band within 0.5 to 18.0 VDC of control. The product will further enhance the performance of any phase-locked loop by providing a clean sprectral signal of –96 dBc/Hz (typ) at 10 kHz from the carrier and is guaranteed to operate over the extended commercial temperature range of –40° to +85°C. This VCO typically draws only 22 mA from a 5 V DC supply and offers the end user 7±3 dBm of output power into a 50 Ω load. It pushes less than 5 MHz/V within 5 percent from the nominal supply voltage and pulls less than 15 MHz with a 14 dB return loss, any phase. Size: 0.50" × 0.50" × 0.13". Price: $15.95.

Z-Communications,
San Diego, CA (858) 621-2700.

Circle No. 248

TEST EQUIPMENT

■ Test Portfolio

The model E1966A 1xEV-DO test application provides a bench-top tool for designing access terminals (AT) and helps high volume manufacturers speed production and reduce the risk of returns and recalls. The E1966A improves test times significantly by testing 1xEV-DO ATs at the packet level instead of at the frame level used in cdma2000 receiver testing. At the highest data rate, the 1xEV-DO has up to 16 packets per frame, making packet-error-rate measurements 12 times faster than typical frame-error-rate measurements. The test application also incorporates a new spectrum monitor feature with three independent markers that allow quick and easy visual inspection of relative measurements, which are especially useful for production repair benches, quality assurance testing, and general purpose research and development tasks.

Agilent Technologies,
Test and Measurement Organization,
Santa Clara, CA (800) 452-4844.

Circle No. 250

NEW LITERATURE

■ 2002 Product Catalog

This catalog includes high power passive resistive components featuring terminations, resistors and attenuators in various configurations. Packages are available in chip form, chips with leads and flanged devices. Also featured are standard precision chip packages, thick film circuit capabilities, cable load and power sensors.
Barry Industries Inc.,
Attleboro, MA (508) 226-3350.
Circle No. 200

■ SAW Products

This folder provides a variety of brochures and application notes detailing the company's SAW foundry services, SAW filters and "How to Specify a Custom SAW Filter." Photographs, detailed product descriptions, specifications, performance graphs and ordering information is provided.
COM DEV SAW Products,
Cambridge, Ontario, Canada
(647) 887-7297.
Circle No. 201

■ Product Brochure

This four-page brochure details the company's Merlin Mobile™ Bluetooth protocol analyzer. The first full-featured Bluetooth protocol analyzer in PC card format, it offers Bluetooth developers a portable, low cost solution for quickly pinpointing problems to reduce the time needed to debug device or piconet operation.
Computer Access Technology Corp.,
Santa Clara, CA (800) 909-2282.
Circle No. 202

■ Foundry Services Brochure

This six-page brochure details the company's range of foundry services. The company's My SOI Foundry is a complete foundry solution for all RFIC manufacturing needs. The company is equipped to develop customer's state-of-the-art ICs using proven SOI technology, service and expertise.
Honeywell, Solid State Electronic Center,
Plymouth, MN (800) 323-8295.
Circle No. 203

■ Catalog and Interactive CD-ROM Update

This new product catalog and interactive CD-ROM showcases custom products with actual test data as well as product software downloads. Product descriptions, performance graphs, specifications, outline drawings, applications and ordering information is all provided.
K & L Microwave Inc.,
Salisbury, MD (410) 749-2424.
Circle No. 204

■ Product Catalog

This new product catalog features the latest RF passive product releases for the company's isolators, circulators and customer waveguide assemblies. The products cover frequency ranges from 300 MHz to 40 GHz and offer best in industry performance for insertion loss, SWR, temperature performance and IMD.
M2 Global Technology Ltd.,
San Antonio, TX (210) 561-4800.
Circle No. 205

■ Product Folder

This folder provides information on the company's MagLatch™ RF MEMS switch. The units are micro magnetic latching switches designed for switching RF signals from DC to tens of GHz in 50 Ω. The device fabrication is based on a surface micromachining approach and is thus a high volume, low cost production method.
Microlab, Chandler, AZ (480) 926-9500.
Circle No. 206

■ LC and Crystal Filter Products

This brochure features an extensive line of cost-effective crystal filters in all polynomials from 100 kHz to 300 MHz with fractional bandwidths of 0.002 to 3 percent, LC RF filters in all polynomials from 1 kHz to 2.5 GHz, and crystal-based frequency discriminators from 100 kHz to 80 MHz. The company also offers a broad selection of monolithic crystal filters.
Network Sciences,
Phoenix, AZ (602) 258-8095.
Circle No. 207

■ E-mail Newsletter

This newsletter offers readers information on the company's product line, such as wireless transceiver products, wireless base station products, switches, isolators and circulators, new product announcements, company news and personnel appointments. For further information, visit rec-usa.com or e-mail: wirelessedge@rec-usa.com.
Renaissance Electronics Corp.,
Harvard, MA (978) 772-7774.
Circle No. 208

■ Product Selection Guide

This eight-page Winter 2002/2003 product guide presents key performance characteristics of the company's low noise amplifiers, power amplifiers, IF amplifiers, fiber-optic transimpedance amplifiers, high power and high linearity discrete components, gain blocks, multi-component modules and CATV amplifiers.
Sirenza Microdevices Inc.,
Sunnyvale, CA (408) 616-5400.
Circle No. 209

■ Passive Components Catalog

This 54-page catalog covers the company's range of passive component products offered to the North American market. The catalog's table of contents lists 19 product sections, including capacitors, inductors, ferrite beads, thermistors, choke coils and EMI suppression filters. Each section provides designers with valuable information on a product group's features, applications and operating temperature ranges.
Taiyo Yuden Co. Ltd.,
Schaumburg, IL (847) 925-0888.
Circle No. 210

■ Diode Mixer Components Catalog

This catalog features the company's line of diode mixer components. The components offer high IP3 performance at an affordable price. Product photographs, features, specifications, performance charts, applications, outline drawings and ordering information is provided.
WJ Communications,
San Jose, CA (800) 951-4401.
Circle No. 211

Compact and Broadband Microstrip Antennas

Kin-Lu Wong
John Wiley & Sons Inc.
327 pages; $94.95
ISBN: 0-471-41717-3

Modern portable communication equipment requires compact, broadband microstrip antennas. This book is intended to present new advanced designs of this type of antenna, offering more than 100 advanced microstrip antenna designs and their detailed experimental results.

The microstrip antenna designs covered in the book are divided into two groups: compact microstrip antennas and broadband microstrip antennas. Chapter 1 presents an introduction and overview of recent advances in the design of both compact and broadband microstrip antennas. Chapter 2 describes recent advances in compact microstrip antennas. Based on recent compact design techniques, such as using a shorted patch, a meandered patch, a meandered ground plane, an inverted U-shaped patch and an inverted-L patch, microstrip antenna designs are discussed and experimental results are presented. Chapter 3 discusses compact broadband microstrip antenna designs. Design techniques for achieving broadband operation with a reduced antenna size are described. Chapter 4 presents designs for compact dual-frequency and dual-polarized microstrip antennas. Advances in compact circularly polarized (CP) microstrip antennas are considered in Chapter 5. Designs for achieving gain-enhanced compact microstrip antennas are included in Chapter 6. Chapter 7 is devoted to recent advances in broadband microstrip antennas. Advances in broadband microstrip antennas with additional microstrip resonators, an air or foam substrate, slot loading, or integrated reactive loading, for example, are presented and discussed in detail. Chapter 8 presents broadband dual-frequency and dual-polarized microstrip antennas. Various design examples are presented and design considerations for achieving high isolation and low cross-polarization for broadband dual-polarized radiation are addressed. Finally, in Chapter 9, advances in broadband and dual-band circularly polarized microstrip antennas are discussed. Related broadband designs with single-feed, dual-feed and four-feed excitations are studied. In addition to obtaining a wide axial-ratio bandwidth, how to improve CP quality in the entire radiation pattern to achieve wide-angle coverage is shown.

"[This book presents] more than 100 advanced microstrip antenna designs and their detailed experimental results."

To order this book, contact: John Wiley & Sons Inc., One Wiley Drive, Somerset, NJ 08875 (800) 225-5945.

Principles of Radar and Sonar Signal Processing

Francois Le Chevalier
Artech House Inc.
397 pages; $109, £76
ISBN: 1-58053-338-9

The recent evolution of radar and sonar has been marked by the rapid increase in information processing capabilities. By enabling the implementation of signal processing procedures that are closer to theoretically optimum ones, this evolution has revealed the arbitrary character of some of the underlying assumptions: single-scatterer target, additive white Gaussian noise. It has also led to the development of multifunction radars and sonars and has enabled various types of processing developed in sonar to be adapted to radar.

The book is divided into three parts. Chapters 1, 2 and 3 deal with processing in the presence of white noise — detection and location, ambiguity function and tracking systems. Chapters 4 and 5 discuss reception in colored noise — various structures of the optimum receiver, adaptive processing, application to jammer suppression and tracking, and location in passive listening mode. Chapters 6 and 7 deal with the physical characteristics of targets and backgrounds, their modeling, and the signal processing taking them into account. The various types of processing are illustrated by examples applied to real or simulated signals, thus giving an immediate feeling of the exact nature of the limitations encountered.

Focused on principles, the book is aimed at students attending specialist courses; it is derived from such a course, given at the Ecole Nationale Superieure des Telecommunications and the Ecole Nationale de l' Aeronautique et de l' Espace in France. However, the sections on adaptive processing, and target and background signatures should also be of interest to engineers working on the design and evaluation of radar and sonar systems. A detailed, equation-free summary is given at the beginning of each chapter, which describes in the simplest terms the key points of the approach and the main results described in the chapter. This format should help readers understand the general approach of each chapter, allowing them to turn directly to the next, without getting lost in the detailed calculations of the previous ones.

"The recent evolution of radar and sonar has been marked by the rapid increase in information processing capabilities."

To order this book, contact: Artech House Inc., 685 Canton St., Norwood, MA 02062 (781) 769-9750 ext. 4002; or 46 Gilligham St., London SW1V 1HH, UK +44 (0) 207 596-8750.

Dan Massé

Dan Massé is a member of the Microwave Journal *staff.*

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Charles A. Ayotte (signature) Charles A. Ayotte, Vice President
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Tim O'Callaghan
Eastern Reg. Sales Mgr.
685 Canton Street
Norwood, MA 02062
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tocallaghan@mwjournal.com

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Australia
James Tonkin
Tonkin Media Pty. Ltd.
300 Scenic Highway
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Juergen Wissling
Wissling Marketing Services
Riedstrasse 5
72813 St. Johann-Wuertingen
Germany
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FAX: +49 7122 828145
j.wissling@wms-marketing.de

Israel
Oreet Ben Yaacov
Oreet International Media
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51201 Bene-Berak, Israel
Tel: +972 3 570 6527
FAX: +972 3 570 6526
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Japan
Kenichi Takefusa
Appeal Inc.
Arpus Hiroo, 2F, 5-17-11
Hiroo, Shibuya-ku,
Tokyo 150, Japan
Tel: +81 3 3443 0895
FAX: +81 3 3443 0871
takefusa@mxb.mesh.ne.jp

Korea
Young-Seoh Chinn
JES Media International
2nd Floor, ANA Bldg.
257-1, Myungil-Dong
Kangdong-Gu
Seoul, 134-070 Korea
Tel: +82 2 481-3411
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jesmedia@unitel.co.kr

Singapore
Hoong-Mun Koo
Publicitas
02-01 Wellington Bldg.
20 Bideford Road
Singapore 229921
Tel: +65 6836 2272
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Taiwan
Hiram Lai
Arco Infocomm
4 F, No. 5, Sec. 1,
Pa-Te Road
Taipei, Taiwan ROC
Tel: +886-2 396 5128
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