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# Microwaves & RF

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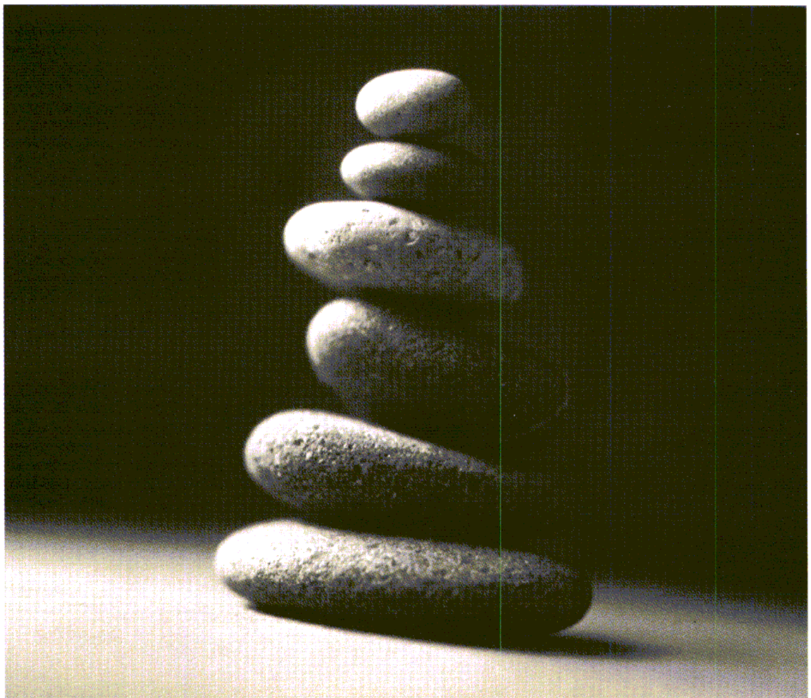
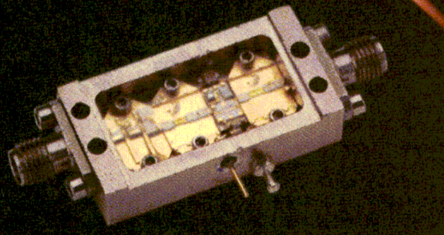
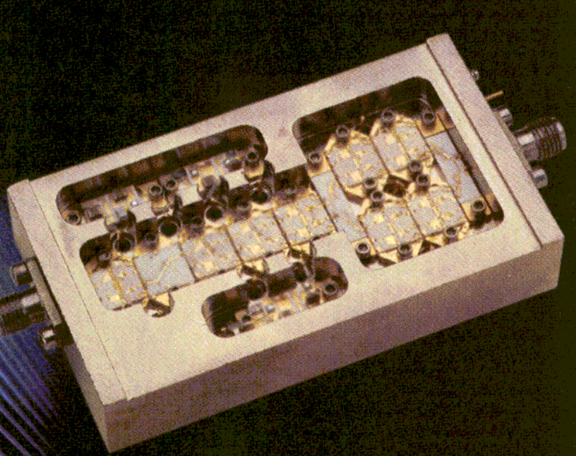
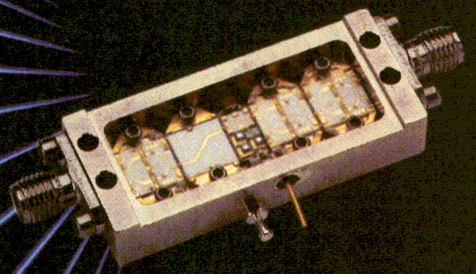


# AMPLIFIERS

## for every application







### ULTRA BROAD BAND

| Model | Freq. Range GHz | Gain dB min | N/F dB max | Gain Flat +/-dB | 1 dB Comp. pt. dBm min | 3rd Order ICP typ | VSWR In/Out max | DC Current mA |
|---|---|---|---|---|---|---|---|---|
| JCA018-203 | 0.5-18.0 | 20 | 5.0 | 2.5 | 7 | 17 | 2.0:1 | 250 |
| JCA018-204 | 0.5-18.0 | 25 | 4.0 | 2.5 | 10 | 20 | 2.0:1 | 300 |
| JCA218-506 | 2.0-18.0 | 35 | 5.0 | 2.5 | 15 | 25 | 2.0:1 | 400 |
| JCA218-507 | 2.0-18.0 | 35 | 5.0 | 2.5 | 18 | 28 | 2.0:1 | 450 |
| JCA218-407 | 2.0-18.0 | 30 | 5.0 | 2.5 | 21 | 31 | 2.0:1 | 500 |

### MULTI OCTAVE AMPLIFIERS

| Model | Freq. Range GHz | Gain dB min | N/F dB max | Gain Flat +/-dB | 1 dB Comp. pt. dBm min | 3rd Order ICP typ | VSWR In/Out max | DC Current mA |
|---|---|---|---|---|---|---|---|---|
| JCA04-403 | 0.5-4.0 | 27 | 5.0 | 1.5 | 17 | 27 | 2.0:1 | 550 |
| JCA08-417 | 0.5-8.0 | 32 | 4.5 | 1.5 | 17 | 27 | 2.0:1 | 550 |
| JCA28-305 | 2.0-8.0 | 22 | 5.0 | 1.0 | 20 | 30 | 2.0:1 | 550 |
| JCA212-603 | 2.0-12.0 | 32 | 5.0 | 3.0 | 14 | 24 | 2.0:1 | 550 |
| JCA618-406 | 6.0-18.0 | 20 | 6.0 | 2.0 | 25 | 35 | 2.0:1 | 600 |
| JCA618-507 | 6.0-18.0 | 25 | 6.0 | 2.0 | 27 | 37 | 2.0:1 | 800 |

### MEDIUM POWER AMPLIFIERS

| Model | Freq. Range GHz | Gain dB min | N/F dB max | Gain Flat +/-dB | 1 dB Comp. pt. dBm min | 3rd Order ICP typ | VSWR In/Out max | DC Current mA |
|---|---|---|---|---|---|---|---|---|
| JCA12-P01 | 1.35-1.85 | 35 | 4.0 | 1.0 | 33 | 41 | 2.0:1 | 1000 |
| JCA34-P02 | 3.1-3.5 | 40 | 4.5 | 1.0 | 37 | 45 | 2.0:1 | 2200 |
| JCA56-P01 | 5.9-6.4 | 30 | 5.0 | 1.0 | 34 | 42 | 2.0:1 | 1200 |
| JCA812-P03 | 8.0-12.0 | 40 | 5.0 | 1.5 | 33 | 40 | 2.0:1 | 1700 |
| JCA1218-P02 | 12.0-18.0 | 22 | 4.0 | 2.0 | 25 | 35 | 2.0:1 | 700 |

### LOW NOISE OCTAVE BAND LNA'S

| Model | Freq. Range GHz | Gain dB min | N/F dB max | Gain Flat +/-dB | 1 dB Comp. pt. dBm min | 3rd Order ICP typ | VSWR In/Out max | DC Current mA |
|---|---|---|---|---|---|---|---|---|
| JCA12-3001 | 1.0-2.0 | 40 | 0.8 | 1.0 | 10 | 20 | 2.0:1 | 200 |
| JCA24-3001 | 2.0-4.0 | 32 | 1.2 | 1.0 | 10 | 20 | 2.0:1 | 200 |
| JCA48-3001 | 4.0-8.0 | 40 | 1.3 | 1.0 | 10 | 20 | 2.0:1 | 200 |
| JCA812-3001 | 8.0-12.0 | 32 | 1.8 | 1.0 | 10 | 20 | 2.0:1 | 200 |
| JCA1218-800 | 12.0-18.0 | 45 | 2.0 | 1.0 | 10 | 20 | 2.0:1 | 250 |

### NARROW BAND LNA'S

| Model | Freq. Range GHz | Gain dB min | N/F dB max | Gain Flat +/-dB | 1 dB Comp. pt. dBm min | 3rd Order ICP typ | VSWR In/Out max | DC Current mA |
|---|---|---|---|---|---|---|---|---|
| JCA12-1000 | 1.2-1.6 | 25 | 0.75 | 0.5 | 10 | 20 | 2.0:1 | 80 |
| JCA23-302 | 2.2-2.3 | 30 | 0.8 | 0.5 | 10 | 20 | 2.0:1 | 80 |
| JCA34-301 | 3.7-4.2 | 30 | 1.0 | 0.5 | 10 | 20 | 2.0:1 | 90 |
| JCA56-401 | 5.4-5.9 | 40 | 1.0 | 0.5 | 10 | 20 | 2.0:1 | 120 |
| JCA78-300 | 7.25-7.75 | 27 | 1.2 | 0.5 | 13 | 23 | 2.0:1 | 120 |
| JCA910-3000 | 9.0-9.5 | 25 | 1.2 | 0.5 | 13 | 23 | 1.5:1 | 150 |
| JCA910-3001 | 9.5-10.0 | 25 | 1.2 | 0.5 | 13 | 23 | 1.5:1 | 150 |
| JCA1112-3000 | 11.7-12.2 | 27 | 1.1 | 0.5 | 13 | 23 | 1.5:1 | 150 |
| JCA1213-3001 | 12.2-12.7 | 25 | 1.1 | 0.5 | 10 | 20 | 2.0:1 | 200 |
| JCA1415-3001 | 14.4-15.4 | 35 | 1.4 | 1.0 | 14 | 24 | 2.0:1 | 200 |
| JCA1819-3001 | 18.1-18.6 | 25 | 1.8 | 0.5 | 10 | 20 | 2.0:1 | 200 |
| JCA2021-3001 | 20.2-21.2 | 25 | 2.0 | 0.5 | 10 | 20 | 2.0:1 | 200 |

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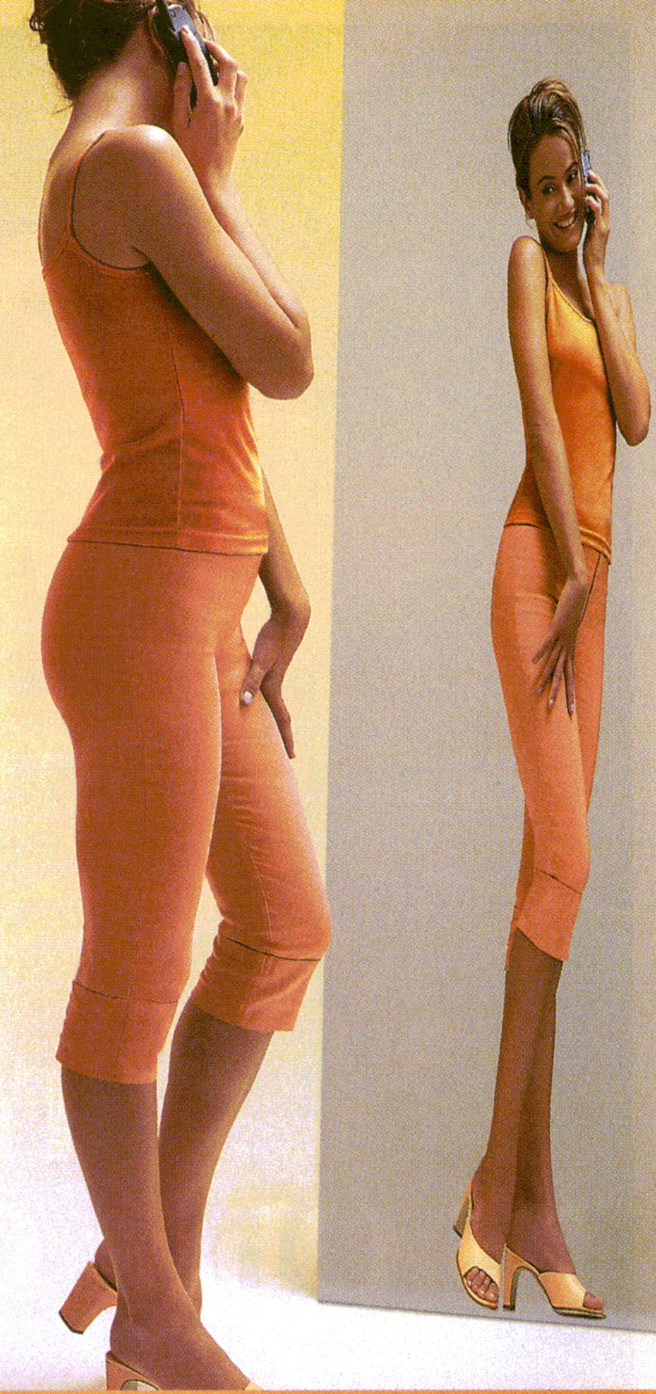


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Application Profile

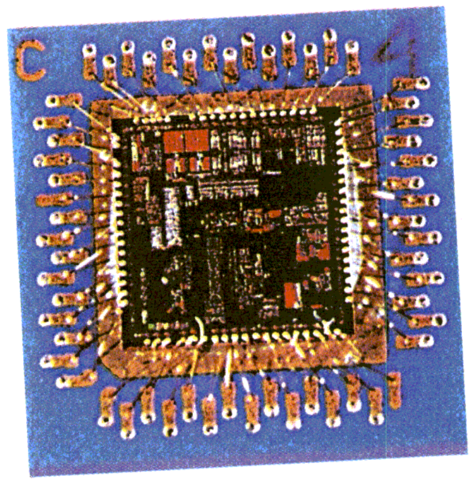
# How to make Cell Phones Smaller and Lighter? BGA with Integrated Components using DuPont Green Tape™.

National Semiconductor is a leader in applying the LTCC advantages of high-density interconnect capability, ability to integrate passive components and functions, and low-loss performance. In a recent design, National chose to combine its advanced ICs for wireless communications with Green Tape™, DuPont's brand of LTCC tape dielectric material, to provide optimum performance in the smallest possible package.

## Challenge: Decreased Size and Cost, Improved Performance for Wireless Devices

Portable wireless applications have quickly become the main driver for smaller, more cost-effective packaging and interconnects. For example, in the last few years, cell phones have evolved into lightweight, palm-size devices with a host of new functions. Their weight has decreased by a factor of 10, and the wholesale selling price by 75 percent.

OEM designers are now learning that integrating IC and package design to take advantage of the



unique properties of Low Temperature Co-fired Ceramic (LTCC) technology can yield decreased size and improved performance in wireless devices.

## Solution: Green Tape™ LTCC Allows for High I/O Counts in Chip Scale Package

National's newest chipsets use Green Tape™ packaging capabilities to provide a chip scale package that can accommodate the high I/O counts of a highly integrated RF analog front end using micro BGA (ball grid array) technology. The current package, only 9 x 9 mm, can provide 81 I/Os in a micro BGA array, plus topside pads for wirebonding that interconnects to the BGA pads on the backside. The high number of I/Os allows for multiple grounds to improve RF performance, while the embedded multilayer structure contains 14 RF bypass capacitors constructed using a combination of high-K and low-K dielectrics.

The performance of the frequency synthesizer function can be enhanced through the use of an embedded VCO resonator that provides a high Q, and therefore lower phase noise, than that available using a VCO resonator located on the silicon.

This approach, co-designing the silicon and LTCC elements to achieve optimized size and performance, demonstrates the use of co-integration for wireless applications requiring smaller package size and higher performance at the lowest possible cost.

**For more information, call DuPont at 1-800-284-3382, press 3, or visit the DuPont Microcircuit Materials website at http://www.dupont.com/mcm.**

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CIRCLE NO. 234

## CORRECT EQUATIONS

**To the editor:**

In my article "Calculate Intercept And Compression Points" that appeared in the May 2000 issue starting on page 141, I discovered some equation errors. The corrected equations should read:

$$P_{2IP}\,(out) = V^2_{2IP}\,(out)/2/R_0 \quad (3b)$$

$$P_{2IP}\,(out) = 2\,I_{DC}^{\,2}\,R_0 \quad (3c)$$

$$P_{3IP}\,(out) = V^2_{31IP}(out)/2/R_0 \quad (4b)$$

$$P_{3IP}\,(out) = 4\,I_{DC}^{\,2}\,R_0 \quad (4c)$$

$$P_{3IP}\,(out)/P_1\ dB = 4/0.4 = 10 = 10\ dB \quad (13)$$

I apologize for the errors.

**David Rosemarin**
*Consultant*
*Israel*

## INCORRECT PROFILE

**To the editor:**

It was recently noticed in July's Crosstalk section (p. 37) that Eugene R. Brannock's photo caption was incorrect. The incorrect profile read, "As a member of Fujitsu's Board of Directors, Brannock leads FCSI's Marketing Team and Engineering Design Centers." The caption should have read, "As a member of the FCSI Board of Directors, Brannock leads FCSI's Marketing Team and Engineering Design Centers."

Thank you for this clarification.

**Agnes Toan**
*Fujitsu Compound Semiconductor*

## CONTACT INFO

**To the editor:**

I think that a big improvement to this magazine would be to list contact names with their respective phone numbers. This may mean a decrease in print, but companies that do not want to list this information, I feel, do not need the news print.

**Harvey Pollack**
*Forset Hills, NY*

## ERRATUM

*In last month's issue, several figures were left out of a Design Feature. They were not purposely left out nor were they lost.*

*The article "Create Transmission-Line Matching Circuits For Power Amplifiers (p. 109)" omitted Figures 7 through 12 even though they were cited in the text. One of the figures that was included, Figure 3, was not shown in its entirety (the bottom part of the drawing was cut off).*

*In our haste to meet September press deadlines, the editors overlooked these major mistakes that usually would have been caught. We have high editorial standards here at* Microwaves & RF *and we take mistakes such as this one very seriously.*

*We apologize not only to our readers and advertisers but also to Andrei Grebennikov, who is the author of the story.*

*As one of the biggest omissions that this magazine has ever made, we have decided to re-run the article in this issue starting on p. 113. This time, however, all of the figures will appear in the article. This is the least that we can do as a responsible magazine.*

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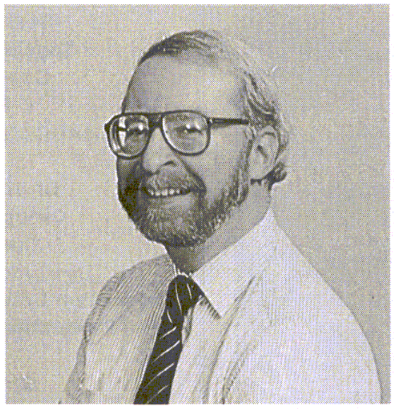
**CIRCLE NO. 358**

# THE INEVITABILITY OF A WIRELESS INTERNET



Internet-related messages are everywhere. Turn on a radio, and the majority of the sponsors are touting their electronic (E) business capabilities or the contents of their websites. Turn on the television, and sponsors urge us to visit their websites. Flip through a magazine, and we are directed to the advertiser's website. The power of the Internet was felt quite strongly at the latest trade show sponsored by the Personal Communications Industry Association (PCIA), the PCIA Global Xchange, where the concept of the wireless Internet was everywhere.

Held recently (September 26-29, 2000) at McCormick Place South (Chicago, IL), the PCIA's giant annual event has gone through some drastic transformations in recent years as it struggles for differentiation with the large annual event sponsored by the Cellular Telephone Industries Association (CTIA). At one time, the two events seemingly battled toe to toe for the attention of cellular equipment and services suppliers. And during the rapid growth in the cellular and personal-communications-services (PCS) markets during the mid-1990s, there was enough business for both shows. But with the maturation of cellular/PCS networks, many of the larger suppliers have chosen to focus on only one of the shows—the CTIA's event—leaving the PCIA to attempt to "reformat" itself.

The latest "look" of the PCIA Global Xchange was noticeably leaner than events of years past. And the focus on the show floor was different than in the past, having moved away from being a purely cellular/PCS event and getting closer to becoming a wireless Internet show.

Some design content was still apparent at the PCIA Global Xchange, given that the Fall version of the Wireless Symposium/Portable By Design Conference & Exhibition is co-located with the larger event. The exhibitors for Wireless/Portable promoted traditional products, including computer-aided-engineering (CAE) software, power amplifiers (PAs), and circuit boards. But many of the exhibitors on the main show floor were hawking Internet content and services aimed at transforming existing websites into wireless websites.

Even in the PCIA Global Xchange's general sessions, the "wireless web" was ubiquitous. The lead session, called the "Globalization of the Wireless Web," which was hosted by Jay Kitchen, President and Chief Executive Officer (CEO) of the PCIA and featured speakers from Sun Microsystems and Intel (Santa Clara, CA), was a sharing of visions and plans for the wireless Internet.

In fact, an entire three-day track of technical papers (Track 1) was devoted to the wireless web, and included speakers from Phone.com, AT&T Wireless Services, America Online, Yahoo!, and Texas Instruments. Another technical-paper track (Track 2) was devoted to content and how to make websites designed for desktop computers usable over wireless devices.

The Internet is a powerful tool, and already firmly embedded in our culture. (Both Vice-President Gore and Governor Bush made key references to the Internet in their first presidential debate.) Coupled with a growing wireless infrastructure, is there any doubt that the wireless web will one day be everywhere?

Jack Browne

*Publisher/Editor*

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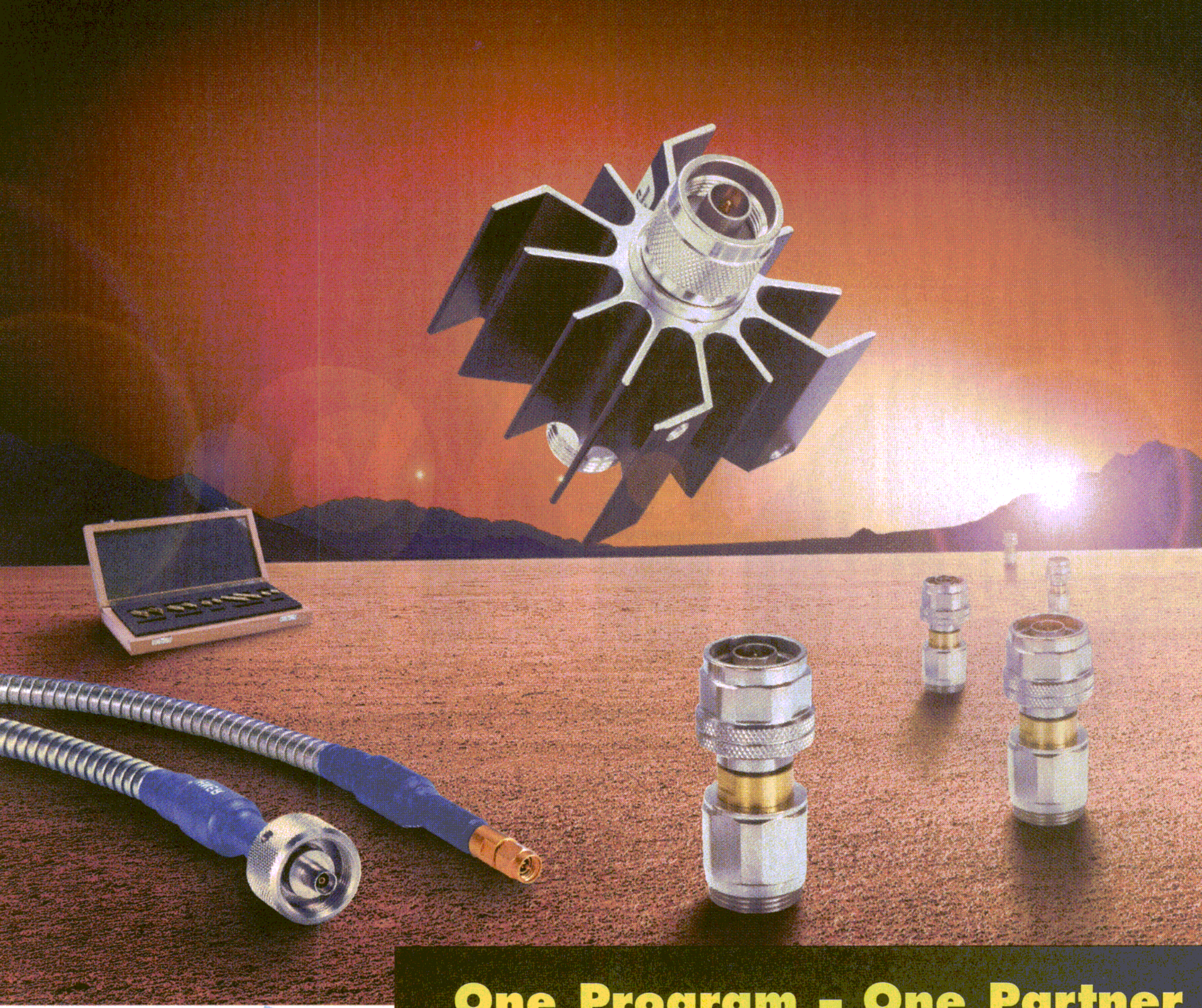


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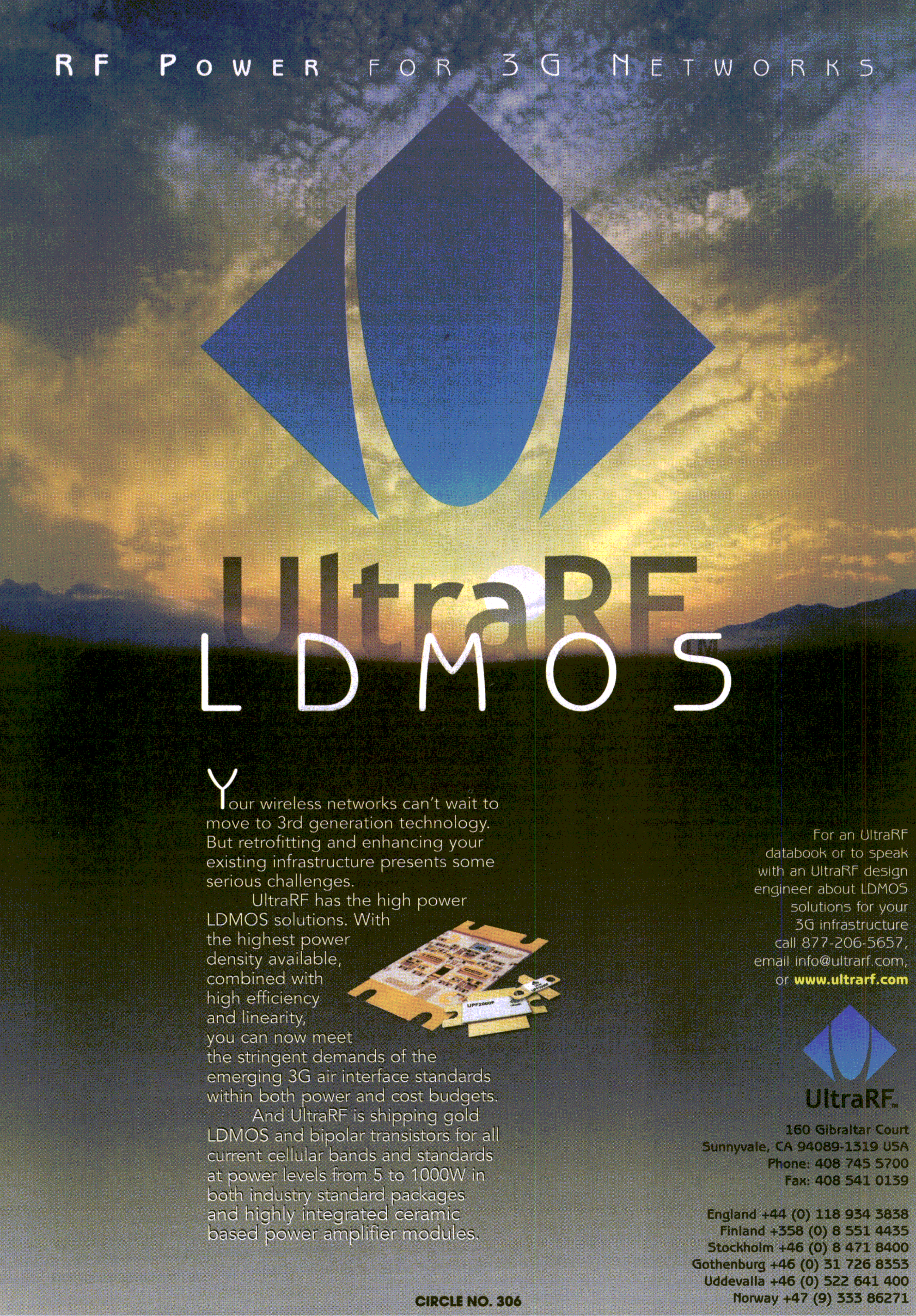
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CIRCLE NO. 306

## Gap Closes On Japan In Car-Navigation Services Market

*AUSTIN, TX*—In a report entitled "The Worldwide Market for Car Navigation, Multimedia, & Emergency-Only Systems," Intex Management Services (IMS), an electronics research firm, projects that the worldwide market for car-navigation, multimedia, and emergency systems will grow from 2.13 million units in 1999 to 12.18 million units in 2004 at a compound annual-growth rate (CAGR) of 41.7 percent. Strong growth is particularly forecast for the European and North American markets over the period to 2004.

In 1999, Japan was the largest market with 61.6 percent of worldwide shipments occurring in this region. Autonomous navigation systems represented the largest product category by far in Japan, although multimedia systems (which incorporate the navigation function) have made significant strides in the last year.

By 2004, Europe is forecast to have grown its share of the market considerably to 37.7 percent of shipments. Growth factors are projected to include the acceptance of advisory systems and emergency-only systems and the widespread penetration of multimedia systems (which include navigation) into mid-sized cars. Market growth within the latter category will be assisted by the advent of general-packet radio services (GPRS) and other cellular networks which permit higher data-transmission rates.

Growth in the North American market is also forecast to be driven by advisory and emergency-only situations (e.g., from OnStar) which are forecast to penetrate luxury to mid-sized cars. Autonomous car-navigation systems are forecast to continue to struggle in gaining significant penetration in the American market.

## RFID Technology International Standard Reaches Milestone

*DALLAS, TX and EINDHOVEN, THE NETHERLANDS*—13.56-MHz RF-identification (RFID) technology used in contactless cards and RFID smart labels reached a milestone on the way toward full international standardization with the recent unanimous vote by the International Standards Organization (ISO) to make ISO/IEC 15693-2 an international standard (IS).

ISO 15693-2, a communications protocol proposed by Texas Instruments and Philips Semiconductors in November 1998 which defines the way that data are exchanged between the RF tag and the reader, officially became an IS when it was published by the *ISO Secretariat.*

For end users implementing 13.56-MHz vicinity cards in different applications, and smart-label technology in parcel shipping, airline-baggage tagging, supply-chain management, and other item-management applications, the ISO vote means that RF tag and reader integrated circuits (ICs) using the ISO 15693-2 protocol will be compatible. ISO 15693-compatible smart labels from any manufacturer can be identified by any ISO-15693 reader.

"This is an important step in our goal of advancing standards and open systems for radio-frequency identification technology and is the catalyst that the industry has been waiting for," says Dave Slinger, vice president of Texas Instruments, Inc. "It's also good news for end users who want an even wider range of choices when implementing 13.56-MHz smart-label solutions.

## Stock-For-Stock Transaction Is Valued At $7.6 Billion

*DALLAS, TX*—Texas Instruments (TI), Inc. announced that it will acquire Burr-Brown Corp. in a stock-for-stock transaction valued at approximately $7.6 billion. The acquisition strengthens TI's position in the data-converter and amplifier segments of the analog semiconductor market.

"We are as serious about analog as we are about DSP. The people of Burr-Brown are elite developers of high-performance analog products. This combination means that TI will have a leading position in essentially every high-performance analog category, and the ability to offer almost any analog component that touches a DSP," says Tom Engibous, TI's chairman, president, and chief executive officer (CEO).

Burr-Brown designs data converters at the highest end of the precision range, including 24-b products. "Burr-Brown's product position accelerates our data-converter product roadmap by several years. Together we will extend and expand the data-converter portfolio much faster than either company could alone," Engibous says.

## RadHaz Safety Services Program Is Announced

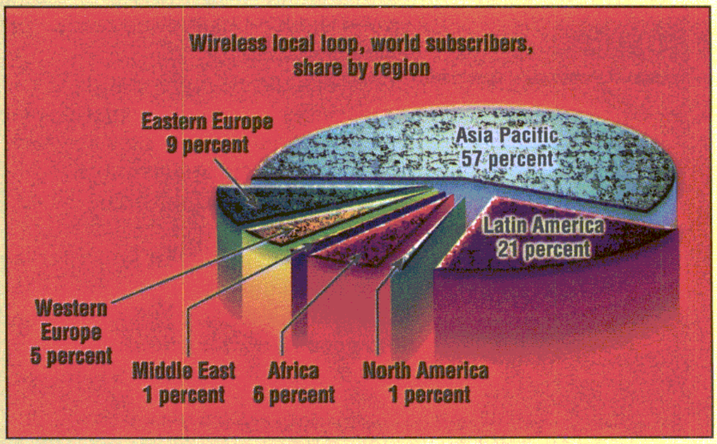
*RESTON, VA*—Comserch, an RF engineering company, announced that it has launched a comprehensive eight-step Radiation Safety Services Program designed to help wireless telecommunications operators and site owners to fully comply with the Federal Communications Commission's (FCC) and Occupational Safety and Health Administration's (OSHA) stringent guidelines and standards for radiation-hazard (RadHaz) safety which were mandated for compliance by September 1.

According to the FCC, operators and owners must implement and document site-wide compliance efforts to prove that they have a validated radiation-safety program in place. Comsearch has assisted facility managers in determining the necessary steps and services for this site evaluation: in-house computer modeling of the RF environments, on-site RF measurements, staff training, development of safety plans, regulatory support, as well as licensing and lease-agreement support.

"The proliferation of RF-transmitting facilities worldwide is staggering," states Lester Polisky, director of Comsearch's Radiation Safety Services Program. "When Comsearch entered the wireless industry more than two decades ago, there were only a few satellite and point-to-point microwave facilities worldwide, serving primarily government and corporate-communications needs. Today, virtually everyone who owns a pager, cell phone, or Palm Pilot(TM) is transmitting radio frequencies at some level. Antenna facilities can be found on hillsides, on rooftops, and even hidden in the steeples of churches—all emitting ever-increasing levels of RF electromagnetic (EM) energy. With every exciting new technology comes safety concerns and proposed regulations. The FCC felt it was time to impose and strictly enforce compliance with national guidelines for RF radiation safety."

## WLL Technology Will Fulfill Promise Despite Limited Growth In 1999

*OYSTER BAY, NY*—The wireless-local-loop (WLL) market has taken a back seat to the mobile wireless market for the time being, but WLL technologies will still provide for a large marketplace by 2006, according to a study from Allied Business Intelligence (ABI). During this decade, many of the improvements made for third-generation (3G) cellular service will be ported over to fixed systems, making fixed systems more robust and more affordable. In addition to cellular and personal-communications-services (PCS) technologies, local multipoint distribution services (LMDS) and multichannel multipoint distribution services (MMDS) will also play an increasingly important role in the deployment of robust voice and data services in underserved areas.




By year-end 2006, there will be more than 100 million narrowband WLL subscribers worldwide using close to 500,000 base stations. Eastern Europe is one of the leading regions of use for WLL narrowband technologies in 2000, but this area will only account for 10 percent of all WLL systems by 2006, according to ABI findings. Asia-Pacific will continue to be a leading area for WLL infrastructure deployment, with 38 percent of the world market in 1999, rising to 57 percent in 2006. Industrialized nations such as the US will realize little growth comparatively (see figure). Wideband technologies, such as LMDS, MMDS, and air-interfaces using fixed bands, will account for more than 10 million subs during the study period as businesses turn to the emerging technologies to build robust voice and data systems needed for establishing growth.

"Wideband systems are doing well now comparatively as many are awaiting next-generation narrowband technology to be produced in commercial quantities," says Larry Swasey, ABI's vice president of communications research and the report's author.



## New Standard Is Created For 3G Wireless Development Platform

*SANTA CLARA, CA*—Xpedion Design Systems, Inc. and Cadence Design Systems, Inc. recently announced the integration of the Xpedion GoldenGate family of RF simulation and modeling products with the Cadence® signal-processing worksystem (SPW) and Cadence analog design environment. The new product integration provides the designers of third generation (3G), Bluetooth, and RF integrated circuits (RF ICs) with a unified bottom-up and top-down design methodology.

"This integration provides an automatic way to bring very accurate, high-performance, parametric models into the SPW and analog design environment," says Les Wilson, director of marketing at Cadence. Xpedion's GoldenGate/Neural Network Model Compiler (NN-Model Compiler) accepts input data from the Analog Design Environment or from lab-measured data, and generates fully parametric SPW simulation models automatically. This provides users with a rapid way of developing complex models in a bottom-up design flow. The integration of GoldenGate/Sim, within the analog design environment, provides users with multi-tone and complex 3G modulation-analysis capability to rapidly check those models in a top-down flow. Together, the products augment analog design environment's existing RF and wireless capabilities.

Combining the power of GoldenGate with SPW empowers designers to make critical design trade-offs at the circuit and system level, throughout the wireless communication system-development cycle. From the bottom-up flow, the RF design team can simulate at the circuit level with GoldenGate/Sim.

"This integration is another powerful tool for communication-system architects and RF circuit designers," says Richard Curtin, senior vice president of sales and marketing at Xpedion. "The interface gives users of SPW and Analog Design Environment access to flexible top-down and bottom-up development flows in order to keep pace with the time to market and complexity of today's evolving wireless standards."

## Kudos

Ensure Technologies recently announced that the US Patent and Trademark Office has issued patent number 6,070,240 for its wireless security technology used in its XyLoc wireless personal-computer (PC) security system...Excellon Automation Co. announced the first shipment of its new-generation high-velocity-precision (HVP) drilling machines to Winonics Corp. Winonics received the multiple machines as part of their Southern California capacity expansion...Analog Devices announced that it is the fastest-growing supplier of V.90 modem semiconductors, according to a recently published industry report by International Data Corp. (IDC)...Discovery Semiconductors, Inc. has been named to Deloitte & Touche's "Fast 50" Program for New Jersey, a ranking of the 50 fastest-growing technology companies in the area. Rankings are based on the percentage of growth in revenues from 1995 to 1999 (five-year period)...Cree, Inc. has been ranked number 11 in *Fortune* magazine's list of the "Top 100 Fastest Growing Companies in America"...In its seventh ranking, *Individual Investor* magazine announced that ANADIGICS, a supplier of wireless- and broadband-communications solutions, is among this year's list of 100 fastest-growing companies in the US. In addition, *Individual Investor* gave ANADIGICS a "thumbs up" vote, meaning that its stock is expected to continue to grow...RF Micro Devices has again been recognized as one of the fastest-growing technology companies in North Carolina. During the past fiscal year, RFMD posted the second-highest five-year growth rate in the state. In recognition of that growth, the company has received a "North Carolina Technology Fast 50" award as part of a national program sponsored by Deloitte & Touch, LLP...Motorola outranked all other mobile-telephone manufacturers by a significant margin to lead the fifth customer-loyalty survey conducted by Brand Keys®, the New York-based strategic-brand-planning and customer-loyalty research firm. Motorola also received high marks for its leading-edge technologies such as Internet access, e-mail, and messaging. "The first thing people look for in a mobile phone is good value for the money, and Motorola does an excellent job meeting or exceeding their expectations," says Dr. Robert Passikoff, president of Brand Keys.

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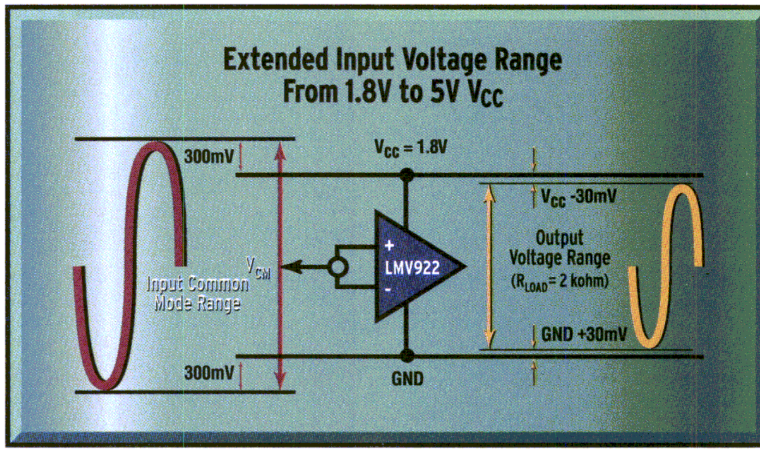


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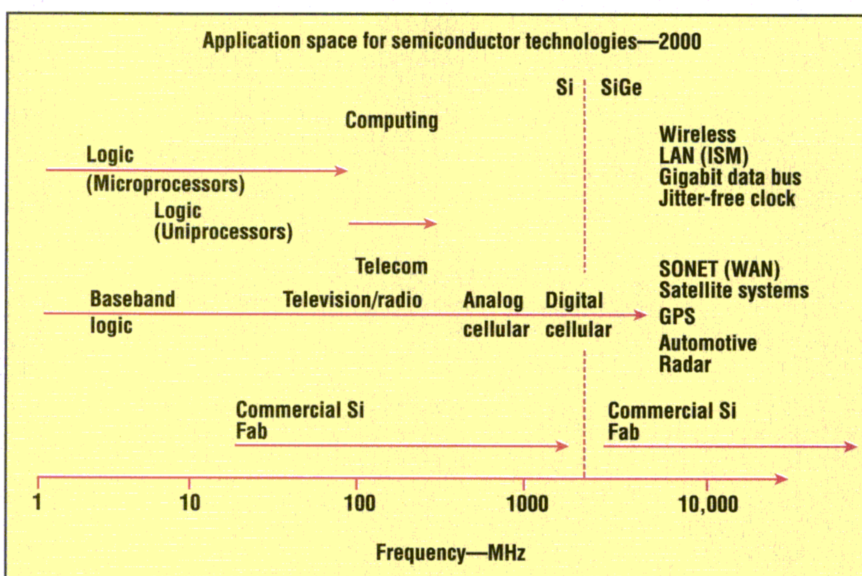
# SiGe's Influence Spreads To Many Applications

**GENE HEFTMAN**
*Senior Editor*

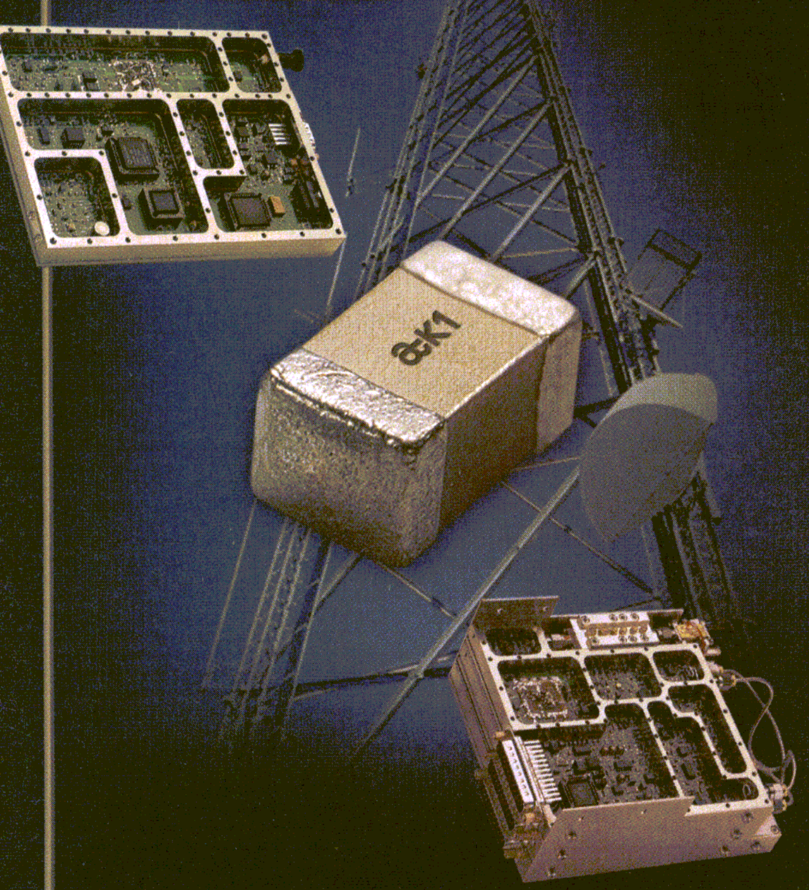
**APPROXIMATELY a decade ago, semiconductor experts were writing the obituary for the silicon (Si)-bipolar technology that had been the bedrock of the transistor-based electronics industry from its inception. High-frequency communications applications in the microwave range—above 1 GHz—they thought, would bury the Si dynasty which could not match the raw speed of the up-and-coming gallium-arsenide (GaAs) processes. But Si was not ready to roll over and die because researchers at IBM Microelectronics (Yorktown Heights, NY) had been working for years on adding germanium (Ge) to the base of an Si transistor to make it operate faster. That effort was the foundation of the heterojunction bipolar transistor (HBT), and more specifically the SiGe type. When engineers examined the results, Si had a new life. SiGe HBTs operate faster than their Si fathers at the same power level.**

The current crop of SiGe devices can operate at frequencies well into the tens of gigahertz, making them candidates for virtually every important high-frequency application. Indeed, the technology is making its mark in wireless communications, fiber optics [Synchronous Optical Network (SONET)], the Global Positioning System (GPS), test instrumentation, computer-disk drives, and wireless local-area networks (WLANs) to name some of the major applications (Fig. 1). All of the space above 1000 MHz in the figure was formerly the exclusive domain of GaAs.

Most of today's SiGe technology got started in the early 1980s at IBM's Communications Research and Development Center (CRDC) under the leadership of Dr. Bernard Meyerson, an IBM fellow and vice president of the CRDC. Meyerson says, "The common thread within CRDC is that all of our people work on parts that have to do with very high-speed communications applications. The result is that we have dramatically extended the frequency space in which silicon can play." Another common theme espoused by Meyerson is that all SiGe applications, whether in transceivers, disk drives, or test instruments are communications because the whole idea is to move information from one place to another at high speeds. The types of transistors needed, however, are not the same because a 40-GHz clock-rate for fiber optics (SONET) is a dif-




1. SiGe technology is moving into the application space—above 1000 MHz—formerly dominated by GaAs. Since it is Si based, existing fabs can be used to manufacture devices.



# CIRCUIT DESIGNER'S NOTEBOOK

## *Effective Capacitance vs Frequency*

It is generally assumed that the capacitance value selected from a vendor's catalog is constant over frequency. This is essentially true for applications with applied frequencies that are well below the capacitors self-resonant frequency. However as the operating frequency approaches the capacitors self-resonant frequency, the capacitance value will appear to increase resulting in an effective capacitance ($C_E$) that is larger than the nominal capacitance. This article will address the details of effective capacitance as a function of the application operating frequency. In order to illustrate this phenomenon, a simplified lumped element model of a capacitor connected to a frequency source operating in a network will be considered, as depicted in Figure 1.

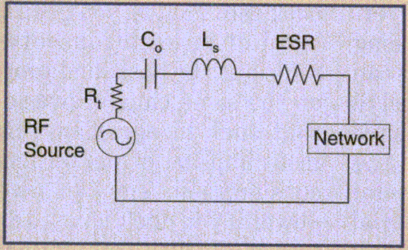



***Figure 1***
***Lumped Element Equivalent Model***

This model has been selected because the effective capacitance is largely a function of the net reactance developed between the capacitor and its parasitic series inductance ($L_S$). The equivalent series resistance 'ESR' shown in this illustration does not have a significant effect on the effective capacitance.

### Effective Capacitance:

The nominal capacitance value ($C_O$) is established by a measurement performed at 1MHz. In typical RF applications the applied frequency is generally much higher than the 1MHz measurement frequency, hence at these frequencies the inductive reactance ($X_L$) associated with the parasitic series inductance ($L_S$) becomes significantly large as compared to the capacitive reactance ($X_C$). Figure 2 illustrates that there is a disproportionate increase in $X_L$ as compared to $X_C$ with increasing frequencies. This results in an effective capacitance that is greater than the nominal capacitance. Finally at the capacitors series resonant frequency the two reactance's are equal and opposite yielding a net reactance of zero. The expression for $C_E$ becomes undefined at this frequency.

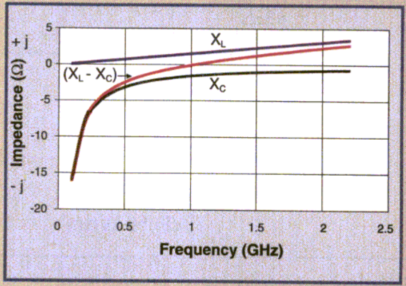



***Figure 2***
***Net Impedance vs. Frequency***

As illustrated in Figure1, the physical capacitor can be represented as $C_O$ in series with $L_S$. The impedance of the series combination of $C_O$ and $L_S$ can then be set equal to $C_E$, which may be referred to as an "ideal equivalent" capacitor. This will yield the following equation:

This will yield the following equation:

$$j(\omega L_s - 1/\omega C_o) = -j\, 1/\omega C_E$$

$$\omega^2 L_s - 1/C_o = -1/C_E$$

The relationship between the operating frequency $F_O$ and the effective capacitance $C_E$ can then be stated as:

$$C_E = C_o/(1 - \omega^2 L_s C_o)$$

$$C_E = C_o/1 - (2\pi F_o)^2 L_s C_o$$

Where:

$C_E$ = Effective Capacitance at the application frequency, ($F_o$)
$C_o$ = Nominal Capacitance at 1 MHz
$L_s$ = Parasitic Inductance, (Ω)
$F_o$ = Operating Frequency, (Hz)

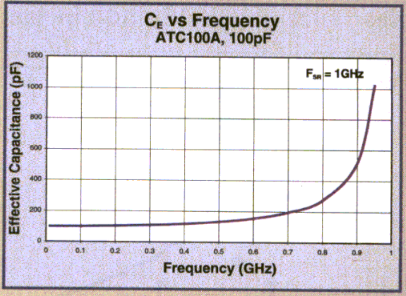



***Figure 3***
***Effective capacitance ($C_E$) vs. frequency vs $F_O$***

From this relationship it can be seen that as th applied frequency increases the denominat becomes smaller thereby yielding a larger effectiv capacitance. At the capacitors series resonant fre quency the denominator goes to zero and th expression becomes undefined. The relationshi of $C_E$ vs frequency is a hyperbolic function as illus trated in Figure 3.

### Example:

Consider an ATC 100A series 100pF capacitor.

Calculate the effective capacitance ($C_E$) at 10MHz 100MHz, 500MHz, 900MHz, 950MHz.

Solution: Calculate by using the relationshi $C_E = C_O/1 - (2\pi F_O)^2 L_S C_O$. Refer to Table 1.

***Table 1***

| Operating Frequency (MHz) | Effective Capacitance ($C_E$), pF | Impedance, (Ω |
|---|---|---|
| 10 | 100.01 | 0.013 – j 159.1 |
| 100 | 101.01 | 0.023 – j 15.76 |
| 500 | 133.34 | 0.051 – j 2.38 |
| 900 | 526.29 | 0.069 – j 0.337 |
| 950 | 1025.53 | 0.070 – j 0.168 |

***Relationship between $F_O$, $C_E$ and Z***

### Application Considerations:

Impedance matching and minimum drift applica tions such as filters and oscillators require specia attention regarding $C_E$. For applications below th capacitors self-resonant frequency the net imped ance will be capacitive (-j) whereas for applied frequencies above resonance the net impedanc will be inductive (+j). Operating above series res onance will correspondingly place the impedanc of the capacitor on the inductive side of the Smith chart (+j). When designing for these applications both $C_E$ and the sign of the net impedance at th operating frequency must be carefully considered

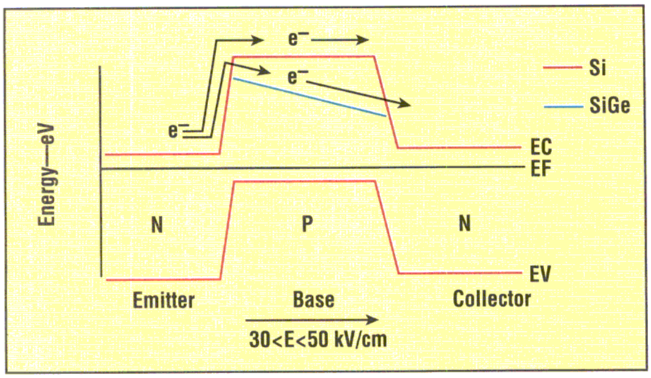
In contrast, the majority of coupling, bypass and DC blocking applications are usually not sensitive to the sign of the impedance and can be capacitive or inductive, as long as the magnitude of th impedance is low at the applied frequency. The effective capacitance will be very large and the ne impedance will be very low when operating close to resonance. At resonance the net impedance wil be equal the magnitude of ESR and the capaci tance will be undefined.


**Richard Fiore**
*Sr. RF Applications Enginee*
*American Technical Ceramics Corp*

ferent application than what is used for a 2-GHz RF front-end in a wireless radio.

## SiGe UNVEILED

A key to the search for a higher-speed semiconductor arises from the traditional view that if a device is made smaller—scaled—it will run faster. But there are practical limits to how small it can be. There is another way to increase speed, known as bandgap engineering. Si alone does not have the physical characteristics to produce transistors with the required performance, but if a second semiconductor material is introduced into the device, higher speeds can be attained through mobility, velocity saturation, or variations in the bandgap energy. It turns out that Ge, the first semiconductor material used to fabricate a transistor half a century ago, is such a material. Adding Ge grading to the base of an Si transistor boosts its speed significantly. The SiGe combination provides higher mobility than plain Si. In a SiGe HBT, the graded Ge material produces a built-in electric field that pulls electrons through the base region faster than in Si (Fig. 2).




**2. SiGe's fast switching speed stems from adding Ge grading to the base of an Si transistor. This produces an electric field that pulls electrons through the base region much faster than in conventional Si.**

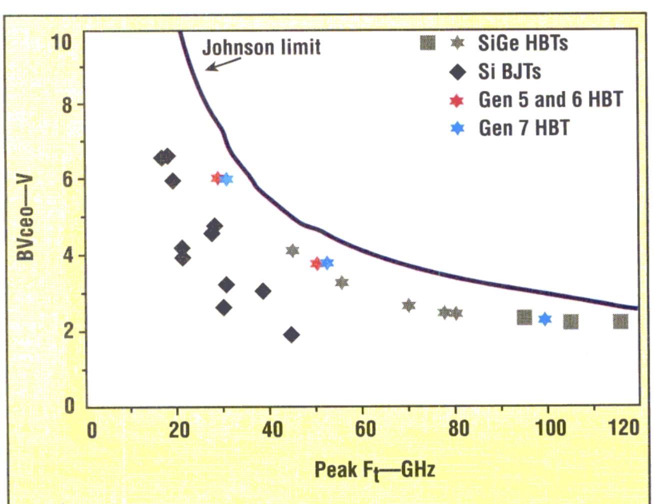
The smaller base bandgap of SiGe, when compared to Si, enhances electron injection which produces a higher current gain for the same doping level as a Si device. This permits the base to be heavily doped in order to lower the total base resistance. In addition, advanced processing techniques allow the Ge to be graded across the base in a controlled fashion that introduces a carrier-drift velocity that increases the device's speed (improving its frequency response). It should be pointed out that while bandgap engineering seems straightforward, creating the SiGe alloys without ruining the material took a long time to develop, which held up SiGe technology for several years.

From a cost standpoint, SiGe offers advantages over a process such as GaAs. As IBM's Meyerson points out, there is no such thing as a SiGe fab. That is, the devices can be manufactured in a conventional Si foundry using techniques that have existed for decades. GaAs, on the other hand, requires a special foundry. While the basic material cost of SiGe is greater than that of standard Si complementary metal-oxide semiconductor (CMOS), it is lower than that of GaAs material.

## SiGe AND MORE

As much as the low-power, high-speed features of SiGe will appeal to designers, the fact that it can be easily integrated with other processes such as CMOS and bipolar CMOS (BiCMOS) puts it in a different league from GaAs. At SiGe Microsystems (Ottawa, Ontario, Canada), a fabless chip company, Stephen Kovacic, director of research and development, reports that low power, low noise, and integrability are the key elements in the mobile-communications-system devices his company manufactures. "Integrating SiGe with CMOS gives you a huge advantage over GaAs," Kovacic says. "You can bring the front end of the radio into intimate contact with the back end. That is, you get a very high level of integration between the radio and the baseband processor, all on the same chip." Kovacic believes that such integration, "Has enormous implications in terms of power consumption. You can save a lot of power if you don't have to drive 50-Ω traces off chip. You just have to get the signal to the baseband processor which is a few hundred microns away on the same substrate."

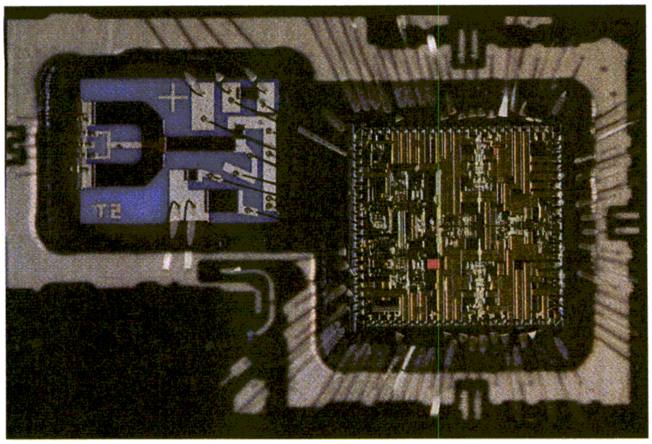
David Osika, chief scientist at ANADIGICS (Warren, NJ), agrees with Kovacic on some points, but he finds some compelling reasons for the continuing use of GaAs. "If integration of functions, including digital, is a strong driver for a product, then BiCMOS or SiGe BiCMOS makes more sense (than GaAs). If it's a high-performance socket, then GaAs, whether its HBT or pseudomorphic high-electron mobility transistor (PHEMT) or metal semiconductor field-effect transistor (MESFET) may outshine, on a raw performance basis, SiGe." One of the drawbacks of SiGe, and Si-based technologies in general, in Osika's view, is that parasitics can degrade performance in some applications. "Broadband applications have a difficult time dealing with parasitics, whereas in narrowband applications such as (per-




**3. A weakness of Si semiconductors is that they must obey the Johnson Limit, which shows that higher operating frequencies are accompanied by lower breakdown voltages.**

**CIRCLE NO. 249**

sonal-communications-services) PCS/cellular receivers (Rxs), they can be overcome by tuning, resonating out the parasitics." Thus, there will always be a need for GaAs and SiGe will not replace it in all applications.



**4. Complex ASICs such as the Tektronix device shown here rely on SiGe because it enables the integration of many functions on a single chip. This type of ASIC is used in the company's new TDS7000 series of real-time oscilloscopes.**

Another consideration when using SiGe revolves around the Johnson Limit. This physical law says that there is an inverse relationship between frequency and breakdown voltage in Si transistors including SiGe types. That is, the higher the operating frequency, the lower the breakdown voltage (Fig. 3). A SiGe device operating at 100 GHz, for example, would have a breakdown voltage between +2.0 and +2.5 VDC. If the operating frequency drops to 30 GHz, the breakdown rises to approximately +6.0 VDC. Put another way, the Johnson limit determines the speed (operating frequency) at which a Si device can operate.

Low breakdown voltage is not necessarily a drawback for SiGe. As SiGe Microsystems' Kovacic points out, the power rails in most equipment are on the way down. Many of today's devices operate at +3.3 VDC, but that could drop to +1.8 VDC in the near future and possibly to +1 VDC in a few years. The lower supply voltages bode well for SiGe's future since it could easily operate to 100 GHz at +1.8 or +1.0 VDC power rails, and as transistor suppliers suggest, there are not many current applications at that frequency.

## TESTING WITH SiGe

An area having significant growth potential for SiGe is test instrumentation. Two of the major oscilloscope manufacturers, Tektronix (Beaverton, OR) and LeCroy (Chestnut Ridge, NY), have been working for years with IBM to develop transistors for the front ends of their high-perfor-

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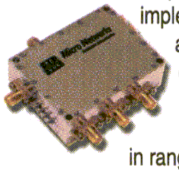

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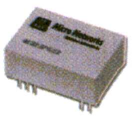

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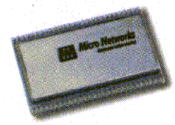
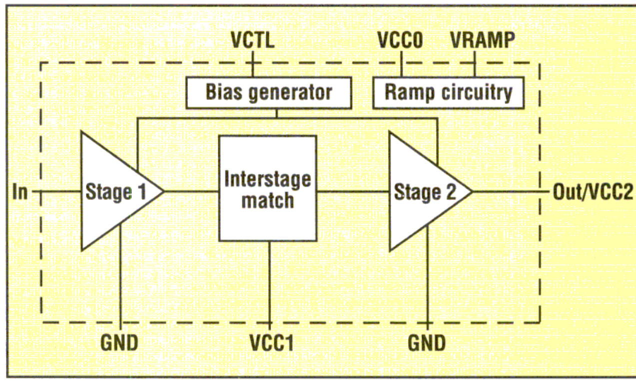
The M250 provides synchronous output frequencies of 622.08 MHz, 311.04 MHz, and 155.52 MHz. The input frequency is 19.44 MHz. Two output signals indicate an "Out of Lock" and "Loss of Reference Clock" condition, respectively. The device contains an internal 19.44 MHz reference clock which can be activated by an "Input Mode Select" control signal. Suitable for clock generation/distribution in fibercom systems and other applications requiring coherent frequency sources.



**www.micronetworks.com**

CIRCLE NO. 368

**Micro Networks**
**Andersen Laboratories**



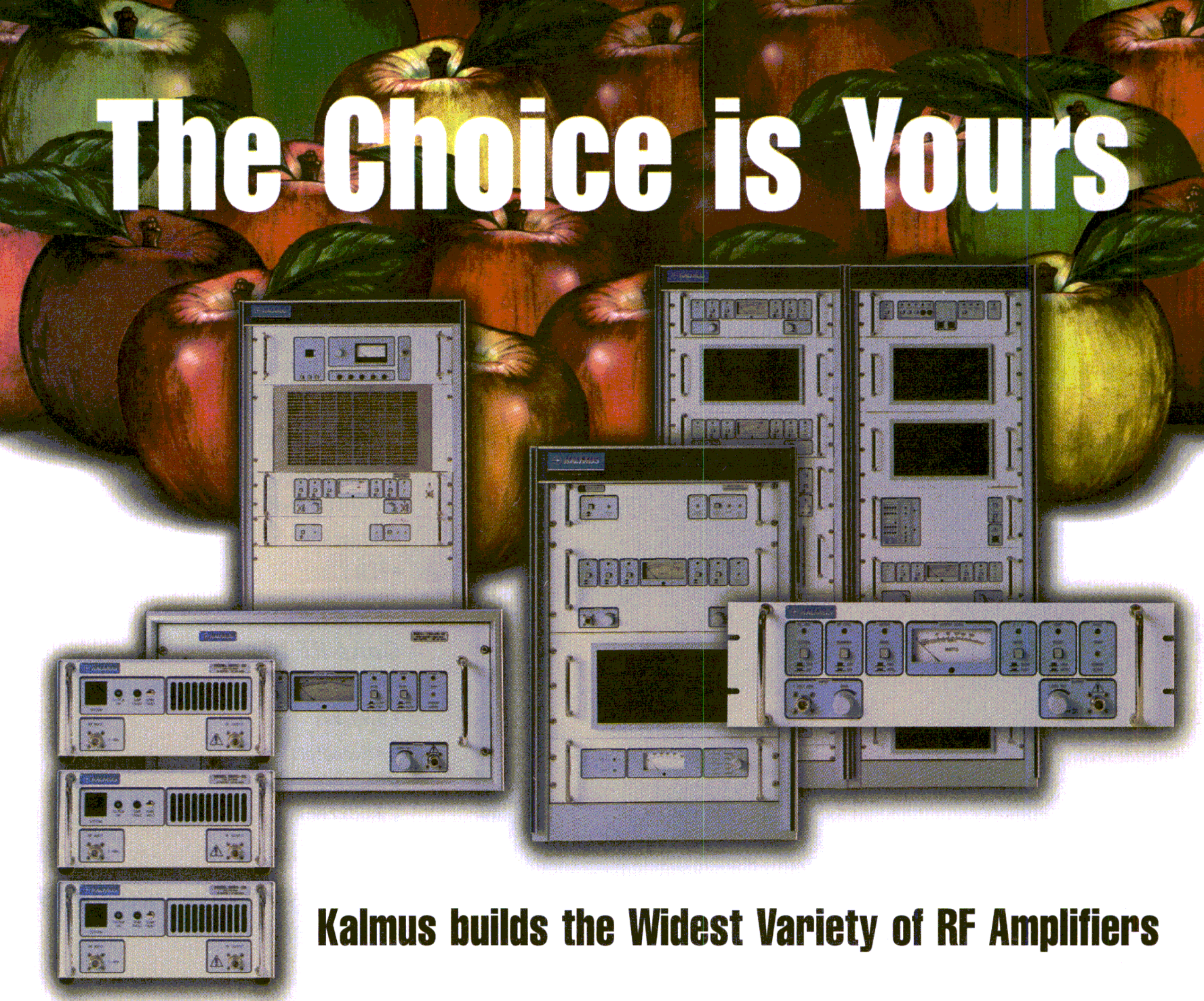

**5. Applications for SiGe are expanding as shown by this Bluetooth PA from SiGe Microsystems which meets Class-1 requirements.**

mance oscilloscopes.

According to Marv LaVoie, a Tektronix fellow and strategic technology manager for the Instrumentation Business Unit, " Tektronix has always had to develop its own (integrated circuits) ICs, because they are more complex than for other applications. We needed a technology like bipolar but with higher $f_T$ for analog-to-digital converters (ADCs), amplifiers, track-and-holds (T/Hs), and triggers." His colleague, Bob Woolhiser, engineering manager for the Premium Oscilloscopes division, adds, "We got involved in SiGe early with IBM—about four years ago—because we make large, precision application-specific circuits (ASICs) of around 50,000 devices, which is a large bipolar design (Fig. 4). We can't use GaAs because the level of integration is not high enough. SiGe is good, old silicon, and it is a much more optimal solution because of the integration issue." The company is incorporating the technology in its recently introduced family of TDS7000 series of real-time oscilloscopes. The TDS7404, for example, is a 0-to-4-GHz, 20-GSamples/s sample-rate instrument that requires high-speed SiGe transistors for its front-end preamps, T/Hs and other signal-processing functions. LaVoie and Woolhiser believe that the oscilloscope's performance is closely related to its SiGe technology in the front end.

The same power-saving rationale mentioned earlier for handheld telecommunications products holds true for test instrumentation. Says LaVoie, "You want to integrate as much as possible to eliminate driving off one chip into a lot of package delays. It costs time and wastes power." Despite the advantages of SiGe, it has some weaknesses that have to be designed around. Woolhiser says that it is difficult to laser-trim resistors to obtain the half-percent accuracy required, and the low breakdown voltages of SiGe at high frequency are also an area that requires some design ingenuity.

LeCroy has also developed a relationship with IBM for its SiGe technology. Like other devotees, it prizes SiGe's integration capability. Before going to SiGe, LeCroy used GaAs HBT transistors in oscilloscope front ends. "But," says Dave Graef, vice president of R&D, "we had problems with the HBT's DC stability, its DC performance at low frequencies. The drift was just not good

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**E-mail**: kalmus@kalmus.com • **Web**: www.kalmus.com

**CIRCLE NO. 303**

enough for an oscilloscope. It can be corrected, but with SiGe, you don't have to do that." The primary use of SiGe is for signal integrity in the vertical amplifiers, and it serves in a small-signal amplification capacity. "We don't care about its drive capability," says Graef, "because we have good control over how we drive our ADCs."

Although LeCroy is high on SiGe, it is not used in the company's new WavePro series of scopes because while it has working parts, the company wants to fine-tune the process a little longer. The WavePro 960 is the flagship of the line with a 1.5 GHz bandwidth, a sampling rate of 4GSamples/s on each channel, and combine its ADCs to sample at 16 GSamples/s when acquiring one channel. But SiGe will come into play in the company's next generation of instruments. Graef is sold on SiGe's prospects. "SiGe will be in the front ends of all high frequency measuring instruments in the future," he says.

**ALL SiGe APPLICATIONS, WHETHER THEY ARE IN TRANSCEIVERS, DISK DRIVES, OR TEST INSTRUMENTS, ARE COMMUNICATIONS.**

"GaAs is not the solution for LeCroy because there is no advantage over the wide frequency range from DC on up."

Some semiconductor experts believe that SiGe has virtually unlimited potential—operating frequencies in excess of 100 GHz. As Graef says, it is not only the process itself: "We don't know how high SiGe can go because there are factors such as the interconnects that affect performance." Indeed, IBM's development of copper (Cu) metallization in an SiGe process [instead of aluminum (Al)] has been found to have a significant impact on device speed. Cu has a lower resistance than Al, which enables more current to flow to a device in a shorter time, hence increasing its speed. IBM's Meyerson has the last word on SiGe's future when he says, "You can't win just by having a rocketship technology. You've got to have a good total solution—front end, back end, the metal, to achieve leadership, not SiGe alone."

## SiGe ARRIVES

Meanwhile, there is no shortage of chipmakers vying for shares of the SiGe pie. SiGe Microsystems is aiming at the potentially huge Bluetooth market with Class 1 power amplifiers (PAs) such as the PA2423F, a 22.5-dBm device that operates with a 50-percent power-added efficiency (PAE) at a supply voltage of +3.3 VDC. The IC's output power can be controlled continuously through a low-current control pin to accommodate settings in increments between 2 and 8 dB as required for Class-1 operation (Fig. 5).

Another competitor in Bluetooth is Atmel Corp. (San Jose, CA) which this spring introduced an RF front-end IC that contains a low-noise amplifier (LNA) and a PA. The T7024 is designed for Class-1 Bluetooth applications and can be used with the company's T2901 single-chip transceiver. When the ICs are used together, the company claims that the range of a Bluetooth system extends beyond 100 m.

Maxim Integrated Products (Sunnyvale, CA) is using SiGe to manufacture ICs for the cellular/digital-phone market. The MAX2323 includes all the circuitry from the antenna to the intermediate-frequency (IF) filter in a package that measures $5 \times 5$ mm. It is designed for dual-band, trimode IS-98A/B/C-based code-division-multiple-access (CDMA) phones, but it can also be applied in other access system such as wideband CDMA (WCDMA), Global System for Mobile Communications (GSM), and time-division multiple access (TDMA).

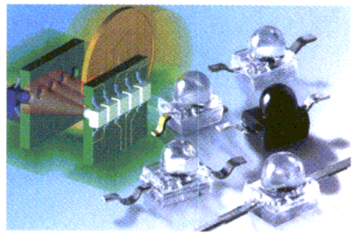
A family of low-current SiGe amplifiers that feature low power consumption ($I_d$ between 8 and 13 mA) is available from Stanford Microdevices (Sunnyvale, CA). The monolithic microwave integrated circuits (MMICs) are 50-$\Omega$ cascadeable amplifiers for front-end stage use in low-power applications—(P1dB) ranges from 1.0 to 5.0 dBm. The SGA-0x series operate from a single-supply voltage and contain on-chip bias circuitry to reduce system parts counts.

Early this year, Texas Instruments (Dallas, TX) introduced a BiCMOS process that incorporates SiGe for the purpose of fabricating wireless communications devices. The process delivers a peak $f_T$ of more than 50 GHz and an optimum $f_T$ of 25 GHz, which is ten times the operating frequency that the technology is designed for. TI claims that less than 20 $\mu$A is required for 25-GHz operation, providing power consumption said to be one-third less than that of competitive processes.

The speeds attainable from SiGe fit perfectly with the requirements of fiber-optics data technology which has been moving steadily into the tens of Gb/s territory. Last year, Lucent Technologies (Allentown, PA) introduced a process optimized for 10 Gb/s OC-192 SONET applications. The process involves a bipolar npn transistor together with a super self-aligned transistor structure for the SiGe. This combination can be added as a selective epitaxial process to the company's standard COM-1 CMOS process which, at the time, was running at 0.25 $\mu$. Lucent claims that the transistors have an $f_T$ switching frequency of 74 GHz and an $f_{max}$ of 116 GHz. Even in low-current applications, using smaller transistors, cutoff frequencies are above 60 GHz. Other companies have developed SiGe transistors for 40 Gb/s OC-768 SONET.

Motorola (Phoenix, AZ) has jumped aboard the bandwagon with a SiGe process that includes carbon (SiGe:C), and which can be folded into its BiCMOS platform. Behind the carbon (C) initiative is the need to integrate more passives on RF devices, and this technology supports simpler integration than do others. ••

## IR devices suit communications

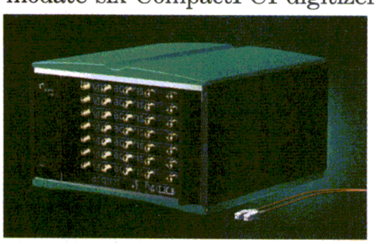


Four new series of discrete infrared (IR) devices include two sets of receivers (Rxs) and emitters. The TEMT10xx series phototransistor and the TSML10xx series emitter operate at 950 nm and are designed for photo-interrupter, miniature-light-barrier, optical-counter, and IR-sensor applications. The TEMD10xx series positive-intrinsic-negative (PIN) photo diode and the TSMF10xx series emitter operate at 870 nm and are designed for IrDA-compliant data transmission, high-speed optical encoder systems, and fiber-optic modules. The TEMT10xx phototransistors have a turn-on time of 2 μs. The TMSL10xx emitters provide a radiant power of 35 mW. The TEMD10xx PIN diodes provide 4-ns rise and fall times. The TSMF10xx emitters provide 30-ns rise and fall times. All four devices are housed in a miniature surface-mount package with integrated lens measuring 2.5 × 2.0 × 2.8 mm. **Vishay Intertechnology, Inc., 63 Lincoln Highway, Malvern, PA 19355; (610) 644-1300, FAX: (610) 296-0657, Internet: http:// www.vishay.com.**

CIRCLE NO. 58 or visit www.mwrf.com

## DAQ system conserves space and power

A multichannel, data-acquisition (DAQ) system can accommodate tens of fully synchronized channels for monitoring signals at frequencies up to 500 MHz while reducing space and power consumption to as little as 1/8th of conventional rack-and-stack modular systems. Each module—called a "crate"—can accommodate six CompactPCI digitizer cards. The system can be configured with an embedded 500-MHz Pentium II processor or can be connected to any benchtop personal computer (PC) using a peripheral component interface (PCI). When configured with the embedded processor, each crate can accommodate up to 24 channels at 1 GSamples/s or 12 channels at 2 GSamples/s. A utility bus supports channel synchronization, and trigger distribution supports triggering on any channel. Several crates can be connected to a main host computer through Ethernet. **Acquiris USA, P.O. Box 2203, 234 Cromwell Hill Rd., Monroe, NY 10950-1430; (914) 782-6544, FAX: (914) 782-4745, Internet: http://www.acqiris.com.**



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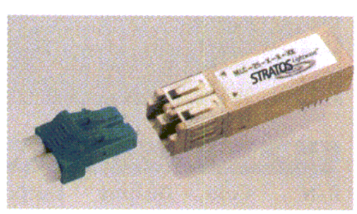
## Optical transceivers reach 2.488 Gb/s




A series of small-form-factor (SFF) optical transceivers can run at speeds up to 2.488 Gb/s for asynchronous-transfer-mode (ATM) OC-48 applications. The transceivers are available in four different versions—one for 850-nm multimode applications, and three for 1300-nm single-mode applications. The 850-nm version uses a vertical-cavity, surface-emitting laser to provide data rates to 2.488 Gb/s. The three 1300-nm versions provide communication over single-mode fiber at maximum transmission distances of 2, 10, and 20 km and meet or exceed existing ATM OC-48 requirements. The transceivers feature a transistor-transistor-logic (TTL) signal-detect output and transmitter (Tx)-disable input. The transceivers have low-profile enclosures to fit mezzanine-card systems and they operate from a single +3.3-VDC power supply. **Stratos Lightwave LLC, Optoelectonic Products, 7444 W. Wilson Ave., Chicago, IL 60706-4549; (708) 867-9600, FAX: (708) 867-0996, Internet: http://www.stratoslightwave.com.**

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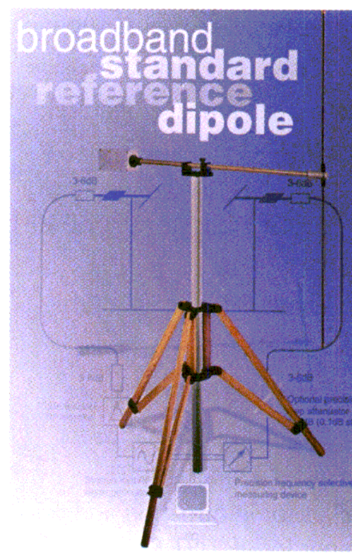
## EMC site-calibration system speeds verification

The BSRD6500 series site-calibration system combines hardware and software to verify electromagnetic-calibration (EMC) test sites. Rather than using the traditional 24-spot-frequency method, the system uses standard reference dipole antennas and a broadband-swept technique to meet the new CISPR-16 requirement. The manufacturer claims it can reduce cost up to 40 percent, and verification times by as much as 80 percent compared to conventional techniques. The software can predict




site attenuation and antenna factors for outdoor sites and for fully anechoic rooms. The software can calibrate test sites with an uncertainty of better than ±0.2 dB in the 30-to-600-MHz range and ±0.3 dB in the 600-to-1000-MHz range. The antennas can handle power levels to 10 W. **Schaffner-Chase EMC Ltd., Broadwood Test Centre, Rusper Rd., Cape, Dorking, Surrey RH5 5HF, United Kingdom; +44 1306 710205, FAX: +44 1306 713303, Internet: http://www.schaffner.com.**

CIRCLE NO. 61 or visit www.mwrf.com

# Bypassing College Can Bring Big Money

Leave it to the technological revolution brought on by computers and communications to turn conventional wisdom on its head. Statistics say that the more education one has, the greater the person's income earning potential, and a college education is an increasingly important ticket to a good job. But some recent high school graduates are able to earn the high salaries promised by an undergraduate degree using self-taught computer skills.

According to an article in *The New York Times*, these 18-to-20-year-olds can command salaries of $30,000 and higher, some with stock options and other perks, to go directly into the job market. Most of them see college as impeding rather than furthering their knowledge of computers and technology. A common observation of these fledgling Bill Gates is that the computer industry is changing so quickly that spending four years studying rather than doing will leave them behind the times technologically when graduation comes. Contrary to what one might expect, none were among the intellectual or social elite of their high-school classes. Rather, they were the so-called computer nerds or geeks who spent most of their free time and a lot of their school time learning about advanced system techniques such as setting up secure e-commerce sites and tracking customer-service performance on the Internet. As one young man observed, "Everyone is giving you the positives of going to college, but when you look around, what you see are the positives of making $80,000 a year."

The fact is, opportunities abound for those with the interest and ability to go into the computer industry in all of its facets. In the 1999 edition of the authoritative *Statistical Abstract of the United States*, US Census Bureau employment projections by occupation are made for the 10-year period from 1996 to 2006. The three fastest-growing occupations listed are database administrators and computer-support specialists, computer engineers, and systems analysts. Each is projected to have a growth rate exceeding 100 percent. Translating the rates into jobs, more than one million systems analysts will be needed, along with approximately 450,000 each of data-base administrators, support specialists, and computer engineers. The Bureau assumes that the minimum educational requirement for each type of position is a bachelor's degree, but obviously, some of today's whiz kids would disagree with that advice.

An interesting side note to the projections provided previously is that they must have been made before the recent surge of Internet, wireless communications, and e-commerce activity. This probably makes the jobs projections on the low side. ••

## Contracts

**Vitesse Semiconductor Corp.**—Announced that it has completed the acquisition of certain assets of the wide-area-network (WAN) product line of Philips Semiconductors for approximately $30 million in cash. The transaction will be accounted for as a purchase.

**Gabriel Electronics, Inc.**—Was recently awarded a $100,000 performance grant by the Maine Technology Institute of Gardiner, ME. The award will assist Gabriel in expanding its SectorWave™ broadband multipoint base-station antenna product line.

**SyChip, Inc.**—Has won a major contract to jointly develop and supply chip-scale modules to CEC Wireless R&D Ltd. of Beijing, China, a producer of wireless handsets. The contract has the potential to be worth several million dollars to SyChip over the next several years.

**RF Micro Devices**—Has received production orders to supply power amplifiers (PAs) for Sanyo's latest dual-band code-division-multiple-access (CDMA) personal-communications-services (PCS) handset. Sanyo's customers include Sprint PCS. Volume shipments have begun and will ramp to a multimillion-dollar level during this calendar year.

**Motorola, Inc.'s Network Solutions Sector (NSS)**—Has signed 10 new digital cellular-network expansion contracts with China Unicom worth a total of approximately $30 million. Under the terms of the contract, Motorola will expand China Unicom's GSM-900 networks in eight of China's major provinces, which together boast a population of 421 million people.

**Trimble**—Announced that Digicore Holdings Ltd., a South African fleet-management supplier, has placed an order for 25,000 Global Positioning System (GPS) receivers (Rxs). The Trimble GPS Rxs will be used in Digicore's C-Track, a real-time fleet-management system. Digicore secured a five-year, $22 million contract to equip 19,000 Telkom Ltd. vehicles.

**Granada Media**—Entered into a 47 million pound (approximately $70.5 million US) strategic partnership with London, England-based soccer club Arsenal FC. The deal includes a 50-percent interest for Granada in AFC Broadband, a joint-venture company that will set up a global portal to exploit Arsenal FC's New Media rights. Over time, the portal will be developed and expanded for international distribution. The portal's services will be delivered to home and personal devices, including personal computers (PCs), digital set-top boxes, personal digital assistants (PDAs), and wireless-application-protocol (WAP) phones.

## Fresh Starts

**TeliSmart**—Launched the TeliSmart.com website, an online global marketplace for the telecommunications industry. Telismart.com brings buyers and sellers together for the real-time exchange of used, surplus, and decommissioned telecom assets.

**Sanders and M/A-COM**—Have signed an agreement under which M/A-COM will add production of monolithic microwave integrated circuits (MMICs) at Sanders' Microelectronics Center (MEC) in Nashua, NH. The facility will be used to support products developed by both companies for the wireless and aerospace markets. Under a Cooperative Support and Technical Assistance Agreement, M/A-COM will manufacture MMICs on 6-in. gallium-arsenide (GaAs) wafers at Sanders' Nashua, NH MEC facility for wireless telecommunications applications, while Sanders will continue to manufacture MMICs at the same facility for aerospace and defense markets.

**Agilent Technologies**—Announced that it is building a dedicated 6-in. wafer-fabrication line in its Newark, CA facility for its innovative film-bulk-acoustic-resonator (FBAR) filter products. Agilient is also outfitting high-volume automated back-end manufacturing capabilities in its Penang, Malaysia facility, and is increasing engineering and production staffing to support this manufacturing ramp.

**Bird Component Products**—Has appointed two companies to represent its line of attenuators, loads, power divider/combiners, and couplers. Matrix Controls Corp. will cover Texas, Arkansas, Oklahoma, and Louisiana. Wes Tech Associates will cover Kansas, Missouri, Iowa, Nebraska, Minnesota, Wisconsin, and Illinois.

**Tyco Electronics Corp.**—Has reached an agreement for a major patents license with British Telecommunications, plc. (BT). This license agreement will permit implementation of BT's technology. The license agreement strengthens Tyco's current line of optical cutters and splitters, and complements their other passive product offerings including wavelength-division multiplexers (WDMs), dense WDMs (DWDMs), attenuators, isolators, switches, and integrated optoelectronic modules.

**Link Microtek Ltd.**—Has been appointed as the exclusive representative in the United Kingdom and Ireland for Narda Safety Test Solutions. As the UK representative, Link Microtek will offer customers an expanded portfolio of products, including the EMR series survey equipment and the popular Rad Man personal monitors.

**Excellon Automation Co. and PhotoMachining, Inc.**—Have established a cooperative technology-sharing agreement in the field of laser micromachining. The working agreement is designed to build on each company's strengths in the laser micromachining arena. While the terms of the agreement are not disclosed, PhotoMachining will take delivery of an LVD 2001 diode-pumped dual-laser infrared and ultraviolet (IR and UV) system.

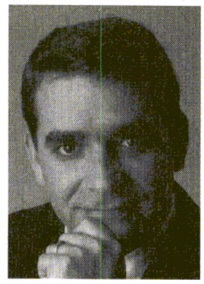
**Gowanda Electronics**—Recently held a groundbreaking ceremony to signal the start of construction of its new state-of-the-art, 33,000-sq.-ft. facility at a new site, near the company's existing location, in Gowanda, NY. It is anticipated that construction will be completed by the end of the year.

**Analog Devices, Inc. and Minebea Co. Ltd.**—Announced that the two companies have formed an agreement to jointly develop resolver systems for the automotive market. This joint-development agreement will result in new resolver systems that will be offered at a lower price than commonly found in the industry, making the systems affordable and easy to use for high-volume automotive applications. These applications include electric power steering (EPS), starters/generators, and hybrid electric vehicles.

**Advanced Hardware Architectures (AHA)**—Gary Hahnert to vice president of sales and marketing; formerly North American business-development manager for the cable and network operation at Intel.

**Sanders**—Randal E. Morger to director of communications and administrative services; formerly public information manager.

MORGER



PERROTTA

**Quad Systems Corp.**—John Perrotta to vice president of technical services; formerly senior director of quality assurance. Also, Joel A. Schoubert to international service manager; formerly international technical trainer and service engineer.

**Microcosm Technologies, Inc.**—Glenn E. Harder to executive vice president and chief financial officer (CFO); formerly executive vice president and CFO of Carolina Power & Light Co. (CP&L).

**CPI Wireless Solutions**—Mike Cheng to president of the Eimac Division; formerly vice president of operations at the MPP Division.

**AIR2LAN**—Larry Oaks to vice president of national sales; formerly vice president of sales for NEXTLINK Communications, Inc.

**Micro Networks Corp.**—Dr. Franz Z. Bi to director of surface-acoustic-wave research and development (SAW R&D); formerly held a senior SAW technical position at Vectron International USA.

**Varian, Inc.**—Frederick C. Campbell to operations manager at the Vacuum Technologies facility in Lexington, MA; formerly vice president of research and development (R&D) for Mentor Corp.

**JMS Worldwide, Inc.**—R.J. (Rocco) LaPenta to chief operating officer (COO); formerly vice president of telecom services.

**Aegis Broadband, Inc.**—Holly Bertz to director of marketing communications; formerly was responsible for worldwide marketing communications at ComStream Corp.

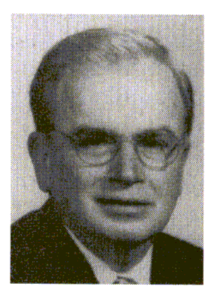
**Cidera, Inc.**—David A. Gollob to vice president of product strategy and management; formerly director of programming development for Road Runner.

**Tekelec**—Paul J. Pucino to vice president and chief financial officer (CFO); formerly CFO at Scour.com.

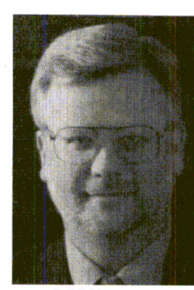
**TriPoint Global Communications, Inc.**—Gregory J. Smith to vice president of operations at VertexRSI, a TriPoint Global company; formerly vice president of worldwide operations for AVEX Electronics, Inc.

**Microchip Technology, Inc.**—Larry Ross to vice president of Far East sales; formerly manager of the Asia Pacific sales organization.

**Cherry Electrical Products**—Robert Ralston to national distributor marketing manager; formerly vice president of product merchandising with the Newark Electrical Products Division of Premier Farnell.



RALSTON



BERNHARDT

**Stellex Technologies, Inc.**—Christopher C. Bernhardt to president and chief executive officer (CEO); formerly president of Litton Data Systems.

**CTS Corp.**—James M. LaNeve to controller for CTS' RF Integrated Modules business in West Lafayette, IN; formerly planning and budgeting manager at CTS' corporate offices.

**MDeverywhere**—David Morris to the position of vice president of business development; formerly senior vice president of business development at ProxyMed.

## Short Courses

**Principles of Modern Radar**
October 30-November 3 (Atlanta, GA)
Georgia Institute of Technology—Continuing Education
Atlanta, GA 30332
(404) 385-3502, FAX: (404) 894-7398
e-mail: conted@gatech.edu
Internet: http://www.conted.gatech.edu.

**Next Generation Networks—Technical Overview**
November 7-8 (Phoenix, AZ)
Telcordia Technologies
Lisle, IL 60532-3198
(800) 832-2463, FAX: (630) 960-6230
e-mail: teachme@telecordia.com

**Integrated Wide Area Networks for the Enterprise**
November 7-10 (Washington, DC)
Learning Tree International
Reston, VA 20190-5630
(800) 843-8733
Internet: http://www.learningtree.com

**Wideband CDMA Communications**
November 8-9 (Mountain View, CA)
Besser Associates
201 San Antonio Circle, Bldg. E
Mountain View, CA 94040
(650) 949-3300, FAX: (650) 949-4400
e-mail: info@bessercourse.com

**Fundamentals of Cellular and PCS Wireless Communications**
November 13-15 (Madison, WI)
Dept. of Engineering Professional Development
University of Wisconsin-Madison
Madison, WI 53706
(800) 462-0876, FAX: (608) 263-3160
e-mail: custserv@epd.engr.wisc.edu/
Internet: http://epd.wisc.edu/

**Multirate Signal Processing in Transmitter and Receiver Designs**
November 15-17 (Los Angeles, CA)
UCLA Extension, Dept. of Engineering, Information Systems and Technical Management
10995 Le Conte Ave.,
UCLA Extension Bldg.
Los Angeles, CA 90024
(310) 825-3344, FAX: (310) 206-2815
Internet: http://www.uclaextension.org/shortcourses

## Meetings

**2000 IEEE GaAs IC Symposium**
November 5-8 (Westin Hotel, Seattle, WA)
IEEE Electron Devices Society, Solid-State Circuits Society and Microwave Theory and Techniques Society
Mary Clemente
445 Hoes Lane
Piscataway, NJ 08855
(732) 562-5350, FAX: (732) 981-1203
e-mail: meclemen@ieee.org.

**2000 IEEE APS Conference on Antennas and Propagation for Wireless Communication**
November 6-8 (Westin Hotel, Waltham, MA)
IEEE Boston Section
240 Bear Hill Rd., Suite 20B
Waltham, MA 02451-1017
(781) 890-5290
Internet: http://www.eece.unm.edu/awpc2000

**Managing Bandwidth in the Optical Internet**
November 8-10 (Fairmont Hotel, Dallas, TX)
Institute for International Research
New York, NY 10017-4103
(888) 670-8200, FAX: (941) 365-2507
Internet: http://www.iir-ny.com

**IMAPS Advanced Technology Workshop on Packaging and Integration of MEMS and Related Microsystems**
November 10-12 (Orlando Marriott Hotel, Orlando, FL)
International Microelectronics and Packaging Society (IMAPS)
Internet: http://www.imaps.org

**56th ARFTG Microwave Measurements Conference**
November 28-December 1 (Omni Hotel, Boulder, CO)
Automatic RF Techniques Group (ARFTG)
David Walker
National Institute of Standards and Technology (NIST)
Boulder, CO 80305
(303) 497-5490
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## Digital radio broadcasting is clearer with FEC

Commercial-free music, digital-quality sound and seamless coast-to-coast coverage are enjoyed by motorists in the US through the digital satellite-to-vehicle radio-broadcasting system. But the system is susceptible to fading and other impairments that interfere with reception. To make the system more robust, Hui-Ling Lou and Vijitha Weerackody of Bell Telephone Laboratories, Lucent Technologies (Murray Hill, NJ), and M.J. Fernández-Getino García of the Department of Signals, Systems, and Radio Communications, Polytechnic University of Madrid (Spain) propose a forward-error-correction (FEC) scheme. It is based on two time-division-multiplexed (TDM) satellites transmitting different streams of source-coded bits subject to different interleaver depths. This allows one TDM satellite transmission to be used to minimize tuning delays and the other to obtain a better quality signal. The source-coded bit stream can be FEC coded, interleaved, and transmitted using different satellites. See "FEC Scheme For A TDM-OFDM Based Satellite Radio Broadcasting System," *IEEE Transactions on Broadcasting*, March 2000, Vol. 46, No. 1, p. 60.

## Reflector improves cell-phone antenna efficiency

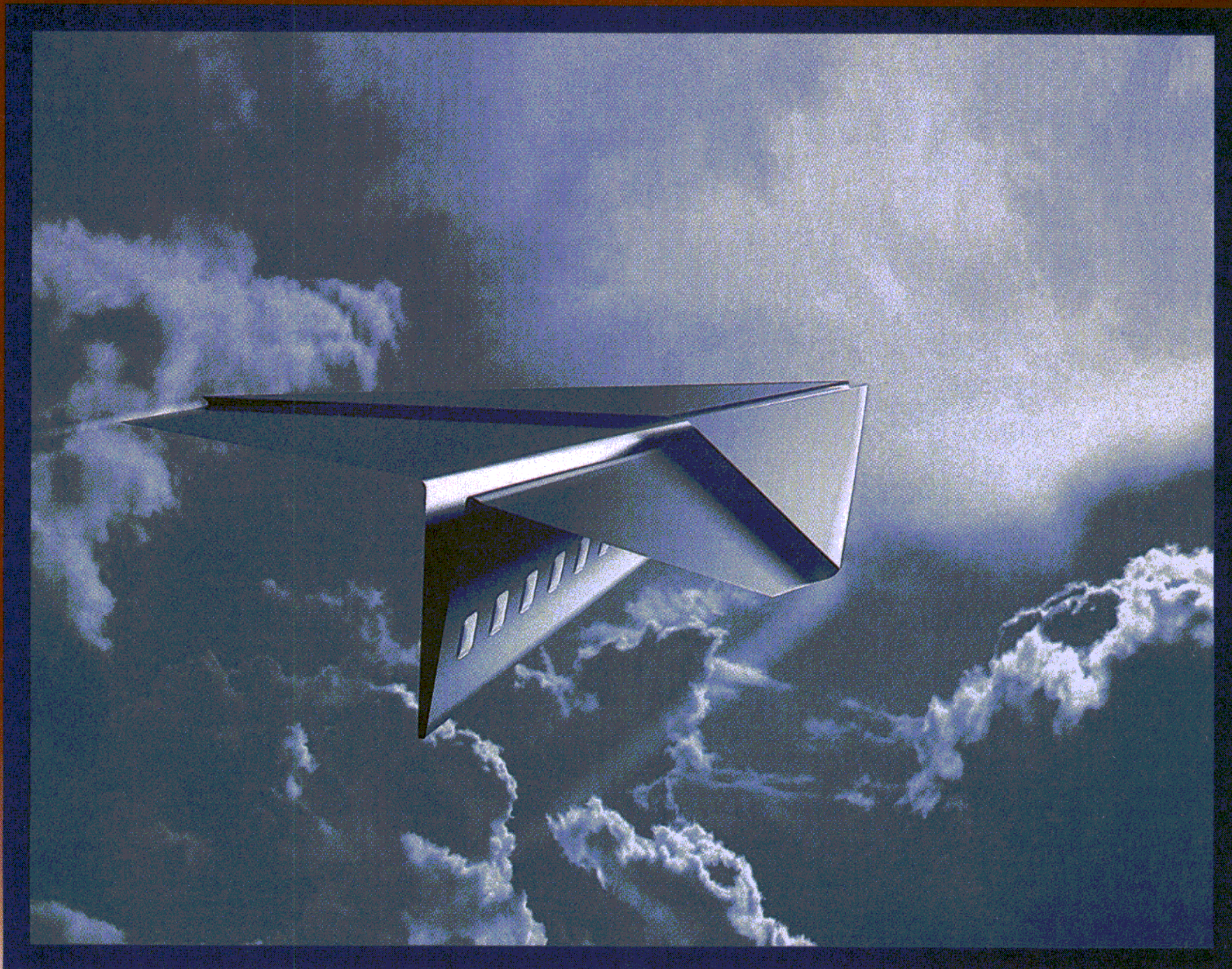
Health concerns aside, cellular-phone antennas lose efficiency because some of their power is radiated into the head of a human speaker. Antenna efficiency can be improved using a reflector between the antenna and the head, according to Eiji Hankui of the NEC Corp. EMC Engineering Center (Kanagawa, Japan) and Osamu Hashimoto and Shinichiro Nishizawa of Aoyama Gaukin University (Tokyo, Japan). The researchers experimented with a number of different reflector materials and measured radiation efficiency and power loss using finite-difference-time-domain (FDTD) methods. After an extensive analysis and measurement program, they arrived at the basic conclusion that a magnetic material is more effective for increasing radiation efficiency than either conductive or dielectric materials. A specially constructed model was used to simulate the cell-phone antenna and the human head. See "An Application of a Reflector for Increase of the Radiation Efficiency of Cellular Telephone Antennas," *International Journal of RF and Microwave Computer-Aided Engineering*, July 2000, Vol. 10, No. 4, p. 253.

## Spatial techniques open wireless ATM spectrum

In the quest for greater transmission bandwidth, wireless asynchronous-transfer-mode (ATM) systems such as HIPERLAN/2 can take a page from the book of cellular mobile-communication systems by using space-division-multiple-access (SDMA) techniques. Researchers Ulrich Vornefeld, Christoph Walke, and Bernhard Walke of the Aachen University of Technology (Germany) claim that the spatial dimension helps to increase transmission capacity, reduces cell interference, and leads to more efficient use of the scarce radio spectrum. Two types of spatial signal-processing algorithms are required for SDMA—spatial filtering to separate signals impinging on the antenna array during reception, and beamforming algorithms to control radiation directions during transmission. Although the SDMA technique was developed for HIPERLAN/2, the authors claim that their ideas are applicable in general and can be transferred to various systems. They describe a time-division-multiple-access (TDMA)/SDMA system that has features such as a reservation-based media-access-control (MAC) protocol and automatic-repeat-request (ARQ) error recovery at the radio interface. See "SDMA Techniques for Wireless ATM," *IEEE Communications Magazine*, November 1999, Vol. 37, No. 11, p. 52.

## Leaky local oscillator does double duty

RF transceiver designers are always on the hunt to reduce the bill-of-materials, making the end product less expensive, smaller, more reliable, and with lower power. A method used to make an oscillator perform two functions comes from Jiazong Zhang and Zhizhang Chen of the Department of Electrical and Computer Engineering of Dalhousie University (Halifax, Canada) and Yunyi Wang of the Department of Radio Engineering, Southeast University (Nanjing, People's Republic of China). The authors offer a design where the power of a local oscillator (LO) in the transmitter (Tx) portion of a transceiver is intentionally leaked into the receiver (Rx) circuitry through an antenna, and is used as the LO for the Rx. Thus, just a single LO serves the transmitting and receiving channels and no duplex switch is required, eliminating an expensive component. A rectangular patch antenna permits virtually independent transmitting and receiving frequencies. See "Use of Antenna In Constructing a Single LO For Transmitting And Receiving Duplex Circuitry," *Microwave And Optical Technology Letters*, July 20, 2000, Vol. 26, No. 2, p. 122.





## Introduction to Wireless Local Loop, 2nd Ed.: Broadband and Narrowband Systems

**William Webb**

Wireless local loops (WLLs) can be thought of as fixed wireless links in place of copper (Cu) cables in telephone networks, usually as a means of providing telephone service to the home. In contrast to other wireless applications, WLLs do not benefit from standardization; as a result, the use of technology varies greatly across different WLL systems. *Introduction to Wireless Local Loop, 2nd Ed.: Broadband and Narrowband Systems* is a thoughtfully written attempt to clear up the misunderstandings surrounding WLLs.

The opening chapter explains the role of WLL in modern communications systems, and provides a justification for the writing of the book. Chapter 2 discusses the importance of WLL in terms of converging communications markets and customer demands, while Chapter 3 briefly reviews communications access technologies, such as digital subscriber line (DSL) and cellular systems.

Chapter 4 explores the international telecommunications environment for WLLs, including Eastern Europe and developing nations, and Chapter 5 compares the economics of using wireless links and traditional wired communications systems. Chapter 6 briefly explains why initial predictions of WLL markets were overly optimistic, and what can be expected of future markets.

Chapter 7 provides a cursory look at radio propagation, including the effects of attenuation, slow and fast fading, and reflections, while Chapter 8 offers a quick overview of radio systems, including descriptions of speech-coding, error-correction, modulation, multiple-access techniques such as time-division multiple access (TDMA) and code-division multiple access (CDMA). As might be expected, Chapter 9 asks the decade-old question: Is it better to use TDMA or CDMA?

Chapter 10 offers an overview of narrowband and broadband technologies. Chapter 11 provides a closer look at cordless-telephone technologies, while Chapter 12 covers cellular technologies. Chapter 13 highlights proprietary technologies, such as Lucent's AirLoop technology, while Chapter 14 discusses broadband technologies, including microwave-multipoint-distribution-system (MMDS) and local-multipoint-distribution-system (LMDS) approaches. Chapter 15 guides the reader to a proper choice of the correct technology for a given application, while Chapter 16 provides greater details on CDMA technology.

*Introduction to Wireless Local Loop, 2nd Ed.: Broadband and Narrowband Systems* is easy to read and understand. It is a text with much value, including such sections as Chapter 17 in which detailed instructions are provided for obtaining a WLL license, Chapter 18 which explains how to choose a service offering, and in the final chapter (Chapter 19) in which the reader is taught how to develop a business case for WLL. (2000, 403 pp., hardcover, ISBN: 1-58053-071-0, $83.00.) **Artech House, Inc., 685 Canton St., Norwood, MA 02062; (800) 225-9977, (781) 769-9750, FAX: (781) 769-6334, e-mail: artech@artech-house.com, Internet: http://www.artech-house.com.**

# Devise A Millimeter-Wave PHEMT Mixer

***This approach uses harmonic-balance and conversion-matrix optimization methods to produce a drain mixer with good conversion gain.***

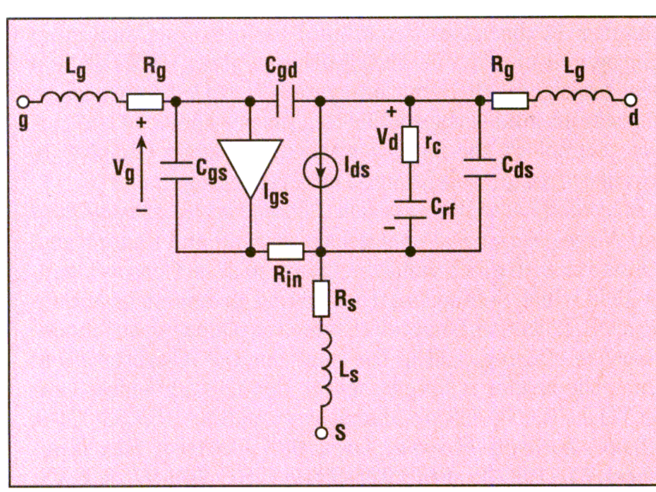
**Xu Ruimin, Xiao Shaoqiu, and Yan Bo**

*Institute of Applied Physics, University of Electronic Science and Technology of China, Chengdu, Sichuan 610054, China; FAX: 86 28 3202526, e-mail: rmxu@uestc.edu.cn.*

**MIXERS, which are key circuits in superheterodyne receivers (Rxs), can be classified in two main categories: diode mixers and three-port-device mixers. In millimeter-wave Rxs, three types of three-port devices are used as mixers—the field-effect transistor (FET), the high-electron-mobility transistor (HEMT), and the pseudomorphic high-electron-mobility transistor (PHEMT). These three-port-device mixers have advantages over diode mixers, including higher conversion gain, wider dynamic range, and better natural isolation between ports. Some of the literature available on FET, HEMT, and PHEMT mixers show that these devices perform well,[1] even when using hybrid integrated-circuit technology in the Ka-band.[2]**

Since FETs, HEMTs, and PHEMTs each have a gate, source, and drain, each can be configured as a gate mixer, a source mixer, or a drain mixer. A gate mixer is complicated because it requires an input directional coupler, which is difficult to fabricate and suffers substantial loss in the millimeter-wave band. A source mixer has a low conversion gain[3] and is difficult to ground. A drain mixer, however, uses the inherent isolation characteristics of the three-port-device terminals and offers good conversion gain. This article describes a Ka-band PHEMT hybrid integrated-circuit drain mixer that provides better performance than its predecessors.

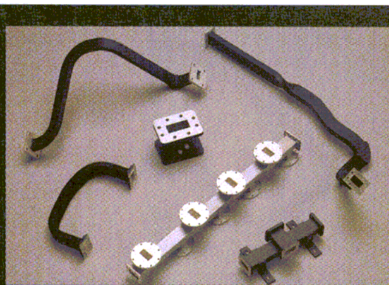
The study of high-frequency equivalent circuits is essential to solid-state-device application, and is an important link to computer-aided design (CAD). The authors chose the PHEMT nonlinear model because it is succinct and accurate, and selected the model A41005C000 PHEMT from Alpha Industries because it is a

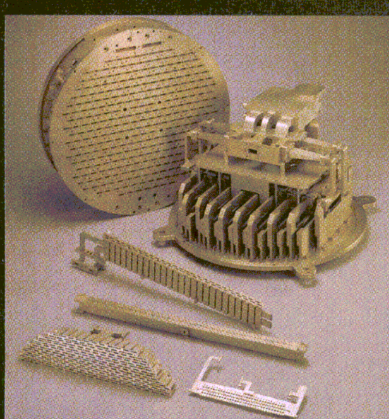


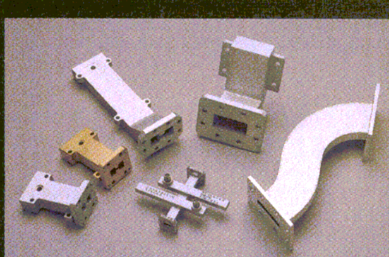

**1. This schematic shows the PHEMT's nonlinear equivalent circuit.**

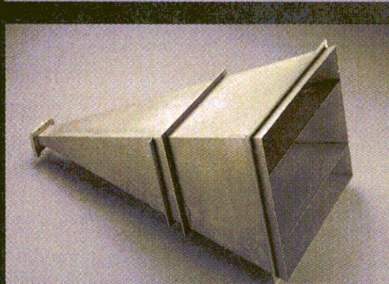
1.5
3.7
5.2

# CMT WAVEGUIDE COMPONENTS AND ASSEMBLIES







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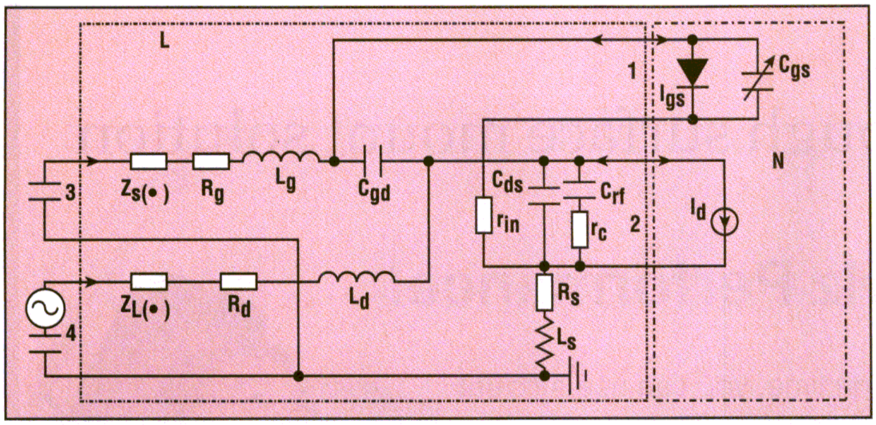
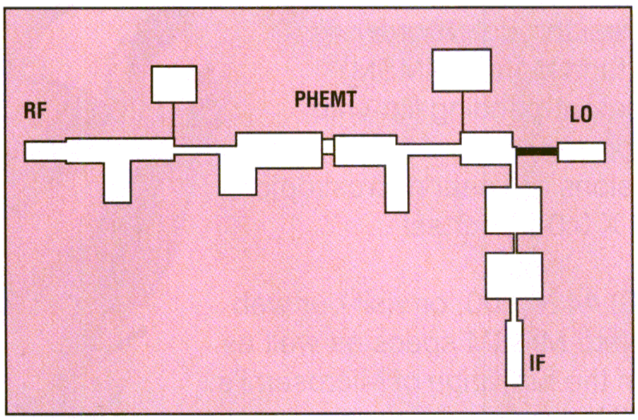



**2. This schematic illustrates the complete drain mixer, including linear and nonlinear networks used for harmonic-balance analysis.**

low-noise device. Figure 1[4] shows the device's nonlinear equivalent circuit.

The equivalent circuit contains three nonlinear elements that depend on voltages $v_g$ and $v_d$, namely $C_{gs}(v_g)$, $I_{gs}(v_g)$, and $I_{ds}(v_g,v_d)$. The equations describing these values appear below.




**3. The drain mixer's microstrip topology is shown here.**

$$C_{gs} = C_{gs_0} / \sqrt{1 - V_g(t)/V_{bi}} \qquad (1)$$

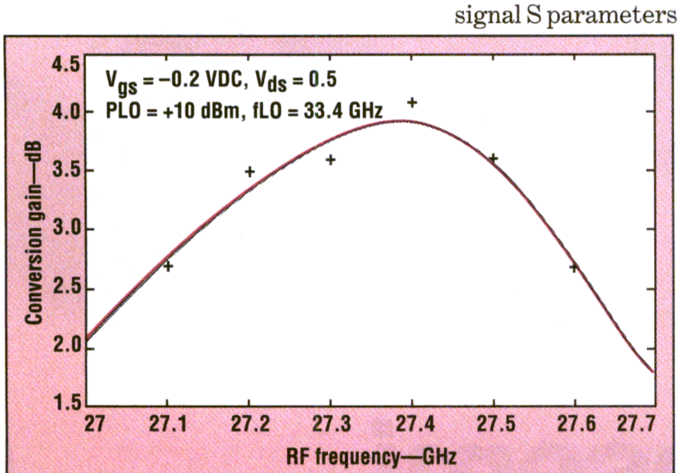
$$I_{gs} = I_{dss}(e^{qV_g(t)/nkT} - 1) \qquad (2)$$

$$V_h = V_g(t-\tau)(1+\beta(V_{ds_0} - V_d(t))) \qquad (4)$$

The $\tau$ is the delay constant of the device. $C_{gs}$ is modeled as an ideal Schottky-barrier capacitance, and $I_{gs}$ is modeled as a diode. The nonlinear equivalent-circuit-element values were extracted from the small-signal S parameters and from DC I/V characteristics.

Harmonic-balance and conversion-matrix methods are accurate and classical methods for analyzing mixers.[5]

To determine the content of harmonic balance, the circuit is separated into linear and nonlinear parts to form linear




**4. This graph illustrates the drain mixer's conversion gain versus RF frequency.**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and nonlinear networks. The linear network contains the matching circuit, source impedance, and linear elements in the PHEMT equivalent circuit. The nonlinear network contains only the nonlinear elements in the PHEMT nonlinear model. Because the local-oscillator (LO) signal is injected at the drain, the network is separated as shown in Fig. 2.

The principle of harmonic balance requires that all port currents between the nonlinear network and the linear network be balanced, so the error function of the currents can be given as:

$$F(V) = Y_{2\times 2}\begin{bmatrix} V1 \\ V2 \end{bmatrix} + Is + \begin{bmatrix} I_{nol1} \\ I_{nol2} \end{bmatrix} \qquad (5)$$

where:

$V_1$ and $V_2$ are the voltage vectors of ports 1 and 2, respectively, $Y_{2\times 2}$ is the conductance matrix of linear network (here, ports 3 and 4 have to be handled properly), $I_s$ is the effect of port 3 and 4 on port 1 and 2, $I_{no11}$ and $I_{no12}$ are the nonlinear network-current vectors at ports 1 and 2, respectively.

The current and voltage waveforms at the nonlinear ports can be obtained by using the Newton method to solve for F(v).

After the voltage waveform at the nonlinear ports is obtained, the conversion-matrix analysis can be conducted. The RF signal is applied to the PHEMT gate.

The conversion matrices of the drain-mixer circuit can be obtained by using straightforward, but rather tedious, matrix manipulations. If all frequencies except for the IF and RF frequencies are short-circuited, the conversion gain can be expressed as:

$$Gt = [4Re(Y_s(\omega_{RF}))Re(Y_l(\omega_{IF}))|y_{2,1}|^2]/[|(y_{1,1}+Y_s(\omega_{RF}))(y_{2,2}+Y_l(\omega_{IF})) - y_{2,1}y_{2,2}|^2] \qquad (6)$$

where:

$Y_s(\omega_{RF})$ is the RF-source admittance, $Y_l(\omega_{IF})$ is the IF-load admittance, and the y matrix is the simplified version of the conversion matrix.

A Ka-band PHEMT hybrid inte-

*(continued on p. 170)*

# Solving A High-Frequency Signal-Integrity Problem

*A full-wave simulator aids the design of an interconnect circuit carrying high-speed digital signals.*

**Linda Walling and Sergey Polstyanko**
*Ansoft Corp., 4 Station Square, Suite 200, Pittsburgh, PA 15212; (412) 261-3200, FAX: (412) 471-9427, e-mail: lwalling@ansoft.com and spolstyanko@ansoft.com*

**DESIGNERS of high-speed digital circuits traditionally use circuit models that employ lumped RLC circuit elements to account for the interconnect structures that carry the signals between components. These lumped-element models are a good approximation of the physics that prevail in interconnect structures at clock frequencies below 1 GHz. However, for frequencies exceeding 1 GHz, this approximation begins to break down because many interconnect structures are a significant fraction of a wavelength, and the "distributed" nature of the signal propagation cannot be ignored.**

Another problem in modern high-speed design practices is that high-speed signals can induce crosstalk in nearby printed-circuit traces. It is now common to design circuit structures with dedicated grounds and "return paths" located close to the signal traces.

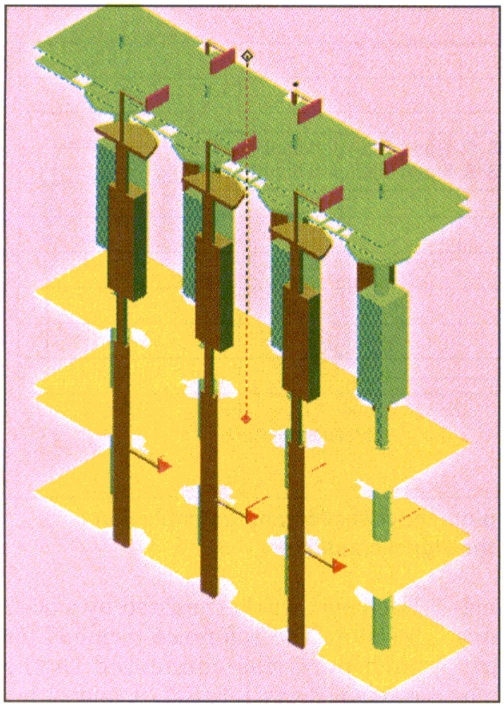
To address these concerns, designers of high-speed digital circuits increasingly rely on simulation techniques that solve the full set of Maxwell's equations. This article describes the analysis of a complex interconnect structure using a full-wave solver called the High Frequency Structure Simulator (HFSS) from Ansoft Corp. (Pittsburgh, PA).

Full-wave solvers have a number of advantages over quasi-static analyzers. The main advantages are listed below:

- At high frequencies, the return path depends strongly upon the excitation-current paths. The more distributed the ground structure is, the stronger the effect will be. With quasi-static analysis, the return paths are dependent upon the placement of the current sinks and sources, and cannot vary in the analysis according to the frequency of the excitation. A full-wave solver calculates all of the coupling terms, including the effect of variable return-current paths.
- By examining field and current

**Results of HFSS noise comparison (normalized to dv/dt = 1 V/ns)**

| | Pitch (mm) | Vnear (mV) | Vfar (mV) |
|---|---|---|---|
| Ansoft HFSS | 2.5 | 4.1 | 1.7 |
| Ansoft Parasitic Parameters | 2.5 | 3.3 | 1.8 |
| Measurement (1983) | 2.5 | 4.5 | 1.8 |



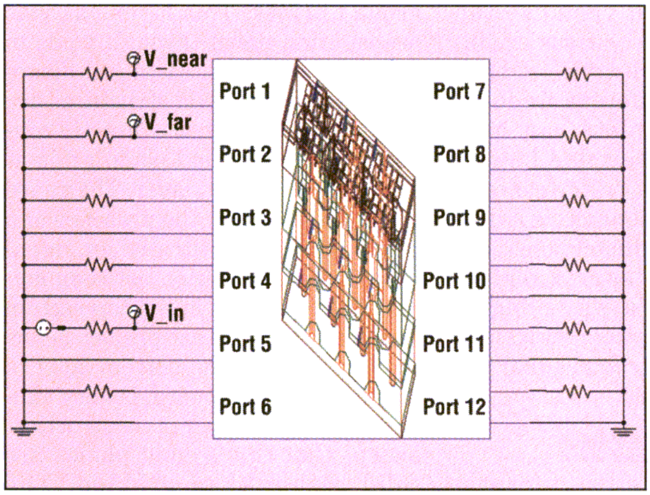
1. This is the simplified connective structure shown with dielectric layers removed.

plots, one can visualize the return path for various excitations (or combinations of excitations) at different frequencies. These plots can help engineers determine how to improve circuit performance.

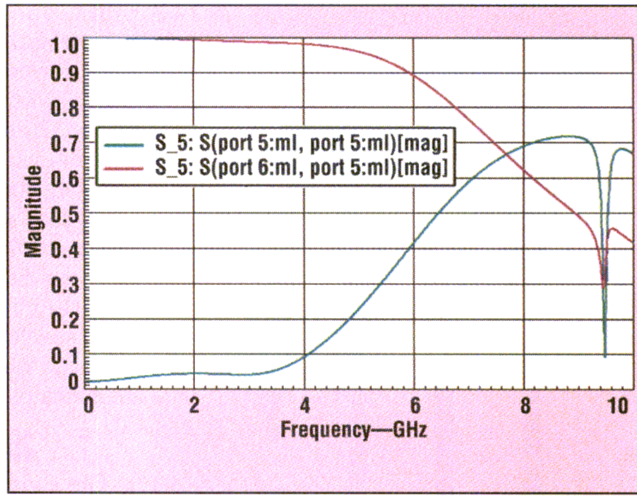
• Quasi-static analysis is only accurate for dimensions that are less than approximately one-tenth of a wavelength, whereas full-wave analysis is accurate for structure that are electrically long.

## ANALYSIS

The complete structure analyzed for this article was originally submitted by IBM (Poughkeepsie, NY). It is an interconnect device that consists of three different sections. The top section is a multilayer multichip module (MCM) with multiple layers of stripline traces separated by gridded ground planes. This layer has vias propagating down through voids in the ground planes. Arranged diagonally between the via structures are "guard" vias, which are directly connected to the ground-plane grids at every ground or DC layer. The guard vias help to ensure the presence of a nearby ground return when signal lines make a transition between layers. All of the vias in the top section terminate in pins that project below the top section. The bottom layer, a multilayer circuit board, is very similar to the MCM layer in construction, except that the ground planes are solid rather than gridded. The ground planes have holes etched in them to allow the signal vias to pass through. The guard vias, interspersed diagonally between the signal vias, are tied to the ground vias at every ground layer. Each via structure (both the signal and ground vias) terminates at the top in connector springs that protrude upward from




**2. This schematic shows the transient measurement setup for Maxwell SPICE.**




**3. This graph plots the reflection and transmission responses of the signal conductor.**

the top of the board. These connector springs are embedded in dielectric material, which constitutes the "carrier" layer. The pins protruding from the bottom of the MCM board mate with the connector springs mounted on the top of the board layers, forming a connection between the MCM and board devices. The spacing between adjacent vias for this structure is 2.5 mm.

The purpose of the simulation is two-fold: to calculate the near- and far-end crosstalk between horizontally-coupled, vertically-coupled, and diagonally-coupled traces, and to compare the measured data with those calculated by several different field solvers.

## STEP 1: HFSS SIMULATION

The structure was modified to speed up the solution process and reduce memory size. Taking advantage of symmetry in the geometries makes it possible to simplify the model without sacrificing accuracy. The top ground-plane grid layers were also simplified (the two upper gridded ground planes were eliminated, and the top input traces were placed at the same level). Figure 1 shows the modified geometry. No significant differences in the simulated S-parameters resulted from these geometry modifications. The results in this article were obtained using this simplified geometry.

The structure was simulated over a frequency range of 0.01 to 10 GHz in HFSS version 8.

## STEP 2: SPICE SIMULATION

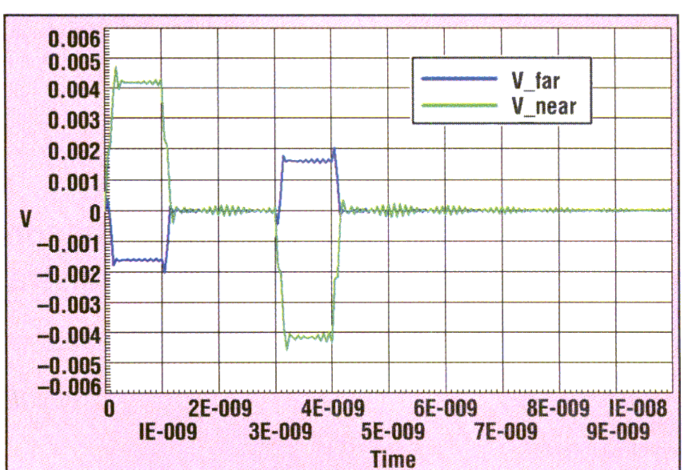
Once the solution process was complete, the data were exported from HFSS to Maxwell SPICE. Next, the device model was inserted into a Maxwell SPICE schematic to perform transient analysis. Figure 2 shows the SPICE circuit created for the transient simulation.

The device simulated was a 12-port device (having six signal conductors). Each port was terminated by a 50-Ω resistor. The reference terminal for each port was connected to the ground. Ports 1, 3, 5, 7, 9, and 11 lay on the top plane, while ports 2, 4, 6, 8, 10, and 12 are on the bottom plane of the device. Port 5 (a middle port) was excited by a pulse with a 1-ns rise and fall time, and a magnitude of +1 VDC. The output signals at ports 1 and 2 (the terminals of a trace adjacent to port 5) were then plotted.

## FULL-WAVE SPICE

Some of the S-parameter results of the HFSS simulations are shown in Fig. 3. It shows the reflection and transmission responses for the excited trace.

The numerical results from Maxwell SPICE are shown in Fig. 4 and agree closely with measurements and results from simulations performed using Ansoft's Parasitic Parameters simulation tool (see table). Parasitic Parameters use the partial-element-equivalent-circuit (PEEC) technique to create an equivalent RLC circuit model. A large number of RLC elements are used in the PEEC circuit in an attempt to capture the distributed nature of the structure. ●●




**4. This graph plots the amount of signal coupling to the adjacent vertical trace (Maxwell SPICE).**

# Design A Ka-Band Integrated Microstrip Isolator

*This approach uses stripline circulator theory to simplify design calculations for a millimeter-wave microstrip isolator.*

**Mengxia Yu, Jun Xu, and Guiping Li**

*Institute of Applied Physics, University of Electronic Science and Technology of China, Chengdu, Sichuan 610054, China.*

**STRIPLINE circulator and isolator theory is sufficiently well-developed to allow engineers to be confident that the design models they use will accurately predict the behavior of the final product. However, the theory behind microstrip circulators and isolators is less reliable due to the complex fringing fields involved. To overcome this, the authors of this article use a simplified version of the stripline model to design a millimeter-wave microstrip isolator.[1,2] The method employs the contour-integrated equations of ferrite-planar circuits described in related literature,[3] and employs electromagnetic (EM) field theory and computer-aided design (CAD) to analyze the parameters of the isolator.**

As part of the preliminary isolator design, the radius of the ferrite circular disk was defined by the low-order resonance condition described in Eq. 1.

$$x_{1,1} = (KR_{1,1}) = 1.84 \qquad (1a)$$

where:

$$K^2 = \omega^2 \varepsilon_r \varepsilon_0 \mu_{eff} \mu_0 = \omega^2 \varepsilon_0 \varepsilon_r \mu_0 (\mu^2 - k^2)/\mu \qquad (1b)$$

$$\mu_{eff} = (\mu^2 - k^2)/\mu \qquad (1c)$$

= the specific effective permeability of the ferrite,

$\mu$ and k = the elements of Polder tensor of the ferrite, and

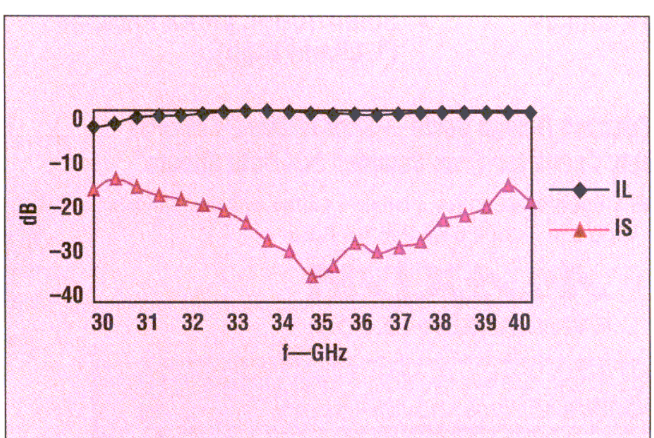
$\epsilon_\gamma$ = the specific permittivity of the ferrite.

Using the contour-integrated method of ferrite planar circuits to analyze the radius of the ferrite circular disk from Eq. 1:

$$f_0 = 35 GHz,\ \varepsilon_d = 13.2,$$
$$R = 0.9\ mm,\ 2\psi = 1\ rad \qquad (1d)$$

To simplify the calculations, the circumference of the ferrite disk is divided into equal 36 sections, and the symmetry of a three-sided, Y-junction circulator is applied.[4] Using Bosma's boundary condition[1] and the EM-field integral Eq. 2.

$$E_z(r_m) = 2\oint_C \left[ \begin{array}{l} j\omega\mu_{eff} G(KR) H_l(r_0) + \\ (j\frac{k}{\mu}\sin\theta_0 - \cos\theta_0)\frac{\partial G(KR)}{\partial R} E_z(r_0) \end{array} \right] dt \qquad (2a)$$




**1. This graph shows the calculated insertion loss and isolation for the microstrip isolator.**

yields the S-matrix of the circular-junction disk and the insertion-loss and isolation parameters of the isolator (Fig. 1).

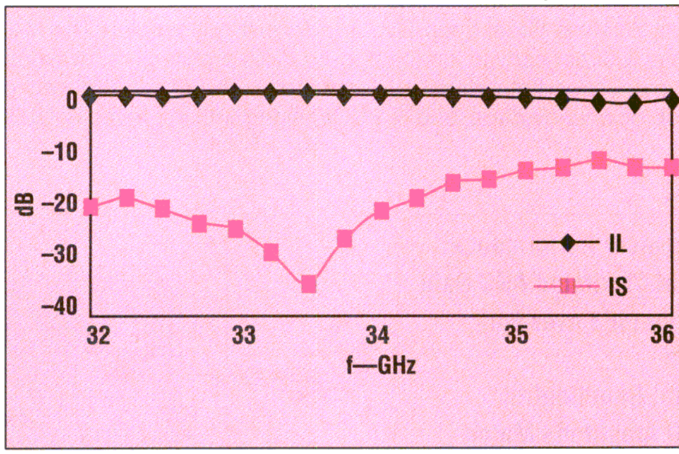
## DESIGN EXAMPLES

According to the method of analysis previously discussed, the Ka-band microstrip isolator is fabricated in duroid-5880 substrate. Nickel-zinc (NiZn) ferrite was used to construct the isolator. The characteristics of NiZn ferrite are listed as follows:

$$4\pi M_s = 5200\,G_s,\ \varepsilon_d = 13.2,$$
$$\Delta H = 80\,O_s,\ \tan\delta_\varepsilon = 1.2\times10^{-4},$$
$$\theta_f = 365^{\circ}C \qquad (2b)$$

First, the ferrite element was metalized by sputtering techniques. A hole smaller than the radius of ferrite disk was punched in the duroid substrate so that the disk would fit tightly in the hole. The disk was then electrically connected to the top conductor and the bottom ground plane. A small, high-field-intensity biasing magnet was installed under the circulator to make it non-reciprocal. In cases where the dielectric constant of the substrate is different from that of the ferrite, a matching transformer is required on the three ports. The performance of the isolator, shown in Fig. 2, was measured on a GT101/GT201-2013 scalar network-analyzer system.

The measurements indicate that the device had a center frequency is 33.2 GHz. The bandwidth over which isolation loss is greater than 20 db and insertion loss of less than 1 dB is 1.3 GHz. ●●




2. The graph shows the measured insertion loss and isolation of the experiment microstrip isolator.

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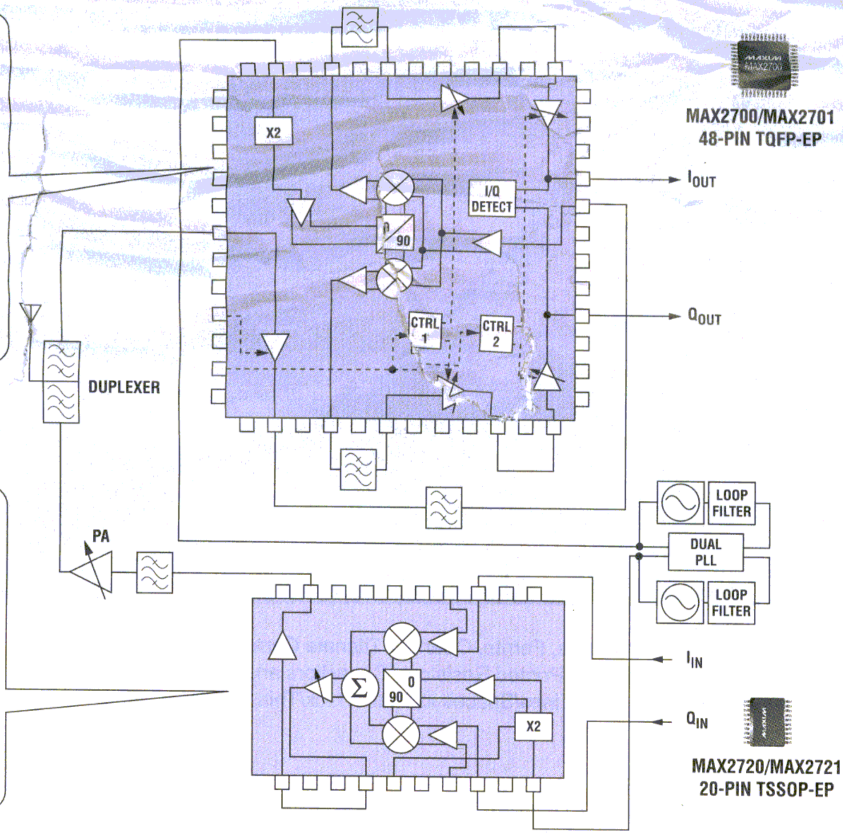



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# Understanding WAP: Wireless Applications, Devices, and Services

**Marcel Van Der Heijden and Marcus Taylor, Eds.**

Internet mania has taken over most businesses and households that are properly equipped for access to this giant network. For now, this obsession with the World Wide Web has been largely confined to stationary access points, such as desktop and laptop computers. But what happens when the Internet is freely available through handheld wireless devices, such as cellular telephones and personal digital assistants (PDAs)? The Wireless Application Protocol (WAP) was developed for that purpose—to bring the Internet to portable devices. *Understanding WAP: Wireless Applications, Devices, and Services* tells the story of the WAP and what kind of user interfaces and services can be expected from future WAP-enabled devices. The WAP is designed to provide seamless mobile Internet access.

WAP operates through a microbrowser, which can be compared to a standard Internet browser, although optimized for use with mobile devices. The microbrowser accesses applications written in a specialized markup language known as the wireless markup language (WML). The WML is, in turn, supported by the scripting language, WMLScript.

The opening chapter of *Understanding WAP: Wireless Applications, Devices, and Services* discusses some of the possible applications using WAP, along with the benefits of the protocol and business opportunities awaiting WAP implementers. Chapter 2 details the wireless application environment for creating WAP services and applications, with a close look at WML and WMLScript. Chapter 3 covers the design of effective user interfaces for WAP services, while Chapter 4 explores the possibility of telephony services in WAP. Chapter 5 explains how to integrate WAP gateways, which act as intermediaries to connect a WAP-enabled device to the Internet world as well as existing and emerging wireless networks. Chapter 6 provides an introduction to WAP "push" services, where content providers can direct their messages and information to users of WAP-enabled devices. Chapter 7 explores security issues for WAP users, while Chapter 8 investigates what WAP means for network operators. Chapter 9 highlights extended unified-messaging solutions using WAP, while Chapter 10 covers mobile financial services and applications that will be available for users of WAP-enabled devices. The well-written text is informative without being highly technical. It includes a list of acronyms useful in understanding WAP-related terminology, as well as a copy of the complete text on a CD-ROM. (2000, 256 pp., hardcover, ISBN: 1-58053-093-1, $79.00.) **Artech House, Inc., 685 Canton St., Norwood, MA 02062; (800) 225-9977, (781) 769-9750, FAX: (781) 769-6334, e-mail: artech@artech-house.com, Internet: http://www.artech-house.com.**

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# Biphase Modulators Serve Wireless And Radar Systems

*This power-efficient design uses PHEMT variable-resistance reflection terminations for precise control of phase and amplitude.*

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**I.D. Robertson**

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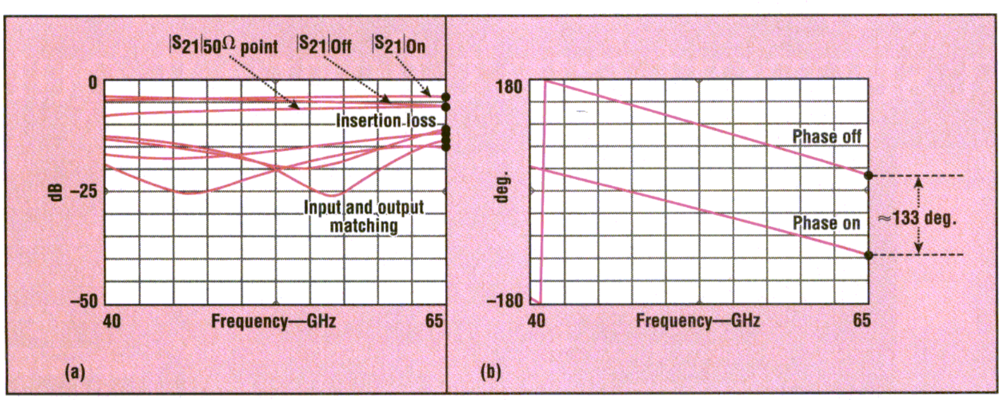
**MILLIMETER-WAVE transmitters (Txs) offer generous bandwidth and high precision in digital communications systems and radar systems. Direct modulation of the carrier signal provides an ideal solution to this challenge, since it can minimize the overall complexity, size, and cost of the millimeter-wave subsystem. In this article, the design and performance of a V-band balanced-biphase-amplitude modulator, employing low-power pseudomorphic-high-electron-mobility-transistor (PHEMT) variable-resistance reflection terminations are presented. The monolithic-microwave-integrated-circuit (MMIC) chip, which measures 1.1 × 1.2 mm, can be used to implement power-efficient binary-phase-shift-keying (BPSK), amplitude-shift-keying (ASK), and biphase amplitude-modulation schemes with very-high precision directly at a carrier frequency of 60 GHz. It has been found that by employing the proposed balanced topology, the substantial amplitude and phase imbalances of the standard single-stage modulator (caused by the parasitics of the switching devices) can be removed, resulting in near-perfect responses even at high millimeter-wave frequencies. Moreover, the state-of-the-art responses obtained with the balanced configuration can be sustained over a wide bandwidth. As a result, the broadband-monolithic 60-GHz balanced modulator can be an important component for a wide range of emerging wireless millimeter-wave applications.**

The past decade witnessed considerable growth in the use of the millimeter-wave frequency range for new civil applications. Wireless systems using millimeter-wave frequencies benefit from a number of advantages, such as high levels of security, more flexibility, and wideband operation, and are thus highly attractive for a broad range of communications applications. Examples of applications at V-band (used in this article) include high data-rate wireless local-area networks (WLANs), line-of-sight (LOS) links, vehicle-to-vehicle and vehicle-to-infrastructure communications, and radar-sensor systems for

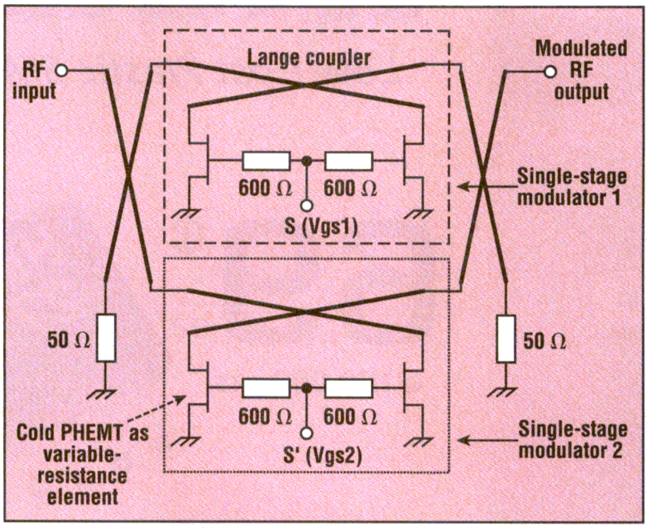



**1. These plots show the simulated performance of the single-stage modulator for (a) insertion loss and matching and (b) phase-shift response.**


AWR

adaptive-cruise-control (ACC) and collide-protection systems, while the 60-GHz band is also of immense significance for micro/pico cellular-mobile communications systems.

The major drawback of millimeter-wave technology is the high cost of the components. This can be overcome with MMIC technology. MMIC technology is becoming increasingly influential as a means of reducing the cost and improving the performance of many advanced microwave and millimeter-wave systems. This technology offers a major step forward for future commercial markets. The requirement for systems operating in higher millimeter-wave bands is expected to become more widespread in the near future, resulting in the need to investigate the design and applications of various V- and W-band components and circuits. This article describes the design and performance of a 60-GHz MMIC balanced biphase amplitude modulator for use in V-band wireless data-communications systems.

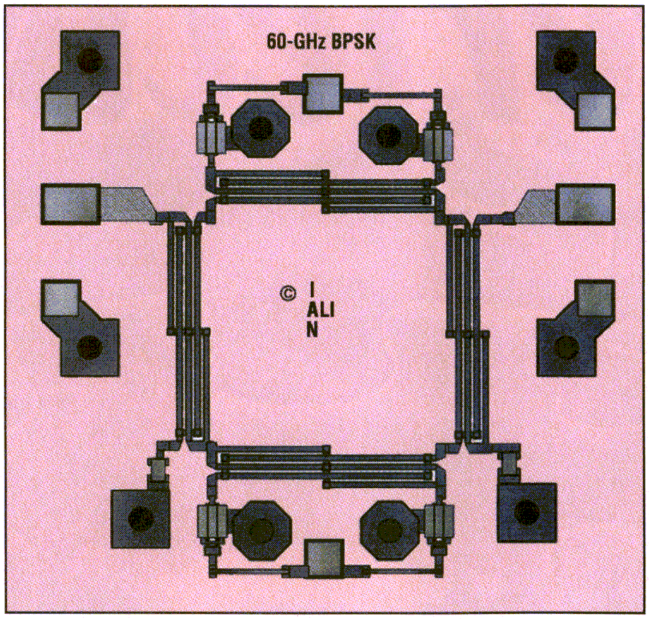



**2. The direct 60-GHz reflection-type balanced-biphase amplitude modulator employs this circuit topology.**

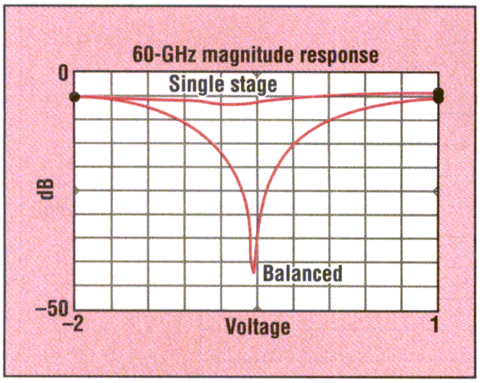
## SINGLE-STAGE MODULATOR

Biphase (or BPSK) modulation is frequently used for power-efficient applications since it is robust and spectral efficiency is not of primary importance. Direct-carrier modulation[1-5, 8, 9] has been shown to be an attractive means of reducing the overall hardware complexity, assembly, time, size, and cost of millimeter-wave subsystems. This is attained by modulating directly at the carrier frequency, thus removing the need for an upconversion process from intermediate frequency (IF) to RF, which is used in the more conventional systems, and would otherwise consume a substantial MMIC chip area. This approach is therefore highly desirable for millimeter-wave

applications, where size and cost remain a key factor restricting the widespread use of wireless communications systems. A common technique for implementing BPSK modulation directly at the carrier frequency is to employ a reflection topology, with cold field-effect transistors (FETs)/PHEMT devices acting as variable-resistance elements,[6-8] connected to the coupled and direct ports of a 3-dB 90-deg. Lange coupler. With analog control,[8] it is possible to achieve a continuous range of resistance values, which, in turn, results in a continuous reflection-coefficient magnitude range. When the switches are on, the resistance would be zero and the ideal voltage-reflection coefficient would be −1 (i.e., a short circuit). When the switches are off, the resistance would be infinite and the ideal voltage-reflection coefficient would be +1 (i.e., an open circuit), with a 180-deg. phase difference between the on and off states. At the 50-Ω bias point, the signals are fully absorbed with no reflections, and the ideal reflection coefficient would be zero. The reflected signals cancel at the input port of the coupler and add at the output port only. It has been shown,[8] however, that the parasitic-circuit elements of the switching devices start to dominate at higher frequencies—the series inductance in the on state and the shunt capacitance in the off state. Moreover, the unwanted parasitics elements of the switching elements become more substantial as the frequency of operation increases. The overall performance of the single-stage topology is severely degraded in the V-band frequency range, as




**3. The fabricated 60-GHz MMIC balanced-biphase amplitude modulator has a chip size of 1.1 × 1.2 mm.**

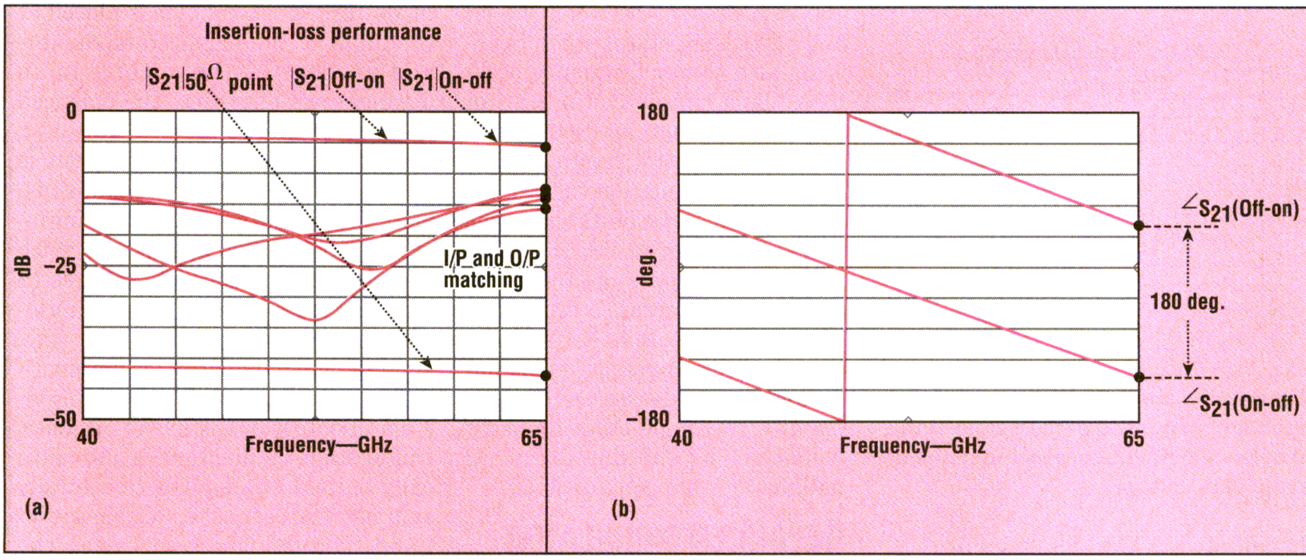



**4. These plots show the simulated 60-GHz magnitude responses of the single-stage and balanced modulators as functions of varying bias voltage.**

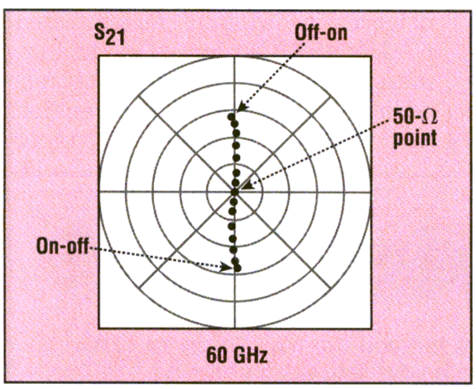
substantiated by the simulated magnitude and phase responses illustrated in Figs. 1a and b, respectively. It can be seen that the single-stage modulator suffers from considerable amplitude- and phase-error characteristics, with an amplitude imbalance of 1.3 dB and a substantial phase deviation of 47 deg. from the perfect 180 deg. present between the on and off biasing states when operating at a center frequency of 60 GHz. This simple configuration also suffers from an insufficient magnitude-control range, with a maximum attenuation value of 6.5 dB attained at the 50-Ω bias point when operating at 60 GHz. The poor performances observed are also sustained across a broad frequency range. It is difficult to tune out these parasitic elements in a wideband design and, subsequently, the single-stage topology would not be suitable for many millimeter-wave applications.

## BALANCED MODULATOR

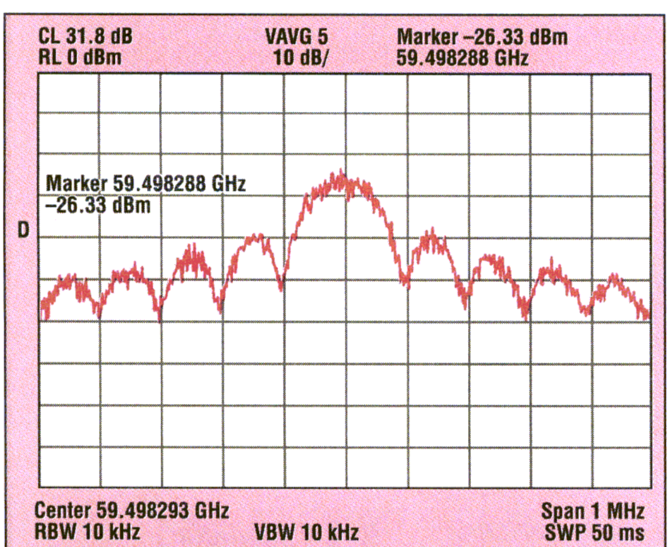
The large amplitude and phase errors of the standard modulator, caused by the parasitic elements of the PHEMT devices at V-band, can be overcome by employing a balanced configuration.[8,9] Here, two analog reflection-type single-stage modulators are operated in push-pull mode (Fig. 2). A coupler is placed at the input of the balanced modulator and another at the output, in an arrangement similar to the balanced amplifier, to realize the 180-deg. operation. The combined output is the vector sum of the two transmission coefficients.[8] The two single-stage modulators are driven with complementary control voltages (S and S|). As a result, when single-stage modulator 1 is in the on state, single-stage modulator 2 is automatically turned off, and when single-stage modulator 1 is in the off state, single-stage modulator 2 is automatically turned on, leading to a totally symmetrical response, even in the V-band frequency range. Since the Lange couplers are nearly ideal, the balanced configuration can provide near-perfect amplitude and phase balance. The layout of the fabricated 60-GHz MMIC-balanced modulator, measuring 1.1 × 1.2 mm, is illustrated in Fig. 3. Low power consumption has been addressed by using




**5. These plots show the performance of the balanced modulator from 40 to 65 GHz for (a) insertion loss and matching and (b) phase-shift response.**




**6. This is an $S_{21}$ polar representation of the monolithic-balanced modulator at a center frequency of 60 GHz.**




**7. This BPSK spectral response was measured at 59.5 GHz for a data rate of 100 kb/s.**

PHEMT devices with low DC-voltage requirements, which is a major advantage for battery-operated and/or handheld applications. The 0.25 × 2.0 × 60-µm cold-PHEMT reflection terminations of the monolithic chip are biased with complementary-control analog baseband signals through high-value resistors, which are employed to prevent RF leakage and act as forward-bias current limiters. The essential parameters of the couplers, the finger length, finger width, and spacing between the fingers have been optimized for operation at V-band to provide 3-dB power coupling and 90-deg. phase difference between the direct and coupled ports. The four Lange couplers, which have a length of approximately 375 µm corresponding to a center frequency of 60 GHz, can be replaced by miniaturized microstrip couplers [9] to save expensive chip space and to gain a compact layout. This approach would subsequently result in a reduced monolithic chip size and cost, which is vital for emerging commercial millimeter-wave applications.

## MEASURED RESULTS

The simulated reflection-coefficient magnitude response of the monolithic-balanced-biphase amplitude modulator, as a function of the varying bias voltage, is illustrated in Fig. 4 at a fixed-center frequency of 60 GHz. The simulated voltage-wave magnitude response of the standard single-stage topology is also illustrated for comparison. It can be seen that the amount of amplitude variation and tuning-curve symmetry around the 50-Ω point are greatly improved with the balanced topology, with virtually zero amplitude imbalance between the on and off biasing states. A maximum attenuation value of 42 dB is also obtained at the 50-Ω bias point with the balanced configuration, which is a major improvement compared to the maximum attenuation value of 6.5 dB obtained with the single-stage modulator. The insertion loss and input/output (I/O)-matching response of the balanced biphase-amplitude modulator from 40 to 65 GHz are illustrated in Fig. 5a, while the phase-shift performance of the circuit from 40 to 65 GHz is illustrated in Fig. 5b. It can be seen that

good impedance matching, with minimal amplitude or phase errors, can be achieved over a wide bandwidth by employing this topology. An input and output return-loss performance of better than 15 dB is obtained at 60 GHz, with a worst-case value of 12 dB across the entire 40-to-65-GHz frequency range. The overall balanced circuit also achieves a mean insertion-loss level of approximately 5 dB in the V-band frequency range. Moreover, with reference to Fig. 5, it can be seen that the amplitude and phase imbalances between 'on-off' and 'off-on' are reduced to less than ±0.1 dB and ± 0.12 deg., respectively, when operating at a center frequency of 60 GHz, while a maximum voltage-wave attenuation value of 42 dB is also attained at the 50-Ω bias point, as previously described.

The state-of-the-art responses achieved with the balanced configuration demonstrate a significant improvement compared to the single-stage responses obtained in the V-band, which is critical for the implementation of accurate biphase-amplitude (and BPSK) constellations at such high frequencies. This is substantiated by the $S_{21}$ polar response of the overall balanced circuit shown in Fig. 6, generated directly at 60 GHz for 15 bias points between 'on-off' and 'off-on', illustrating a near-perfect constellation with virtually zero amplitude or phase errors. The state-of-the-art responses of the balanced-design approach are also maintained over a wide bandwidth, endorsed by a mean amplitude imbalance of only ± 0.1 dB and a phase error of less than ± 2.5 deg., together with a maximum attenuation value of better than 40 dB within the entire 45-to-65-GHz frequency range. This demonstrates that the balanced modulator can provide high precision with a near-perfect amplitude and phase balance over a wide bandwidth. According to detailed simulations, the circuit is usable to more than 65 GHz. The measured output BPSK-spectral response at 59.5 GHz is illustrated in Fig. 7. The data rate was 100 kb/s and no data prefiltering was employed. The spectrum exhibits a near-ideal profile, with deep nulls and negligible carrier breakthrough.

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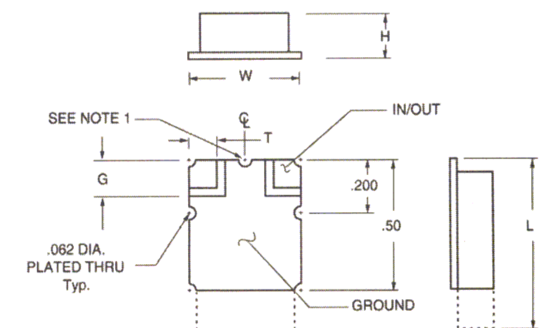


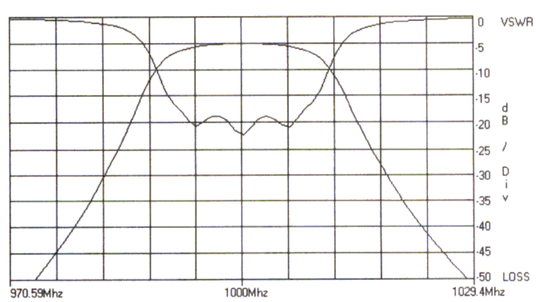


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The performances of monolithic 60-GHz single-stage and balanced biphase amplitude modulators for power-efficient applications at V-band have been presented. It has been shown that the balanced configuration can be used to remove the substantial amplitude and phase imbalances of the standard single-stage modulators, caused by the parasitic elements of the cold-PHEMT switching devices when operating in the 60-GHz frequency range. The magnitude-control range and tuning-curve symmetry are also improved with this topology, which is vital for the implementation of accurate biphase-amplitude/BPSK constellations at V-band. It has been shown that this design approach can provide near-perfect, state-of-the-art performances, with virtually zero amplitude or phase-error characteristics, over a wide frequency range, with minimum hardware complexity or design effort.

This simple low-power technique has been found to be very robust, and the resulting MMIC-balanced modulator is highly suitable for the realization of broadband, affordable, high-performance millimeter-wave Txs for a wide range of power-efficient wireless applications, including portable data-communications systems, WLANs, collision-avoidance systems, and LOS-communications links. ●●

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# Measuring Power Levels In Modern Communications Systems

***A choice of video bandwidths and time-gating capabilities can increase the accuracy and effectiveness of power measurements on modern wireless-communications systems.***

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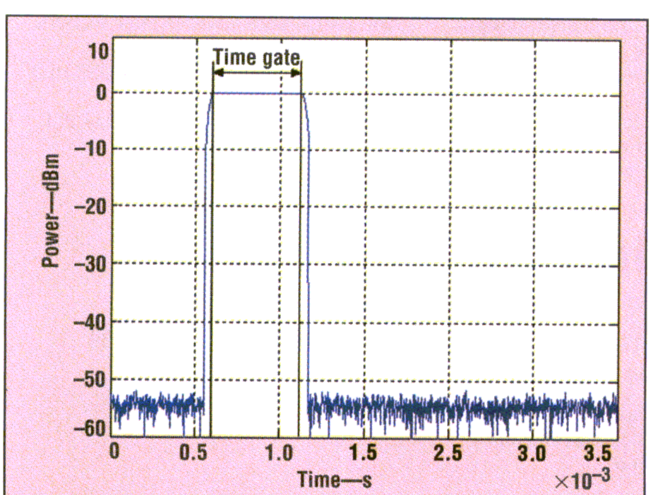
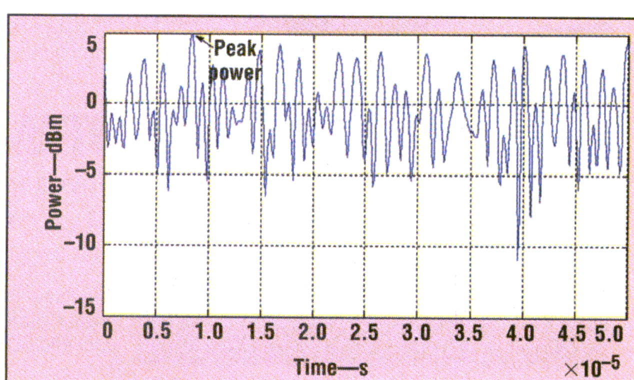
**POWER-MEASUREMENT needs can vary greatly among different communications systems. Distinctions can be made, for example, between the power-measurement requirements of the two prevalent second-generation (2G) wireless formats, time-division-multiple-access (TDMA) and code-division-multiple-access (CDMA) systems. But as the industry moves toward third-generation (3G) wireless-communications systems, and TDMA and CDMA systems merge, the power-measurement needs of these new systems will also merge. The capability to make the required measurements is a key factor in choosing a power-measurement system, but this must also be supported by the capability to make measurements in a fast, repeatable, and accurate manner. This article will explore the types of power measurements required by these emerging systems and highlight the contributions to measurement accuracy that can be made by a properly designed test system.**

Signals in TDMA-based systems involve bursts of modulated RF signals within an allotted time slot. Systems using the TDMA format include the Global System for Mobile Communications (GSM), the North American Digital Cellular (NADC) system, and the Bluetooth personal-connectivity system. The main requirement for power measurements of this type of signal is the capability to measure the average power within the burst, or in gated sections within the burst (Fig. 1).

Traditionally, power measurements of TDMA signals were achieved by performing an average power measurement of the periodic signal and then calculating the power in the whole burst using the duty




1. Burst average power is illustrated here.



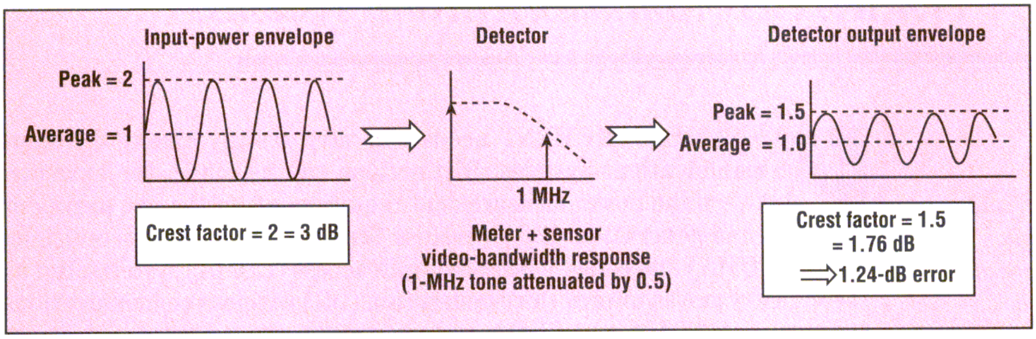

2. This figure shows CDMA power envelope.

cycle. However, this pulse-power measurement is reliant on the assumption of a perfectly pulsed RF signal. It is not very accurate, because it does not take into account the slope of the rise and fall time and variations in the pulse period. Secondary power-measurement requirements for TDMA may include measurements of the maximum RF power in a burst, or the power at different sections of the burst. Neither of these measurements can be achieved using the traditional method.

For CDMA applications, the main requirement is to measure the average power of a modulated signal. Further signal characterizing is achieved using either a peak-to-average ratio or peak-power measurements (Fig. 2). CDMA signals are intended to appear as random, "noise-like" signals, and a succinct measurement such as the crest factor can be used to give an immediate indication of whether a system is performing to specification.

Time-gated average power, peak power, and peak-to-average-ratio measurements of TDMA and CDMA signals are the core test requirements of a power-measurement system for today's wireless standards. As these wireless standards evolve, the capability to make peak, peak-to-average ratio, and average measurements on a single burst is required. This supports the fuller characterization of signals such as the time-division-duplex (TDD) signals used in wideband CDMA (WCDMA) and in EDGE systems, which have amplitude-varying modulation formats within an RF burst.

Underpinning the requirement for additional measurement capability is the need to make the previously mentioned measurements in a fast, accurate, and repeatable manner. This depends on a combination of traditional power-meter




3. The effect of insufficient video bandwidth can be seen here.

**Digital control of video bandwidths**

| | DSP video-bandwidth setting | | |
|---|---|---|---|
| **Sensor** | **Low** | **Medium** | **High** |
| E9321A/E9325A | 30 kHz | 100 kHz | 300 kHz |
| E9322A/E9326A | 100 kHz | 300 kHz | 1.5 MHz |
| E9323A/E9327A | 300 kHz | 1.5 MHz | 5 MHz |

and sensor specifications such as VSWR/mismatch, linearity, and instrumentation accuracy. The impact these make on measurement accuracy and repeatability is covered in Ref. 1. In addition, new specifications relating to peak and burst measurements become necessary. Specifications that were not previously apparent from data sheets of peak and average power meters, can still have a bearing on the accuracy of the expected results.

The following is an example of one error that arises with peak measurements. For the case of a signal comprising two RF tones separated by 1 MHz and of equal amplitude, the power envelope is a 1-MHz sine wave superimposed on the average power. The average power is equal to the sum of the power in each tone, and the peak power is twice this average, so the sine wave varies between zero and twice the average. The signal has a 3-dB peak-to-average ratio. When a diode-power sensor measures this signal, a DC component corresponding to the average power and a component at 1 MHz corresponding to the sine-wave variation in power envelope is generated. Any video bandwidth roll-off will directly effect the peak measurement and leave the average measurement unaffected (Fig. 3). If the video bandwidth of the power-measurement equipment is quoted as having a 3-dB bandwidth of 1 MHz, the sine-wave tone is attenuated to half the amplitude it should be. So, although this might be sufficient to measure average power, a 1.24-dB error will be present in peak-to-average ratio measurement. (Note that video bandwidth represents the ability of the power sensor and meter to follow the power envelope of the input signal. The power envelope of the input signal is, in some cases, determined by the signal's modulation bandwidth and, hence, video bandwidth is sometimes referred to as modulation bandwidth.)

Errors in peak-power measurements can be significant. Visibility of these errors needs to be apparent in the power-measurement system's specifications. A general statement of a megahertz value is insufficient for specifying the video bandwidth without defining its tolerance or variation.

The effect of measurement noise can also impact the accuracy of a peak measurement. As a peak-power measurement is by definition a single occurrence, the measurement cannot benefit from additional averaging to reduce noise. For average-power measurements, averaging repeated measurements can reduce the variation of a measurement due to noise and allow

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| Type | Case | Ratings | | | | | Characteristics, typical | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | $V_{CEO}$ (V) | $I_C$ (mA) | $P_{TOT}$ (mW) | $f_T$ (GHz) | $I_T$ (mA) | F (dB) | $G_{um}$ (dB) | @ (MHz) | F (dB) | $G_{um}$ (dB) | @ (MHz) |
| PMBTH10 | SOT23 | 25 | 40 | 400 | 0.6 | 1-20 | | | | | | |
| PMBTH81 | SOT23 | 20 | 40 | 400 | 0.6 | 1-20 | | | | | | |
| BFS17W | SOT323 | 15 | 50 | 300 | 1.6 | 2-20 | 4.5 | | 500 | | | |
| BFR92AT | SC-75* | 15 | 25 | 300 | 5 | 3-30 | 2 | 14 | 1000 | 3 | 8 | 2000 |
| BFT92W | SOT323 | 15 | 35 | 300 | 4 | 3-30 | 2.5 | 17 | 500 | 3 | 11 | 1000 |
| BFR93AT | SC-75* | 12 | 35 | 300 | 5 | 5-40 | 1.5 | 13 | 1000 | 2.1 | 8 | 2000 |
| BFQ67T | SC-75* | 10 | 50 | 300 | 8 | 3-30 | 1.3 | 13 | 1000 | 2.2 | 8 | 2000 |
| PBR941 | SOT23 | 10 | 50 | 360 | 8 | 3-30 | 1.4 | 15 | 1000 | 2 | 9.5 | 2000 |
| PRF947 | SOT323 | 10 | 50 | 250 | 8 | 3-30 | 1.5 | 16 | 1000 | 2.1 | 10 | 2000 |
| PRF949 | SC-75* | 10 | 50 | 150 | 8 | 3-30 | 1.5 | 16 | 1000 | 2.1 | 10 | 2000 |
| PRF957 | SOT323 | 10 | 100 | 270 | 8 | 5-50 | 1.3 | 15 | 1000 | 1.8 | 9.2 | 2000 |
| BFR505T | SC-75* | 15 | 18 | 150 | 9 | 1-10 | 1.2 | 17 | 900 | 1.9 | 10 | 2000 |
| BFR620T | SC-75* | 15 | 70 | 300 | 9 | 3-30 | 1.1 | 15 | 900 | 1.9 | 9 | 2000 |
| BFC520 | SOT353 | 8 | 70 | 1000 | 9 | 3-30 | 1.3 | 31 | 900 | 1.5 | 19 | 2000 |
| BFE520 | SOT353 | 8 | 70 | 100 | 9 | 3-30 | 1.2 | 17 | 900 | 1.9 | 10 | 2000 |
| BFM520 | SOT363 | 8 | 70 | 100 | 9 | 3-30 | 1.1 | 15 | 900 | 1.9 | 9 | 2000 |
| BFG520W/X | SOT343 | 15 | 70 | 500 | 9 | 3-30 | 1.6 | 17 | 900 | 1.8 | 11 | 2000 |
| BFG540W/X | SOT343 | 15 | 120 | 500 | 9 | 10-60 | 1.9 | 16 | 900 | 2.1 | 10 | 2000 |
| BFG11W/X | SOT343 | 8 | 500 | 760 | 9 | 50-150 | | | | | 7 | 1900 |
| BFG403W | SOT343R | 4.5 | 3.6 | 16 | 17 | 5-5 | 1 | 20 | 900 | 1.6 | 22 | 2000 |
| BFG410W | SOT343R | 4.5 | 12 | 54 | 22 | 2-15 | .9 | | 900 | 1.2 | 22 | 2000 |
| BFG425W | SOT343R | 4.5 | 30 | 135 | 22 | 3-30 | .8 | | 900 | 1.2 | 20 | 2000 |
| BFG480W | SOT343R | 4.5 | 250 | 360 | 18 | 30-150 | 1.2 | | 900 | 1.8 | 16 | 2000 |
| BFG21W | SOT343R | 4.5 | 200 | 600 | 18 | 50-250 | | | | | 12 | 1900 |

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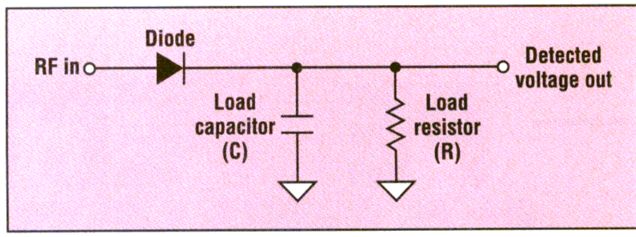
lower powers to be measured. So it needs to be recognized that the specified minimum measurement power for a peak measurement is typically 15 dB above that achieved for time-gated average measurements.

To make time-gated power measurements on TDMA-type pulses, the measurement system must have sufficient rise and fall times to ensure that the gated section being measured is actually the signal and not a latent effect of a slow rising or falling edge caused by insufficient measurement-response times. If overshoot is to be characterized, the power-sensor rise-and-fall time specifications must be fast enough to follow the rising and falling edges of the signal under test. It is generally recommended that the power sensor have a rise time of approximately one-eighth of the expected signal's rise time to obtain accurate results. This will minimize the error introduced by the sensor and meter responses. However, if only the burst average power is being measured, a rise time similar to that of the signal can be used. Although the rising edge of the burst will delay the quickness of the measured pulse rise time, the start of the time-gated measurement can be delayed to account for this. In this manner, accurate burst-average-power measurements can be made.

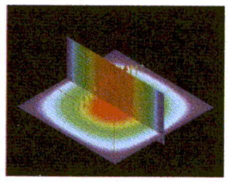



**4. This is a basic RF-power detection circuit.**

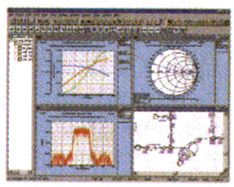
Confusion sometimes arises from the video-bandwidth definition of a sensor and its capability to perform accurate pulse profiling. If a pulsed signal has a pulse-repetition rate of X Hz, then it will also have significant signal components at odd-harmonics of the X Hz (3X, 5X, etc.). Therefore, if a sensor only had a video bandwidth of X MHz, these additional harmonics would not be accurately measured, resulting in an inaccurate pulse profile. So when accurate pulse profiling is required, it is best to follow the one-eighth rise-time rule.

The causes of peak-measurement errors are inherent in the diode-detection method of power measurement, yet by careful design they can be reduced to acceptable levels. Thermistors and thermocouplers are accurate for average power measurements, but for accurately following a fast-changing power envelope, diode detection is the most suitable approach.

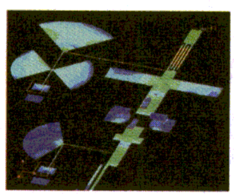
To introduce the causes of these errors, a brief overview of the diode-detection process is outlined, allowing the cause of these errors to then be



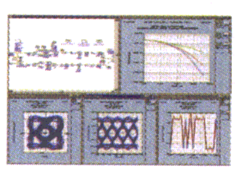




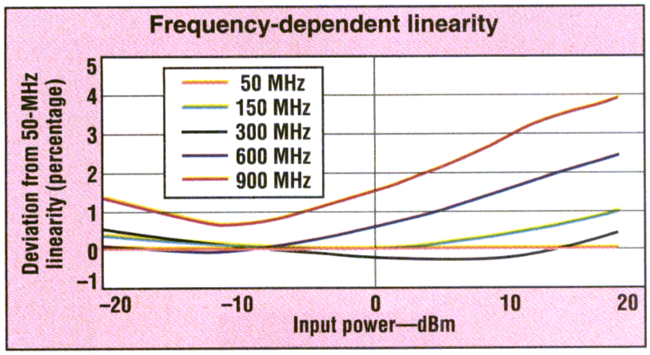








highlighted. At RF, the diode conducts for approximately half of a cycle, shorting the RF to ground through the load-capacitor C (Fig. 4). As the signal passes through the non-linear diode, the RF signal is mixed down to baseband and generates a voltage across the load-resistor R. The voltage generated is proportional to the input power as the diode has a $V^2$ (square law) response ($P = V^2/R$) at low power levels (below −20 dBm). For a signal with a fast-changing power envelope, the baseband signal changes at the same rate as the signal power. To have the capability to make peak measurements, sufficient video bandwidth is required to allow this change to be followed through.




**5. The frequency-dependent linearity of a wide video-bandwith sensor at spot frequencies from 50 to 900 MHz is depicted here.**

The filtering effects of R, C, and video-resistance of the diode in the first instance dictate the video bandwidth. For the types of diode used in power detection, the value of the video-resistance can vary from 200 Ω to 4 kΩ over power and temperature, leaving the video bandwidth susceptible to power and temperature changes and, therefore, introducing errors.

The two components in the measuring system, R and C, have conflicting demands placed upon them as they have a large impact on the performance of the power sensor. For example, for maximum sensitivity and dynamic range, R should be large, and for maximum RF-frequency range, C should also be large. However, for maximum flatness and stability of the video bandwidth, C and R should both be small. In addition there are optimum values of R and C for linearity, temperature stability and mismatch.

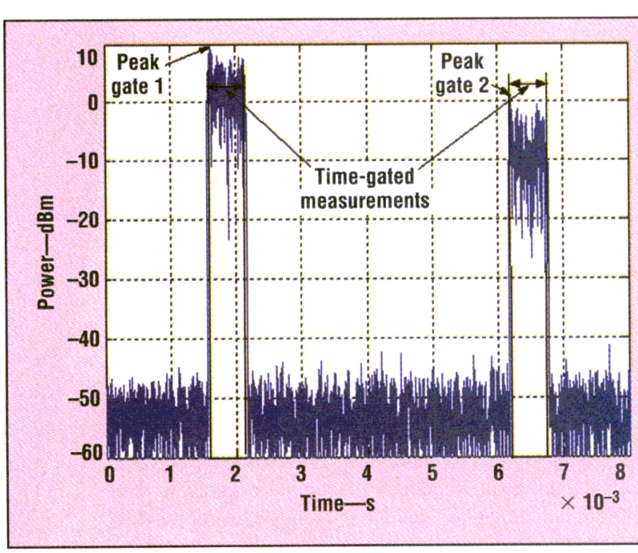
So while it is possible to practically eliminate any significant error due to insufficient video bandwidth, this would have the effect of severely impacting other key specifications such as linearity, mismatch, dynamic range, and temperature stability. Conversely, optimizing for the other key power-sensor specifications and ignoring the error introduced in peak-power measurements is also undesirable.

Agilent Technologies (Palo Alto, CA) has taken the approach to introduce a range of sensors (the E9320 family of peak- and average-power sensors) with three video bandwidths of 300 kHz, 1.5 MHz, and 5 MHz. Each sensor type has been specifically optimized for maximum dynamic range and minimum errors for specific wireless standards:

- 300-kHz bandwidth for GSM and EDGE.
- 1.5-MHz bandwidth for IS-95 CDMA.
- 5.0-MHz bandwidth for WCDMA.

One instance where the conflicting demands on load-circuit components becomes apparent is when one considers the trade-off between the low-frequency RF range and wide video bandwidth. In order to remain sensitive at low frequencies (<100 MHz) high load resistance and capacitance are optimal. However, wide video bandwidth requires low values for these components. This leads to a compromise between variations in video bandwidth and the frequency dependence of sensor linearity. If the video bandwidth is not sufficiently flat, errors will be introduced in peak-power measurements. The wider the video bandwidth, the greater the variation in linearity. In traditional wide-dynamic range sensors with minimal video bandwidth, only a 50-MHz linearity calibration is required. Figure 5 shows how only one frequency-linearity correction at 50 MHz will lead to linearity errors for sensors with a wide video bandwidth.

To achieve a wide, flat video bandwidth without compromising linearity, a frequency-dependent linearity correction (FDLC) has been implemented in the E9320 sensors. These sensors are factory-calibrated at 50 MHz and at key frequency points up to 900 MHz in order to provide the correction data for the FDLC. This addresses the trade-off in linearity and video bandwidth, but it still leaves the bandwidth susceptible to power and temperature variations in the detection diode's resistance. To stabilize this, a fifth-order filter has been designed to replace the simple resistive-capacitive (RC) detection circuitry. This filter improves the accuracy of peak measurements and has the additional benefit of increasing the roll-off of the video bandwidth, helping to further reduce the frequency dependence of linearity.




**6. Gated-measurement capability makes it possible to focus on peak power levels.**

The E9320 peak and average sensors operate with the new Agilent EPM-P-series power meters (E4416A single-channel and E4417A dual-channel). The EPM-P meters are also compatible with the entire range of 8480 and E-series sensors. Based on industry-standard E4418B/19B

# Understanding Thermal Basics For Microwave Power Devices

***Some straightforward calculations can be made to predict the flange and channel temperatures of transistors at different input and output-power levels.***

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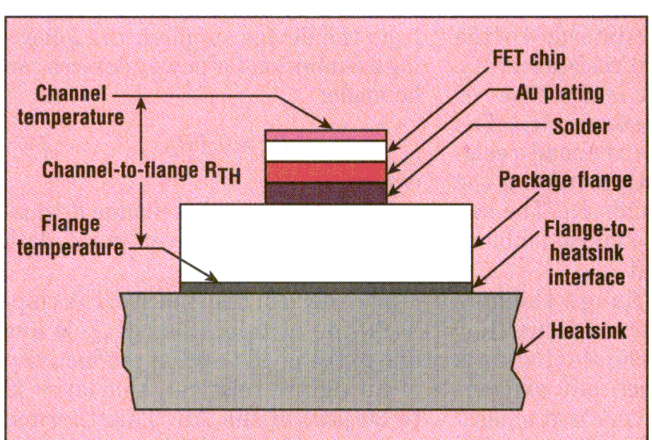
**HIGH temperatures usually mean short lifetimes for electronic devices, such as gallium-arsenide (GaAs) field-effect transistors (FETs). For that reason, it is useful to know the operating temperatures that can be expected for a particular GaAs-FET device under various operating conditions. Since the majority of GaAs-microwave power-transistor failures occur in the channel or junction area, all life-test data are referenced to the channel or junction temperature. In this article, the term "channel" will be used for simplicity. The importance of accurately determining the channel temperature of each device cannot be overstressed. With a knowledge of the case or flange temperature, DC-bias conditions, and RF input- and output-power levels, the channel temperature can be calculated using the thermal resistance of each device.**

The problem for power-amplifier (PA) designers who use GaAs devices is that the data sheets, in general, provide the device-thermal resistance with the measurement conditions for only one set of conditions (case temperature and power dissipated or channel temperature). Most data sheets do not provide any information on calculating the device-thermal resistance versus flange and channel temperature or power dissipated. A common error is to assume that the thermal resistance for GaAs devices is a constant and to use that value for different device-operating conditions. The thermal conductivity of GaAs material is a fairly strong function of temperature, which means that the channel and case or flange temperature should be considered in order to calculate the thermal resistance of GaAs devices. This article is meant to provide PA designers with a simple methodology to accurately calculate the device-thermal resistances versus their operating temperatures




**1. The definitions for GaAs-FET thermal resistance and heat flow are based on this cross-section diagram.**

from information provided on data sheets.

Before analyzing a particular device, it is necessary to define key terms. The thermal resistance, $R_{TH}$, may be used to compute the channel temperature of a device under a particular set of operating conditions, such as the case or flange temperature ($T_F$, the DC bias, and the RF input- and output-power levels. The thermal resistance (Fig. 1) can be defined as:

$$R_D = (T_{CH} - T_F) / PD \quad (1)$$

where:

$R_D$ = the channel-to-flange device-thermal resistance (K/W or °C/W),

$T_{CH}$ = the channel temperature (K or °C),

$T_F$ = the case or flange temperature (in K or °C),

$P_D$ = the power dissipated by the device, which is equal to the DC-input power plus the RF-input power minus the RF-output power ($P_D = V_{DS} \times I_{DS} + P_{IN} - P_{OUT}$ (in W).

Heat flow can be analyzed with a thermal circuit in the same way that current flow can be analyzed with an electrical circuit. The dissipated power ($P_D$) is analogous to a current-source driver and temperatures are analogous to voltages, and each element in the thermal path has a thermal resistance ($R_{TH}$) associated with it (Fig. 2). The first element in the heat-flow path is the channel-to-chip-bottom thermal resistance, $R_C$. The second element (not shown in Fig. 2) is the back-side gold (Au) metallization (Au-plated heat sink) of the chip-thermal resistance, $R_{Au}$. The third element is the solder (interface of the chip to case) thermal resistance, $R_S$. The fourth element is the case- or flange-thermal resistance, $R_F$. The next element is the thermal resistance associated with mounting the device on the heat sink, $R_I$. The last element is the heat-sink-to-ambient thermal resistance, $R_H$.

The channel-to-flange thermal resistance of the device alone ($R_D$) which is given in the device data sheet consists of practically constant thermal-resistance-versus-temperature parameters: $R_{Au}$, $R_S$, and $R_F$, and the thermal resistance of the GaAs chip, $R_C$, which is a function of the temperature of the channel ($T_{CH}$) and the temperature of the chip bottom ($T_{CB}$):

$$R_D = R_F + R_S + R_{Au} + R_C(T_{CH}, T_{CB}) \quad (2)$$

For the case of power-GaAs devices, the chips are physically long and narrow and the thermal resistance of the package flange can be approximated with the following formula:

$$R_F = [L_N(16t / \pi W)] / (\pi\sigma_F NL) \quad (3)$$

where:

$R_F$ = the case- or flange-thermal resistance (in K/W),

L = the length of the chip (in cm),

$\sigma_F$ = the thermal conductivity of the flange material (in W/cm-K),

t = the flange thickness (in cm),

W = the width of the chip (in cm), and

N = the number of chips used in the device.

If these parameters cannot be obtained by the amplifier designer from the device supplier, the following assumption for power devices can be made:

$$R_F = 0.3\, R_D \quad (4)$$

where:

$R_D$ = the channel-to-flange device-thermal resistance given in the data sheet.

The thermal resistance of the chip-back-side metallization ($R_{Au}$) is low due to the good thermal conductivity of Au and the relatively thin layer (2 to 30 μm) of the Au. This thermal resistance can be disregarded for the purpose of this analysis.

When the transistor chips are correctly mounted to the flange (with no voids underneath the chips), the interface chip-to-flange thermal resistance ($R_S$) is low in comparison with the flange and chip $R_{TH}$ and can be ignored. It is approximately 2 to 3 percent of the global device-thermal resistance ($R_D$).

If needed, it can be approximated by assuming uniform heat flow through the interface of the chip to flange:

$$R_S = t / (\sigma_s NLW) \quad (5)$$

where:

t = the solder thickness (in cm),

$\sigma_S$ = the solder-thermal conductivity (in W/cm-K),

N = the number of chips,

L = the chip length (in cm), and

W = the chip width (in cm).

If these parameters cannot be obtained by the amplifier designer from the device supplier, the following assumption for power devices can be made:

$$R_S = 0.025\, R_D \quad (6)$$

The value of these parameters ($R_{Au}$ and $R_S$) will be assumed to be negligible for the remainder of this article.

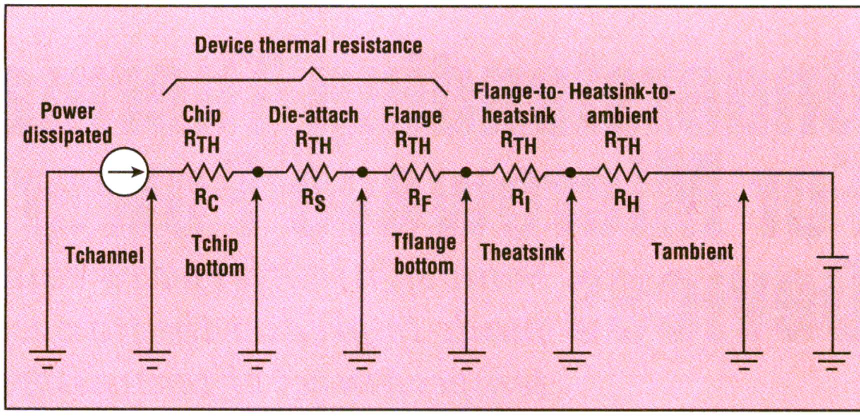
The thermal conductivity of the GaAs material can be expressed [1,2] as:

$$\sigma GaAs = 0.44[(T + 273.2) / 300]^{-1.25} \quad (7)$$

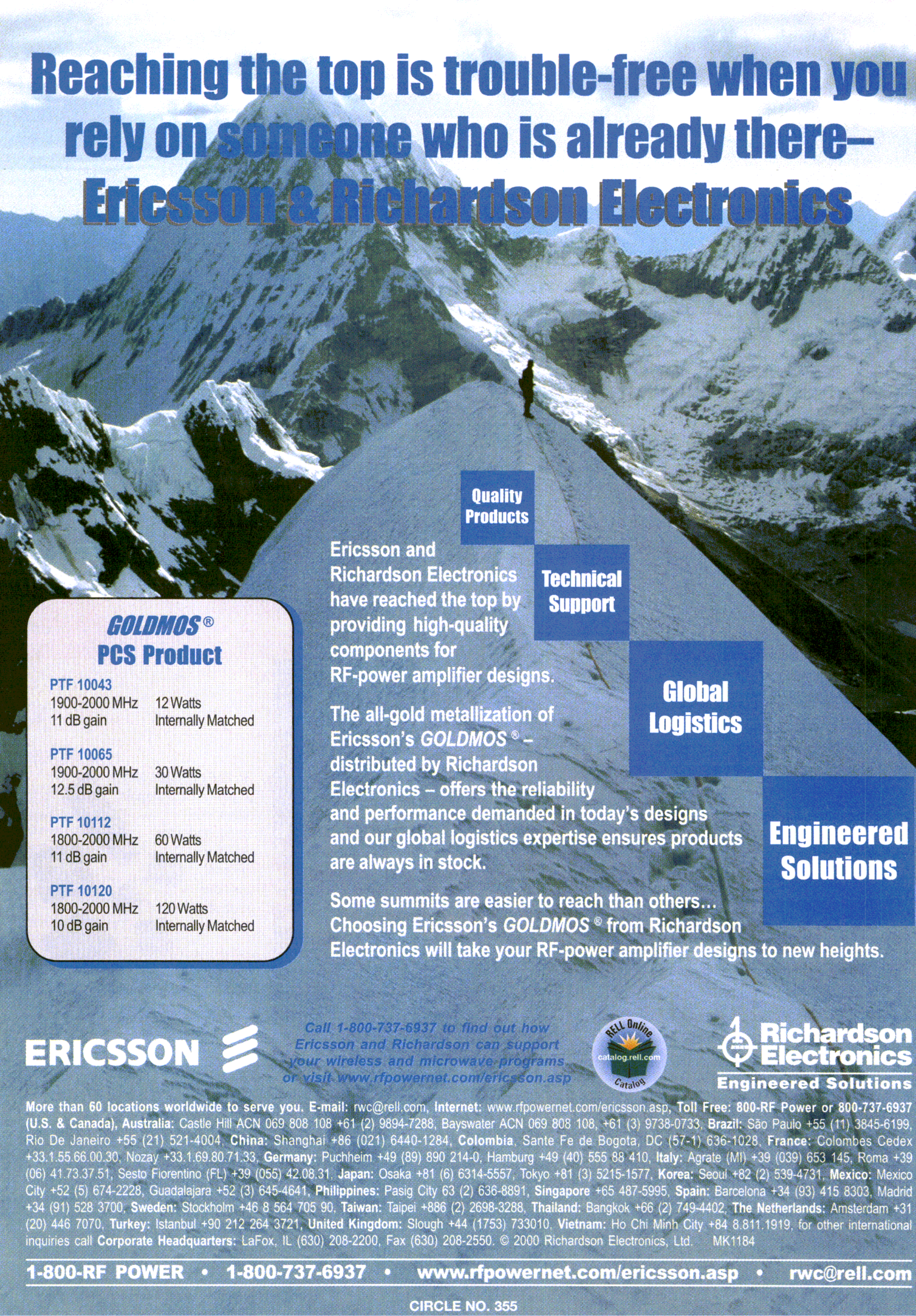
where:

$\sigma_{GaAs}$ = the thermal conductivity of the GaAs material at T (W/cm-K) and

T = the temperature of the GaAs substrate (in °C).




**2. The thermal flow through a circuit can be thought of in terms of an electrical circuit.**



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From Eq. 7 and Kirchoff's transformation,[3,4] the following formula can be derived:

$$R_{c2} = R_{c1} A / B \qquad (8a)$$

where:

$$A = [(T_{CH1} + 273.2)^{-0.25} - (T_{CB1} + 273.2)^{-0.25}] / (T_{CH1} - T_{CB1}) \qquad (8b)$$

$$B = [(T_{CH2} + 273.2)^{-0.25} - (T_{CB2} + 273.2)^{-0.25}] / (T_{CH2} - T_{CB2}) \qquad (8c)$$

where:

$R_{C1}$ = the chip-thermal resistance for a chip-bottom temperature, $T_{CB1}$ and a channel temperature $T_{CH1}$}(in °C/W or K/W),

$R_{C2}$ = the chip-thermal resistance for a chip-bottom temperature, $T_{CB2}$ and a channel temperature $T_{CH2}$(in °C/W or K/W),

$T_{CH1}$ and $T_{CH2}$ = the channel temperatures (in °C), and

**MOST DATA SHEETS DO NOT PROVIDE ANY INFORMATION ON CALCULATING THE DEVICE THERMAL RESISTANCE VERSUS FLANGE AND CHANNEL TEMPERATURE OR POWER DISSIPATED.**

$T_{CB1}$ and $T_{CB2}$ = the chip-bottom temperatures (in °C).

From the known thermal resistance ($R_{C1}$) of a chip for one set of temperatures, $T_{CH1}$ and $T_{CB1}$, its thermal resistance ($R_{C2}$) can be calculated for any set of temperatures ($T_{CH2}$ and $T_{CB2}$).

Figure 3 shows the methodology used to calculate device-thermal resistance versus temperature from the data-sheet information. The seven steps include defining parameters and calculating values from data-sheet information. Step 1 involves compiling the value of the device-thermal resistance ($R_{D1}$) for a defined set of conditions, as well as the drain-source voltage ($V_{DS}$) and the drain current ($I_{DS1}$), which can then be used to calculate the power dissipated ($P_{D1}$), which is $P_{D1} = V_{DS} \times I_{DS1}$.

Step 2 involves calculating the power dissipated, $P_{D2}$, from the new operating conditions. The new operating conditions are defined by the values of $I_{DS2}$, $T_{F2}$, $P_{IN}$, and $P_{OUT}$ for the application of interest. The value of the power dissipated can be found from:

$$P_{D2} = V_{DS} I_{DS2} + P_{IN} - P_{OUT} \qquad (9)$$

where:

$V_{DS}$ = the drain voltage (in V),

$I_{DS2}$ = the drain current (in A),

$P_{IN}$ = the input power of the device (in W), and

$P_{OUT}$ = the output power of the device (in W).

It should be noted that $V_{DS}$ is assumed to be the voltage defined in the data sheet since $R_D$ is a function of $V_{DS}$.

Step 3 involves calculating the thermal resistance of the flange ($R_F$). In order to do this, several parameters are needed from the device supplier, including the flange thickness (t), the thermal conductivity ($\sigma_F$), the width (W) of the chip, the length (L) of the chip, and the number (N) of chips used in the application. Equation 3 is then used to calculate the flange-thermal resistance ($R_F$). If these parameters are not available from the device supplier, or cannot be measured from the device itself, then it is necessary to use Eq. 4.

The next two steps involve the calculation of the chip-thermal resistance for $T_{F1}$ and $T_{CH1}$:

$$R_{C1} = R_{D1} - R_F \qquad (10)$$

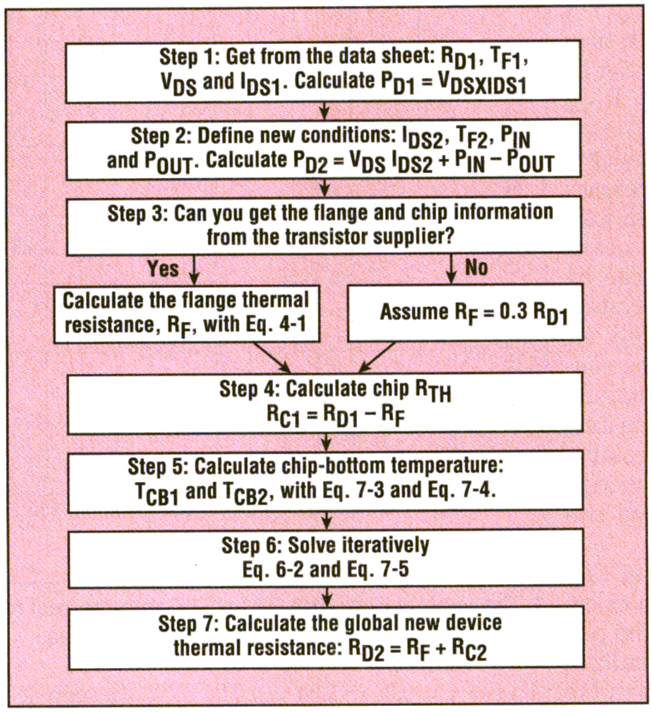
and the calculation of the chip-bottom temperatures

$$T_{CB1} = T_{F1} + R_F P_{D1} \qquad (11)$$

$$T_{CB2} = T_{F2} + R_F P_{D2} \qquad (12)$$

Step six involves finding a new chip-thermal resistance ($R_{C2}$) and channel temperature ($T_{CH2}$), which requires solution of:

$R_{C2} = R_{C1}(A/B)$




**3. This flowchart shows how to calculate the thermal resistance.**

where:

$A = [(T_{CH1} + 273.2)^{-0.25} - (T_{CB1} + 273.2)^{-0.25}]/(T_{CH1} - T_{CB1})$

$B = [(T_{CH2} + 273.2)^{-0.25} - (T_{CB2} + 273.2)^{-0.25}]/(T_{CH2} - T_{CB2})$

and

$$T_{CH2} = T_{CB2} + R_{C2} P_{D2} \qquad (13)$$

This can be solved by using an iterative method. The nth approximation of the chip-thermal resistance, $R_{C2/n}$, and the channel temperature, $T_{CH2/n}$, are obtained from Eqs. 8 and 13, respectively, in the nth step by using the previous approximation of $R_{C2}/(n - 1)$ and $T_{CH2}/(n - 1)$ obtained in the (n - 1)th step.

The initial assumed values of $R_{C2}$ and $T_{CH2}$ for starting the first iteration, n = 1, are $R_{C2/0} = R_{C1}$ and $T_{CH2/0} = T_{CB2} + R_{C1} \times P_{D2}$, respectively.

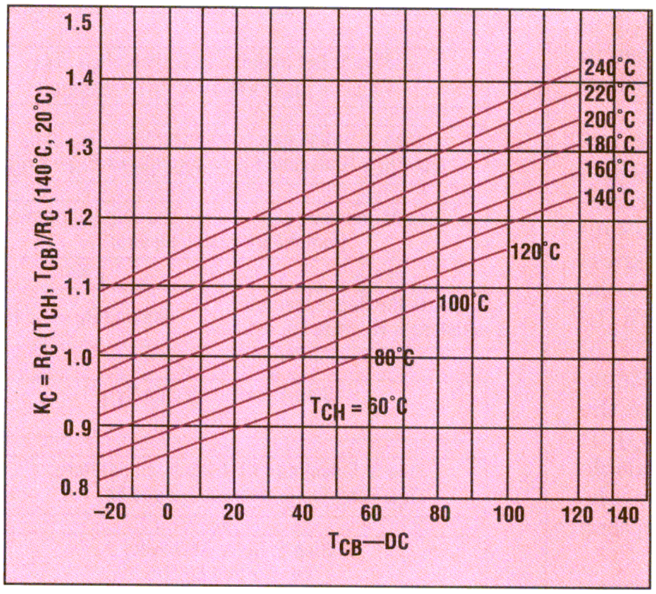
From the set of values $R_{C1}$, $T_{CB1}$, $T_{CH1}$, and $T_{CB2}$, and $T_{CH2/(n-1)}$ and Eq. 8, it is possible to calculate the nth approximation of $R_{C2/n}$. It is then possible to calculate the nth approximation of $T_{CH2/n}$ from Eq. 13.

If the following double inequality

is satisfied: $-0.01 < (R_{C2/n} - R_{C2/(n-1)})/R_{C2/n} < 0.01$, then the iteration process can be stopped. Here, $R_{C2} = R_{C2/n}$ and $T_{CH2} = T_{CH2/n}$.

If the double inequality is not satisfied, it is necessary to perform an (n + 1)th iteration with the same process.

A small program can be written[4] to solve this system of equations iteratively or, because this sequence converges to the solutions rapidly (in two or three iterations), a simple calculator with an exponential function can be used to solve the equations.




**4. These plots show the GaAs-FET chip-thermal resistance versus channel and chip bottom temperatures.**

The graphical plot of Fig. 4 can also be used to calculate $R_C$ from Eq. 8. The plot represents the normalized chip thermal resistance, $K_C = R_C/R_{CO}$, with $R_{CO}$ being the chip thermal resistance for $T_{CHO} = +140°C$ and $T_{CBO} = +20°C$, versus $T_{CH}$ and $T_{CB}$. To determine $R_{C2}$ from the graphical plot of Fig. 4, proceed as follows: First determine $K_{C1} = R_{C1}/R_{C0}$ for $T_{CB1}$ and $T_{CH1}$. Then, compute $K_{C2} = R_{C2}/R_{C0}$ for $T_{CB2}$ and $T_{CH2}$. The new thermal resistance value at $T_{CB2}$ and $T_{CH2}$ is provided by the equation $R_{C2} = R_{C1} \times K_{C2}/K_{C1}$.

The final step in the process is to calculate the new device channel-to-flange thermal resistance, $R_{D2}$, at the new base and channel temperatures, $T_{CB2}$ and $T_{CH2}$:

$$R_{D2} = R_F + R_{C2} \qquad (14)$$

## FIRST EXAMPLE

An example may help to understand how to apply this process. In this example, the first step calls for extracting information from the device data sheets. The data sheets offer the following information: $R_{D1} = 1.04$ K/W at $T_{F1} = +25°C$, $V_{DS} = +10$ VDC, and $I_{DSQ1} = 6$ A. The power dissipated is $P_{D1} = 10 \times 6 = 60$ W.

Step 2 involves finding the thermal resistance, based on 60-W power dissipation, and calculating the device-channel temperature. In the following step, since the flange and chip parameters are unknown, it can be assumed in this case that $R_F = 0.3R_{D1} = 0.3 \times 1.04 = 0.312$ K/W. In step 4, the chip-thermal resistance can then be found by $R_{C1} = 1.04 - 0.312 = 0.728$ K/W.

In step 5, the chip-bottom temperatures can be found from $T_{CB1} = 25 + (0.312 \times 60) = +43.7°C$ and $T_{CB2} = 150 + (0.312 \times 60) = +168.7°C$.

In step 6, several iterations may be needed to find the desired values. In the first iteration, from Eq. 8 and $R_{C1} = 0.728$ K/W, $T_{CB1} = +43.7°C$, $T_{CH1} = 25 + (1.04 \times 60) = +87.4°C$, $T_{CB2} = +168.7°C$, $T_{CH2/0} = 168.7 + (0.728 \times 60 = +212.4°C$. Following this, $R_{C2/1}$ can be calculated as equal to 1.08 K/W, which corresponds to a difference of 48 percent from the initial assumed value.

Therefore, a second iteration is necessary, where $T_{CH2/1} = +168.7°C/(1.08 \times 60) = +233.5°C$, which gives $R_{C2/2} = 1.109$ K/W. This value corresponds to a difference of 2.7 percent from the assumed value of $R_{C2/1}$.

Since the value still lacks the desired accuracy, a third iteration is necessary. In the third iteration, $T_{CH2/2} = 168.7 + (1.11 \times 60) = +235.2°C$, which yields $R_{C2/3} = 1.111$ K/W. This value corresponds to a difference of 0.2 percent from the assumed value of $R_{C2/2}$, which is close enough to the assumed value of $R_{C2/2}$ so as to not require a fourth iteration.

In the final step of this example (step 7), the new device-thermal resistance is found from $R_{D2} = 1.111 + 0.312 = 1.423$ K/W and $T_{CH2} = 168.7 + (1.111 \times 60) = +235.4°C$. If the original value of $R_{D1}$ were used, it would have introduced an error of –23°C for the channel temperature for this device.

## SECOND EXAMPLE

In a second exercise, the flange and chip parameters are known. For this example, the data sheet provides the following information: $R_D = 0.66$ K/W at $T_{F1} = +65°C$, $V_{DS} = +10$ VDC, and $I_{DS1} = 12$ A. The power dissipated is $P_{D1} = 10 \times 12$ A = 120 W.

In step 2, since the device is used at $T_{F2} = +20°C$ and with $P_{IN}$ of 3 W, the output power ($P_{OUT} = 60$ W, $I_{DS2} = 15$ A, and $V_{DS} = +10$ VDC. The power dissipated is $P_{D2} = (15 \times 10$ V) + 3 – 60 W = 93 W.

In step 3, the flange and chip parameters are known: the flange is 0.178 cm thick and is comprised of Cu/W-15 material with thermal conductivity of 1.90 W/cm-K. Four chips are used and each chip is 0.42 cm long and 0.10 cm wide. From Eq. 3, the flange-thermal resistance can be calculated as $R_F = 0.220$ K/W.

The next step involves finding the chip-thermal resistance, as $R_{C1} = 0.660 - 0.220 = 0.440$ K/W. The following step involves the calculation of the chip-bottom temperatures, as $T_{CB1} = +65 + (0.220 \times 120) = +91.4°C$ and $T_{CB2} = +20 + 0.220 \times 93) = +40.5°C$.

As with step 6 in the first example, this step is a matter of finding how many calculation iterations are needed to find relatively accurate values of the chip-thermal resistance. From Eqs. 8 and $R_{C1} = 0.440$ K/W, $T_{CB1} = +91.4°C$, $T_{CH1} = +91.4 + (0.440 \times 120) = +144.2°C$, $T_{CB2} = +40.5$, $T_{CH2/0} = +40.5°C + (0.440 \times 93) = +81.4°C$. From this, it is possible to calculate $R_{C2/1}$ as 0.362 K/W for the first iteration. This corresponds to a difference of –1.8 percent from the previous assumed value, this requiring a second iteration.

# 4 - 40 GHz BLOCK DOWNCONVERTER

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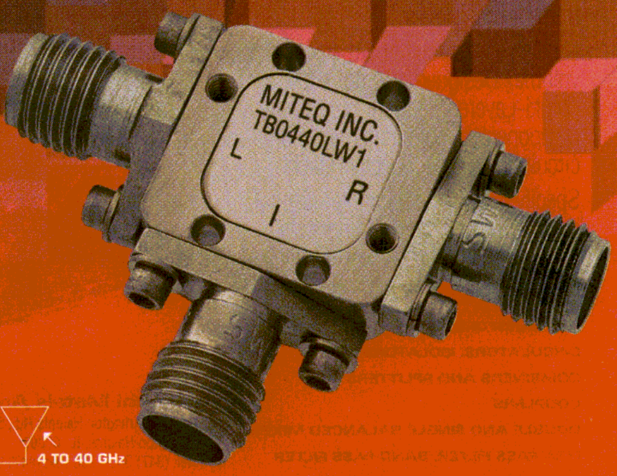
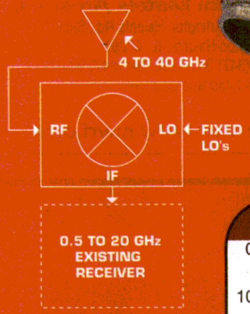
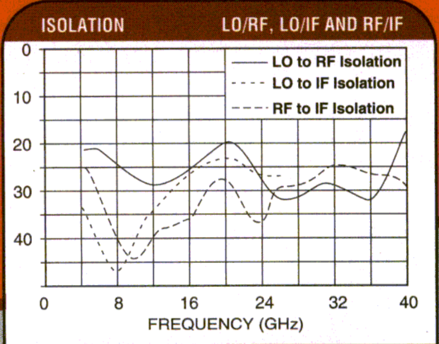


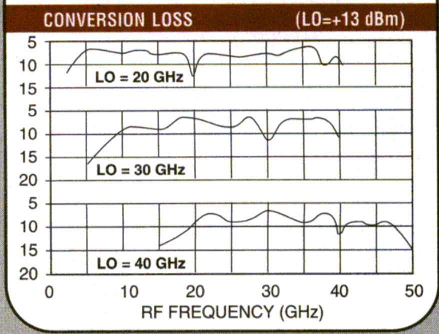

## FEATURES:

- RF/LO Coverage....... 4 to 40/4 to 42 GHz
- IF Operation............. 0.5 to 20 GHz
- LO Power Range ...... +10 to +15 dBm (usable at +7 dBm)
- RF to IF Isolation...... 25 dB
- Removable K Connectors
- From Stock




| INPUT PARAMETERS | MIN. | TYP. | MAX. |
|---|---|---|---|
| RF frequency range (GHz) | 4 | | 40 |
| RF VSWR (RF = -10 dBm, LO = +13 dBm) | | 2.5:1 | |
| LO frequency range (GHz) | 4 | | 42 |
| LO power range (dBm) | +10 | +13 | +15 |
| LO VSWR (RF = -10 dBm, LO = +13 dBm) | | 2.0:1 | |
| **TRANSFER CHARACTERISTICS** | **MIN.** | **TYP.** | **MAX.** |
| Conversion loss (dB) | | 10 | 12 |
| Single sideband noise figure (dB, at +25° C) | | 10.5 | |
| Isolation - LO to RF (dB) | 18 | 20 | |
| Isolation - LO to IF (dB) | 20 | 25 | |
| Isolation - RF to IF (dB) | 20 | 30 | |
| Input power at 1 dB compression (dBm) | | +5 | |
| Input two-tone 3rd order intercept point (dBm) | | +15 | |
| **OUTPUT PARAMETERS** | **MIN.** | **TYP.** | **MAX.** |
| IF frequency range (GHz) | 0.5 | | 20 |
| IF VSWR (RF = -10 dBm, LO = +13 dBm) | | 2.5:1 | |

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CIRCLE NO. 263

The second iteration is performed with $T_{CH2/1} = +40.5 + (0.362 \times 93) = +74.2°C$. This yields $R_{C2/2} = 0.357$ K/W, or a difference of –1.4 percent from the previous assumed value, requiring a third iteration.

The third iteration is performed with $T_{CH2/2} = +40.5 + (0.357 \times 93) = +73.70°C$, which gives $R_{C2/3} = 0.357$, which is a difference of approximately –0.1 percent from the previous assumed value. This is close enough so as to not require another iteration.

The new device-thermal resistance can be found as $R_{D2} = 0.357 + 0.220 = 0.577$ K/W. Has the original value of $R_{D1}$ been used, it would have introduced an error of +7.5°C for the channel temperature.

It should be noted that sometimes the following formula is used for the calculation of device-thermal resistance versus channel temperature:

$$R_{D2} = R_{D1}(T_{CH2})/T_{CH1})^{-1.25}$$

where:

$T_{CH1}$ and $T_{CH2}$ are the channel temperatures in K.

but this is only correct if $T_{CB1} = T_{CH1}$ and $T_{CB2} = T_{CH2}$, or if $P_{D1} = P_{D2} = 0$ W, which is a rare case. This formula does not take into account the fact that the flange-thermal resistance, which is constant versus temperature, must first be subtracted from $R_{D1}$ before applying any temperature correction.

## TECHNIQUES AND ANALYSES

The techniques and analyses previously mentioned should be useful for power-amplifier (PA) designers working at higher frequencies, especially in the frequency ranges requiring impedance-matched or unmatched GaAs FET devices. Device data sheets do not always provide thermal parameters for the desired operating conditions, and the procedures presented here allow designers to calculate the channel temperatures of a device of interest under virtually any set of input and output power conditions.

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1. Robert Anholt, *Electrical and Thermal Characterization of MESFETs, HEMTS, and HBTs*, Artech House, Norwood, MA, Chap. 4, p. 62.
2. P.D. Maycock, "Thermal Conductivity of Silicon, Germanium, III-V Compounds and Alloys," *Solid-State Electronics*, Pergamon Press, London, 1967, Vol. 10, pp. 161-168.

### For Further Reading

W.B. Joyce, "Thermal Resistance of Heat Sinks with Temperature-Dependent Conductivity," *Solid-State Electronics*, Pergamon Press, London, 1975, Vol. 18, pp. 321-322.
CEL Application Note #AN1030, "Microwave Power GaAs Device Thermal Resistance Basics," www.cel.com.

# Create Transmission-Line Matching Circuits For Power Amplifiers

***A simplified approach to the design of transmission-line matching circuits uses analytical equations to calculate the circuits elements for L, π, and T transformers.***

**Andrei V. Grebennikov**

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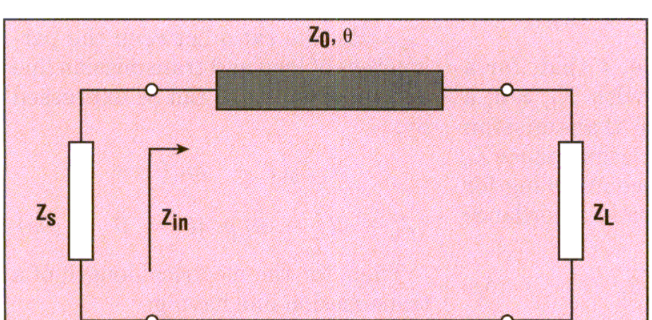
**MATCHING circuits allow power amplifiers (PAs) to attain high gain, high linearity, high efficiency, and high stability over a narrow or wide frequency band. The techniques for designing PA-matching circuits are especially important for circuits that use transmission lines, where determining circuit parameters is more complicated than for lumped-element matching circuits. An effective approach is to use a theoretical analysis to define the appropriate matching circuit, then calculate the matching-circuit parameters to minimize final tuning and attain a stable operation mode. This article presents a transmission-line matching-circuit design based on analytical equations to calculate the circuits elements for L, π, and T transformers. The introduction of a principle of equal quality factors for multisection L transformers simplifies the calculation procedure and requires only one Q circle on the Smith chart. The article demonstrates these matching principles—analytically and by Smith chart—using two examples: a narrowband, microwave, bipolar PA, and a broadband, ultra-high-frequency (UHF), laterally-diffused-metal-oxide-semiconductor-field-effect-transistor (LDMOSFET), high-power amplifier.**

Figure 1 shows an impedance-matching circuit in the form of a transmission-line transformer between source impedance $Z_S$ and load impedance $Z_L$. Equation 1a provides the input impedance as a function of a length of transmission line with arbitrary load impedance:

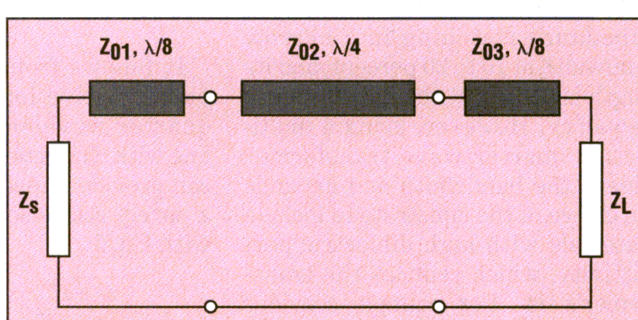
$$Z_{in} = Z_0 \frac{Z_L + jZ_0 tan\theta}{Z_0 + jZ_L tan\theta} \quad (1a)$$

where:

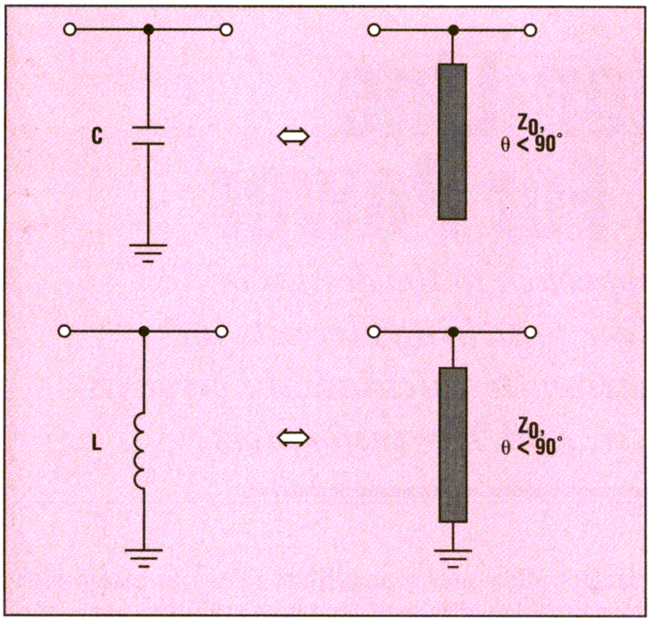
$Z_O$ = the characteristic impedance



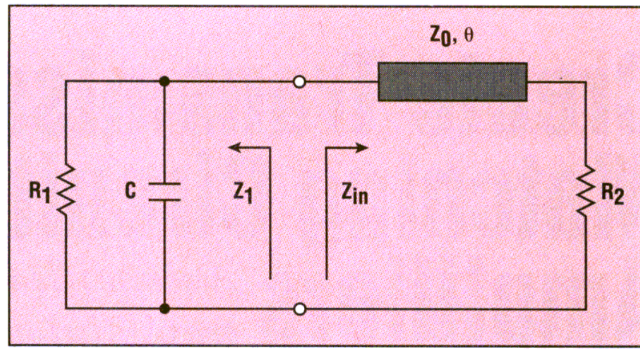

1. This schematic shows a matching circuit in the form of a transmission-line transformer.




2. This schematic illustrates a matching circuit using two λ/8 transformers and one λ/4 transformer.




**3. These are the lumped elements and transmission line for single-frequency equivalence.**




**4. This schematic shows a basic L transformer with a series transmission line.**

$$\theta = \beta l \qquad (1b)$$

= the electrical length of transmission line

$$\beta = \frac{\omega}{c}\sqrt{\mu_r \varepsilon_r} \qquad (1c)$$

= the phase constant,
c = the speed of light in free space,
$\mu_I$ = the substrate permeability,
$\epsilon_I$ = the substrate permittivity,
$\omega$ = the operating frequency, and
l = the length of the transmission line.[1]

It follows directly from Eq. 1a that, for a quarter-wavelength transmission line with:

$$\theta = \pi / 2 \qquad (1d)$$

the expression for $Z_{in}$ simplifies to:

$$Z_{in} = Z_0^2 / Z_L \qquad (2)$$

Usually, this quarter-wavelength impedance transformer is used for impedance matching in a narrow bandwidth of 10 to 20 percent, and its length is chosen at the band's center frequency. However, using a multisection quarter-wave transformer widens the bandwidth and expands the choice of the substrate to include materials with high dielectric permittivity, which reduces the transformer's size. For example, consider the case of a 15-W, gallium-arsenide (GaAs), metal-semiconductor-field-effect-transistor (MESFET) PA.[2] It uses a transformer composed of seven quarter-wavelength transmission lines of different characteristic impedances whose lengths are selected for the highest bandwidth. This design achieved a gain flatness of ±1 dB over 5 to 10 GHz.

To provide a conjugate matching of input transmission-line impedance $Z_{in}$ with a source impedance $Z_S = R_S + jX_S$ when $R_S = R_e Z_{in}$ and $X_S = -I_m Z_{in}$, Eq. 1a can be rewritten as:

$$R_S - jX_S = Z_0 \frac{R_L + j(X_L + Z_0 tan\theta)}{Z_0 - X_L tan\theta + jR_L tan\theta} \qquad (3)$$

For a quarter-wavelength transformer, Eq. 3 can be divided easily into two equations representing the real and imaginary parts of source impedance $Z_S$:

$$R_S = Z_0^2 \frac{R_L}{R_L^2 + X_L^2},$$
$$X_S = -Z_0^2 \frac{X_L}{R_L^2 + X_L^2} \qquad (4)$$

It follows from Eq. 4 that, for a purely active load with $X_L = 0$, a quarter-wavelength transmission line with characteristic impedance $Z_0$ can provide impedance matching for a purely active source in accordance with Eq. 5:

$$Z_0 = \sqrt{R_S R_L} \qquad (5)$$

For any values of load and source impedance, Eq. 3 can be divided in two equations representing the real and imaginary parts as follows:

$$\begin{cases} R_S(Z_0 - X_L tan\theta) - R_L(Z_0 - X_S tan\theta) = 0, \\ X_S(X_L tan\theta - Z_0) - Z_0(X_L + Z_0 tan\theta) + R_S R_L tan\theta = 0. \end{cases} \qquad (6)$$

Solving the two parts of Eq. 6 for the two independent variables $Z_0$ and $\theta$ provides:

$$Z_0 = \sqrt{\frac{R_S\left(R_L^2 + X_L^2\right) - R_L\left(R_S^2 + X_S^2\right)}{R_L - R_S}} \qquad (7)$$

$$\theta = tan^{-1}\left(Z_0 \frac{R_S - R_L}{R_S X_L - X_S R_L}\right) \qquad (8)$$

As a result, the transmission line with characteristic impedance $Z_0$ and electrical length $\theta$, determined by Eqs. 7 and 8, respectively, can match any source and load impedance when the impedance ratio provides a positive value under the square-root expression in Eq. 7.

For a purely active source when $Z_S = R_S$, the ratio between the parameters of load and transmission line derived from Eq. 6 can be expressed by:

$$X_L Z_0\left(1 - tan^2\theta\right) + \left(Z_0^2 - X_L^2 - R_L^2\right)tan\theta = 0 \qquad (9a)$$

Then, for the electrical length of a transmission line having:

$$\theta = \pi / 4 \qquad (9b)$$

$$Z_0 = \left| Z_L \right| = \sqrt{R_L^2 + X_L^2} \quad (10)$$

Consequently, any load impedance can be transformed to a real source impedance by a λ/8 transformer whose characteristic impedance equals the load impedance magnitude.[3]

By applying the same approach to match a purely active load with a source impedance, the total matching circuit, including two λ/8 transformers and a λ/4 transformer (as in Fig. 2) can provide impedance matching between any source impedance $Z_S$ and load impedance $Z_L$.

In some simple cases, the input impedance of a transmission line at a particular frequency can be expressed as that of a lumped element, as shown in Fig. 3. So, when $Z_L = 0$, it follows directly from Eq. 1a that:

$$Z_{in} = jZ_0 tan\theta \quad (11a)$$

which corresponds to the inductive input impedance for:

$$\theta < \pi/2 \quad (11b)$$

The equivalent inductance at the operating frequency ω is:

$$L = \frac{Z_{in}}{\omega} = \frac{Z_0 \tan\theta}{\omega} \quad (12)$$

Similarly, when $Z_L = \infty$,

$$Z_{in} = -jZ_0 cot\theta \quad (13a)$$

which corresponds to the capacitive-input impedance for:

$$\theta < \pi/2 \quad (13b)$$

The equivalent capacitance at the operating frequency ω is:

$$C = -\frac{1}{\omega Z_{in}} = \frac{tan\theta}{\omega Z_0} \quad (14a)$$

To calculate the parameters of parallel open-circuited or short-circuited stubs, it is therefore advisable to use the matching-circuit technique with lumped elements. The Smith chart is particularly useful for graphical admittance solutions.[4] When the appropriate parallel capacitance or inductance is determined analytically or by Smith chart, Eqs. 12 and 14a can be used to calculate the parameters of parallel open-circuited or short-circuited stubs, the characteristic impedance $Z_0$, and the electrical length θ.

Microwave PA design often employs a simple matching circuit in the form of an L transformer with a series transmission line as the basic matching section. It is convenient to analyze the transforming properties of this circuit by substituting the equivalent transformation of the parallel Rx circuit into the series one. $R_1$ and

$$X_1 = -1/\omega C \quad (14b)$$

are the real and imaginary components of the impedance,

$$Z_1 = jR_1X_1/\left(R_1 + jX_1\right) \quad (14c)$$

is for parallel capacitive circuits, and

$$R_{in} = ReZ_{in} \quad (14d)$$

and

$$X_{in} = ImZ_{in} \quad (14e)$$

are the real and imaginary parts of the impedance:

$$Z_{in} = R_{in} + jX_{in} \quad (14f)$$

for the series circuit presented in Fig. 4. As a result, these two circuits are equivalent at some frequency where:

$$Z_1 = Z_{in} \quad (14g)$$

that is, where:

$$R_{in} + jX_{in} = \frac{R_1X_1^2}{R_1^2 + X_1^2} + j\frac{R_1^2X_1}{R_1^2 + X_1^2} \quad (15)$$

The solution for Eq. 15 can be written as two expressions:

$$R_1 = R_{in}\left(1 + Q^2\right) \quad (16)$$

$$X_1 = X_{in}\left(1 + Q^{-2}\right) \quad (17a)$$

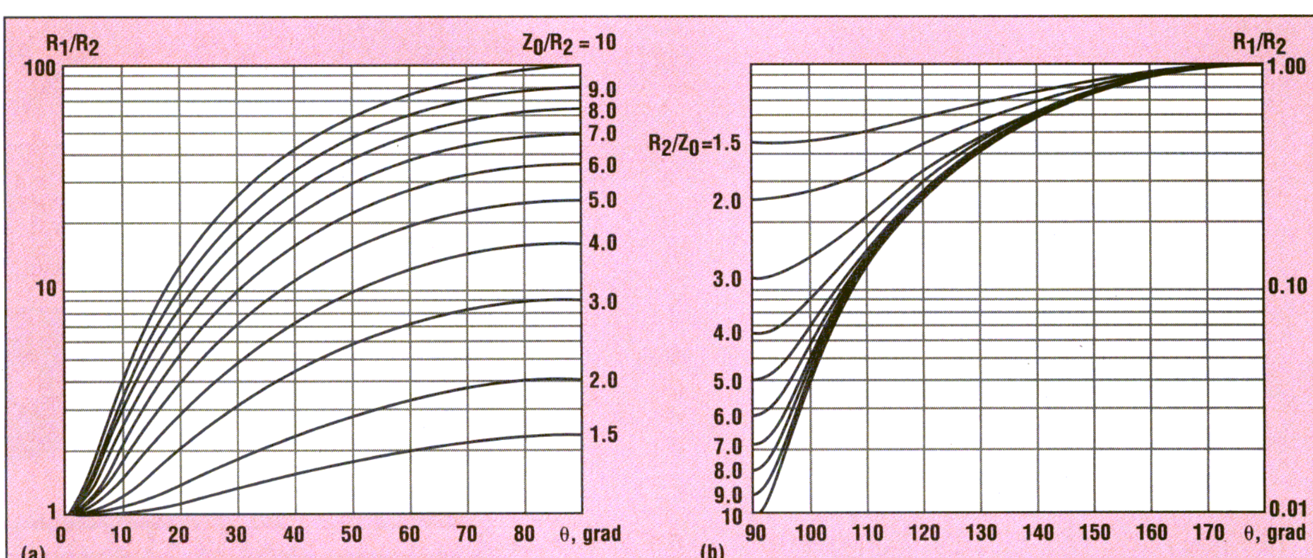
where:

$$Q = R_1/\left|X_1\right| = X_{in}/R_{in} \quad (17b)$$

is a quality factor equal for parallel-capacitive and series-transmission-line circuits. From Eq. 1a, the real and imaginary parts of the input impedance $Z_{in}$ can be written as:

$$R_{in} = Z_0^2 R_2 \frac{1 + tan^2\theta}{Z_0^2 + \left(R_2 tan\theta\right)^2} \quad (18)$$

$$X_{in} = Z_0 \, tan\theta \, \frac{Z_0^2 - R_2^2}{Z_0^2 + \left(R_2 tan\theta\right)^2} \quad (19a)$$




**5. These two nomographs can be used to calculate L transformer parameters (a) for $Z_0/R_2 > 0$ and (b) for $Z_0/R_2 < 0$.**

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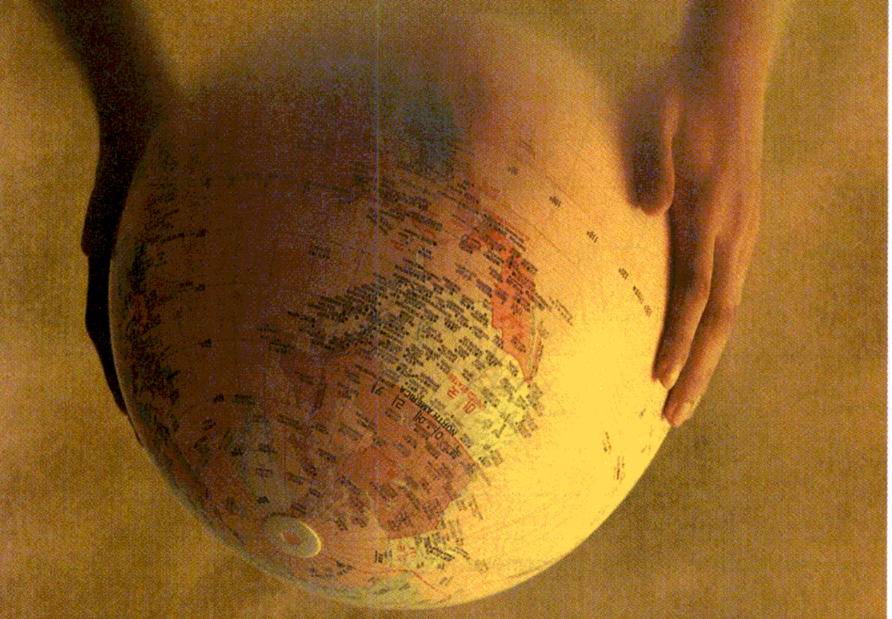

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| VSWR (max.) | 1.15:1 | 1.3:1 | 1.4:1 | 1.5:1 |
| Switching Time (max.) | 10 ms | | | |
| Operating Mode | TTL Latching with Self Cut-Off & Indicating Circuitry | | | |
| Dimension (W*D*H) | 34.0*13.4*40.0mm | | | |
| I/O Port Connector | SMA (F) / SMA(F) | | | |

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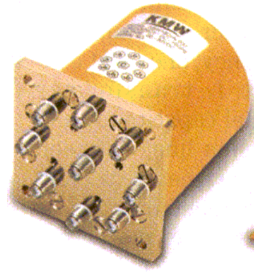

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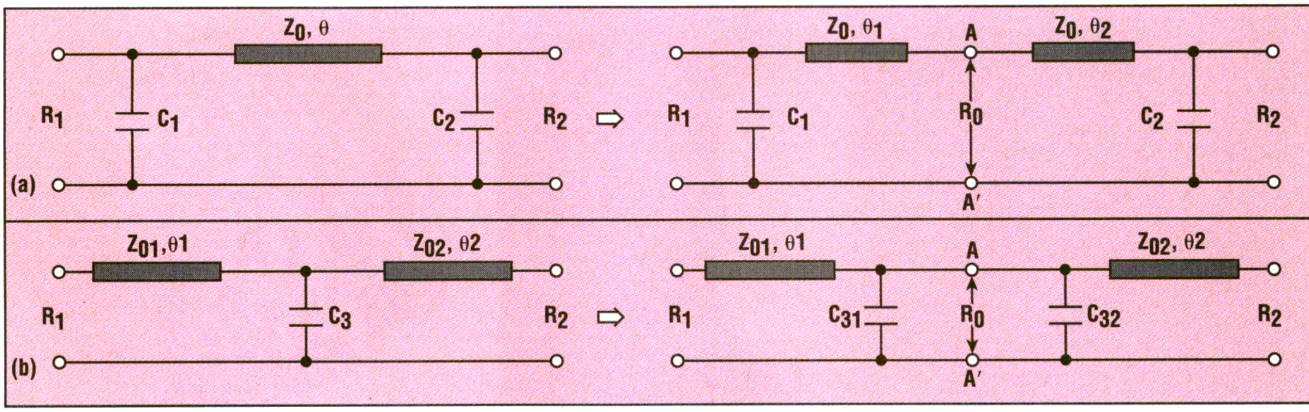



**6. These schematics represent the equivalent representation of (a) a π transformer and (b) a T transformer.**

From Eq. 19a, it follows that an inductive input impedance, which is necessary to compensate for the capacitive-parallel component, is realized when:

$$Z_0 > R_2 \quad (19b)$$

for

$$\theta < \pi/2 \quad (19c)$$

and

$$Z_0 < R_2 \quad (19d)$$

for

$$\pi/2 < \theta < \pi \quad (19e)$$

As a result, to transform resistance $R_1$ into another resistance $R_2$ at the given frequency, one must connect a two-port L transformer (including a parallel capacitance and a series transmission line) between them. When one parameter (usually the characteristic impedance $Z_0$), is decided beforehand, the matching-circuit parameters can be calculated from the following two equations:

$$C = Q/\omega R_1 \quad (20)$$

$$\sin 2\theta = 2Q/\left(\frac{Z_0}{R_2} - \frac{R_2}{Z_0}\right) \quad (21)$$

where:

Q = a function of resistances $R_1$ and $R_2$ as well as the parameters of the transmission line, the characteristic impedance $Z_0$, and electrical length θ, by the following equation:

$$Q = \sqrt{\frac{R_1}{R_2}\left[\cos^2\theta + (R_2/Z_0)^2 \sin^2\theta\right] - 1} \quad (22)$$

It follows from Eqs. 21 and 22 that the calculation of θ is a result of the numerical solution of a transcendental equation with one unknown parameter. However, it is more convenient to combine these two equations and to rewrite them in the form:

$$\frac{R_1}{R_2} = \frac{1 + \left(\frac{Z_0}{R_2} - \frac{R_2}{Z_0}\right)^2 \sin^2\theta \cos^2\theta}{\cos^2\theta + (R_2/Z_0)^2 \sin^2\theta} \quad (23)$$

Figure 5 shows the resistance ratio $R_1/R_2$ as a function of the parameter $Z_0/R_2$ and electrical length θ in the form of two nomographs—one for the case of $Z_0/R_2 > 0$, and another for the case of $Z_0/R_2 < 0$. Thus, when input resistance $R_1$ and output resistance $R_2$ are known in advance and when the value of transmission-line characteristic impedance $Z_0$ is chosen, evaluating the required value of θ is easy using these nomographs. These graphical results show that, in contrast to a lumped L transformer, a simple L transformer with a transmission line can match a purely resistive source and load impedance with any values of the ratio $R_1/R_2$.

One can realize a matching circuit in the form of a π transformer by the appropriate connection of two L transformers when, through each L transformer, the resistances $R_1$ and $R_2$ are transformed to some intermediate resistance $R_0$, as shown in Fig. 6a. In this case, to minimize the length of the transmission line, the value of R0 should be smaller than that of $R_1$ and $R_2$, (i.e., $R_0 < R_1, R_2$). The same procedure for the T transformer shown in Fig. 6b provides a value of $R_0$ that is larger than that of $R_1$ and $R_2$ (i.e., $R_0 > R_1, R_2$). Then, in the case of a T transformer, two parallel adjacent capacitances are combined. For a π transformer, two adjacent series transmission lines are combined into one transmission line.

For a π transformer, the lengths of each part of the combined transmission line can be calculated by equalizing the imaginary parts of the impedances from both sides at the reference plane A-A′ to zero, which means that the intermediate impedance, $R_0$, is real. This leads to two quadratic equations for each electrical length of the combined series transmission lines:

$$\tan^2\theta_1 - \frac{R_1}{Z_0 Q_1}\left[1 - \left(1 + Q_1^2\right)\left(\frac{Z_0}{R_1}\right)^2\right]\tan\theta_1 - 1 = 0 \quad (24)$$

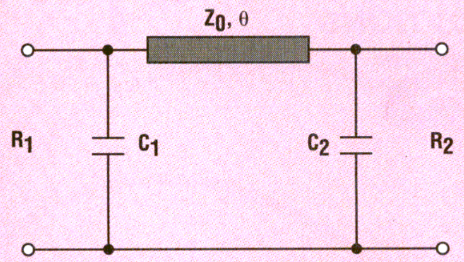
$$\tan^2\theta_2 - \frac{R_2}{Z_0 Q_2}\left[1 - \left(1 + Q_2^2\right)\left(\frac{Z_0}{R_2}\right)^2\right]\tan\theta_2 - 1 = 0 \quad (25a)$$

where:

$$Q_1 = \omega C_1 R_1 \quad (25b)$$

$$Q_2 = \omega C_2 R_2 \quad (25c)$$

One can simplify this analytical calculation procedure by using the no-




$$\omega C_1 = Q_1 / R_1,\ \omega C_2 = Q_2 / R_2,\ \theta = \theta_1 + \theta_2,$$

$$Q_2 = \sqrt{\frac{R_2}{R_1}(1+Q_1^2) - 1},\ Q_1 > \sqrt{\frac{R_1}{R_2} - 1}$$

**7. A two-port π transformer with design formulas is shown here.**

mographs in Fig. 5. So if the values of $R_1$ and $R_2$ are selected in advance to set the intermediate resistance, $R_0$, and the characteristic impedance of the transmission line, $Z_0$, the values of $\theta_1$ and $\theta_2$ can be easily determined from one of these nomographs.

Figure 7 shows the most-common two-port π-transformer design, along with its formulas. This type of transformer is widely used in the output-matching circuits of high-power amplifiers in Class B operation, where it is necessary to isolate the fundamental drain or collector waveform by suppressing the appropriate harmonic(s). Additionally, it is convenient to use this transformer as an input-matching circuit in high-power balanced amplifiers where capacitances can be connected between series transmission lines. And finally, the use of π transformers can be very important for high-efficiency, Class E operation where a capacitive output load provides the appropriate transistor impedance.

Figure 8 shows some of the matching-circuit configurations of two-port T transformers, along with the analytical formulas to calculate the parameters of each transformer. Here, it is assumed that the electrical lengths of the transmission lines are smaller than a quarter wavelength. That is:

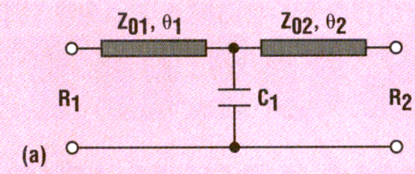
$$(\theta_1,\ \theta_2,\ \theta_3) < \pi/2 \qquad (25d)$$

T transformers are usually used in the high-power amplifiers as input-, interstage-, and output-matching circuits, especially in matching circuits with two capacitances and a series transmission line. By using this type of circuit as an output-matching circuit for a PA, it is possible to realize a high-efficiency, Class F operation. This is possible because the series transmission line that is adjacent to the active device's drain or collector creates the appropriate impedance through open- or short-circuit harmonic conditions.[5]

The first of two practical illustrations of transmission-line matching-circuit techniques can be seen in the design of an output-matching circuit for a 5-W, 1.6-GHz bipolar amplifier that operates from a +24-VDC supply voltage and provides a gain of approximately 10 dB. The mentioned requirements can be provided with an NPN silicon (Si) transistor in Class AB operation that is intended for microwave-transmitter applications over the frequency range of 1.5 to 1.7 GHz. The manufacturer usually states the values of the complex input and output impedances or admittances at the nominal operation point on the data sheet for the device. At the operating frequency of 1.5 GHz, let

$$Z_{out} = (5.5\ - j6.5)\ \Omega \qquad (25e)$$

correspond to a series combination of transistor-output resistance and

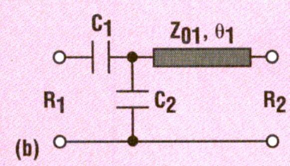



$$\sin 2\theta_1 = 2Q_1 \Big/ \left(\frac{Z_{01}}{R_1} - \frac{R_1}{Z_{01}}\right),\ \sin 2\theta_2 = 2Q_2 \Big/ \left(\frac{Z_{02}}{R_2} - \frac{R_2}{Z_{02}}\right),$$

$$\omega C_1 = \frac{Q_1 + Q_2}{1+Q_1^2}\ \frac{\cos^2\theta_1 + (R_1/Z_{01})^2 \sin^2\theta_1}{R_1},$$

$$Q_1 = \sqrt{\frac{R_1}{R_2}\,\frac{\cos^2\theta_2 + (R_2/Z_{02})\sin^2\theta_2}{\cos^2\theta_1 + (R_1/Z_{01})\sin^2\theta_1}\left(1+Q_2^2\right) - 1},$$

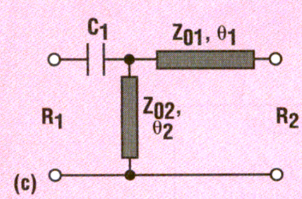
$$Q_2 > \sqrt{\frac{R_1}{R_2} - 1}$$




$$\omega C_1 = 1/(R_1 Q_1),\ \sin 2\theta_1 = 2Q_2 \Big/ \left(\frac{Z_{01}}{R_2} - \frac{R_2}{Z_{01}}\right),$$

$$\omega C_2 = (Q_2 - Q_1) / [R_1(1+Q_1^2)],$$

$$Q_1 = \sqrt{\frac{R_2}{R_1}\,\frac{1+Q_2^2}{\cos^2\theta_1 + (R_2/Z_{01})\sin^2\theta_1} - 1},\ Q_2 > \sqrt{\frac{R_1}{R_2} - 1}$$

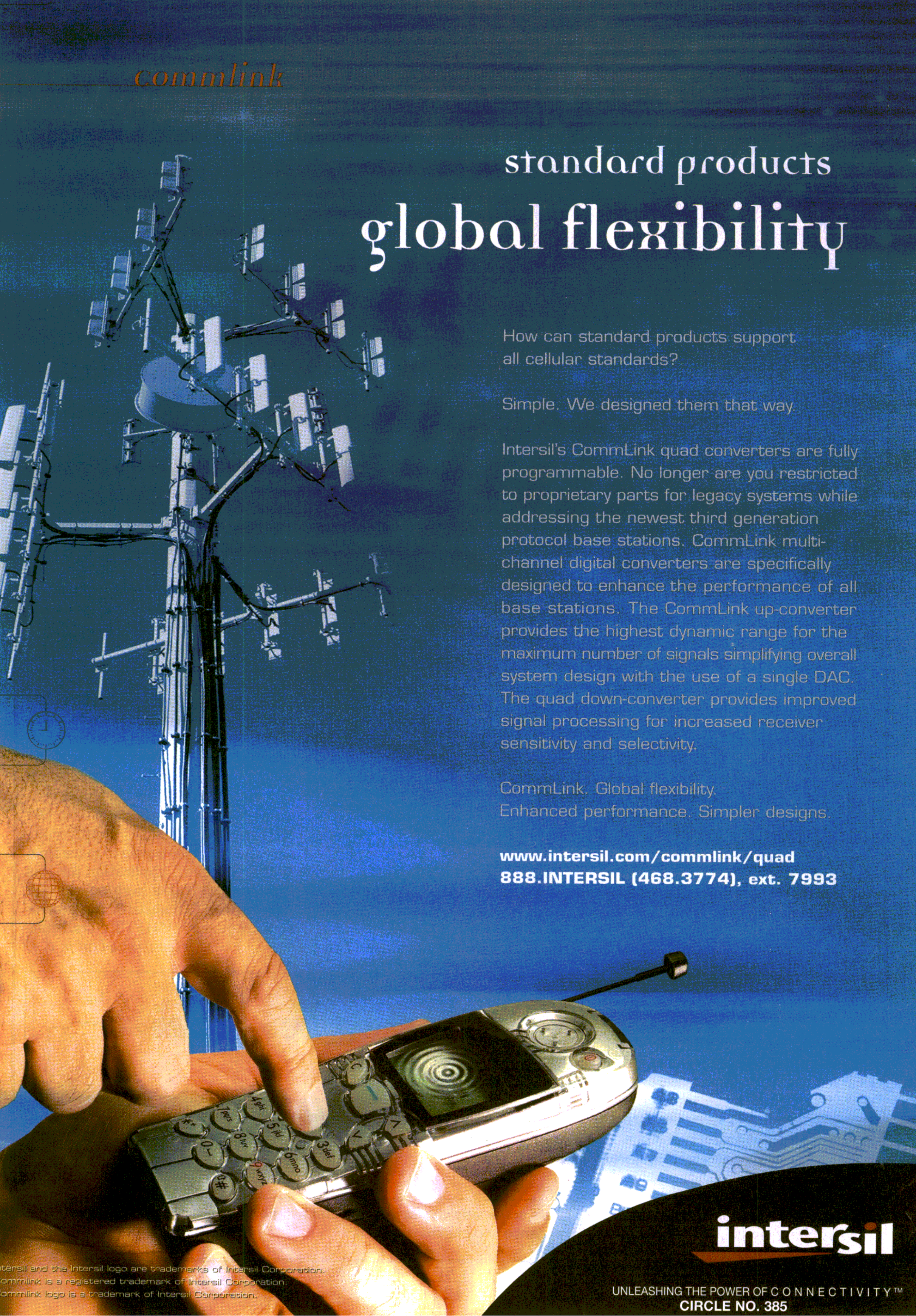



$$\omega C_1 = 1/(R_1 Q_1),\ \sin 2\theta_1 = 2Q_2 \Big/ \left(\frac{Z_{01}}{R_2} - \frac{R_2}{Z_{01}}\right),$$

$$Z_{02} \tan\theta_2 = R_1(1+Q_1^2)/(Q_1 - Q_2),$$

$$Q_1 = \sqrt{\frac{R_2}{R_1}\,\frac{1+Q_2^2}{\cos^2\theta_1 + (R_2/Z_{01})\sin^2\theta_1} - 1},\ Q_2 > \sqrt{\frac{R_1}{R_2} - 1}$$

**8. A two-port T transformer with design formulas is shown here.**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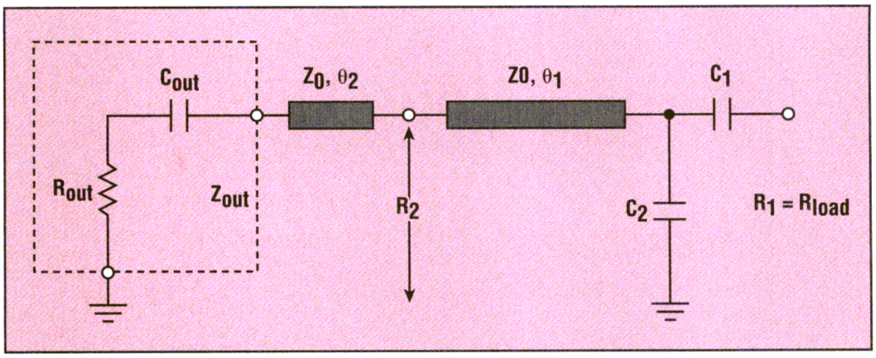
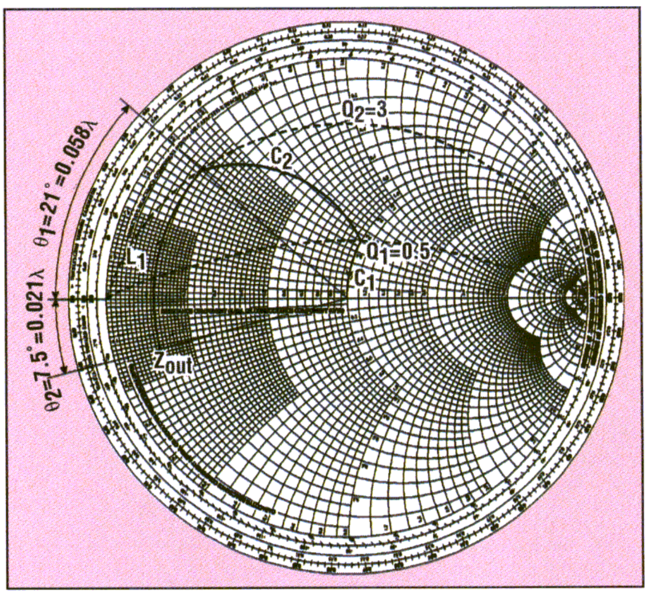



**9. This schematic shows a two-port network including output-device impedance and matching circuit.**

capacitance. To match this capacitive impedance to the standard 50-Ω load resistance, it is advisable to use a matching circuit in the form of a T transformer shown in Figure 8b. Figure 9 shows the complete network, including output-device impedance and matching circuit.

First, the circuit should compensate for the series capacitance inherent in output impedance. For a small electrical length where:

$$\tan\theta \cong \theta \qquad (25f)$$




**10. This Smith chart shows the parameters used in the schematic in Fig. 9.**

and characteristic impedance $Z_1 > R_{out}$, one can deduce from Eqs. 18 and 19a that:

$$\begin{cases} R_{out} \cong R_2 \\ \theta_2 \cong -X_{out} / Z_0 = 1/\omega C_{out} Z_0 \end{cases} \qquad (26)$$

where:

$\theta_2$ = a part of the total transmission line required to compensate for the output capacitance reactance. If one selects the value $Z_0$ to be 50 Ω, then $\theta_2$ = 6.5/50 = 0.13 radian, which is equal to approximately 7.5 deg. of electrical length. Then, the value of quality factor $Q_2$, which one must choose to calculate the parameters of the matching circuit, is defined by:

$$Q_2 > \sqrt{R_1/R_2 - 1} = 2.84 \qquad (27)$$

Consequently, the value of $Q_2$ must be larger than 2.84. For example, a value of $Q_2 = 3$ can be chosen to yield a 3-dB bandwidth of 1.6 GHz/3 = 533 MHz. As a result, the values of the output-matching circuit parameters are as follows:

$$\theta_1 = \frac{1}{2}\sin^{-1}\left[2Q_2 / \left(\frac{Z_0}{R_2} - \frac{R_2}{Z_0}\right)\right] = 21° \qquad (28)$$

$$Q_1 = \sqrt{\frac{R_2}{R_1}\frac{1 + Q_2^2}{\cos^2\theta_1 + (R_2/Z_0)^2\sin^2\theta_1} - 1} = 0.5 \qquad (29)$$

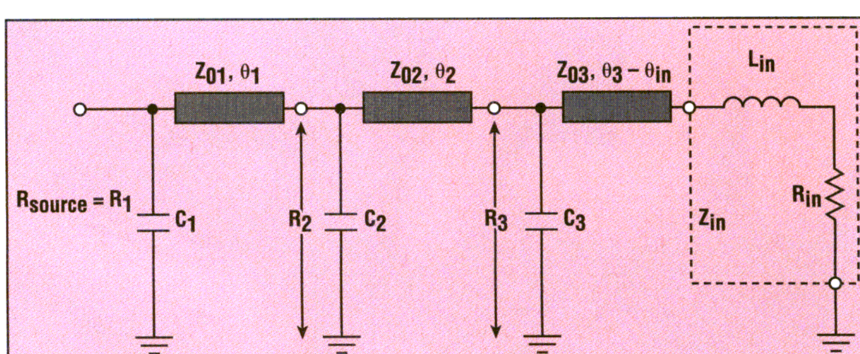
$$C_1 = 1/(\omega Q_1 R_1) = 4\ pF \qquad (30)$$

$$C_2 = (Q_2 - Q_1)/\omega R_1(1 + Q_1^2) = 4\ pF \qquad (31a)$$

Figure 10 shows the function of each element traced on the Smith chart. The easiest and most convenient way to plot the traces of the matching-circuit elements is to first plot the traces of $Q_1$ and $Q_2$. Then plot the trace of the series transmission line as far as the intersection point with the $Q_2$ circle, and plot the trace of capacitance $C_2$ as far as the intersection point with the $Q_1$ circle.

The second practical illustration of transmission-line matching-circuit techniques can be seen in the design of a 150-W amplifier that operates over a frequency band of 470 to 860 MHz, uses a +28-VDC supply voltage, and provides a gain of more than 10 dB. A typical application for this type of circuit is a high-power balanced LDMOS transistor in an ultra-high-frequency (UHF) TV transmit-

*(continued on p. 172)*




**11.This schematic shows a broadband two-port network including input-device impedance and matching circuit.**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# Browsers Can Access Free Spurious-Response Calculator

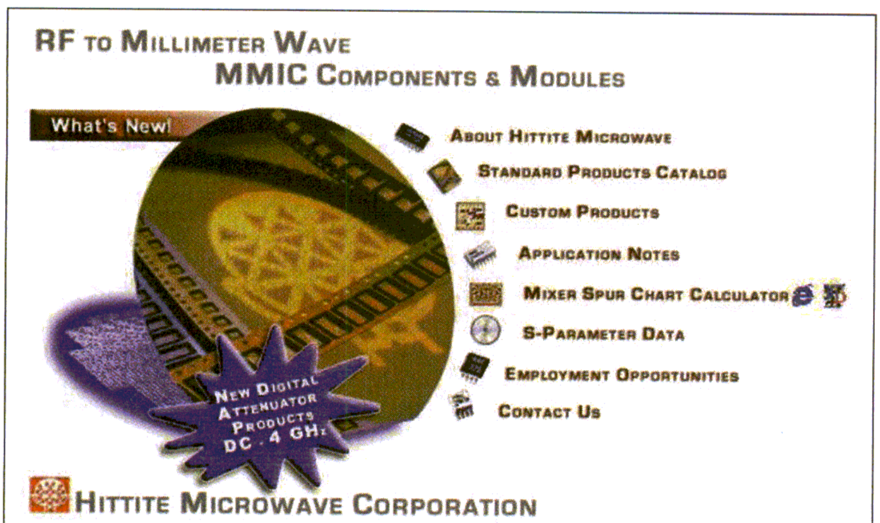
This high-frequency manufacturer's site features ICs through 40 GHz along with a series of application notes and an interactive calculator for mixer users.

ALAN CONRAD
Special Projects Editor

Hittite Microwave Corp. (http://www.hittite.com) is well-known as a quality supplier of microwave and millimeter-wave mixers, switches, and attenuators for wired and wireless applications. The firm's more than 90 monolithic microwave integrated circuits (MMICs) and MMIC assemblies are available in chip and packaged formats, and are described in detail at the company's website. But the site is worth visiting for more than just product data sheets, since it also contains a wide variety of technical application notes and a powerful interactive-mixer spurious-response calculator.

Browsers of the Hittite site *(see figure)* will find company-authored papers on topics such as metal-oxide-semiconductor-field-effect-transistor (MESFET) switches, mixers and frequency converters, wideband amplifiers, and power amplifiers (PAs). Other topics include monolithic phase shifters, dual-gate MESFET applications, active filters, and circulators.

Application notes include guidelines for electrostatic-discharge (ESD) protection of gallium-arsenide (GaAs) MMIC devices, GaAs-MMIC standard-product process flows, layout guidelines for MMIC components, and handling, mounting, and bonding of MMIC die. Other notes include positive voltage control of MMIC multithrow switches with floating-ground techniques, MMIC amplifier-biasing procedures, and mixer applications.




The Hittite Microwave Corp. website contains a variety of product information and useful application notes.

Browsers can also download measured small-signal S-parameter data for many of the firm's low-cost plastic and ceramic mixers, switches, and attenuators. All data are in standard American Standard Code for Information Interchange (ASCII) text format and can be read directly by most linear circuit simulators.

The site boasts an easy-to-use, interactive-mixer spurious-response calculator. An operator provides various defining frequencies for the RF and intermediate-frequency (IF) passband ranges, along with the appropriate local-oscillator (LO) frequency, the multiplication or harmonic factor (N) for the LO, and the harmonic factor (M) for the RF input signals. The program then calculates and displays the in-band spurious responses that will appear at the IF output port.

A visit to the site is time well-spent for specifiers of microwave and millimeter-wave MMICs, and for receiver (Rx) designers interested in quickly calculating the spurious responses for a particular architecture. ●

*Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, FAX: (978) 250-3373.*
**Internet: http://www.hittite.com.**

# Site Holds Engineering Tools and Links

This outstanding website contains links to analysis tools and engineering resources for designing wireless and military systems.

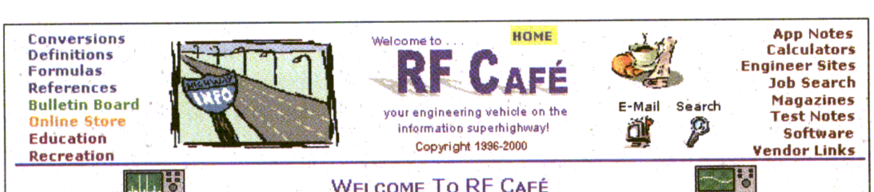
ALAN ("PETE") CONRAD
*Special Projects Editor*

Visitors to RF Café (www.rfcafe.com) will find information on RF design and more. This well-stocked site contains useful information on circuit and system design, with links to hundreds of useful sites covering a host of disciplines. These links will transport visitors to engineering design tools, engineering references and standards, application notes, calculators, engineering sites, magazines, software, test equipment, component manufacturers, and more.

The RF Café site features a page on calculations and formulas that covers more than 60 applications. Calculations apply to analog-to-digital-converter (ADC) parameters, attenuators, bit-error rate (BER), Boolean algebra, and capacitance. Other calculations include damped responses, filters, Fourier series of periodic signals, inductance, second- and third-order two-tone intercept points, link budgets, and noise figure.

Other pages include tables, lists, and graphs of often-needed references. They include amplitude modulation (AM), atmospheric absorption, atmospheric refraction, antennas, batteries, wireless communication specifications, wireless local-area-network (WLAN) specifications, and dielectric constants. A page on mechanical disciplines includes bolts, nuts, and washers in metric and society of automotive engineers (SAE), coefficient of expansion, lettered and numbered drill sizes, as well as SAE and metric tap and drill sizes.

Links to free downloadable software sites will send browsers to an antenna-design program, an RF/microwave capacitor calculator, a program that handles Butterworth filter design, software for active-filter designs, wire-line calculators, and a program to calculate transmission-line parameters. Links to vendor application notes include companies such as Agilent Technologies, Anaren Microwave, Anritsu, Belden Cables, International Rectifier, Mini-Circuits, Motorola, and Trak Microwave. Links to major publications include *Microwaves & RF*, *Wireless System Design*, and *Electronic Design*, along with other national and international sources.




More than 25 links to engineering-related sites are available to browsers including component distributors, the Institute of Electrical Engineers (IEE) in the UK, the Institute of Electrical and Electronics Engineers (IEEE) in the US, an electronic-warfare (EW) and radar-systems engineering handbook, and a variety of online training sites. Links are also provided to all major software programs, including linear and nonlinear simulators and electromagnetic (EM) simulators, each with a list of Internet addresses and phone numbers.

Links are provided to sources with applications on testing of components. Application notes on ADCs, two-tone and three-tone mixer intermodulation (IM) products, and voltage-controlled oscillators (VCOs) are included, along with measuring peak-to-peak and average power of digitally modulated signals.

In addition to the links of many sites and applications, RF Café is home to several outstanding shareware-design tools. They include RF Workbench, RF Cascade Workbook, and RF stencils for Microsoft Visio Technical. RF Workbench serves well for performing cascaded chain and spurious analysis of receivers (Rxs) and transmitters (Txs). ●

*For more information, visit the site at http://www.rfcafe.com.*

# Site Provides Powerful Search Functions

Those who are in need of a customized search engine will find what they need here in terms of engineering topics and applications.

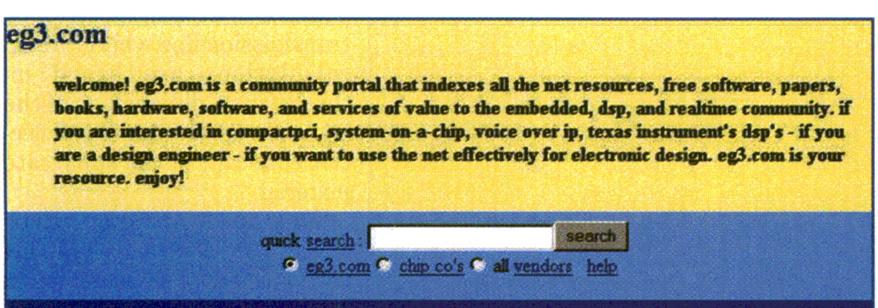
ALAN ("PETE") CONRAD
*Special Projects/Editor*

Search engines on the Internet are many. But there are few that can actually save high-frequency engineers time when they need to quickly find design or application information. While some work better than others, they all lack the dedication to a specific field or discipline. But the size at www.eg3.com is a portal that serves as an edited Yahoo link for design engineers. The site indexes a variety of engineering net resources by simply typing in a single word topic of interest and click search.

For example, searching the www.eg3.com site for the term DSP (for digital signal processor) provides a browser with links to books, vendors, application notes, and white papers *(see figure)*. As an example, application notes from Xilinx feature DSP-related topics such as DSP implemented with field-programmable gate arrays (FPGAs), signal processing with FPGAs, how to use FPGAs to design custom DSP functions, and how to use programmable logic to accelerate DSP functions.

Entering the word "software" opens links to ADA-language publications, resources, and application notes, as well as C-language resources and applications plus software tools and free software downloads.

Searching for DSP wavelets brings up a window containing the following brief description. Wavelets are mathematical functions that cut up data into different frequency components, and then study each component with a resolution matched to its scale. They have advantages over traditional Fourier methods in analyzing physical situations where the signal contains discontinuities and sharp spikes. Wavelets were developed independently in the fields of mathematics, quantum physics, electrical engineering, and seismic geology.




The site at www.eg3.com provides rapid links to a host of engineering tools and resources, including application notes, white papers, and downloadable software.

Searching the word "amplifiers" opens links to online learning and training tools such as those offered by Besser Associates (Mountain View, CA), plus a link to the Internet guide to electronics for beginners. This site contains many tutorials and examples of circuit analysis and design.

Searching the term FFT (for Fast Fourier transform) reveals links to a variety of locations dedicated to FFTs. One link provides the browser with a FFT analysis tutorial. This tutorial clearly explains the Fourier transform and Fourier series, both key parts of advanced mathematics and essential for many science and engineering tasks. Another link describes the Fourier transform and the frequency domain. This tutorial teaches the user the applications of Fourier series, Fourier transforms, discrete Fourier transforms (DFTs), and FFTs.

A link to FFT information presents the browser with a comprehensive table that lists key attributes of FFT chips such as speed, power, and word size. It also contains links to all sorts of FFT processors such as special-purpose chips, board-level products, and programmable DSP chips with links to manufacturers. ●

*For more information, visit the site at http://www.eg3.com.*

# Site Offers Free RF Design Suite

Visitors to this site can download an easy-to-use collection of design tools for RF, microwave, and wireless applications.

ALAN "PETE" CONRAD
*Special Projects Editor*

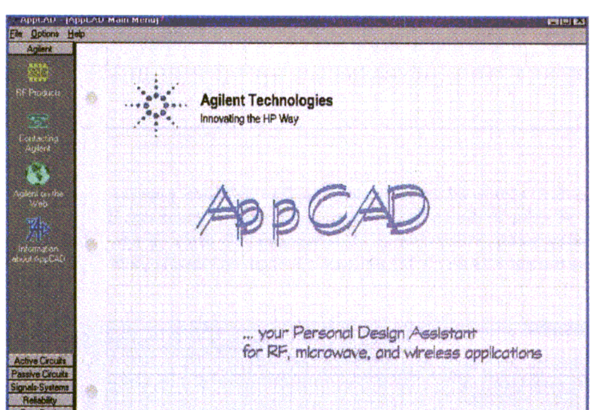
Software enthusiasts will fondly remember AppCAD, a collection of application notes and RF/microwave design tools developed largely by engineers within the semiconductor sector of Hewlett-Packard Co. (Palo Alto, CA). With the spin-off of Agilent Technologies from Hewlett-Packard last year, AppCAD could have become part of history. But, thanks to a relatively obscure branch of the massive Hewlett-Packard website at http://www.hp.woodshot.com, Web surfers can download the latest version (Version 2.0) of the AppCAD collection for Windows 95 and later personal-computer (PC)-based operating systems (the original AppCAD was a DOS-based suite of programs). AppCAD is useful for the design and analysis of circuit, signals, and systems using Agilent Technologies' silicon (Si) and gallium-arsenide (GaAs) integrated circuits (ICs) and discrete semiconductors.

AppCAD is a suite of RF design tools and application notes based on the use of discrete semiconductors and ICs *(see figure)*. The latest version of the software suite features active and passive circuit/system analysis modules. The active-component modules include a cascaded chain analysis routine, mixer spurious response analysis, and third-order intermodulation calculation. The cascaded chain analysis module analyzes a multistage system to calculate cascaded performance parameters such as noise figure, gain, signal-to-noise ratio (SNR), noise floor, spurious-free dynamic range, minimum detectable signal (MDS), intermodulation, and other parameters. The routine allows a sensitivity analysis to be performed on each stage of the system, to compute that stage's contributions to overall system noise figure, gain, and third-order intermodulation distortion. A cascaded multistage system can also be analyzed over several different temperature ranges.

AppCAD's mixer spurious-response module calculates RF ranges that can generate spurious responses within a particular intermediate-frequency (IF) range. The local oscillator (LO) in these analyses can be either fixed or variable; the user selects the order of the harmonics for the RF and LO sources. The third-order intermodulation calculator predicts third-order intermodulation products for RF devices, particularly useful in determining the overhead and dynamic range of components for wireless communications systems. Additional modules include tools for analysis of voltage-biased RF ICs, current-biased RF ICs, large- and small-signal detectors, field-effect transistors (FETs), and bipolar transistors.

Passive-component analysis modules include a microstrip transmission line calculator that computes the characteristic impedance ($Z_0$) of a microstrip transmission line based on the physical dimensions of the line and the dielectric constant of the dielectric material; a coplanar waveguide module that calculates the characteristic impedance of a coplanar waveguide transmission line with or without a ground plane on the backside of the dielectric material; and a stripline transmission line module that calculates the characteristic impedance of a balanced stripline transmission line where the $Z_0$ calculations are based on the physical dimensions of the transmission line and the dielectric constant of the substrate material.

The AppCAD program can be downloaded as a single 7-Mb file or requested for shipment on six 1.44-Mb floppy disks.




AppCAD is a suite of RF design tools and application notes based on the use of discrete semiconductors and ICs.

*For more information, visit the site at www.hp.woodshot.com.*

# High Performance and Low Cost Electromagnetic Simulation and Optimization Tools

From: Zeland Software, Inc., 39120 Argonaut Way, Suite 499, Fremont, CA 94538, U.S.A.,
Phone: 510-797-8109, Fax: 510-797-8241, E-mail: zeland@zeland.com, Web: http://www.zeland.com

## Products:

**IE3D** Planar and 3D Electromagnetic Simulation and Optimization Package

**FIDELITY** Time-Domain FDTD Full 3D Electromagnetic Simulation Package

## Applications:

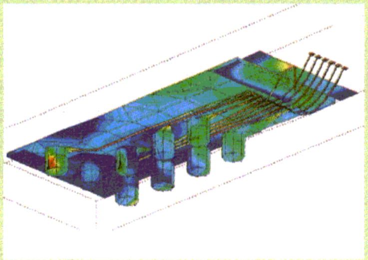
Microstrip, CPW, striplines, suspended-strip lines, coaxial Lines, rectangular waveguides, high speed digital transmission lines, 3D interconnects, PCB, MCM, HTS circuits and filters, EMC/EMI, wire antennas, microstrip antennas, conical and cylindrical helix antennas, inverted-F antennas, antennas on finite ground planes, and other RF antennas.

## Features:

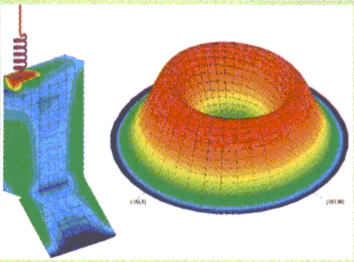
- MS-Windows graphic interface for interactive construction of geometry
- High efficiency and flexibility MOM and FDTD simulation engines
- Non-uniform automatic meshing and de-embedding
- Modeling 3D metallic and dielectric structures
- Built-in robust and efficient electromagnetic optimization
- Accurate modeling of metallic thickness, thin substrate and lossy dielectric materials
- Mixed electromagnetic simulation and nodal analysis for large scale simulation
- Cartesian and Smith Chart display of S-, Y- and Z-parameters, VSWR
- RLC parameter extraction compatible with SPICE
- 2D and 3D display of current distribution, radiation patterns and near field
- Calculation of antenna and scattering parameters including directivity, efficiency and RCS
- Current and near field animation

### IE3D Simulation Examples and Display

The current distribution on an AMKOR SuperBGA model at 1GHz created by the IE3D simulator

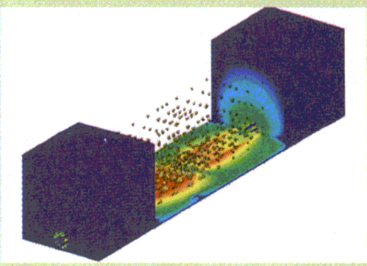


The current distribution and radiation pattern of a handset antenna modeled on IE3D

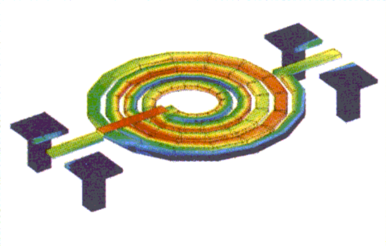


### FIDELITY Examples

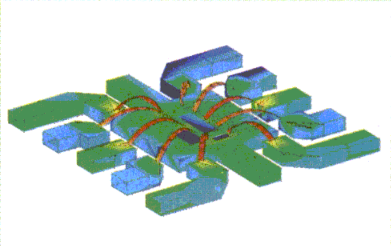
The near field and Poynting vector display on a packaged PCB structure with vias and connectors



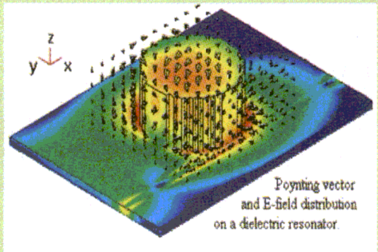
IE3D modeling of a circular spiral inductor with thick traces and vias



IE3D modeling of an IC Packaging with Leads and Wire Bonds



FIDELITY modeling of a cylindrical dielectric resonator and the Poynting vector display




Zeland Software, Inc. provides excellent technical support and services.
Zeland Software, Inc. is also the north American exclusive representative for the LINMIC product from Jansen Microwave GmbH

## Design a PCS base-station amplifier

Amplifiers for personal communications services (PCS) must provide clean output power with high power-added efficiency (PAE). They must also operate reliably, with long mean time to failure (MTTF). An application note (No. 002) from Fujitsu Compound Semiconductor (San Jose, CA), "1930 - 1990 MHz PCS Base Station Applications," describes how to build a 40-W push-pull amplifier using the company's model FLL400IP-2 gallium-arsenide (GaAs) field-effect transistor (FET).

The amplifier provides 40-W output power at 1-dB compression across the full PCS band of 1930 to 1990 MHz. It also meets code-division-multiple-access (CDMA) adjacent-channel-power (ACP) specifications at an output-power level of 8 W. It is easily tuned for optimum third-order-intercept (IP3) or ACP performance, and is rated for a MTTF of more than $10^8$ hours. Since the design employs a push-pull configuration, it provides better PAE than a single-ended design, with simple impedance matching and good rejection of even-order distortion products.

The FLL400IP-2 employs a pair of 20-W gold (Au)-metallized GaAs FETs that are DC and RF connected in a push-pull configuration within a flange package. Impedance-matching networks within the package raise the low impedance of the FETs to the 50 Ω typically seen in PCS base stations. Additional components needed for the amplifier include two Shoshin balun chips and various capacitors and resistors, all to be mounted on a dielectric substrate. The baluns are Shoshin model GSC371-BAL2000 chips while the circuit board is composed of RO3010 material from Rogers Corp. (Chandler, AZ).

The eight-page application note provides input- and output-matching networks, complete with values for capacitors and coupled transmission-line sections. The literature also details input and output circuits designed for optimum CDMA ACP performance, with values for capacitors and resistors and recommended suppliers for each part. Advice is given to properly decouple the gate- and drain-bias networks, dividing the task into several frequency ranges. Sample circuits are presented for the gate and drain bias networks, including forward gate-current protection. The resistor values for this protective circuitry must be carefully chosen, since too large of a resistance will impact the amplifier's PAE while too small of a value will lead to long-term degradation of the transistor and reduced MTTF.

The application note is generously illustrated with schematic diagrams and circuit-layout drawings. It also includes information on simulating the circuit with computer-aided-engineering (CAE) tools and tuning the amplifier under large-signal conditions. Copies of the note can be downloaded from the company's website at: **Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131-1138; (408) 232-9500, FAX: (408) 428-9111, Internet: http://www.fcsi.fujitsu.com.**

**CIRCLE NO. 194 or visit www.mwrf.com**

## Improving IM testing

Measurements of intermodulation distortion (IMD) are fairly standard for components that are used in modern digitally modulated communications systems. These systems require the capability to handle the signals with high peak-to-average power ratios without adding excessive distortion. An application note from Mini-Circuits (Brooklyn, NY), "Stepped Frequency Measurement Improves IM Testing," provides guidance on improving the effectiveness of IMD tests with stepped-frequency techniques.

The three-page note includes a schematic representation of a two-tone IM test system based on separate signal sources. It discusses the importance of providing adequate isolation between the two signal sources (accomplished by careful selection of signal-generator buffer amplifiers). The note also advises that the harmonic and IM levels of the two signal sources be low enough to avoid errors. Methods are then provided for computing acceptable IM levels for the measurement system. Through these calculations, engineers can evaluate the effects of each component in the measurement system, and how to optimize the performance of the test system when selecting those components. Copies of the note can be downloaded from the company's website at: **Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235; (718) 934-4500, FAX: (718) 332-4661, Internet: http://www.minicircuits.com.**

**CIRCLE NO. 195 or visit www.mwrf.com**

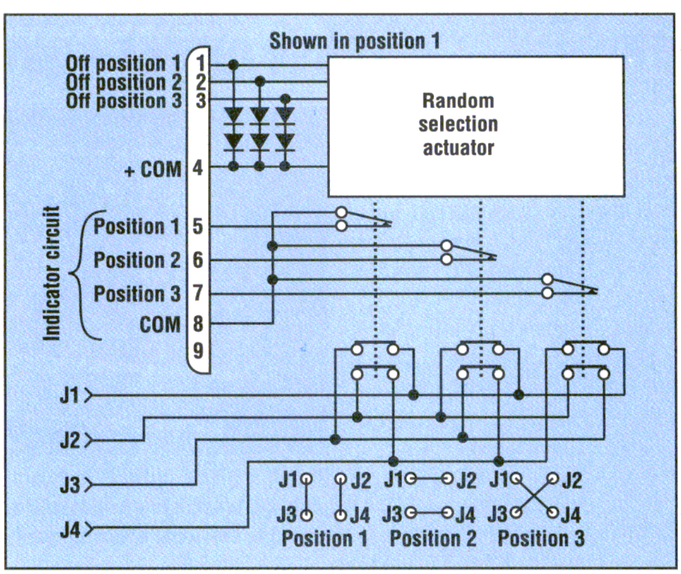
# Space Switch Transfers 140-W Power At S-Band

*Designing a high-power switch with low passive intermodulation distortion required careful design and thoughtful consideration of nonmagnetic materials. The requirement for multipaction-free operation makes the design of RF section and drive mechanism very challenging.*

**JACK BROWNE**
*Publisher/Editor*

**SPACE-BASED applications leave no room for failure. Any component designed for a satellite or the Space Shuttle must operate reliably and with high-performance levels. In the case of an S-band electromechanical T-Switch developed by Dow-Key Microwave Corp. (Ventura, CA), it must operate when needed, over the life of the satellite, and it must handle peak power levels up to 560 W at S-band, and average power levels of 140 W. Quite simply, this is the highest-power T-Switch ever developed for S-band space applications.**

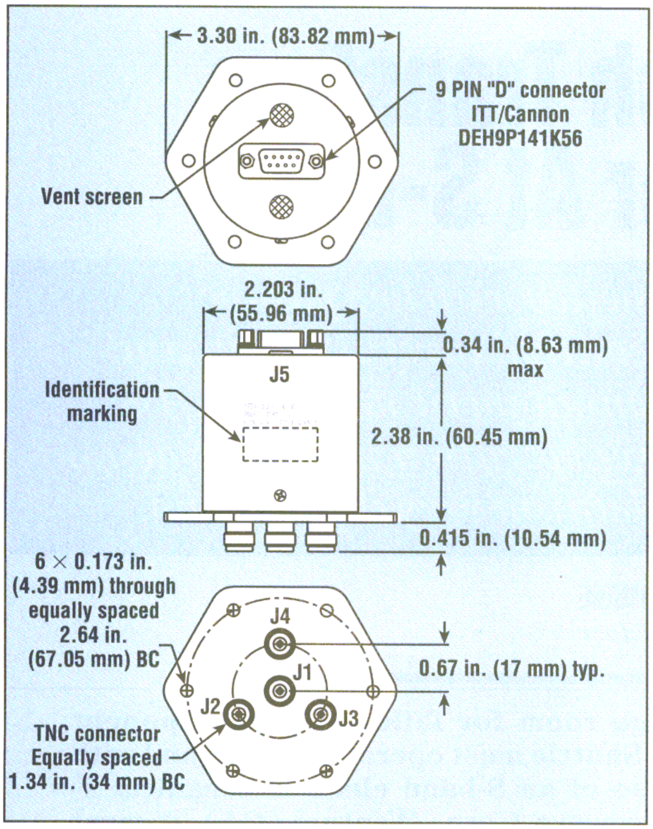
The S-band switch (part number 511HAJ-730322-3) is designed for space-based transmit applications in the frequency band from 2.5 to 4.8 GHz (see sidebar). The function of this T-Switch is to move high-power signals from one port to another. Since the switch has four coaxial ports (J1 through J4), a combination of three positions is possible, with port J1 connecting to ports J4, J3, and J2, respectively, while the two remaining ports are connected (Fig. 1). The switch includes indicator circuitry to alert operators to the switch position selected.




**1. The T-Switch, which can be constructed with long or short reed elements, provides three possible positions where pairs of RF ports are connected in different combinations.**

Since the switch can operate under conditions where power is present in two closed paths, it must be designed to dissipate the heat generated by the average RF power level up to 280 W. Since the switch uses "break-before-make" switch contacts, it is meant for cold-switching operations, where power is removed from the switch before a new signal path is formed. The switch operates with a pulsed latching actuator, requiring 25 ms or less to switch internal signal paths.

Signals are routed through equally spaced TNC connectors (Fig. 2). Switching commands and the power supply are sent to the switch through a 9-pin D-connector. The




**2. The switch is housed in a rugged aluminum (Al) enclosure, with gold (Au)-plated RF cavity and TNC coaxial connectors.**

switch is rated for operating temperatures of −25 to +60°C and storage temperatures of −40 to +85°C. It is designed to operate with nominal voltage of +42 VDC, although it can handle a total operating range of +24 to +43 VDC. It consumes 510-mA maximum switching current at +43 VDC and +25°C.

As can be expected of a high-power component, the T-Switch exhibits extremely low insertion loss. Since insertion loss is magnified at higher power levels, and excessive loss translates into large amounts of power that must be dissipated in the switch, the electromechanical switch was designed for minimal loss. By minimizing the loss and the thermal stress with high insertion loss, the switch can be ensured of meeting its performance limits over the lifetime of a satellite.

The T-Switch is specified for insertion-loss performance of 0.15 dB from 2.5 to 4.0 GHz, and for no more than 0.2 dB from 4.0 to 4.8 GHz (see table), although the typical performance is closer to 0.10 dB across the full operating frequency range.

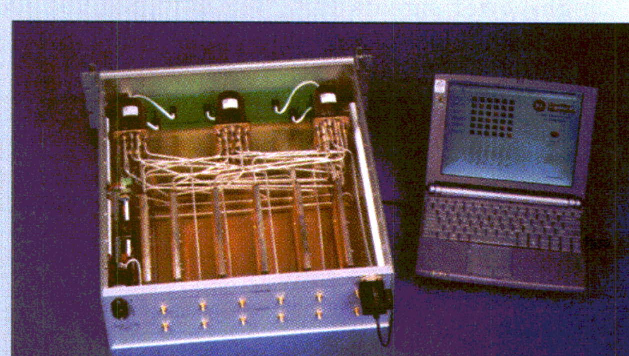
The VSWR and isolation performance levels are similarly impressive. The VSWR is specified for a maximum limit of 1.20:1 across most of the switch's operating frequency range, with a maximum limit of 1.25:1 for frequencies above 4 GHz. The actual performance is well below these maximum limits (Fig. 3), falling below 1.10:1 for much of the operating frequency range, whether the switch is designed with long or short reeds.

The isolation between pairs of connected ports is also good. The maximum specification is 65 dB across most of the switch's operating frequency range, with the minimum limit of 60 dB for frequencies above 4 GHz. The typical performance approaches that maximum limit at only a few frequencies, and is actually better than 85 dB at some points in the band.

The T-Switch features a design lifetime of 100,000 switching operations, which sounds modest when compared to electromechanical switches rated for millions of switch-

## CONTROLLING SWITCHES ON THE CANbus

High-power space-based switches are not Dow-Key Microwave Corp.'s only forte. The firm has also been instrumental in the development of switches and switch matrices based on the controller-area-network (CAN) bus. The CANbus, which is typically used among microcontrollers in embedded networks, can also be used to control electromechanical switch matrices (see figure). The CANbus can link up to 2032 devices on a single network, with communication speeds up to 1 Mb/s. The message-oriented (rather than address-oriented) interface provides simple control functions, and enables easy replacement of switches within a matrix, since each spare switch is provided with a unique identifier code. When a spare switch is installed in place of a defective switch, a self-test program will determine that one of the switches is missing and will identify the spare switch. At that time, the identifier for the spare switch can be changed to that of the switch that it replaces. The firm currently offers single-pole, six-throw (SP6T) and T-Switches for the CANbus, although the interface can be implemented in almost any multiposition switch.



**The CANBus, which is typically used among microcontrollers in embedded networks, can also be used to control electromechanical switch matrices.**

ing operations in automatic-test-equipment (ATE) installations. (The lifetime of 100,000 switching operations is specified mostly for reliability purposes. The useful operating life of a typical Space Qualified Switch is less than a few thousand cycles and, in many cases, the switch is never switched or is switched once.) But the T-Switch must provide a reliable switching operation over a much longer time period—15 years in a typical satellite—than expected of million-operation switches designed for ATE applications. Considerable computer simulations and analyses were performed during the development of the switch to study, for example, the long-term effects of cold-welded joints and local mechanical stresses on long-term performance and operating lifetime. The result is a switch that will meet its specified performance levels even 15 years after it has been installed and sitting in an orbiting satellite. Several patents have been applied for the technology used in the switch.

**THE TYPICAL ISOLATION BETWEEN PAIRS OF CONNECTED PORTS APPROACHES THE MAXIMUM LIMIT AT ONLY A FEW FREQUENCIES, AND IS ACTUALLY BETTER THAN 85 dB AT SOME POINTS IN THE BAND.**

The T-Switch weighs only 190 g. It is tested with 140-W signals passed through two signal paths at +85°C to verify if there is sufficient margin for conditions caused by the in-phase addition of multicarrier signals. Since the switch is vented, the design of the RF section has to provide a minimum 6-dB margin to ensure the multipaction-free operation at 140 W CW. To verify that the unit operation is free of multipaction, the multipaction test is performed in a vacuum with peak power levels up to 560 W.

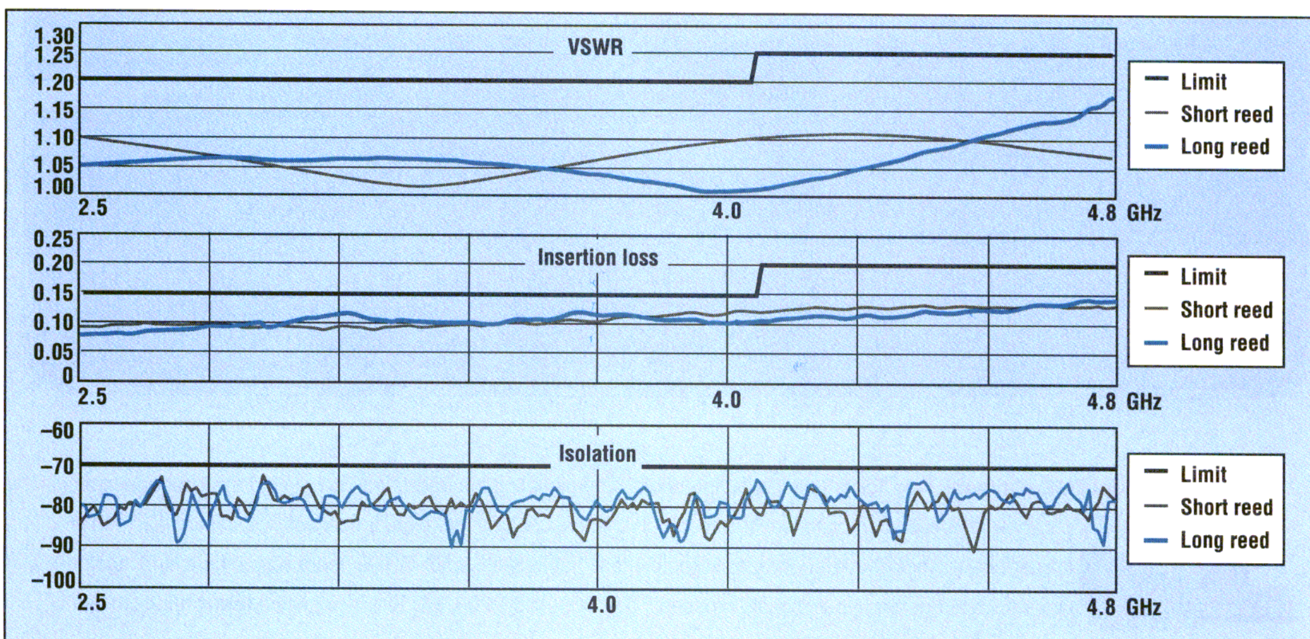
Since it must operate with multicarrier signals, the T-Switch is designed for low passive-intermodulation (PIM) distortion. To achieve low distortion, materials were carefully considered, and mainly nonmagnetic materials were used. For example, isolation barriers had to be created without nickel (Ni) plating, while forming a barrier against corrosion.

The T-Switch is housed in an aluminum (Al) enclosure. The RF cavity, where good power-handling capability is critical, is also Al, with gold (Au) plating (per MIL-G-45204 requirements) to enhance conductivity and minimize signal losses. The switch's connector shell is also Au plated. It is a switch designed to provide high-performance levels without compromise, and without failure for the 15-year typical lifetime of a space-based mission. **Dow-Key Microwave Corp., 4822 McGrath St., Ventura, CA 93003-5641; (805) 650-0260, FAX: (805) 650-1734, Internet: http://www.dowkey.com.**

**CIRCLE NO. 51 or visit www.mwrf.com**

### The S-band switch at a glance

| | | |
|---|---|---|
| Frequency range | 2.5 to 4.0 GHz | 4.0 to 4.8 GHz |
| Maximum VSWR | 1.20:1 | 1.25:1 |
| Maximum insertion loss | 0.15 dB | 0.20 dB |
| Minimum isolation | 70 dB | 70 dB |
| RF power-handling capability | 140 W CW | 140 W CW |




**3. The typical VSWR, insertion loss, and isolation performance levels easily exceed the maximum specified limits across the operating frequency range of 2.5 to 4.8 GHz.**

# Meters And Sensors Simplify Tests On Complex Waveforms

*These meters and sensors feature flexible triggering functions to read peak as well as average power levels for signal bandwidths as wide as 5 MHz.*

**JACK BROWNE**
*Publisher/Editor*

**PEAK power measurements are necessary for gauging the true envelope power of a communications signal with complex digital modulation. The new EPM-P series of single- and dual-channel power meters and the E9320 series of power sensors are designed for these measurements, with generous 5-MHz sensor video bandwidth and meter triggering capabilities for signals from 9 kHz through 110 GHz.**

These power meters and sensors (Fig. 1) are ideally suited for power measurements on complex wireless and wire-line communications systems. The power meters can be specified with one channel (model E4416A) or with two channels (model E4417A). The meters are designed for research and production-line testing; the capability of making up to 1000 corrected readings per second will benefit most manufacturing test environments. Both the single- and dual-channel meters perform continuous sampling of input signals at 20 MSamples/s.

1. The EPM-P series of peak power meters and E9320 series of power sensors provide peak and average power-measurement capability at frequencies to 18 GHz.

The power meters and sensors combine for a total power-measurement range of –70 to +44 dBm. Measured results can be shown in absolute units of watts (W) or decibels referenced to 1 mW (dBm), or in relative units of percent or decibels (dB). The display resolution is selectable with resolutions of 1.0, 0.1, 0.01, and 0.001 dB in logarithmic mode and with one to four significant digits in linear mode.

The sensors determine the frequency and measurement range. The power meters essentially display the measured results and provide control over measurements and sensor triggering functions. Of the six sensors currently in the E9320 series, three provide a frequency range of 50 MHz to 6 GHz while three cover 50 MHz to 18 GHz (see table). The power meter also allows the user to select the sensor's video bandwidth at its high, medium, or low setting.

Models E9321A and E9325A, which operate at frequencies from 50 MHz to 6 GHz and 50 MHz to 18 GHz, respectively, feature video-bandwidth settings of 30, 100, and 300 kHz. Models E9322A and E9326A, which operate at frequencies from 50 MHz to 6 GHz and 50 MHz to 18 GHz, respectively, have video-bandwidth settings of 100 kHz, 300 kHz, and 1.5 MHz. Models E9323A and E9327A, which operate at frequencies from 50 MHz to 6 GHz and 50 MHz to 18 GHz, respectively, offer the widest video-bandwidth settings, at 300 kHz, 1.5 MHz, and 5 MHz.

For each sensor, the maximum setting is the default setting. The maximum setting (widest video bandwidth) makes it possible to capture level information about fast-changing or modulated signals, but with some compromise in dynamic range compared to narrower video-bandwidth settings. By having the choice of three video-bandwidth settings for each sensor, operators can optimize their measurements for dynamic range.

The EPM-P meters, which are com-

**The E9320 series sensors at a glance**

| Model | Frequency range | Power- measurement range Average-only mode | Normal mode | Maximum video BW |
|---|---|---|---|---|
| E9321A | 50 MHz to 6 GHz | -65 to +20 dBm | -55 to +20 dBm | 300 kHz. |
| E9322A | 50 MHz to 6 GHz | -60 to +20 dBm | -45 to +20 dBm | 1.5 MHz |
| E9323A | 50 MHz to 6 GHz | -60 to +20 dBm | -40 to +20 dBm | 5.0 MHz |
| E9325A | 50 MHz to 18 GHz | -65 to +20 dBm | -55 to +20 dBm | 300 kHz |
| E9326A | 50 MHz to 18 GHz | -60 to +20 dBm | -45 to +20 dBm | 1.5 MHz |
| E9327A | 50 MHz to 18 GHz | -60 to +20 dBm | -40 to +20 dBm | 5.0 MHz |

patible with the company's 8480 and E-series power sensors, can be used with the E9320 series sensors for peak, average, and time-gated power measurements. Triggering can be performed continuously, on programmed levels, in response to external transistor-transistor-logic (TTL) signals, and according to general-purpose-interface-bus (GPIB) commands.

Triggering is performed with a time resolution of 50 ns. The trigger delay range is ±1 s. Delays can be set as fine as 50 ns for delays of less than ±50 ms; the delay resolution for longer delays is 200 ns. The internal trigger in each power meter can be set across a range of –20 to +20 dBm with 1-dB level accuracy and 0.1-dB resolution.

The E9320 series of power sensors can be thought of as two sensors in one package. The sensors combine stable, low-level power measurements with modulation measurement capability. Each sensor employs two independent measurement paths (Fig. 2).

Together, the EPM-P series power meters and E9320 power sensors combine for a powerful measurement solution for CW and complex-modulated signals. P&A: $3500 (single-channel EPM-P meter), $5500 (dual-channel EPM-P meter), $1500 to $3200 (E9320 series sensors); stock. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052.**

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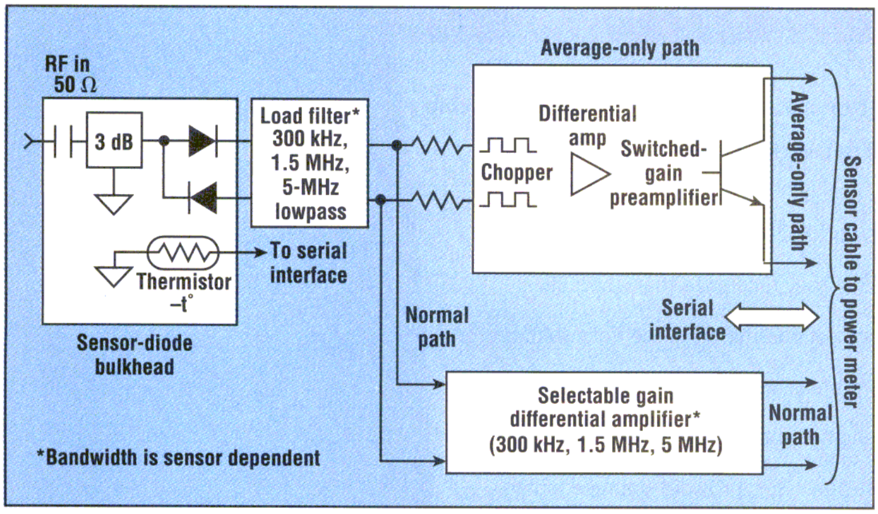


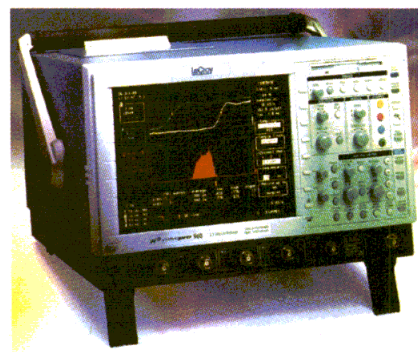

**2. The E9320 series power sensors have two measurement paths—normal for modulated signals and time-gated measurements, and average only for average power measurements.**

# Scopes Offer High Bandwidth And Time-Saving Features

***A series of oscilloscopes helps engineers troubleshoot complex communications signals and high-speed digital designs.***

**DON KELLER**
*Senior Editor*

**HIGH-BANDWIDTH communications devices that process complex signals and high-speed digital processors that operate at gigahertz clock speeds, play an increasing role in today's electronic systems. To design, test, and troubleshoot these devices, engineers need test equipment that can operate at ever-higher frequencies. In response to this need, LeCroy Corp. (Chestnut Ridge, NY) has introduced a series of high-bandwidth oscilloscopes that can cope with these designs. The WavePro 900 series consists of three new oscilloscopes with bandwidths of 500, 1000, and 2000 MHz (see table). They offer a number of features that facilitate the testing and troubleshooting of high-speed devices and systems.**



**Model 960 is the Flagship of the WavePro series of oscilloscopes.**

The flagship of the WavePro series oscilloscopes is the 2-GHz model 960 (see figure), followed by the 1-GHz model 950, and the 500-MHz model 940. Aside from the differences in bandwidths, the three oscilloscopes are similar to each other. All three models have four input channels, a 10.40-in. (26.42-cm), 640 × 480-pixel, thin-film-transistor (TFT) color liquid-crystal display (LCD), and a set of buttons known as WavePilot that provides the operator with quick, one-touch access to commonly used features.

The standard (non-optional) amount of acquisition memory for all models is 250,000 points (250 kpts), and the standard single-shot sampling rate is 4 GSamples/s on each of the four channels. When using only two channels, the standard acquisition-memory length per channel doubles to 500 kpts and the single-shot sampling rate also doubles to 8 GSamples/s. When using only one channel, the acquisition memory doubles again to 1 Mpts and the single-shot sampling rate for models 960 and 950 doubles again to 16 GSamples/s. However, the single-shot sampling rate for model 940 remains at 8 GSamples/s. Three memory options—M, L, and VL—are available for all models to increase the standard acquisition memory by a factor of 4, 16, or 32, respectively. A fourth memory option, XL, is available only for model 960 to increase its standard acquisition memory by a factor of 64.

The timebase on all models can be adjusted from 100 ps/div to 1000 s/div, and has a clock accuracy of 10 PPM or better. For real-time sampling, the interpolator resolution is 5 ps. Vertical resolution is 8 b, or up to 11 b in the enhanced-resolution (ERES) mode. Vertical sensitivity is adjustable from 1 mV to +1 VDC at 50 Ω, and from 1 mV to +2 VDC at 1 MΩ.

Hardware options include an internal printer, Ethernet, and personal-computer (PC) card slots. Numerous software options are available, including several signal-analysis packages. **LeCroy Corp., 700 Chestnut Ridge Rd., Chestnut Ridge, NY 10977-6499; (914) 425-2000, Internet: http://www.lecroy.com.**

**CIRCLE NO. 53 or visit www.mwrf.com**

**The main statistics for the three WavePro oscilloscopes**

| Model | Bandwidth | Acquisition points per channel (1/2/4 Ch.) | Single-shot sample rate (1/2/4 Ch.) |
|---|---|---|---|
| 960 | 2 GHz | 1 M/500k/250k | 16/8/4 GSamples/s |
| 950 | 1 GHz | 1 M/500k/250k | 16/8/4 GSamples/s |
| 940 | 500 MHz | 1 M/500k/250k | 8/8/4 GSamples/s |

# GaAs MMICs Power High-Frequency Communications Systems

***These GaAs MMIC devices and modules provide the output power and gain required for line-of-sight and satellite-communications systems.***

**JACK BROWNE**
*Publisher/Editor*

**SEMICONDUCTOR process technologies emerge almost monthly with bold promises of better performance than existing techniques. But gallium-arsenide (GaAs) technologies have been installed in communications systems for more than 20 years, yielding reliable performance under demanding conditions. Two of the latest examples of high-performance GaAs devices come from Fujitsu Compound Semiconductor. They are aimed at frequencies above 13 GHz and are based on monolithic-microwave-integrated-circuit (MMIC) technology. Model FMM5048GJ is targeted at 14-GHz satellite communications while model FMM5805X targets systems from 17 to 20 GHz.**

The lower-frequency device, model FMM5048GJ is a MMIC-based amplifier module. The module, which is designed for very-small-aperture-terminal (VSAT) applications from 13.75 to 14.50 GHz, contains two stages of amplification and impedance-matching circuitry for 50-Ω systems. It delivers 4-W (+36-dBm) typical output power at 1-dB compression (Fig. 1), with a typical gain of 26 dB. The worst-case gain flatness is specified as ±1.5 dB.

The FMM5048GJ exhibits typical input VSWR of 2.0:1, with typical output VSWR of 3.0:1. It is housed in a hermetic package measuring 12 × 15 × 3.5 mm.

The higher-frequency GaAs MMIC, model FMM5805X, is a three-stage amplifier supplied in chip form. Similar to the FMM5048GJ, the MMIC chip is impedance matched at its input and output ports for 50-Ω systems. The amp delivers +31-dBm output power at 1-dB compression from 17.5 to 20 GHz, with a typical gain of 21 dB at that output power (Fig. 2). The power-added efficiency (PAE) is high at 30 percent.

The FMM5805X exhibits input and output return losses of 12 and 8 dB, respectively. The GaAs MMIC draws typical drain current of 650 mA from a +6-VDC supply. **Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131-1138; (408) 232-9500, FAX: (408) 428-9111, Internet: http://www.fcsi.fujitsu.com.**

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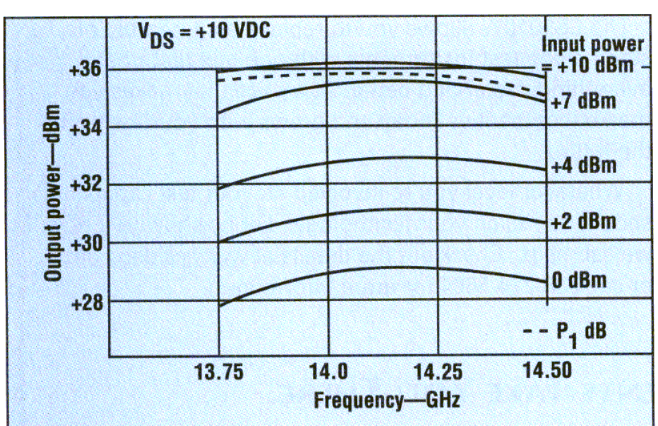
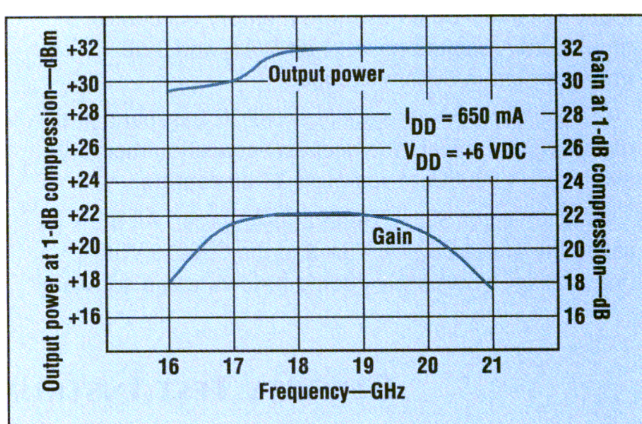



**1. Model FMM5048GJ is a two-stage amplifier module that delivers +36-dBm output power at 1-dB compression and 26-dB typical gain from 13.75 to 14.50 GHz.**




**2. Model FMM5805X is a GaAs PHEMT chip that provides +31-dBm output power at 1-dB compression and 21-dB typical gain from 17.5 to 20 GHz.**

# Frequency Synthesizer Spans 2 To 18 GHz In 10-MHz Steps

***This universal frequency synthesizer is ideal for ground-based and airborne military-system upgrades and new applications.***

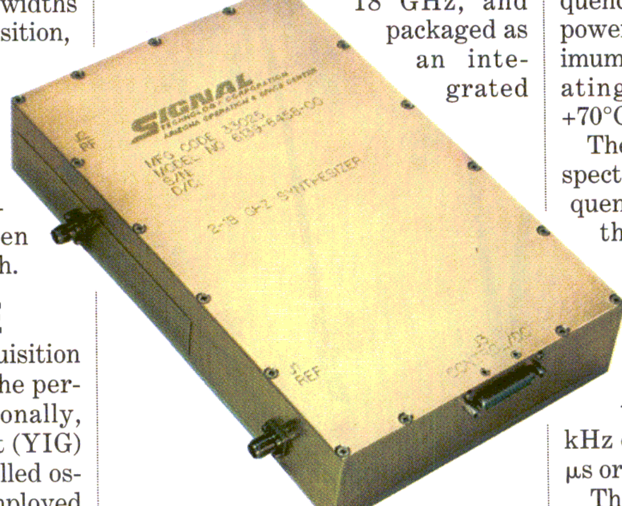
**ALAN "PETE" CONRAD**
*Special Projects Editor*

**STABLE local-oscillator (LO) frequencies are vital to the effectiveness of radar-warning receivers (RWRs) emitter-location systems. By incorporating digital control, Signal Technology Corp. (Chandler, AZ) has developed a highly accurate generic frequency synthesizer that is ideally suited for new or upgraded military-electronics systems. The compact synthesizer module is capable of tuning from 2 to 18 GHz in 10-MHz steps.**

RWRs and precision emitter-location (interferometer-based) platforms (interferometer) typically operate with intermediate-frequency (IF) bandwidths ranging from 10 to 500 MHz, depending upon the IF center frequency. Wider bandwidths are often used for signal acquisition, while narrower bandwidths serve systems where detailed signal analysis is required. For example, signals acquired in a 500-MHz bandwidth exhibit a 26.9-dB improvement in sensitivity when viewed in a 10-MHz bandwidth.

## LO PERFORMANCE

One key to good signal acquisition for any receiving system is the performance of the LO. Traditionally, banks of yttrium-iron-garnet (YIG) oscillators and voltage-controlled oscillators (VCOs) have been employed as the LO sources in military-grade receivers (Rxs). But without phase locking, such sources are limited in frequency accuracy and often subject to post-tuning drift. Additionally, YIG sources are limited in tuning speed. With demands for improved performance, designers developed digitally controlled sources based on these source technologies, typically operating in harmonically related bands of 2 to 4 GHz, 4 to 8 GHz, 8 to 12 GHz, and 12 to 18 GHz, and packaged as an integrated module. Due to the increasing need for reductions in size and operating power, the engineers at Signal Technology Corp. developed the model 6139-6458-00 compact digitally controlled frequency-synthesizer module that serves as a standard replacement for older systems, or as a building-block component for new systems.



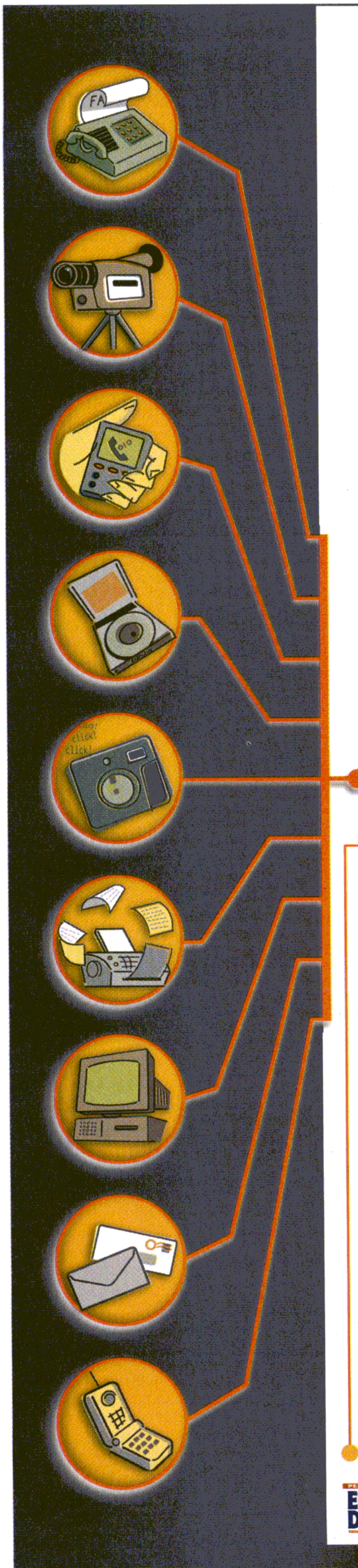

**1. Model 6139-6458-00 is a high-performance frequency synthesizer that operates from 2 to 18 GHz in 10-MHz steps with 11-b digital transistor-transistor-logic (TTL) control.**

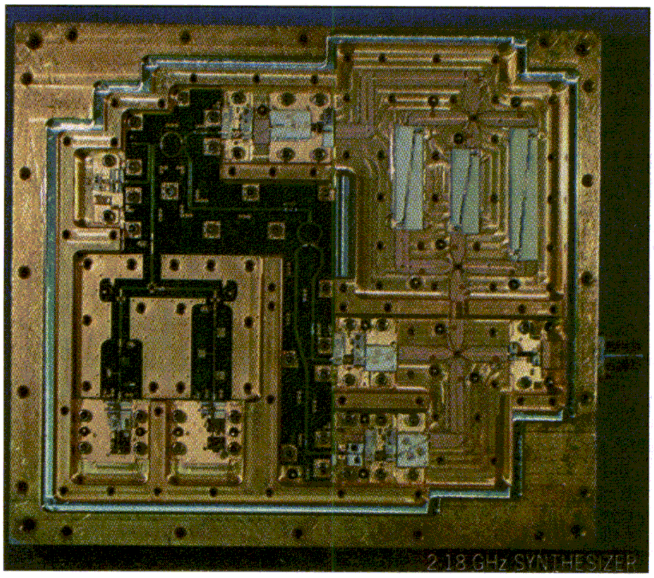
## TTL CONTROL

The model 6139-6458-00 (Fig. 1) operates from 2 to 18 GHz in 10-MHz steps with 11-b transistor-transistor-logic (TTL) compatible tuning control. The synthesizer delivers at least +10-dBm output power across its frequency range, with maximum output power of +16 dBm. It maintains maximum output VSWR of 2.0:1 for operating temperatures from –40 to +70°C.

The harmonic performance is a respectable (considering the broad frequency range) –15 dBc anywhere in the operating frequency range. Subharmonic signals over the operational-frequency range are –35 dBc maximum, with spurious responses of –60 dBc maximum. The frequency synthesizer can switch within ± 100 kHz of a new frequency in only 100 μs or less.

The digitally controlled frequency synthesizer operates from a 10-MHz external-frequency reference (crystal-controlled oscillator) that can supply power levels between –3 to +3 dBm into a 50-Ω load. The frequency stability of the synthesizer is determined by the frequency stability of






**2. The compact frequency module achieves low SSB phase noise from 2 to 18 GHz when operating from supplies of +15 and −15 VDC.**

this reference source.

The synthesizer achieves single-sideband (SSB) phase noise of –68 dBc/Hz offset 100 Hz from a 9-GHz carrier and –62-dBc/Hz offset 100 Hz from an 18-GHz carrier. The phase noise improves to –73 and –65 dBc/Hz, respectively, for offsets of 1 through 100 kHz from carrier frequencies of 9 and 18 GHz. The SSB phase noise is –100-dBc/Hz offset 1 MHz from a 9-GHz carrier and –95-dBc/Hz offset 1 MHz from an 18-GHz carrier.

The frequency synthesizer provides the performance levels needed to update and improve electronic-warfare (EW) and military-communications systems. It delivers the proper trade-off in resolution necessary for these systems, where signals must be detected and identifiedwith high accuracy but also woth fast tuning speed. The phase noise is low enough at the close-in offset frequencies so as to not obscure the detection (by interferometers) of closely spaced signals that are similar in amplitude, and sufficiently low at offsets far from the carrier (1 MHz) to support high-performance levels in radar systems. The basic short- and long-term stability of the frequency synthesizer can be enhanced by the selection of the 10-MHz reference source, with an oven-controlled crystal oscillator (OCXO) typically providing a good compromise among cost, size, and stability.

The frequency-synthesizer module (Fig. 2) measures 7.5 × 4.50 × 1.35 in. (19.05 × 11.43 × 3.43 cm) with an SMA female RF-output connector and DC and tuning control through a 31-pin microminiature D connector. The synthesizer's DC-power requirements are 1.5 A at +15 VDC and 0.5 A at –15 VDC, with maximum peak-to-peak supply ripple of 25 mV from 10 kHz to 1 MHz. **Signal Technology Corp., 340 N Roosevelt Ave., Chandler, AZ 85226; (408) 940-1655, FAX: (408) 961-6297, Internet: http://www.sigtech.com.**

**CIRCLE NO. 55 or visit www.mwrf.com**

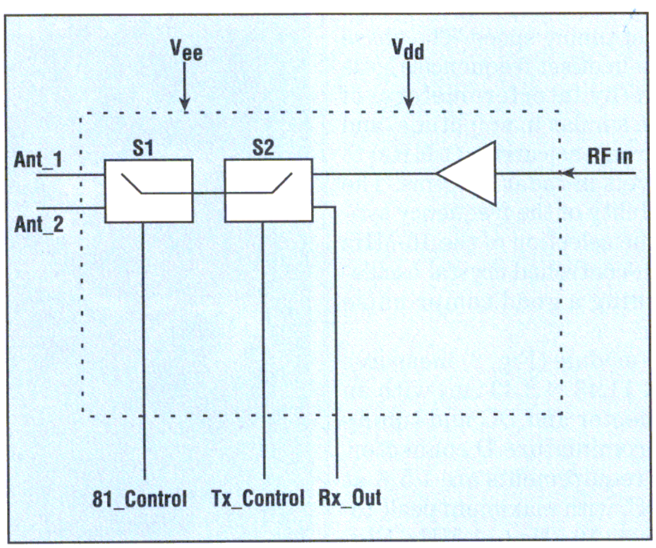
# Module Combines Switches And Amp For 5-GHz WLANs

## *This high-gain 5-GHz PA is aided by a pair of onboard GaAs switches for flexibility in implementing high-data-rate WLAN systems.*

**JACK BROWNE**
*Publisher/Editor*

**WIRELESS data communications is a market waiting to explode. Original equipment manufacturers (OEMs) have been hesitant to develop wireless-local-area-network (WLAN) products prior to the industry's firm adoption of the IEEE's 802.11 standard. Now, with the addition of the wideband, high-data-rate version of the standard (802.11a), data throughput/capacity is no longer an issue for developers of WLAN equipment. The RTPA-5250 monolithic-microwave-integrated-circuit (MMIC) power-amplifier (PA)/switch from Raytheon Microelectronics (Andover, MA) is one of the first building-block components that aims at the latest version of the WLAN standard, with performance that supports data rates to 54 Mb/s at 5 GHz.**

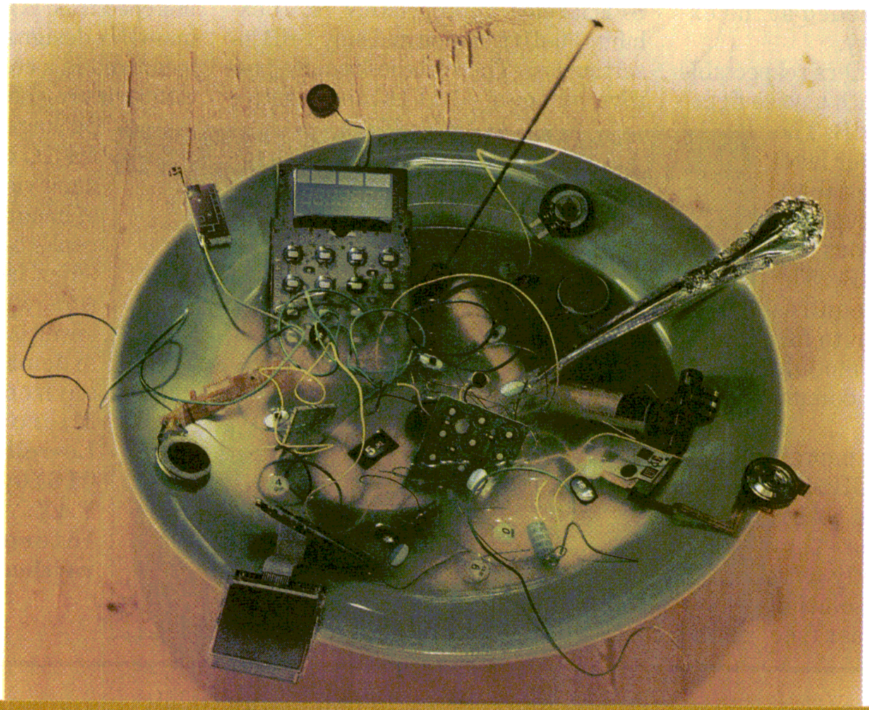
The RTPA-5250 is aimed at WLANs and other applications in the relatively unoccupied 5-GHz Unlicensed National Information Infrastructure (UNII) band. (The RTPA-5250 is actually suitable for use in all communications segments of the 5-GHz band, including 5.15 to 5.25 GHz, 5.25 to 5.35 GHz, and 5.725 to 5.825 GHz.) This is in stark contrast to standard WLAN architectures (IEEE 802.11b) which operate in the relatively crowded 2.4-GHz band at data rates up to 11 Mb/s. The 2.4-GHz band is also home to such appliances as microwave ovens and cordless telephones, and will soon provide the bandwidth for a growing multitude of Bluetooth- and HomeRF-enabled devices and appliances. But, while the higher-frequency band offers the potential for faster data rates with potentially less interfering sources, product development in that band has been viewed as expensive compared to the 2.4-GHz band, where components are less costly. The RTPA-5250 has been introduced to change the perception of high cost associated with the 5-GHz band. It is an economic PA that also incorporates a pair of high-speed switches which can be used for antenna-diversity or switching between transmit and receive functions.




**The RTPA-5250 5-GHz amplifier/switch module integrated impedance-matching circuitry with a monolithic PA and a pair of high-speed PHEMT switches.**

The monolithic amplifier/switch module is designed to be a low-cost component in a cost-effective 5-GHz WLAN system. It is fabricated with a proven process, the company's high-performance 0.5-µm pseudomorphic-high-electron-mobility-transistor (PHEMT) technology, to achieve somewhat higher gain at 5 GHz than possible with conventional gallium-arsenide (GaAs) metal-epitaxial-semiconductor-field-effect-transistor (MESFET) technology. In fact, the PA circuitry achieves typical small-signal gain of 38 dB from 5150 to 5825 MHz, with only minor gain variations with temperature and only –0.5 to +0.5 dB at temperatures from 0 to +85°C. The gain linearity is ruler flat, remaining within a window of –0.5 to +1.5 dB for output-power levels from 0 to +16.5 dBm.

The amplifier generates enough linear output power (typically +16.5 dBm) for most in-building WLAN applications, allowing it to be connected directly to a



WLAN antenna or pair of diversity antennas. The amplifier can also produce approximately a quarter-watt (+23.5-dBm) output power at 1-dB compression, when greater output-power levels are needed at the expense of some linearity.

The RTPA-5250 offers respectable receive capability, with a noise figure (NF) of typically 6 dB. The maximum noise-power level is –128 dBm/Hz (see table). Harmonic levels are typically –30 dBc, while spurious levels add to only –70 dBc. The RTPA-5250's PA requires only 1 μs to ramp on to full output power. The module consumes 230 mA maximum current at +3.3 VDC.

The RTPA-5250 is impressive as a low-cost 5-GHz PA, but its circuitry also boasts two in-line switches that change states quickly with moderate to high isolation and low insertion loss (see figure). The switch closest to the RTPA-5250 package input/output (I/O) connections, labeled S1, exhibits 17-dB isolation with only 0.5-dB typical insertion loss. The switch following S1 is labeled, appropriately, S2. It exhibits 30-dB isolation and only 1-dB insertion loss. The switches require positive voltage of +3.3 VDC and negative voltage of –6 VDC. The switches offer switching time of typically 25 ns.

The RF inputs and outputs of the RTPA-5250 5-GHz WLAN amplifier/switch module are impedance matched to 50 Ω for ease of installation in 5-GHz designs. The MMIC module is housed in a compact, low-loss low-temperature-cofired-ceramic (LTCC) package. The low loss of the MMIC chip and the package itself yields a module capable of handling signals with 7-dB peak-to-average power ratio (crest factor) without distortion, which translates into lower bit-error-rates (BERs) in demanding WLAN office environments. **Raytheon Microelectronics, 352 Lowell St., Andover, MA 01810; (978) 684-8538, FAX: (978) 684-8596, Internet: http://www.raytheon.com.**

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**The RTPA-5250 at a glance**

| | |
|---|---|
| **Amplifier** | |
| Frequency range | 5150 to 5825 MHz |
| Output power (linear) | +16.5 dBm |
| Output power at 1-dB comp. | +23.5 dBm |
| Small-signal gain | 38 dB |
| Gain linearity | –0.5/+1.5 dB |
| Noise figure | 6 dB |
| Noise power level | –128 dBm/Hz |
| Harmonic levels | –30 dBc |
| Input VSWR | 1.20:1 |
| Output VSWR | 4.80:1 |
| Power supply | 230 mA at +3.3 VDC |
| **Switches** | |
| Switch S1 insertion loss | 0.5 dB (typ.) |
| Switch S1 isolation | 17 dB (typ.) |
| Switch S1 switching time | 25 ns (typ.) |
| Switch S2 insertion loss | 1.0 dB (typ.) |
| Switch S2 isolation | 30 dB (typ.) |
| Switch S2 switching time | 25 ns (typ.) |

# Precision SMA Pads Operate From DC To 18 GHz

***These coaxial attenuators provide excellent amplitude accuracy and 5-W power-handling capability at reasonable prices.***

**JACK BROWNE**
*Publisher/Editor*

**AMPLITUDE control is critical in many systems. It is often important for minimizing system costs, requiring a trade-off between the performance and price of amplitude-control components. Fortunately, a line of precision SMA microwave attenuators from Mini-Circuits (Brooklyn, NY) sacrifices little performance when compared to higher-priced components, with a variety of attenuation values for use from DC to 18 GHz.**

The BW-SXW5 coaxial attenuator line (where X is the desired attenuation value) is available with nominal fixed attenuation values of 1 to 10 dB in 1-dB steps, and 12, 15, 20, 30, and 40 dB. The attenuators are supplied with male and female SMA connectors and feature an amplitude accuracy of ±0.40 dB for attenuation values of 1 through 6 dB, measured at +25°C.

As expected, the amplitude accuracy decreases somewhat for higher-valued attenuators, with accuracy of ±0.60 dB for nominal-attenuation values of 7 to 20 dB and accuracy of ±0.85 dB for the 30- and 40-dB attenuators, all measured at +25°C. These accuracy specifications include variations in power and frequency through 12.4 GHz. The specifications are degraded by an additional 0.5 dB above 12.4 GHz due to power and frequency variations.

The performance of these attenuators is particularly beneficial to test applications. Because the attenuators exhibit minimal variations of attenuation across their operating-frequency range and with variations in applied power, they help to minimize the number of recalibration procedures needed with a particular precision-measurement system. The attenuators present low mismatch to the associated test equipment, with typical VSWR performance of 1.15:1, simplifying the task of de-embedding the unwanted electrical contributions of the attenuators from the measurement setup. The attenuators are rated for operating temperatures from −55 to +100°C.


**The 5-W attenuators are available with male- or female-SMA connectors (right) or with type-N connectors (left) for applications from DC to 18 GHz.**

## ROBUST PADS

At +25°C ambient temperature, the bidirectional 50-Ω attenuators are rated for continuous-wave (CW) power levels of 5 W. At that temperature, the attenuators can also handle 125-W peak power with a 5-μs pulse at a pulse-repetition frequency (PRF) of 100 Hz. This enables the attenuators to be used in test systems operating with high-power pulsed signals or with burst communications signals. As ambient temperatures increase, the power-handling rating derates in a linear fashion to 2 W CW at +100°C. Standard attenuators are available with male- or female-SMA connectors (see figure). For a slight premium, the 5-W attenuators are also available with type-N coaxial connectors. The SMA attenuators are useful for a wide range of applications. With their high accuracy at +25°C, they are suitable for amplitude control in manual and automatic test stations, and for improving the impedance match of a device under test (DUT) connected to the test equipment.

P&A: $44.95 (with SMA connectors) and $54.95 [with type-N connectors] (1 to 49 qty.); stock. **Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235; (718) 934-4500, FAX: (718) 332-4661, Internet: http://www.minicircuits.com.**

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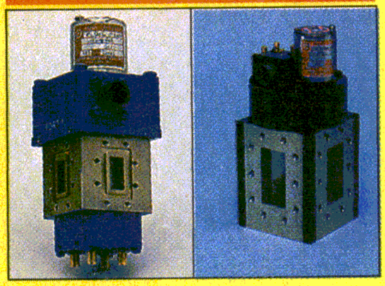

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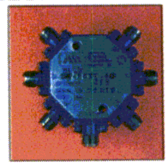

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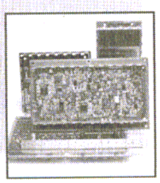

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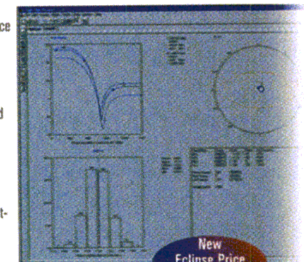

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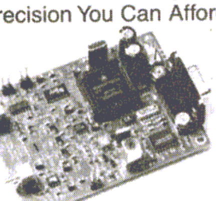

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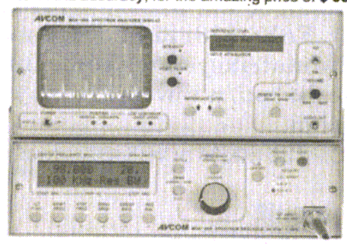

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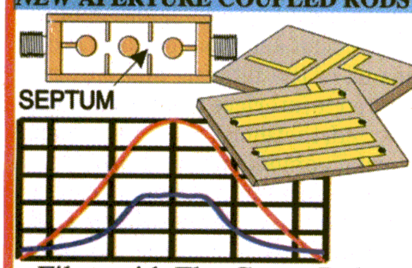


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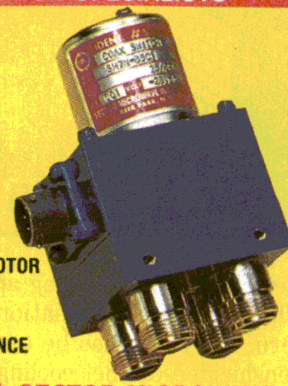

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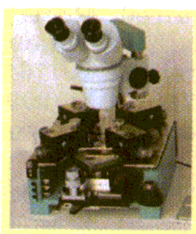

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### Antenna provides broad bandwidth

The Whisper 850 model low-profile directional antenna boasts excellent gain with broad bandwidth and low VSWR. Designed for in-building and outdoor cellular, microcell, and wireless-local-loop (WLL) applications, the antenna provides 8.5-dBi gain and a maximum VSWR of 1.5:1 across the 824-to- 896-MHz band. The antenna comes equipped with a mounting bracket for flush mounting to any surface, and can be enhanced with optional articulated mounting brackets and pole/wall mounts for outdoor and base-station applications. **Xertex Technologies, Inc., 452 Burbank St., Broomfield, CO 80020; (303) 635-2000, FAX: (303) 635-2003, e-mail: jenkins@xertex.com, Internet: http://www.xertex.com.**

CIRCLE NO. 62 or visit www.mwrf.com

### Tuner is capable of high mismatch

The model MT984A is a broadband 50-GHz automated tuner capable of presenting a high mismatch (15:1) over the 8-to-50-GHz frequency range. The unit boasts repeatability of better than 40:1 minimum over the entire frequency range and matching range. Designed to operate with the MT980 series of automated tuner systems (ATS), the unit performs noise, power, intermodulation (IM), and adjacent-channel measurements used for device characterization and circuit design. The MT984A tuner was conceived for measurements where a very high mismatch is required. **Maury Microwave Corp., 2900 Inland Empire Blvd., Ontario, CA 91764-4804; (909) 987-4715, FAX: (909) 987-1112, Internet: http://www.maurymw.com.**

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### Relay module targets DC to 800 MHz

The 7000 series coaxial relay modules maintain reliable switching over a wide range of severe emotional conditions. Applications include RF and intermediate-frequency (IF) switching, baseband video, pulse switching, low-level and low-noise switching, DS-1, DS-3, E-1, T-1, 1553B, and emitter-coupled-logic (ECL) signal switching, and balanced audio switching. The unit is available with two to 24 positions and with low electromagnetic interference (EMI) and insertion loss. The package measures 3.25 × 0.625 × 1.44 in. (8.26 × 1.59 × 3.66 cm). **Matrix Systems, 5177 Douglas Fir Rd., Calabasas, CA 91302; (818) 222-2301, FAX: (818) 222-2304, e-mail: tech@matrixsystems.com, Internet: http://www.matrixsystems.com.**

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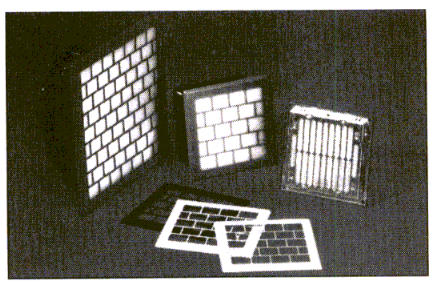
### Fans suit cabinet cooling

The type 3149 patented superflat electromagnetic-interference/RF-interference (EMI/RFI)-shielded filter fans are designed for electronic and



electrical cabinet cooling applications where EMI/RFI radiation is a concern. They can also be used for any nonshielded cabinet cooling needs. A front-grid projection of 0.20 in. (0.51 cm) is provided while unimpeded air flow is 95 cfm (160 $m^3/h$). **Haewa Corp., 4357 B Park Dr., Norcross, GA 30093; (770) 921-3272, FAX: (770) 921-2896, e-mail: haewa@bellsouth.net, Internet: http://www.haewacorp.com.**

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### Detector handles input range

The model STZ-500 is a detector for telecommunications and wireless applications. It handles an input Rengo of 0 to +20 dBm over a temperature range of −55 to +125°C. The detector features a frequency range of 0.01 to 18.0 GHz, with a positive output polarity and a video load ≥1 MΩ. **Signal Technology Corp., 975 Benecia Ave., Sunnyvale, CA 94086; (408) 730-6300, (408) 773-3744, FAX: (408) 773-0254, Internet: http://www.sigtech.com.**

CIRCLE NO. 66 or visit www.mwrf.com

### Digital megohmmeter measures insulation

The model 1026 is a compact, economical digital megohmmeter. The unit measures insulation at +250, +500, and +1000 VDC to 4000 Ω. Resistance is measured to 400 Ω. The 1026 features a continuity test with an audible beeper for resistances under 100 Ω, and a +600-VDC AC/DC range for safety-voltage check. **AEMC Instruments, 99 Chauncy St., Boston, MA 02111; (800) 343-1391, (617) 451-0227, FAX: (617) 423-2952, Internet: http://www.aemc.com.**

CIRCLE NO. 67 or visit www.mwrf.com

### Fiber converter transmits to 7 km

The model PI-005 is a parallel (270-Mb/s) to fiber (1300-nm) converter capable of transmitting over a range of up to 7 km. The PI-005 is a modular solution, enabling easy addition of differing output and input options. It contains all of the filtering and buffering required to interface to external equipment and, therefore, requires minimal external components and provides a complete drop-in solution to conversion. **Faraday Technology Ltd.; Croft Rd. Industrial Estate, Newcastle Staffordshire, ST5 0Qz England; +44 (0) 1782 661501, +44 (0) 1782 630101, e-mail: sales@faradaytech.co.uk, Internet: http://www.faradaytech.co.uk.**

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### Electroplated wire is resistant to oxidation

Electroplated wire is more resistant to heat, corrosion, and oxidation. A company's capabilities include the electroplating of high-speed wire and ribbon with diameters of eight to 10 thousandths to 30 thousandths of an inch. Electroplating materials include nickel (Ni), copper (Cu), silver (Ag), gold (Au), platinum (Pt), aluminum (Al), tin (Sn), tin lead (SnPb), cadmium (Cd), rhodium (Rh), palladium (Pd), and palladium nickel (PdNi) in thicknesses ranging from 3 to 300 millionths of an inch. **California Fine Wire; (805) 489-5144, FAX: (805) 489-5352, Internet: http://www.calfinewire. com.**

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### Couplings suit precision positioning

The SMC line of zero-backlash, flexible shaft couplings is low-inertia, low-windup, precision constant-velocity universal joints. The couplers target coupling to encoders and resolvers in precision positioning applications. Since the flexible element is a bellows, the couplings absorb angular and parallel misalignments in combination with axial movements, while transmitting motion and torque. **Servometer Corp., 501 Little Falls Rd., Cedar Grove, NJ 07009-1291; (973) 785-4630, (973) 267-0756, Internet: http://www.servometer.com.**

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### Epoxy bonds to circuit boards

The TRA-BOND 2153 is a thermally conductive epoxy designed for bonding transistors, diodes, resistors, integrated circuits (ICs), and other heat-sensitive components to circuit boards. This two-part adhesive is available in custom BIPAX packaging, enabling the user to administer small quantities of the epoxy on an as-needed basis, or to use the product for mass production. The adhesive develops durable, high-impact bonds at room temperature, which improves heat transfer while maintaining electrical insulation. The epoxy readily bonds to itself, metals, silica, alumina, steatite, sapphire, and other ceramics, as well as glass, plastics, and other materials. **Tra-Con, Inc., 45 Wiggins Ave., Bedford, MA 01730; (800) TRA-CON1, FAX: (781) 275-9249, Internet: http://www.tra-con.com.**

CIRCLE NO. 71 or visit www.mwrf.com

### Toroid spans 4.7 to 500 µH

The series LLST is a surface-mountable toroid that has 23 standard inductance values spanning 4.7 to 500 µH with rated currents up to 2.6 A-DC. Temperature rating is −40 to +125°C. The low-loss toroidal core provides a self-shielding design plus a small volume size in relation to inductance value. Superior coplanarity is obtained by using a fixed two-terminal base assembly. The package size is 12.07 × 10.67 × 7.37 mm. **API Delevan, 270 Quaker Rd., East Aurora, NY 14052; (716) 652-3600, FAX: (716) 652-4814, e-mail: apisales@delevan.com, Internet: http://www.delevan.com.**

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### Connectors designed for FCC part 15.203

Federal Communications Commission (FCC) part 15.203 requires a non-standard coaxial connector interface for application compliance. To achieve this, more than 65 reverse-polarity (gender), metric-threaded, and left-hand-threaded connectors and adapters are available. These compliant connectors cannot be mated with standard interface connectors. Metric threads are offered on TNC and SMC male crimp, clamp, and bulkhead connectors. **RF Connectors, 7610 Miramar Rd., San Diego, CA 92126-4202; (800) 233-1728, (858) 549-6340, e-mail: rfi@rfindustries.com, Internet: http://www.rfindustries.com.**

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### Thermometer measures −40 to 932°F

The Fluke infrared (IR) thermometer features laser-spot aiming and was designed for the rigors of field and factory applications. It features a shock-absorbing holster and backlit dual liquid-crystal-display (LCD) display, and can measure from −40 to 932°F (−4 to 500°C). The tool allows users to measure temperature on hot, electrically live, or rotating equipment accurately from a safe distance. **Jensen Tools, Inc., 7815 S. 46th St., Phoenix, AZ 85044-5399; (800) 426-1194, (602) 453-3169, FAX: (800) 366-9662, (602) 438-1690, Internet: http://www.jensentools.com.**

CIRCLE NO. 74 or visit www.mwrf.com

### IFM monitors 2 to 18 GHz

Model A15-MH122 operates from 2 to 18 GHz with a nominal noise figure of 5 dB over a dynamic range of −63 to +10 dBm. The effective input at 0-dB signal-to-noise ratio (SNR) is −65 dBm with a 99-percent probability of detection with a 2-dB RF input SNR of −63 dBm. Frequency resolution with 12 b is 4 MHz with a root-mean-square (RMS) frequency accuracy of 5 MHz. Shadow time is 100 ns with an internal throughput time of 350 ns. Signal-processing capability includes pulse, CW, pulse on CW, and pulse on pulse. **Aikon, Inc., 1914 Trade Zone Blvd., San Jose, CA 95131; (408) 432-8039, FAX: (408) 432-1089, Internet: http://www.akoninc.com.**

CIRCLE NO. 75 or visit www.mwrf.com

### Switch controls 18 to 40 GHz

Model F9014 is a single-pole, single-throw (SPST), TTL-controlled switch that spans 18 to 40 GHz with a maximum VSWR of 2.2:1. Minimum isolation is 75 dB from 18 to 26.5 GHz and 70 dB from 26.5 to 40 GHz. Maximum insertion loss is 2.8 dB from 18 to 26.5 GHz and 3.5 dB from 26.5 to 40 GHz. Rise and fall times are 10 ns maximum with a maximum switching speed of 20 ns at a maximum repetition rate of 20 MHz. **General Microwave, Inc., 425 Smith St., Farmingdale, NY 11735; (631) 630-2000, FAX: (631) 630-2066, Internet: http://www.generalmicrowave.com.**

CIRCLE NO. 76 or visit www.mwrf.com

### RF IC offers 1.9-dB noise figure

The model MGA-52543 is an RF integrated-circuit (RF IC) amplifier that operates with a single bias of +5 VDC at 53 mA. Used in base-station receivers (Rxs) as part of the low-noise amplifier (LNA), this RF IC can improve receive performance with its typical 1.9-dB noise figure, 14.2-dB associated gain, and +17.5-dBm third-order intercept point, all at 2 GHz. The unit can also be used in amplifier and driver applications in base-station, wireless-local-loop, wireless local-area-network (WLAN), and other wireless applications. **Agilent Technologies, 5301 Stevens Creek Blvd., Santa Clara, CA 95052; (800) 452-4844, Internet: http://www.agilent.com.**

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## Fiber converter runs to 7 km

The model PI-005 is a parallel (270-Mb/s)-to-fiber (1300-nm) converter which is capable of transmitting over a range of up to 7 km. The unit contains all of the filtering and buffering to interface to external equipment and, therefore, requires minimal external components and provides a complete drop-in solution to conversion. **Faraday Technology Ltd., Croft Rd. Industrial Estate, Newcastle Staffordshire, ST5 0QZ England; +44 (0) 1782 661501, FAX: +44 (0) 1782 630101, e-mail: sales@faradaytech.co.uk, Internet: http://www.faradaytech.co.uk.**

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## Pulse modulator operates to 20 GHz

The Model SWM-DJV-1DT-2ATT is a non-reflective/absorptive single-pole, single-throw (SPST) pulse modulator which operates from DC to 20 GHz with a single power supply, an ultra-high speed of 13 ns on and 15 ns off. The unit's video transient is <20 mV peak-to-peak at 300-MHz bandwidth and 4 mV peak-to-peak at 20-MHz bandwidth. Insertion loss is less than 1.75 dB at 40 MHz and less than 4 dB at 20 GHz. VSWR is 2.0:1 typical. Supply voltage is −5 VDC at 18 mA. Package size is 1.0 × 1.0 × 0.5 in. (2.54 × 2.54 × 1.27 cm. **American Microwave Corp., 7311 G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, e-mail: amcpmi@aol.com, Internet: http://www.amwave.com.**

CIRCLE NO. 79 or visit www.mwrf.com

## Magnetron reaches 915 MHz

The model IMG-915-60KB is a fixed-tuned, electromagnet-focused 60-kW CW magnetron operating to 915 MHz. The unit features ceramic-metal construction, liquid and air cooling, and rugged construction to meet the electrical performance criteria under the temperature and shock/vibration conditions encountered in industrial applications. The unit is plug compatible with Western 30, 50, and 60 kW 915-MHz magnetrons. **Istok Microwave, 8200 S. Memorial Pkwy., Huntsville, AL 35802; (256) 882-1344, FAX: (256) 880-8077, Internet: http://www.istok.com.**

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## Antenna covers 2.5 GHz

The model PCW24-01518-BFL is a high-gain 18-dBi panel antenna for the 2.4-GHz industrial-scientific-medical (ISM) band. The antenna covers frequencies from 2.4 to 2.5 GHz. The unit provides 15 deg. of horizontal beamwidth and 15 deg. of vertical beamwidth. The package measures 15.75 × 15.75 × 4.0 in. (40.01 ×40.01 × 10.16 cm). **HD Communications Corp., Systems Group, 1 Comac Loop, Ronkonkoma, NY 11779; (631) 558-9661, FAX: (631) 588-0487.**

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## Dividers target 1 to 26.5 GHz

A broadband matched-line directional divider boasts superior performance to broadband Wilkinson-designed dividers. Model 6010265 (two way) operates over the entire frequency band of 1 to 26.5 GHz. With a maximum VSWR of 1.60:1, maximum insertion loss is 1.6 dB while minimum isolation is 19 dB. Amplitude tracking is <0.3 dB and phase tracking is <10 deg. **Krytar, 1292 Anvilwood Ct., Sunnyvale, CA 94089; (408) 734-5999, FAX: (408) 734-3017, Internet: http://www.krytar.com.**

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## Terminators suit bus terminations

The DN5 and DNA series of Schottky diode networks is suitable for relieving overshoot and undershoot, ringing, and bus reflections in high-speed input-output (I/O) applications. The DN5 and DNA networks are designed as 18-channel terminators housed in single packages. The supply voltage (VDD) for each series is −0.3 to +7.0 VDC, with a channel-clamp current of ±50 mA. The package-power rating at 70°C is 1 W and the operating temperature is 0 to +70°C. **KOA Speer Electronics, Inc., Bolivar Dr., P.O. Box 547, Bradford, PA 16701; (814) 362-5536, FAX: (814) 362-8883, Internet: http://www.koaspeer.com.**

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## Charger works from −40 to +85°C

The MAX1772 is a highly integrated, multichemistry battery charger which simplifies the construction of accurate and efficient chargers. The unit is specified for the extended temperature range of −40 to +85°C. The charger uses analog inputs to control charge current and voltage, and can be user programmed or hardwired. The MAX1772 can charge two to four lithium-ion (Li+) series cells. **Maxim Integrated Products, 120 San Gabriel Dr., Sunnyvale, CA 94086; (408) 737-7600, FAX: (408) 737-7194.**

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## Switch spans DC to 6 GHz

The model AS196-307 is a GaAs integrated-circuit (IC) high-isolation, nonreflective switch with driver. The unit operates over the DC-to-6-GHz band, and is suitable for Global System for Mobile Communications (GSM), digital communications services (DCS), and personal-communications-services (PCS) base-station synthesizer switching with 55-to-60-dB isolation from 0.5 to 2.5 GHz. **Alpha Industries, 20 Sylvan Rd., Woburn, MA 01801; (781) 935-5150 ext. 0, FAX: (617) 824-4579, e-mail: sales@alphaind.com, Internet: http://www.alphaind.com.**

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## Jumper cables range to 2.4 GHz

A series of jumper cables operates to 2.4 GHz. VSWR ratings are 1.08:1 at 1 GHz and 1.12:1 at 2 GHz (with type-N connectors). The cables feature tight-bending radii and halogen-free flame retardant jackets per DIN VDE 0819-107, soldered inner and outer conductors, and injection-molded boots for mechanical stability. Key applications include cellular and personal-communications-services (PCS) base stations and general indoor and outdoor wireless systems. **Spinner North America, 6611 Bay Circle, Suite 100, Norcross, GA 30071; (770) 263-6326, FAX: (770) 263-6329, e-mail: sales@spinnerna.com, Internet: http://www.spinnerna.com.**

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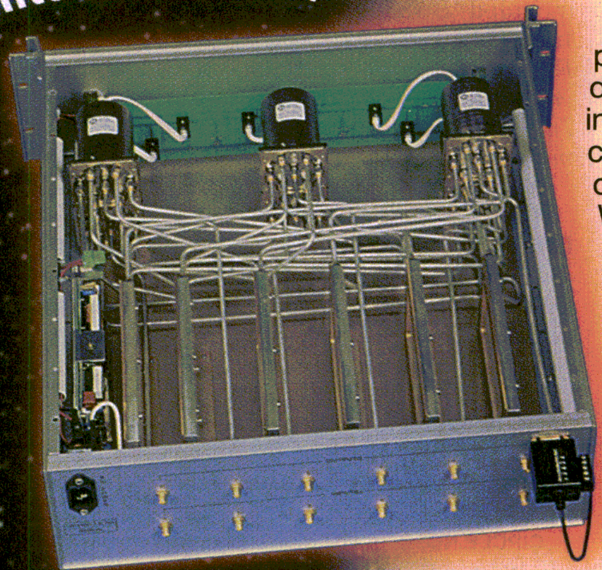

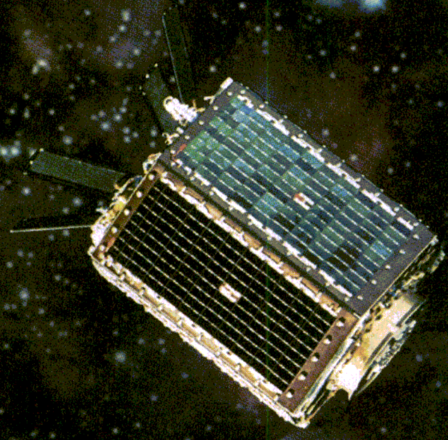

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### Fiber optics

A 13-page catalog highlights test equipment and tools for installing fiber-optic networks of communications systems. Fiber-optic test kits, smart kits, light kits, mini kits, smart meters, and smart sources are presented. Mini fiber-optic instruments, low-cost meters and kits, fiber-optic connectors, visible fiber tracers, connector adapters, bare fiber adapters, reference test cables, and attenuators are offered. Product descriptions are included. **Fotec;** *(781) 396-6155, FAX: (781) 396-6395, e-mail: info@fotec.com, Internet: http://www.fotec.com.*

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### Packaging solutions

A six-page brochure describes plated copper on thick film (PCTF)™. Information on PCTF technology, the advantage of PCTF technology, and solutions are discussed. The company's manufacturing capabilities are also presented. A list of special features for PCTF is included. **REMTEC, Inc.;** *(781) 762-9191, FAX: (781) 762-9777, e-mail: sales@remtec.com, Internet: http://www.remtec.com.*

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### Analog devices

A 40-page magazine details technical information on digital signal processors (DSPs), along with analog and logic devices. Digital control systems (DCS), high-performance and power-efficient DSPs, third-party hardware and software, DSP development tools, video/graphics digitizers, and amplifiers are offered. Audio power amplifiers (PAs), data transmission, and power management are also featured. **Texas Instruments, Inc.;** *Internet: http://www.ti.com/sc/techinnovations4.*

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### Controller ICs

An eight-page brochure provides information on controller integrated circuits (ICs) and other products. High-level-data-link (HDLC) controllers, bridge chips, dual potentiometers, smart-battery monitors, a dual microcontroller, 5-b programmable pulse-width modulators (PWMs), and low-voltage serial timekeeping chips are discussed. Block diagrams are included. Product features are also provided. **Dallas Semiconductor;** *(972) 371-6183, FAX: (972) 371-4370, Internet: http://www.dalsemi.com.*

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### Thermoplastic equipment

An eight-page brochure highlights plastics assembly products. Ultrasonic, vibration, spin, hot-plate welding, and heat-staking equipment are detailed. Product specifications are included. **Sonics & Materials, Inc.;** *(800) 745-1105, (203) 270-4600, FAX: (203) 270-4610, e-mail: info@sonicsandmaterials.com, Internet: http://www.sonicsandmaterials.com.*

**CIRCLE NO. 91 or visit www.mwrf.com**

### Encapsulation systems

A two-page application selector guide details potting and encapsulation systems. Viscosity, setup times, cure schedules, service temperature ranges, volume resistivity, and application recommendations for one- and two-component systems are described. **Master Bond, Inc.;** *(201) 343-8983, FAX: (201) 343-2132.*

**CIRCLE NO. 92 or visit www.mwrf.com**

### Frequency sources

Microwave frequency sources, voltage-controlled crystal oscillators (VCXOs), phase-locked multipliers, ultra-low-noise phased-locked cleanup loops, low-noise frequency multipliers, and ultra-low-noise carrier signature source products are detailed in an eight-page foldout brochure. Gaussian noise calibration standards and low-noise measurement instruments are discussed. Specifications include frequency range, frequency stability, along with temperature range. **Techtrol Cyclonetics, Inc.;** *(717) 774-2746, FAX: (717) 774-6799, e-mail: techtrol@paonline.com, Internet: http://www.tci-ant.com.*

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### Micromachined integration

The benefits and applications of micromachined integration technology are discussed in a brochure. A micromachined isolation integrated-circuit (IC)-based approach to digital isolation and micromachined relay technology is described. A roadmap of future products, as well as additional new technologies being developed for integrated components are included. **Analog Devices;** *(800) ANALOGD, Internet: http://www.analog.com.*

**CIRCLE NO. 94 or visit www.mwrf.com**

### TCM operation

Tandem-connection monitoring (TCM) is examined in a 17-page application note (No. 73). Error checking in synchronous-digital-hierarchy (SDH) networks, information on how TCM works, basic operation of TCM, and information on the N2 byte is provided. Operations, standards, and test solutions are included. **Wavetek Wandel Goltermann Eningen GmbH & Co.;** *+49 7121 86-1616, FAX: +49 7121 86-1333, e-mail: info@wwgsolutions.com, Internet: http://www.wwgsolutions.com.*

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### Frequency signals

Product solutions for applications which require precision frequency and timing signals are described in a 64-page catalog. Quartz instruments and components, space-qualified and military components, as well as distribution and synthesizer modules are specified. A section of application notes is also included. **DATUM;** *(800) 544-0233, (978) 927-8220, FAX: (978) 927-4099, e-mail: fts@datum.com, Internet: http://www.datum.com.*

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### Crystal oscillators

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*(continued from p. 61)*
grated drain mixer was designed, fabricated, and measured to verify the theoretical analysis.

The mixer was designed to accept an RF input of 27.4 GHz and an LO input of 33.4 GHz, to produce an IF of 5 GHz. The RF and LO signals are injected at the gate and drain, respectively. To obtain better conversion gain, the input network and output network are selected so that the circuit can provide RF conjugate matching at the gate, and IF- and LO-conjugate matching at the drain. The bias voltages are $V_{DS}$ = +0.5 VDC and $V_{GS}$ = −0.2 VDC. A software program was used to optimize the circuit for maximum conversion gain. With slight modifications, the program can also be used for designing other HEMT mixers and upconverters. Figure 3 shows the microstrip-circuit topology.

**A DRAIN MIXER USES THE INHERENT ISOLATION CHARACTERISTICS OF THE THREE-PORT-DEVICE TERMINALS AND OFFERS GOOD CONVERSION GAIN.**

The circuit was fabricated on RT-duriod/5880 substrate (h = 0.254mm, $\epsilon_r$ = 2.22). The RF- and LO-input ports are waveguide-microstrip transfer circuits, and the IF output port is a microstrip-coaxial-line transfer circuit. Figure 4 shows the experimental results. The mixer exhibits a conversion gain of 2 dB with a wide frequency band and a maximum conversion gain of 4 dB at RF = 27.4 GHz and LO = 33.4 GHz/ 10 dBm. ••

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5. Stephen A. Maas, "Microwave Mixer," Aretch House, 1986.

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Timing instrumentation is the subject of a 60-page catalog. Time and frequency processors, time-code processors, time-distribution modules, RF pulse analyzers, time-code readers and generators, programmable time systems, switch and distribution units, and time displays are offered. Application notes and specifications are provided. **DATUM;** *(800) 348-0648, (408) 578-4161, FAX: (408) 574-4950, e-mail: salessj@datum.com, Internet: http://www.datum.com.*

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## Test equipment

Component testers, digital multimeters (DMMs), frequency counters, function generators, oscilloscopes, power supplies, instrument/shipping cases, analyzers, as well as Category 5 testers are offered in a 400-page catalog. Continuity testers, bit-error-rate (BER) testers, transmission test sets, optical time-domain reflectometers (OTDRs), and power meters are also presented. Bookmark tabs are provided for referencing purposes. Specifications are included. **Specialized Products Co.;** *(800) 866-5353, FAX: (800) 234-8286, Internet: http://www.specialized.net.*

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A 116-page catalog features equipment for the telecommunications, electrical, and global communications industries. The catalog offers cable, telephone, and electrical test equipment; services aids; and a line of test sets. **Jensen Tools, Inc.;** *(800) 426-1194, (602) 453-3169, FAX: (800) 366-9662, (602) 438-1690, e-mail: jensen@stanleyworks.com, Internet: http://www.jensentel.com.*

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## Cable assemblies

Cable assemblies and wire harnesses are introduced in a brochure. The company's engineering and electrical testing services are discussed. Application information is provided. **Aerospace Systems;** *e-mail: aero@nortechsys.com.*

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## RF connectors

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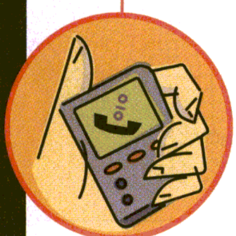

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*(continued from p. 122)*
ter. The center bandwidth frequency $f_c$ is determined as:

$$f_c = \sqrt{470 \cdot 860} = 635\ MHz \quad (31b)$$

For this operating frequency, assume that the manufacturer's stated value of input impedance for each transistor is:

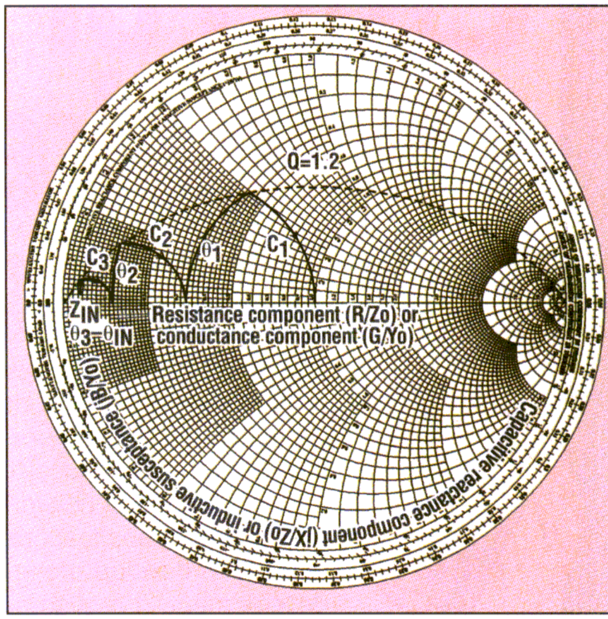
$$Z_{in} = (1.7 \ + j1.3)\ \Omega \qquad (31c)$$

In this case, $Z_{in}$ is expressed as a series combination of input resistance and inductive reactance. To realize the required bandwidth, low-Q matching circuits should be used to reduce in-band amplitude ripple and improve input VSWR. To achieve a 3-dB bandwidth, the value of the quality factor must be less than Q = 635/(860−470) = 1.63. Since the device input quality factor is smaller (i.e., $Q_{in}$ = 1.3/1.7 = 0.76), it is possible to cover the entire frequency range using a multistage matching circuit. It is very convenient to design the input-matching circuit as well as the output-matching circuit by using simple L transformers in the form of series transmission lines and parallel capacitances with a constant value of Q for the balanced portion of each device. Then, the two matching circuits are combined by inserting capacitances whose values are reduced two times between the two series transmission lines. Although the requirement of a constant Q is not strictly necessary, it provides a convenient guide for the analytical calculation of the matching circuit parameters and the Smith chart.

To match the series-input inductive impedance to the standard 50-Ω input-source impedance, it is necessary to use three L transformers, as shown in Fig. 11. At the center frequency of 635 MHz, the input inductance is equal to approximately 0.3 nH. To take this inductance into account, it is necessary to subtract the appropriate value of electrical length $\theta_{in}$ from total electrical length $\theta_3$. Due to the short size of this transmission line, a value of $\theta_{in}$ can be easily calculated in accordance with:

$$\theta_{in} \cong X_{in} / Z_0 \ = \ \omega L_{in} / Z_0 \quad (32)$$

In this case, the input resistance $R_{in}$ can be assumed to be constant.




**12. This Smith chart shows the parameters used in the schematic in Fig. 11.**

According to Eq. 22, there are two simple ways to provide matching using equal quality factors of L transformers. One way is to use the same characteristic impedance values for all transmission lines. The other is to use the same electrical lengths for all transmission lines. Consider the first approach, which also allows direct use of the Smith chart, and choose the value of the characteristic impedance $Z_0 = Z_{01} = Z_{02} = Z_{03}$ = 50 Ω. It is convenient to express the ratio of input and output resistances by:

$$\frac{R_1}{R_2} \ = \ \frac{R_2}{R_3} \ = \ \frac{R_3}{R_{in}} \qquad (33a)$$

This provides the values of $R_2$ = 16.2 Ω and $R_3$ = 5.25 Ω for $R_{source} = R_1$ = 50 Ω and $R_{in}$ = 1.7 Ω. Then, the values of electrical lengths are determined from Fig. 5a as:

$$\theta_1 = 30°,\ \theta_2 = 7.5°,\ \theta_3 = 2.4°.$$

Actually, it is enough to determine an electrical length $\theta_1$ of the first L transformer and to calculate the quality factor Q from Eq. 22. In this case, the remaining two electrical lengths can be directly obtained from Eq. 21. As a result, the quality factor of each L transformer is equal to a value of Q = 1.2. The values of the parallel capacitances are as follows:

$$C_1 = 6\ pF,\ C_2 = 19\ pF,\ C_3 = 57\ pF$$

In the case of a constant Q, one can simplify the design of the matching circuit significantly by using a standard Smith chart. After calculating the value of Q, one must plot a constant-Q circle on the Smith chart. Figure 12 shows an input-matching-circuit design using a standard Smith chart with a constant-Q circle.

Another approach assumes the same values of electrical lengths

$$\theta \ = \ \theta_1 \ = \ \theta_2 = \ \theta_3 \qquad (33b)$$

and calculates the characteristic impedances of transmission lines from Eq. 22 at equal ratios of the input and output resistances according to Eq. 33a. This approach is more convenient in practical designs, because, when using the transmission lines with standard characteristic-impedance $Z_0$ = 50 Ω, the electrical length of the transmission line adjacent to the active device's input terminal is too short. In this case, it is advisable to set the characteristic impedance of the first series transmission line to $Z_{01}$ = 50 Ω. Then, a value of θ should be evaluated directly from the nomograph shown in Fig. 5a. Subsequent calculation of Q from Eq. 22 provides θ = 30° and Q = 1.2. The characteristic impedances of the remaining two transmission lines are calculated easily from Eqs. 21 or 22:

$$Z_{02} = 15.7\ \Omega,\ Z_{03} = 5.1\ \Omega$$

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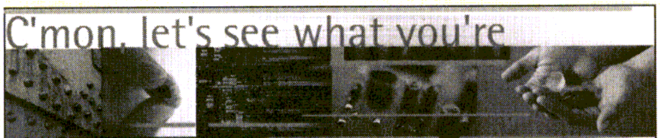

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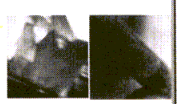

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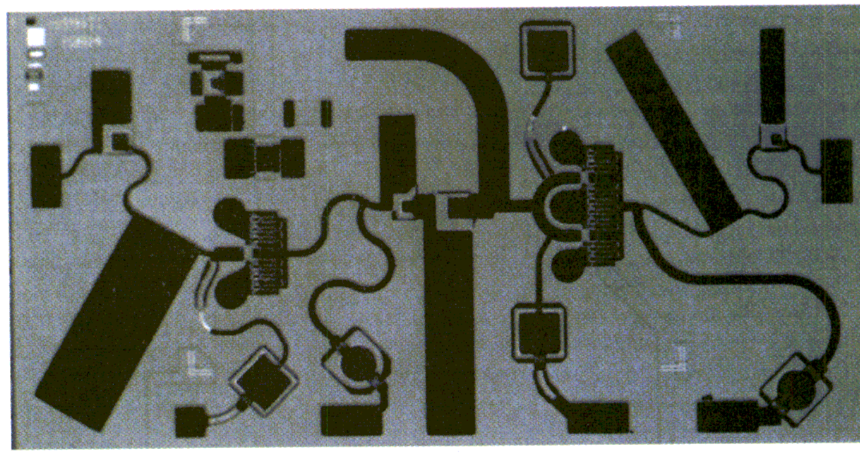

*Nearly 14 years ago, Texas Instruments (Dallas, TX) unveiled a line of commercial gallium-arsenide (GaAs) monolithic-microwave-integrated-circuit (MMIC) amplifiers for use from 6 to 18 GHz. The line included a power-amplifier (PA) MMIC capable of almost 0.5-W full-range output power with 7-dB gain and 14-percent power-added efficiency (PAE).*

# Microwaves & RF November Editorial Preview

## Issue Theme: Amplifiers and Oscillators

### News

Advances in semiconductors pave the way for higher-level advances, in circuits and systems. It is no wonder then that the IEEE's International Electron Devices Meeting (IEDM) is annually one of the better-attended technical conferences in the world. The 2000 IEDM offers details on transistors that operate with single-electron flow, as well as the fastest silicon (Si) transistors ever fabricated, with frequencies reaching 180 GHz. For a glimpse into the future, don't miss this exclusive report.

### Design Features

November leads off with guidelines for constructing a 70-W S-band amplifier for multiuser multipoint distribution system (MMDS) and wireless-data applications. Additional articles cover the simulation of a phase-locked-loop (PLL) frequency source that incorporates a direct digital synthesizer (DDS) and the design of a low-phase-noise frequency synthesizer for mobile communications.

### Product Technology

November marks the launch of a new line of spectrum analyzers from a leading test-instrument supplier. Additional Product Technology articles will report on the use of complementary metal-oxide semiconductor (CMOS) for 5-GHz wireless local-area-network (WLAN) integrated circuits (ICs) and examine a flexible vector modulator that can transform any analog signal generator into a digital signal generator.