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Dual-Use Technology Issue

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Measuring military electronics needs

DESIGN FEATURE
Phase noise in digital communications systems

PRODUCT TECHNOLOGY
Kit solves EMI shielding problems

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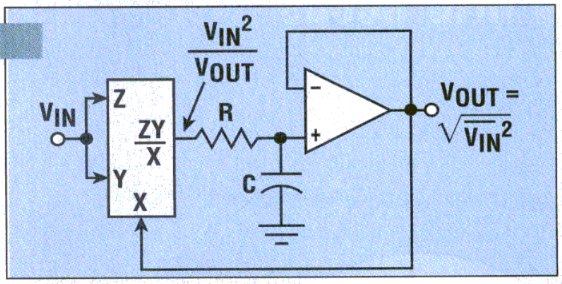
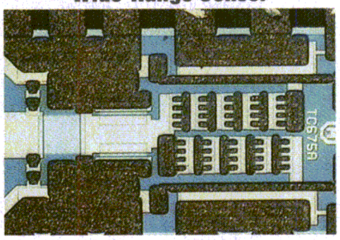


DEPARTMENTS

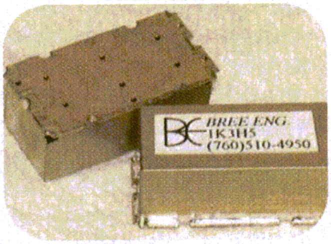
NEWS

SPECIAL REPORT



DESIGN FEATURES

PRODUCT TECHNOLOGY



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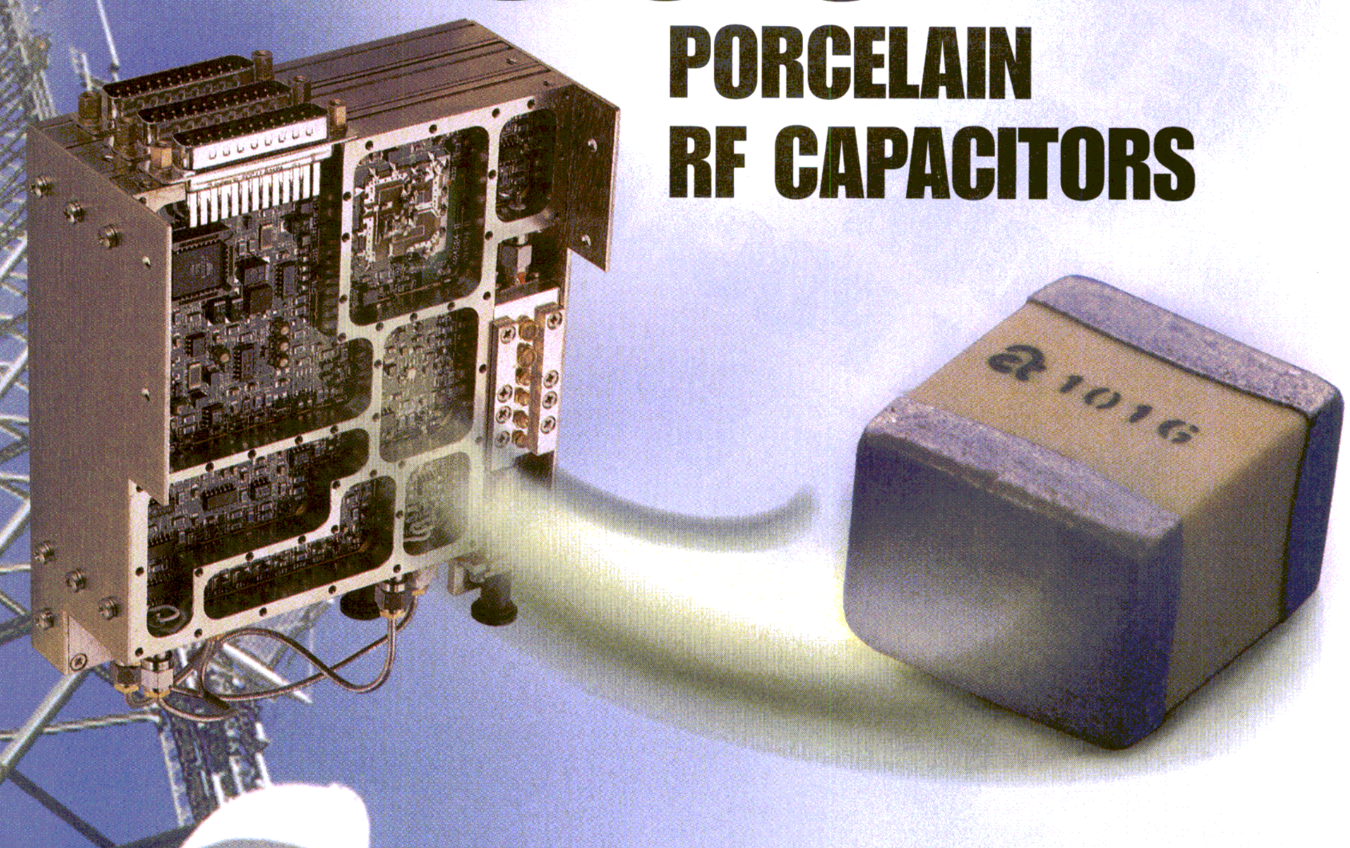

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CIRCUIT DESIGNER'S NOTEBOOK

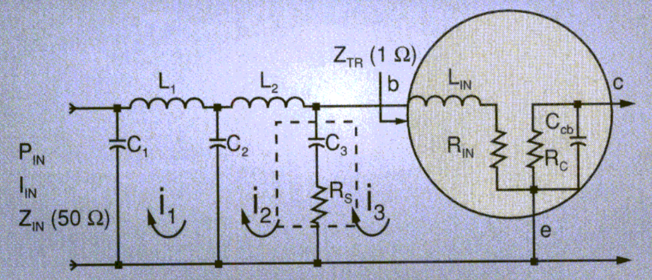
High Q Capacitors in Matching Applications

Capacitor Q is almost always a primary design consideration in RF matching applications. The capacitor's power dissipation is inversely proportional to its Q factor and directly proportional to the equivalent series resistance (ESR).

Capacitor Power Dissipation,
$Pd = i^2\ (Xc/Q)$ or $i^2\ (ESR)$.

FIGURE 1

Matching Network for an RF Power Amplifier



Components Q in Matching Networks Affects Overall Performance

An input matching network is essential for most RF amplifier designs in order to transform the relatively low impedance of the active gain device to the system impedance. The active device's input impedance is typically in the order of 0.5 to 2 ohms and is generally matched to a 50 ohm system. Lets assume that a transistor in a power amplifier has an input impedance of 1 ohm. This will require an impedance transformation of 50:1. Therefore, we must trade off voltage for current as the matching network transforms the signal impedance from 50 ohms to 1 ohm. This will result in circulating current (i_3) to be more than seven times I_{IN}. See Fig 1.

Reasons for designing High Q capacitors into matching networks:

Output Capability – Low loss High Q capacitors in matching network applications will insure maximum effective gain and available output of the amplifier. Losses due to component heating especially in high RF power applications are greatly alleviated with the use of high Q passive components.

Noise Figure – Small signal amplifiers such as LNA's used in satellite receiver applications require capacitors that exhibit high Q. Lossy passive components will add to thermal (KTB) noise and degrade the overall noise figure of the amplifier thereby reducing the signal to noise ratio.

Likewise, MRI imaging coils also require extremely low loss capacitors. These applications utilize capacitors for tuning the coil in a resonant circuit, and must be transparent in that application. The signals being detected by MRI coils are infinitesimally small such that any loss contribution from low Q capacitors would generate increased thermal noise making it difficult or impossible to process the signal.

Thermal Management – (Refer to Fig1). In extreme cases, If C_3 is very lossy, it can get hot enough to melt solder due to high circulating currents. This can easily cause components to de-solder from the board as a result of excessive heat build up. Since C_3 is physically close to the active device, any additional heat generated by the capacitor will be reflected into the transistor thereby reducing reliability and possibly causing early device failure. Although it is desirable to mount matching capacitors physically close to the transistor's device plane for optimal RF performance, thermal management must be judiciously accounted for in these applications. Improper selection of capacitors in critical applications can easily lead to a myriad of circuit performance issues.

Example: Consider the following application:

Power Amplifier @ 150 MHZ

Output Power = 400 W.

System Impedance = 50 ohm.

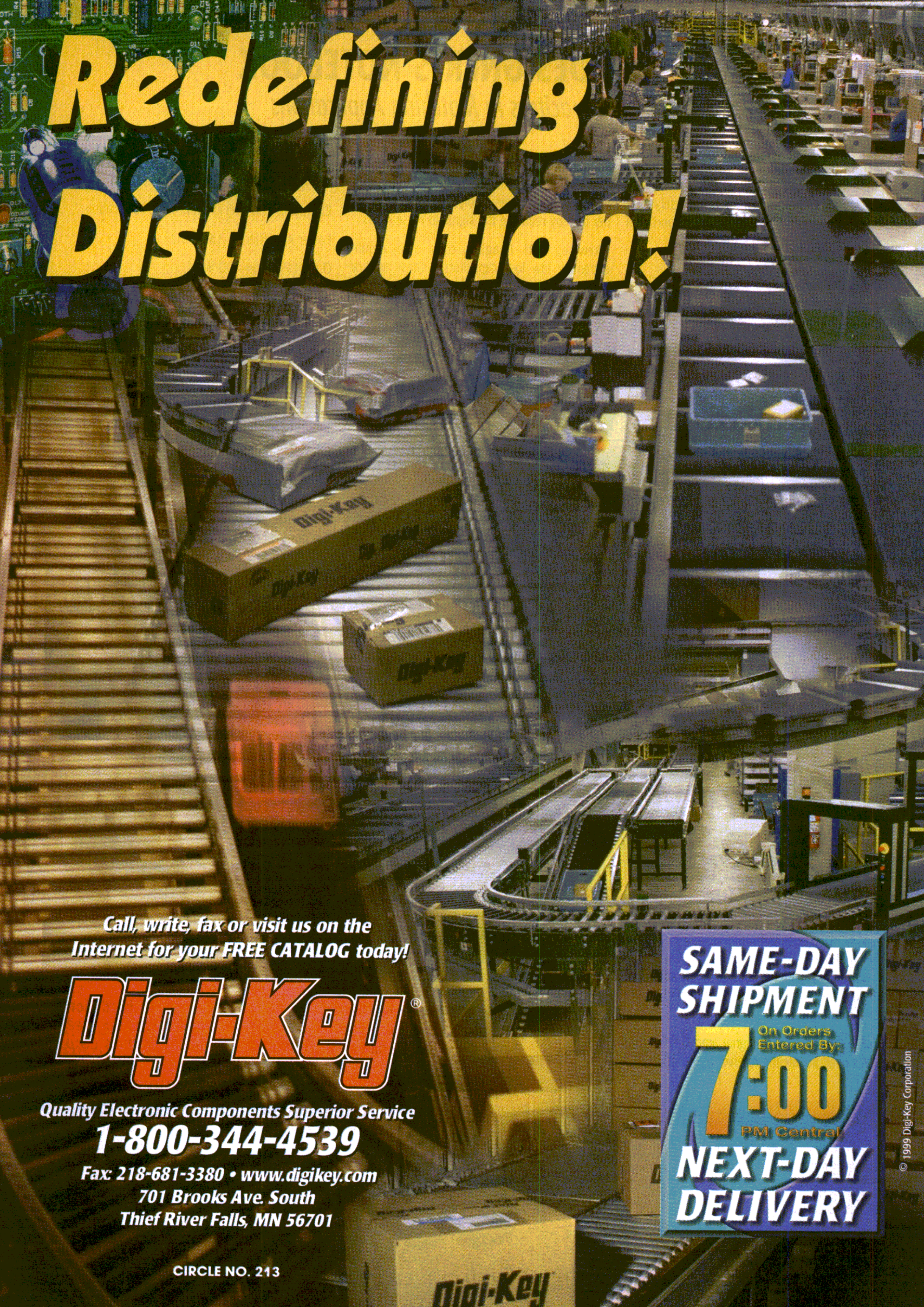
$I = \sqrt{P/Z} = \sqrt{400/50} = 2.83$ A. rms. Assume that an output coupling capacitor in a 400W amplifier has an ESR of 0.022 ohms. Under this condition power dissipation of the capacitor will be i^2 X ESR or 2.83^2 x 0.022 = 176 milli-watt. In this example we see that the power dissipated by the capacitor is directly related to the ESR, making Hi Q low ESR capacitors quintessential for this application. Even small signal amplifiers that do not generating large currents will suffer in effective gain and overall noise figure if losses are not kept to a minimum.

The following table shows typical power dissipation as a function of ESR at octavely related frequencies. The ATC 100B series, 220pF capacitor is compared to a typical 0805 NPO 220pF.

Frequency (MHz)	ESR (ohm) ATC 180R 220pF	Power Dissipation (W) ATC 100B 220pF	ESR (ohm) Typical 0805 NPO 220pF	Power Dissipation (W) Typical 0805 NPO 220pF
150	0.025	**0.200**	0.08	**0.640**
300	0.035	**0.280**	0.113	**0.904**
600	0.049	**0.392**	0.159	**1.272**
1200	0.069	**0.552**	0.224	**1.792**

Reliability – Excessive heat generated by lossy capacitors will affect the reliability of the active device as well as other components associated with or in close proximity to the heat source. Lossy capacitors in coupling, matching, bypass and blocking applications can easily lead to decreased MTBF of the entire circuit.

Richard Fiore
Sr. RF Applications Engineer
American Technical Ceramics Corp.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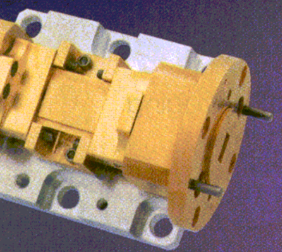
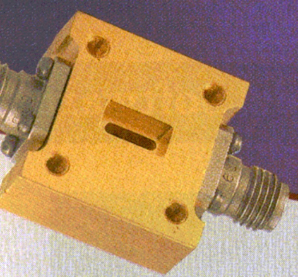
AMPLIFIERS • MIXERS • MULTIPLIERS



AMPLIFIERS

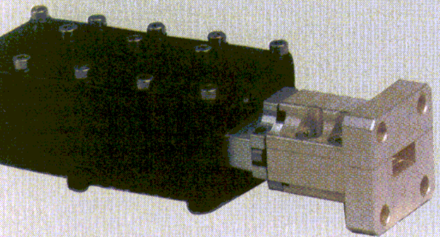
Model Number	Frequency (GHz)	Gain (dB, Min.)	Gain Flatness (±dB, Max.)	Noise Figure (dB, Max.)	I/O VSWR (Max.)	Output Power at 1dB Comp.* (dBm, Typ.)
JSW4-18002600-18-5A	18-26	28	1.0	1.8	2.0:1/2.0:1	5
JSW4-26004000-25-5A	26-40	25	2.5	2.5	2.0:1/2.0:1	5
JSW4-18004000-32-8A	18-40	21	2.0	3.2	2.0:1/2.5:1	8
JSW4-30005000-45-5A	30-50	21	2.5	4.5	2.5:1/2.5:1	5
JSW4-40006000-65-0A	40-60	16	2.5	6.5	2.5:1/2.5:1	0

* Higher output power options available



MIXER/CONVERTER PRODUCTS

Model Number	Frequency (GHz)			Conversion Gain/Loss (dB, Typ.)	Noise Figure (dB, Typ.)	Image Rejection (dB, Typ.)	LO-RF Isolation (dB, Typ.)
	RF	LO	IF				
LNB-1826-30	18-26	Internal	2-10	42	2.5	20	45
LNB-2640-40	26-40	Internal	2-16	42	3.5	20	45
ARE3436LC1	34-36	15.5-16.5	2.7-3.3	25	4	20	60
SBW3337LG2	33-37	33-37	DC-4	-7.5	8	N/A	25
TB0440LW1	4-40	4-42	.5-20	-10	10.5	N/A	20
DB0440LW1	4-40	4-40	DC-2	-9	9.5	N/A	25
SBE0440LW1	4-40	2-20	DC-1.5	-10	10.5	N/A	20



MULTIPLIERS

Model Number	Frequency (GHz)		Input Level (dBm, min.)	Output Power* (dBm, min.)	Fundamental Feed Through Level (dBc, min.)	DC current @+15VDC (mA, nom.)
	Input	Output				
MAX2M260400	13-20	26-40	10	12	18	160
MAX2M200380	10-19	20-38	6	14	18	200
MAX2M300500	15-25	30-50	10	8	18	160
MAX4M400480	10-12	40-48	10	8	18	250
MAX3M300300	10	30	10	10	60	160
MAX2M360500	18-25	36-50	10	8	18	160
MAX2M200400	10-20	20-40	10	10	18	160
TD0040LA2	2-20	4-40	10	-3	30	N/A

* Higher output power options available

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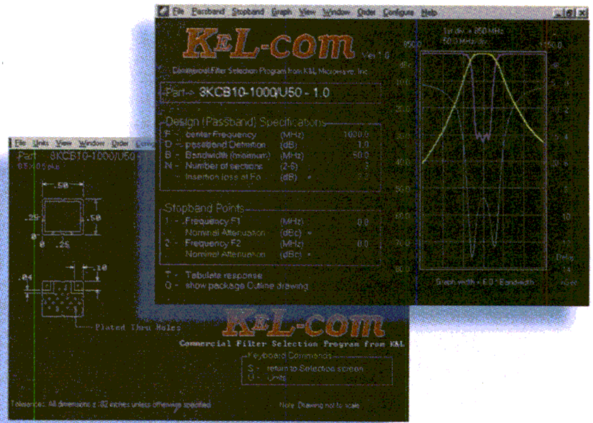

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CIRCLE NO. 277

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CALIBRATION KIT

To the editor:

It is my opinion that the term "calibration kit" is in error with respect to the function that these kits perform. According to the dictionary, to calibrate is "to standardize (as a measuring instrument) by determining the deviation from a standard so as to ascertain the proper correction factors." The actual function that these kits performs is not only to measure these factors, but to correct them. To my surprise, even the IEEE dictionary only applies the term to radiology, but in effect says the same. These kits are, therefore, measurement error-correction kits, which, of course, first ascertain the factors. However, the giants of the industry evidently could not find a convenient term to describe this function. I personally describe them as vector-correction kits [for vector network analyzers (VNAs)]. A true calibration kit is what the measurement industry calls a "verification kit," which checks an instrument's performance versus its specifications.

John Zorzy
Micro-Mode East
Concord, MA

NO MENTION

To the editor:

Thank you for forgetting to mention the design engineer of the PTE32003 high-impedance transformer (HIT) from Ericsson in the July issue of *Microwaves & RF* (July 1999, p. 105). I would appreciate a mention in the September issue.

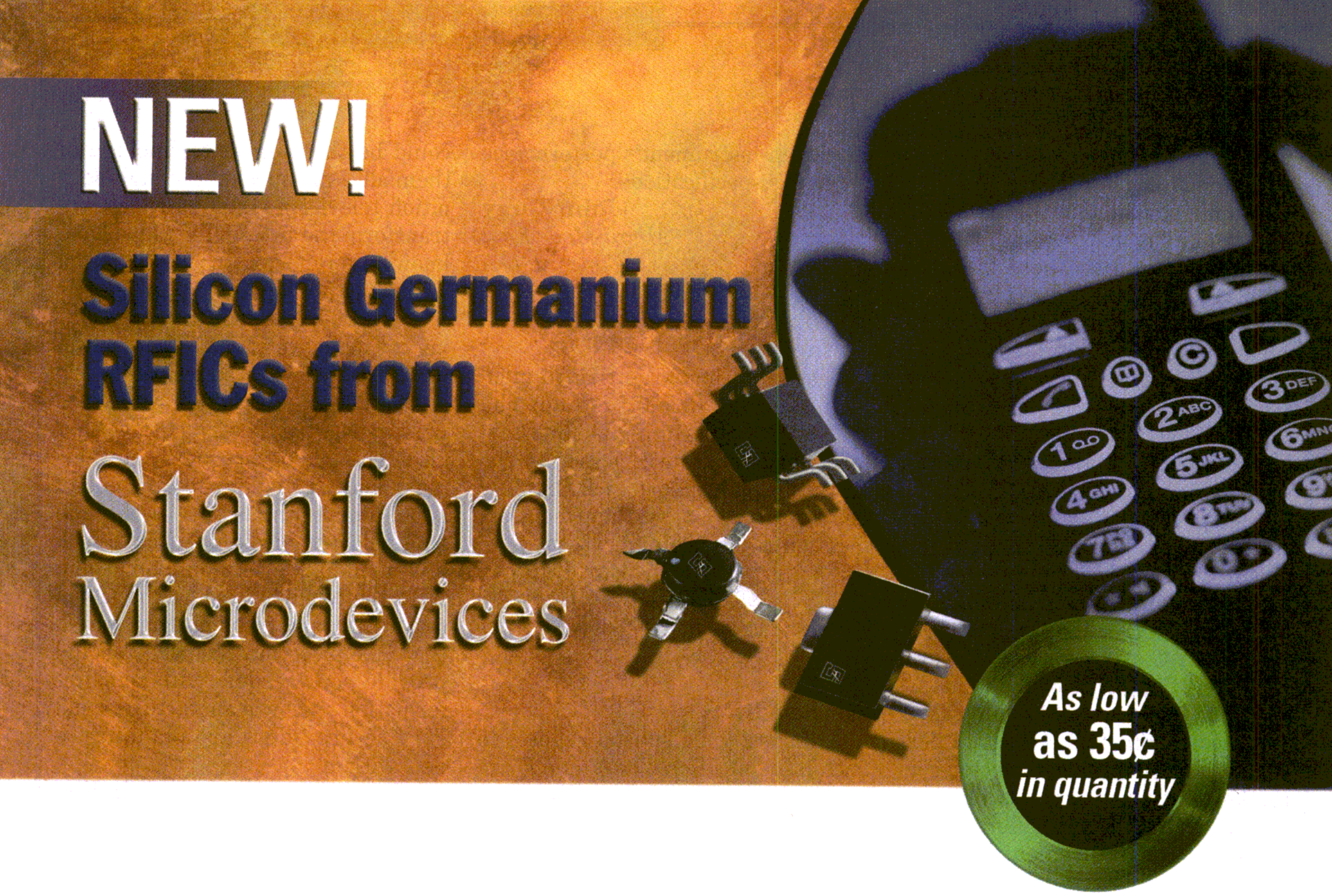
Bob Bartola
Senior Staff Engineer
Ericsson Hybrid Design Center

HDTV

To the editor:

Although your special reports in each issue of *Microwaves & RF* are very informative and usually cover the most recent developments in the communications industry, I have not seen anything on what potentially could be the biggest technology of them all. I am speaking of high-definition television (HDTV) which will transform the face of TV as we have known it for more than 50 years, from an analog to a digital medium. I know it is somewhat premature to be beating the drums for HDTV, but bear in mind that the top TV markets are supposed to begin digital broadcasts next month. And do not forget that all of the analog broadcast stations must convert to digital by the end of 2006. The changes that will come from this new medium will have such a tremendous effect on everything from the home TV set to the signal generating and transmission system that I believe your magazine should provide your readers with some information on how manufacturers and service providers are planning to deal with it.

Pedro Ceballos
Mexican Institute of Communications
Mexico City



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SGA-2186	2.2	20	DC-5.0	+7.0	+20.0	10.5	10.2	4.1
SGA-2286	2.2	20	DC-3.5	+7.0	+20.0	15.0	14.0	3.2
SGA-2386	2.7	20	DC-2.8	+7.0	+20.0	17.4	16.4	2.9
SGA-2486	2.7	20	DC-2.0	+7.0	+20.0	19.6	18.0	2.5
SGA-3286	2.7	35	DC-3.6	+12.0	+26.0	14.8	13.4	3.5
SGA-3386	2.5	35	DC-3.6	+12.0	+25.0	17.4	16.2	3.0
SGA-3486	2.9	35	DC-2.0	+12.0	+25.0	21.5	19.4	2.6
SGA-4186	3.2	45	DC-6.0	+15.0	+29.0	10.4	10.2	4.6
SGA-4286	3.2	45	DC-3.5	+15.0	+29.0	13.8	12.6	3.3
SGA-4386	3.3	45	DC-2.5	+15.0	+29.0	17.0	15.2	2.8
SGA-4486	3.2	45	DC-2.0	+15.0	+29.0	19.0	16.8	2.5
SGA-5286	3.5	60	DC-4.0	+17.0	+30.0	13.5	12.7	4.1
SGA-5386	3.6	60	DC-3.2	+17.0	+31.0	17.3	16.0	3.5
SGA-5486	3.5	60	DC-2.4	+17.0	+31.0	19.7	18.0	2.8
SGA-6286	4.2	75	DC-3.5	+20.0	+34.0	13.8	12.4	3.9
SGA-6386	5.0	80	DC-3.0	+20.0	+34.5	15.4	13.8	3.8
SGA-6486	5.2	75	DC-1.8	+20.0	+34.0	19.7	16.7	2.9

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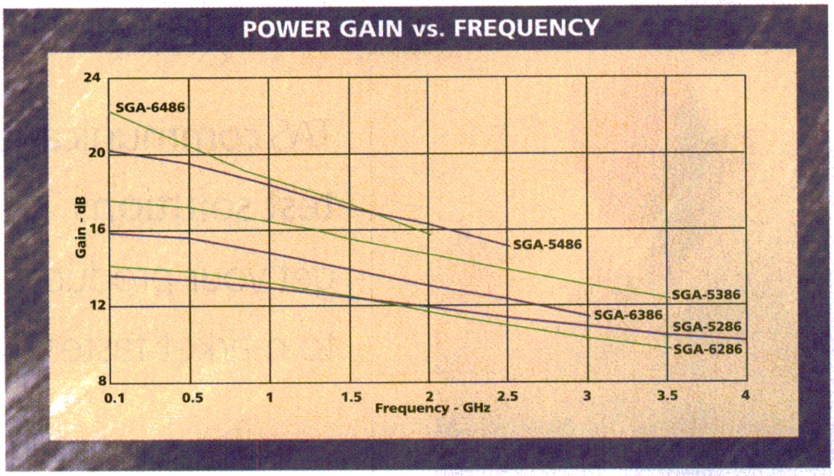

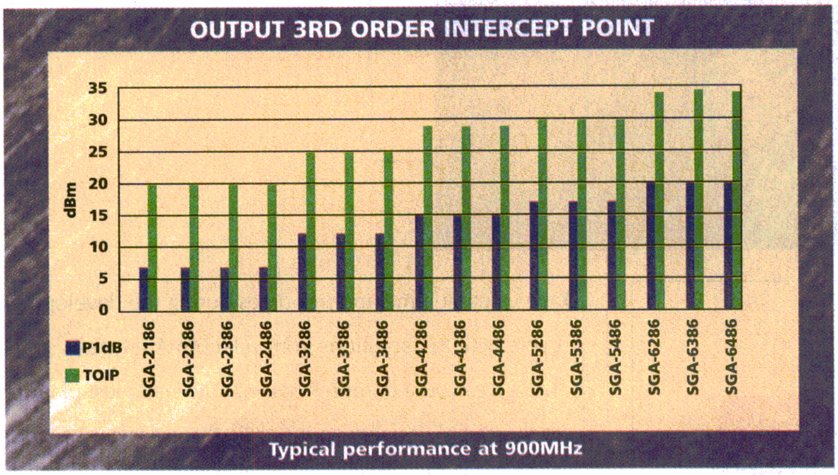

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TESTING THE WATERS OF THE NEXT GENERATION

Test equipment is a form of validation. It can reinforce a device or component manufacturer's claims about their products. Test gear can provide invaluable information about the performance of the product under a variety of conditions. Yet, the test equipment must provide performance that meets or exceeds that of the devices that it tests, even when the applications for those devices are constantly changing. For that reason, test-equipment suppliers must not only keep up with change, they must anticipate it.

Because of the importance of test equipment to high-frequency design, the "Trends in Test Equipment" section which was born last September has been revisited in this issue to provide readers with a quick glimpse of the latest developments in measurement technology. It offers several new instruments as well as new measurement strategies, which may impact high-frequency design engineers.

In gathering information for the "Trends in Test Equipment" section, a note from an old friend, Ben Zarlingo of Hewlett-Packard Co.'s Lake Stevens Division (Everett, WA), served as a reminder of the challenges faced by even the largest of measurement-equipment suppliers. In Ben's words, "It's no news that major new wireless standards represent complex technologies that manufacturers need to roll out quickly and in large numbers, yet typically at low cost. Manufacturers need test equipment (either new equipment or enhancements to existing solutions) to perform this feat. The tough thing is to keep up with the leading edge of the developing technologies, providing solutions to several complex, multifaceted emerging standards at the same time. Sometimes the answer is new equipment, sometimes new versions of existing equipment, and sometimes new test techniques using existing equipment. The emphasis is not on perfect solutions but on getting research and development (R&D) engineers productive solutions as soon as possible."

Ben included information on the company's advanced HP 89400 series of vector signal analyzers (an example of a new version of existing equipment). Not ironically, these analyzers were originally developed for military and surveillance work (a true dual-use product), but have seen greater acceptance recently for studying signals with complex modulation. The newest versions are upgraded for examining such emerging standards as wideband code-division multiple access (WCDMA) signals and Bluetooth designs. Bluetooth, for example, combines digital modulation with frequency-hopping and burst techniques. Yet, the end products are meant for low-cost consumer markets, implying that any measurements developed at the R&D stage must inevitably map to high-volume, high-speed (thus low-cost) tests at the production stage.

The "Trends in Test Equipment" section attempts to document the challenges faced by the measurement community. These challenges are worth noting because they pave the way for next-generation devices, components, and systems. And if you, the reader, would like to see this type of measurement "trend tracking" increase in the future, send your thoughts electronically to jbrowne@penton.com. ••

Jack Browne

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Ceramic Industry Roadmap Addresses Electronic-Packaging Issues

RESTON, VA—The Ceramic Interconnect Initiative (CII) of the International Microelectronics And Packaging Society (IMAPS) has developed a 10-year ceramic industry "roadmap," a document that describes how the features of ceramic technology deliver benefits for a wide variety of electronic-packaging and interconnect applications. The roadmap analyzes the current state and forecasts future requirements of ceramic use in electronic packaging.

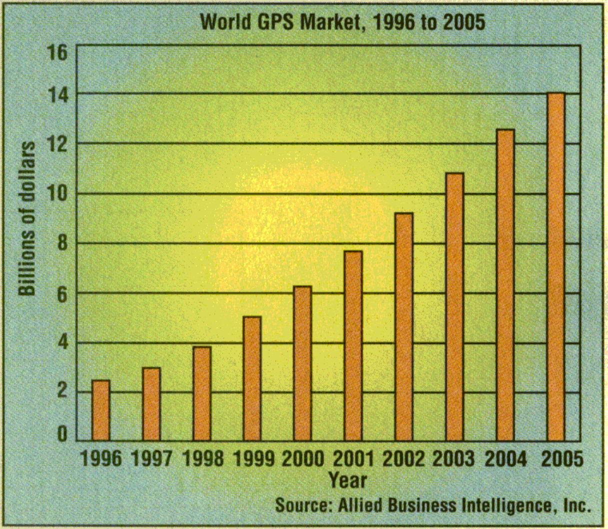
The purpose of the roadmap is to demonstrate the competitive advantages of ceramic technologies and their potential uses in the design, development, manufacturing, and marketing of microelectronics applications over the next decade. It is intended to serve as a reference document to guide investment in research, development, and deployment for the ceramic technology infrastructure.

"Ceramic substrate interconnection technologies are important, timely, and continuously improving to meet changing market requirements, says Samuel J. Horowitz, chairman of the CII. "Packaging and system designers have limited awareness of the benefits and capabilities of ceramic technology which include reduced costs, improved reliability, and faster time to market."

Copies of the roadmap are available by contacting the National Electronics Manufacturing Initiative (NEMI) at http://www.nemi.org or the IPC at http://www.ipc.org.

GPS Market To Be Worth $14 Billion By 2005

OYSTER BAY, NY—While Global Positioning Systems (GPS) have been available since the bicentennial, the technology will finally be used in a commercial market expected to be worth $14 billion by 2005 (see figure), according to Allied Business Intelligence.



In the US, GPS has largely been adopted by two user segments which represent extremes in the technology's development. At the one end lie applications which require precise measurements such as surveying and reference timing. At the other are consumer applications such as small-craft navigation systems and handhelds for outdoor enthusiasts.

To date, only Japan has a mass market that has been identified and exploited—In-Vehicle Navigation Systems (IVNS). But that is about to change. In the US, IVNS is moving from rental cars and luxury vehicles to mid-priced vehicles and will be an option for most domestic automobiles and trucks within a few years.

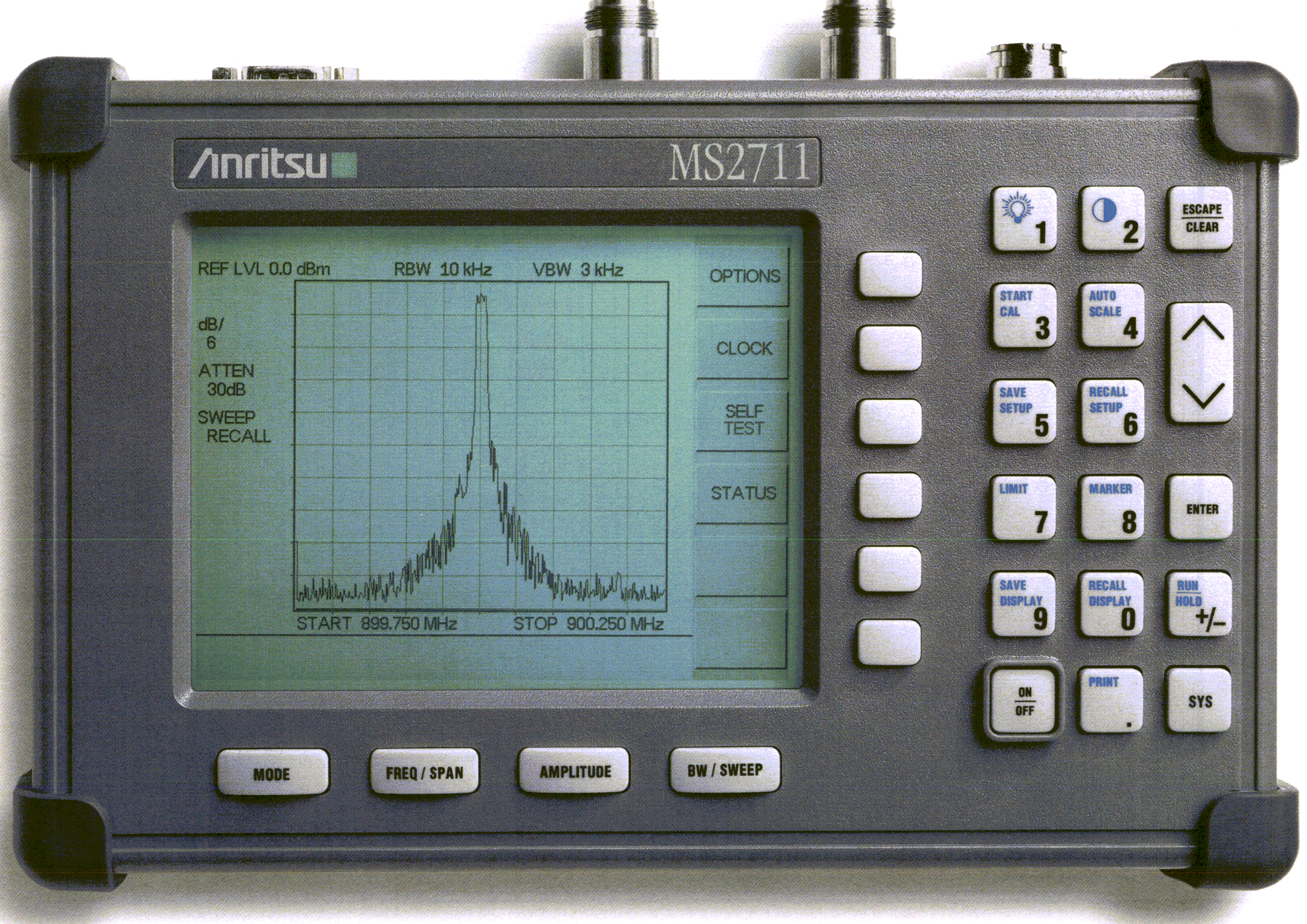
IVNS systems can now be found for as little as $200 although multifeatured systems sell at four times that. By 2005, IVNS will account for one-third of the world market for GPS applications.

Pager Location Through Web Provides Safety

SAN FRANCISCO, CA—A new Seattle-based company will soon make the recovery of lost children and missing Alzheimer's patients easier.

At the recent Wireless Developer's Conference, Loc8.net unveiled a wireless Internet location service that will revolutionize how people manage personal safety. The Loc8.net pager will provide users with access to emergency services at the touch of a button wherever ReFLEX wireless network coverage exists. Currently, it covers 90 percent of the US population.

Loc8.net plans to incorporate SnapTrack's enhanced Global Positioning System (GPS) location technology to locate subscribers' devices, even indoors. XYPOINT, a provider of wireless intelligent network services, will provide the location-related information and Internet services. Glenayre will add the necessary network enhancements to support Loc8.net's services.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CIRCLE NO. 219

High-Frequency Power Transistors Make Debut

SANTA CLARA, CA—An assortment of high-power microwave and RF transistors intended for applications in advanced avionics, radar, and microwave communications equipment has been developed by GHz Technology, Inc. The transistors are designed for long- and short-pulsed operation—from 1 μs to hundreds of microseconds—at power levels in the hundreds of watts. Fabricated in bipolar technology, the devices use a number of processing enhancements to provide high performance and extended reliability. For example, diffused ballast resistors increase a transistor's tolerance to load mismatches and polysilicon deposition enables better performance in the L- and S-bands (1.0 to 3.9 GHz) where the devices are designed to operate. The company uses gold (Au) topside-metal and hermetic solder seals to reduce transistor junction temperature, which, in turn, improves reliability. Low-inductance ceramic packaging yields better performance and reliability.

The company recently announced 10 new transistors for pulsed avionics and radar applications. The ITC1000 is a 1-kW, short-pulse (7-μs) device for pulsed interrogator systems in the 1030-MHz frequency band. A long-pulse (250-μs) transistor, the MDS350L is rated at 350 W and is slated for avionics systems in the 1030-to-1090-MHz range.

Scholarships Awarded To Competition Winners

ARLINGTON, VA—As part of an ongoing effort to address the shortage of qualified technicians in the consumer-electronics (CE) industry, the DeVry Institute is continuing its support of the Skills USA Championships, Electronics Applications Competition. The two first-place winners of this year's championship, which was held in Kansas City, MO on July 1, each received a $30,000 scholarship donated by DeVry.

The Consumer Electronics Manufacturers Association (CEMA) sponsors the annual competition to attract and foster talented technicians. "The CE industry relies heavily on skilled technicians," says Don Hatton, vice president of CEMA Technical Education and Services. "Skills USA and the DeVry scholarships provide contest winners with an opportunity to develop their skills through education."

A long-time supporter of the competition, DeVry has donated approximately $250,000 in scholarships to the contest winners during the last four years, and has made continuing education a possibility for many students who may not have had the opportunity.

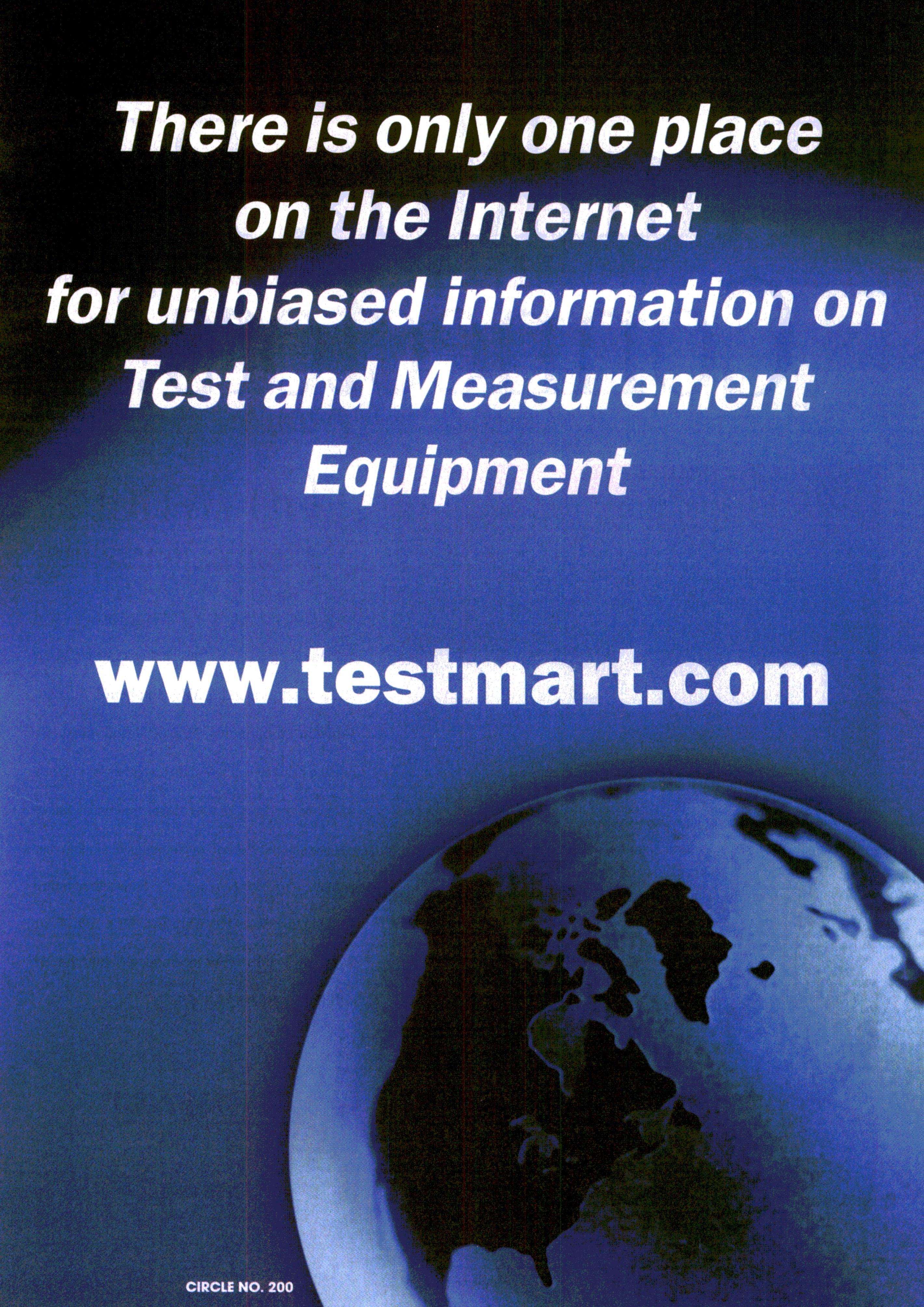
The Skills USA Championships are held annually to recognize the achievements of vocational students and encourage them to strive for excellence and pride in their chosen occupations. Working against the clock and each other, the contestants prove their expertise in job skills for occupations such as electronics, technical drafting, precision machining, medical assisting, and culinary arts. The Electronics Applications competition consists of five troubleshooting sections, a soldering/desoldering section, and a written exam. The exercises are designed to simulate real-world conditions and emphasize the quality, efficiency, and safety of the competitor's work.

Semiconductor Industry Is Challenged To Close "Test Gap"

SAN JOSE, CA—Edward W. (Ned) Barnholt, chief executive officer of Agilent Technologies—a company that will be composed of Hewlett-Packard Co.'s test-and-measurement, components, chemical-analysis, and medical businesses—has challenged the semiconductor test-equipment industry to close the "test gap" between revolutionary system-on-a-chip (SOC) technology for consumer electronics and the production test equipment now being used to test SOC devices.

In the keynote speech at the Semicon West '99 Manufacturing Test Conference recently, Barnholt implored test-equipment manufacturers to design lower-cost test systems that will meet high-volume SOC production demands and offer chip producers platform-independent software that accommodates multiple test technologies.

"Today, there looms a significant test gap—a gulf between the kinds of SOCs that can be designed and what can be tested during high-volume production," Barnholt told the conference attendees. "There are tough problems that will require new thinking, new architectures, and new tools. The answer to testing SOCs isn't bigger, better stand-alone testers or cobbling together some combination of analog, digital, and memory testers with hardware and software hooks." Barnholt urged the semiconductor test industry to move toward flexible, general-purpose tester systems that incorporate multiple test methodologies while continually driving down the cost of the test.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CIRCLE NO. 200

Partnership Announced For Component Information Portal

ARLINGTON, VA—The Electronic Industries Alliance (EIA) And Parametric Technology Corp. (PTC) recently announced an initiative to create a new and innovative Internet-based portal designed to provide access to a data center containing thousands of models and millions of technical specifications created and maintained in partnership with leading electronic component manufacturers.

This program, known as *InPart Electronic*, will make it easy for electronic-design engineers to locate components matching the functional characteristics that are required in product designs. The program is designed to be a companion to the already-operational *InPart Mechanical* portal, which provides comparable data on mechanical components. EIA will be the technical advisor in the design, development, and support of the new electronic component portal and will assist PTC in recruiting EIA members to participate in *InPart Electronic*. The Alliance will also assist PTC in marketing *InPart Electronic* to potential end users. Using a standard web browser, subscribers to *InPart Electronic* will be able to access information on thousands of components from EIA member companies. *InPart Electronic* subscribers will be ensured that supplier component selections are current, resulting from the InPart portal's ability to provide immediate access to supplier updates, engineering changes, and new product releases. Subscribers and participating companies will be able to access *InPart Electronic* seven days a week, 24 hours a day.

Studies Show Consumer Demand For Wireless-Location Services

BOULDER, CO—SignalSoft Corp. recently announced that it has completed two separate market-research studies into consumer needs and preferences for types of Wireless Location Services™ and the benefits of these applications.

The studies were designed to determine what consumers think about location-based services such as SignalSoft's local.info™, which brings subscribers personalized, localized information instantly through their wireless phone and Location Sensitive Billing, which enables reduced rate plans based on a caller's location.

The survey revealed several quantitative findings:

- Twenty-three percent of current cellular users were interested in location-based information services, also known as wireless 411.
- Fifty-four percent of current wireless users were interested in location-enabled emergency roadside assistance.
- Thirty-two percent of current wireless users would definitely or probably switch wireless providers to get a Location Sensitive Billing plan.

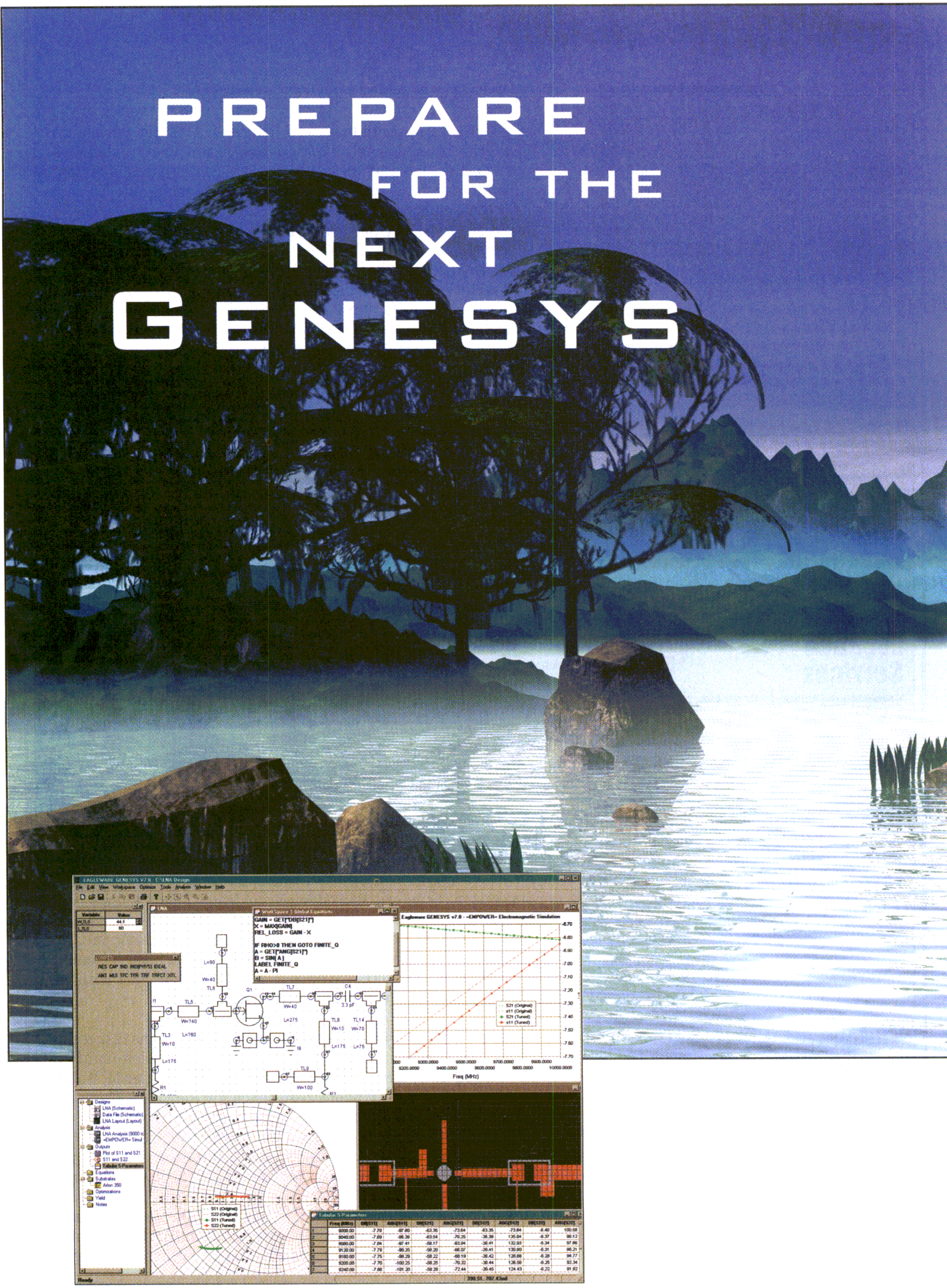
Van Buskirk Joins Stanford Microdevices As CEO

SUNNYVALE, CA—Robert Van Buskirk has joined Stanford Microdevices, Inc. as chief executive officer. In his new capacity, Van Buskirk will be responsible for Stanford's corporate positioning and long-term strategic direction, as well as the management of the company's executive team.

In making the announcement, Stanford Microdevices chairman John Ocampo commented that "Bob's leadership and managerial skills combined with his extensive financial and marketing experience will be key as SMI continues its expansion into the wireless handset and infrastructure markets." Ocampo added, "With his telecommunications industry and technical experience, we expect he will provide SMI with the vision, energy, and experience necessary to take us to the next level."

Kudos

Telecom Analysis Systems (TAS), a provider of wireless communications test equipment, has been selected as the vendor of choice by CTIA for code-division-multiple-access (CDMA) mobile-handset evaluation. TAS' CDMA-ATS Automatic Mobile Phone Test System is the specified mobile-phone test bench for the CTIA Certification Program...Last month, one of Silicon Valley's original technology companies celebrated its 40th anniversary. Founded in 1959, California Eastern Laboratories (CEL) was among the first American companies to introduce Japanese technology to the North American market. In 1962, the company added NEC to the list of the Japanese manufacturers it represented. Today, CEL represents NEC's RF, wireless, and optoelectronic semiconductor products exclusively...Melvin W. Boldt, founder, chairman, and CEO of Palantine, IL-based Boldt Metronics International (BMI) was named "1999 SME Marketer of the Year" by the Sales and Marketing Executives of Chicago.


PREPARE
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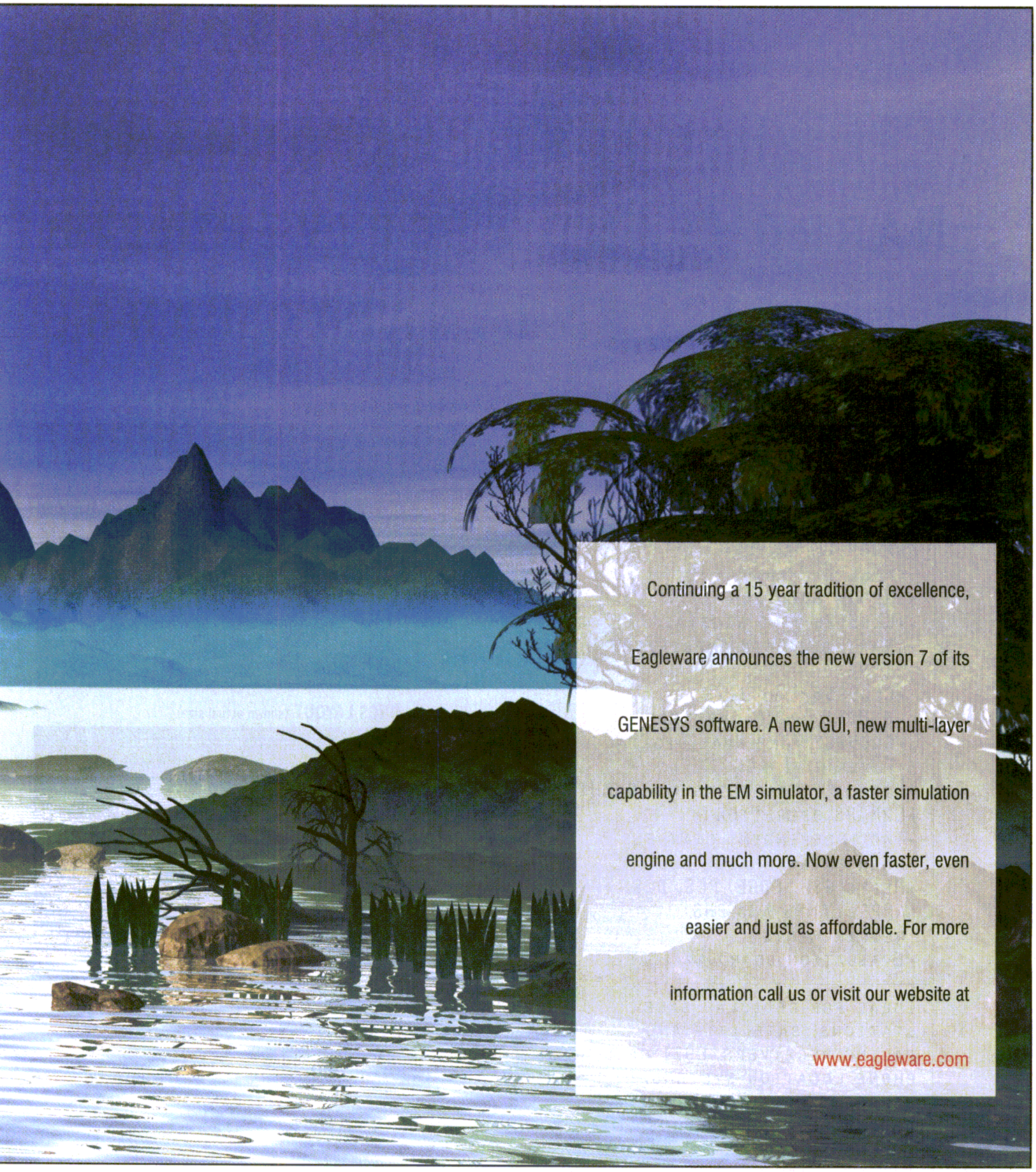
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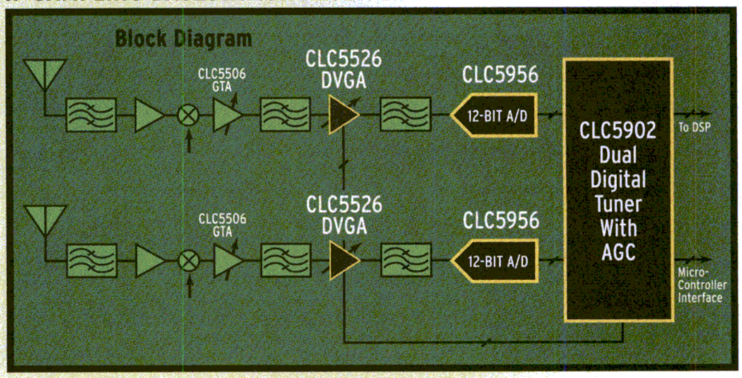
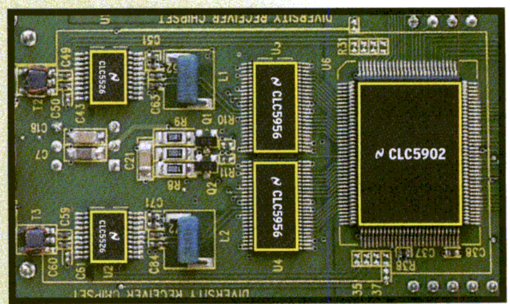


DRCS LAYOUT (shown actual size)



National Semiconductor

Global wireless communications is getting a cautious boost from the growing number of commercial satellites going up into low earth orbits.

Communications Lifts Off Into Inner Space

GENE HEFTMAN

Senior Editor

GO back approximately 25 years and most of the satellites floating around in space were owned by governments that spied on each other for military purposes. Military applications still abound as the urge to snoop is strong, even in peacetime. But inner space (approximately 1000 miles above the earth) is filling up with commercial satellites that promise to expand the communications options for millions of users around the globe. Most of the technology of today's satellites was developed for the nation's defense, but the growth of wireless communications will propel it further than military uses ever could. As the world's citizens demand more instantaneous and varied communications, satellite technology has become the only long-haul medium with the reach and signal-handling capacity to provide it across oceans. Driving the launch of space platforms are the applications that will define communications in the first decade of the 21st century—telephone, television, high-quality voice, digital data, e-mail, two-way paging, and the Internet. Most observers predict that the satellite business will soar—it is a global industry of $60 billion and climbing at approximately 13 percent a year—but a recent spate of mishaps and financial problems has taken some of the lift out of its rockets.

During the writing of this article, for example, Iridium LLC (Washington, DC), owner of the world's largest satellite network for telephone and messaging services filed for protection from its creditors under Chapter 11 of the Bankruptcy Code after defaulting on more than $1.5 billion in loans. A few weeks earlier, America Online (AOL) [Dulles, VA], the world's leading interactive services company made a $1.5 billion investment in Hughes Electronics Corp. (El Segundo, CA), the nation's largest satellite company for the purpose of providing Internet service through satellite. At the same time, however, Hughes was forced to write off $125 million due to problems manufacturing satellites, and it has suffered other setbacks due to the failure of a satellite in orbit and the explosion of another on the launch pad.

Despite these kinds of reversals, high-technology companies such as The Boeing Co. (Seattle, WA) and Lockheed Martin (Bethesda, MD), the nation's two biggest defense firms, are moving strongly toward entering commercial satellite communications. A prime reason is that the commercial market has better long-term potential than military contracting. The key applications are television, the Internet, and broadcasting. TV has already made extensive use of satellites in the form of DirecTV (owned by Hughes), which has 7.5 million customers who receive dozens of digital channels on 18-in. (45.72-cm) home dishes, and Echostar Communications Corp. (Littleton, CO), with a customer base of more than 2.5 million direct-broadcast-satellite (DBS) users. The Internet market, largely untapped to this point, could be much greater.

COMMUNICATION BIRDS

While the recent woes of the satellite industry have put some launch plans on hold, most industry watchers see problems such as short term, and the overall thrust of the satellite-communications business is headed upward. A recent launch by Hughes of its Astra 1H satellite from Kazakhstan in the former Soviet Union will bring wideband satellite return-path capabilities to Europe on Ka-band (26.5-to-40-GHz) transponders (Fig. 1). One of the goals of the Société Européenne des Satellites (SES) of Luxembourg, owner of the bird, is to provide services comparable to emerging terrestrial communications such as XDSL and cable modems. Another goal of SES with the Astra 1H is to enable analog and digital radio and TV services across Europe.

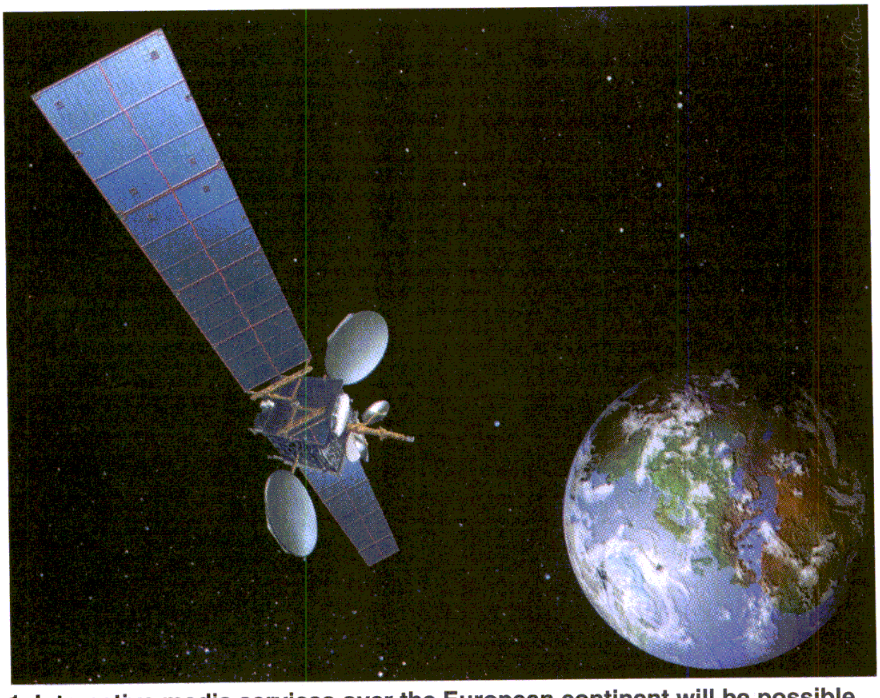
Europe will be gaining some Ku-band (12.4-to-18-GHz) capability in approximately a year when the first of the Europe*Star satellites is launched. Europe*Star 1 will carry 30 transponders of 36-MHz bandwidth each, intended to deliver satellite telecommunications to five coverage areas which are Europe, the Middle East, South Africa, India, and Southeast Asia. This satellite will be supported by a second bird—Europe*Star 2—scheduled for launch in 2002. Other services planned for the satellites are multimedia broadcasting, multicasting, and the Internet. Europe*Star is owned by Alcatel Space Industries (Middlesex, England)—with a 51-percent share—and Loral Space and Communications (Palo Alto, CA)—with a 49-percent share. It is a member of the Loral Global Alliance, a group created to commercialize satellite transmission services worldwide. Users of the service will be offered one-stop shopping and the ability to access the entire network from any entry point.

A more high-profile project than Europe*Star, also with a major Loral contribution, is Globalstar (San Jose, CA), a planned series of 48 low-earth-orbit (LEO) satellites intended to provide wireless telephone service everywhere in the world, particularly in areas underserved by basic telecommunications services (when authorized by local telecommunications authorities). In addition to telephony, the company will offer other digital communications services such as data transmission, paging, facsimile, and position location. Globalstar's plan is not to sell services directly to users, but rather through a network of regional and global telecommunications providers. These providers will market and distribute the service in their areas, obtaining all regulatory approvals and operating the gateways necessary to provide service to customers. All calls, including international, enter the service provider's existing land-based network from a local gateway and do not bypass the existing telephone-network infrastructure.

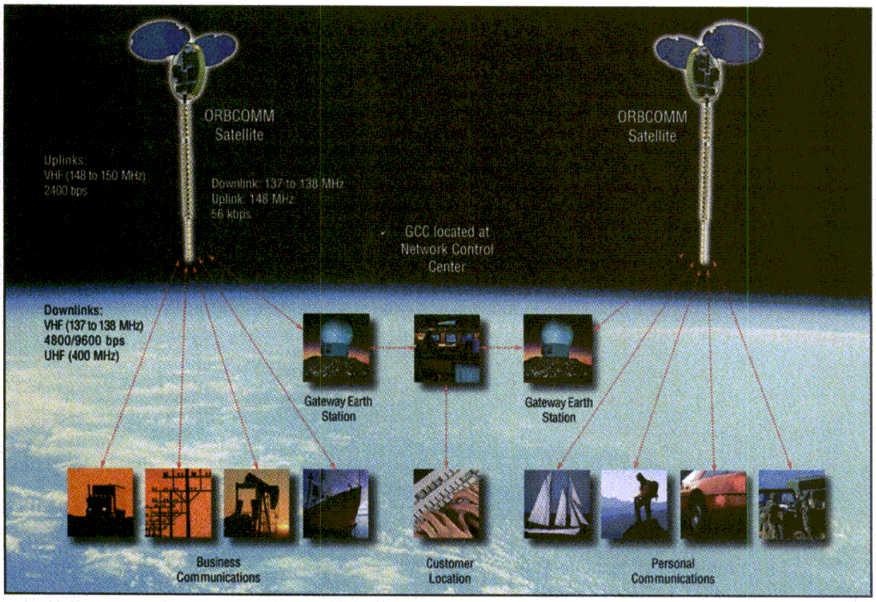
Globalstar is what is called a "bent-pipe" satellite-relay system which is essentially a frequency-translating repeater. A signal received by the satellite's antenna is amplified in an LNA, then translated in frequency, amplified in a high-power amplifier, and retransmitted to earth through the satellite's transmit antenna. Features of the system include a so-called soft handoff which is the transfer of a circuit from one beam or satellite to another without interruption of the call. The system also uses path diversity, a method that combines multiple signals of varying power into a single coherent signal. A subscriber can communicate with a single satellite in view, but typically, two to four birds will be overhead. The subscriber equipment has a rake receiver to combine signals into one static-free, coherent signal. Globalstar is getting ready to inaugurate services now. It recently brought its number of satellites in orbit to 36, and plans to launch another 16 by year's end. It will then have the 48 satellites originally planned plus four birds that are in-orbit spares.

The concept of a global, broadband, Internet-In-The-Sky is the goal of Teledisic LLC (Bellevue, WA) based on a constellation of LEO satellites. The network, still approximately four years away from deployment, is intended to deliver "fiber-like" access speeds for telecommunications services such as computer networking, broadband Internet access, high-quality voice, and other digital data applications. When the satellites are up and running, Teledesic claims they will support millions of users simultaneously, with most users having a two-way connection that will provide up to 64 Mb/s on the downlink side and 2 Mb/s on the uplink. Higher-speed terminals will offer two-way capacity of 64 Mb/s. Communications will take place in the Ka-band on the uplink side and over the 18.8-to-19.3-GHz range on the downlink side. The company recently signed a major launch contract with Lockheed Martin for use of the latter's Proton M and Atlas V launch vehicles to put the satellite constellation in orbit. Each launch vehicle can carry multiple satellites into orbit. Teledesic also inked an agreement with Motorola, Inc. (Schaumberg, IL) which will act as the prime contractor for engineering and construction of the network. In addition, Motorola has made a $150 million investment in Teledesic.

A satellite network already up and running is that of ORBCOMM Global L.P. (Dulles, VA), which is designed for global wireless data and messaging services. The constellation of LEO satellites currently numbers 28



1. Interactive media services over the European continent will be possible using the Astra 1H communications satellite built by Hughes Space and Communications for the Société Européenne des Satellites (SES) of Luxembourg. SES will use the satellite and others to provide analog and digital radio and TV across Europe.



2. LEO satellites in the ORBCOMM system serve as relay repeaters to direct information to and from an extensive ground segment infrastructure that includes GCCs, GSSs, and NCs. The system sends and receives two-way alphanumeric packets, similar to two-way paging or e-mail.

with another seven planned for deployment this year. Unlike most satellite systems which operate at microwave frequencies, ORBCOMM runs in the ultra-high-frequency (UHF) range at 137 to 138 MHz and 400 MHz for transmissions down to mobile or fixed data-communications devices and at 148 to 150 MHz for transmissions up to the birds. The system comprises three main components (Fig. 2). The space segment consists of the satellites themselves, while the ground segment comprises Gateways called Gateway Control Centers (GCCs), Gateway Earth Stations (GESs) and Network Control Centers (NCCs). Finally, the Subscriber Communicators (SCs) are handheld devices for personal messaging, as well as fixed and mobile units for remote monitoring and tracking applications. The system can send and receive two-way alphanumeric packets, similar to two-way paging or e-mail.

Two types of services available with this system are weather information and tracking of long-distance trucks. Recently, ORBCOMM, in conjunction with a value-added reseller, introduced a satellite-based weather information service for general aviation pilots flying over the US and several hundred miles off both coasts. It is based on NEXRAD radar images of weather activity, and enables a pilot to receive hourly weather updates at remote destinations with additional weather data while in flight. Other functions permissible are the transmission of position reports and the sending and receiving of in-flight e-mail. The system offers an advantage over the well-known Global Positioning System (GPS), quickly becoming the standard for navigation. The GPS can show a pilot the shortest distance between two points, but a satellite-based system of weather information can provide the data that describe the safest and most-efficient route of flight that is based on the actual weather situation.

Trailer tracking services use sensor technology embedded in a trailer or container coupled with satellite-communications technology that permits significant information to be monitored by the fleet owner. For example, in the ORBCOMM system now coming online, it is possible to detect when a trailer is connected or disconnected from a tractor, whether it is loaded or empty, its GPS position, and other key data. The information gathered is used in a comprehensive data-management and logistics system.

Active electronic device [integrated-circuit (IC)] operation and performance are far more difficult to control in space than on earth due to the radiation that continuously bombards a satellite.

RAD HARD COTS

As is well-known, radiation from cosmic rays, protons, and heavy ions are responsible for IC faults such as single-event upsets (SEUs), single-event latchups (SELs), and total burnouts of electronic devices. These types of faults have the greatest effect on digital ICs that contain memory elements such as microprocessors, microcontrollers, and memory devices themselves. Upsets and latchups can sometimes be overcome by resetting or rebooting a device through ground control, but burnouts and frozen-bit problems cause device failure, a potentially catastrophic occurrence in a space vehicle a thousand miles above the earth. During the days when the military accounted for a significant percentage of the semiconductor market, special radiation-hardened (rad-hard) devices were in ample supply for aerospace applications and were manufactured on dedicated production lines. Today, however, defense spending has been cut and with it the economic motivation for dedicated, high-reliability manufacturing facilities for rad-hard ICs. The need for these devices still exists because the commercial satellite industry is picking up where the military left off.

Virtually all of the ICs used in today's military, space, and avionics applications are of the commercial-off-the-shelf (COTS) variety. For background on COTS, see "Military Retrenchment Spells Electronics Growth," *Microwaves & RF*, June 1998, p. 37, and "Electronics Are A Secret Weapon In The Military Budget," *Microwaves & RF*, June 1999, p. 27. The same components are employed in commercial equipment, but only a small percentage need to be radiation hardened for space. The additional processing for hardening COTS products includes a cost that is not as readily absorbed in commercial satellites as it is in military types. For commercial satellite technology to be successful, it requires a combination of advanced electronic processing and

manufacturing devices at low cost. And this requires COTS manufacturers to develop a technique for economical production of rad-hard ICs.

According to a report entitled "Radiation Hardening Of Commercial CMOS Processes Through Minimally Invasive Techniques," by J.M. Benedetto and D.B. Kerwin of UTMC Microelectronic Systems (Colorado Springs, CO) and J. Chaffee of Vitesse Semiconductor Corp. (Colorado Springs, CO), commercial silicon (Si) foundries could run a rad-hard product line along with their standard product line, but most manufacturers are very reluctant to alter their commercial product flow in any significant manner. UTMC has developed two different minimally invasive radiation hardening techniques that can be used to meet the requirements of commercial foundries. In both techniques, the rad-hard product is run along with the standard product up to the field-isolation stage. The rad-hard version is then segregated out for additional patterning and processing operations and returned to the commercial flow for final processing.

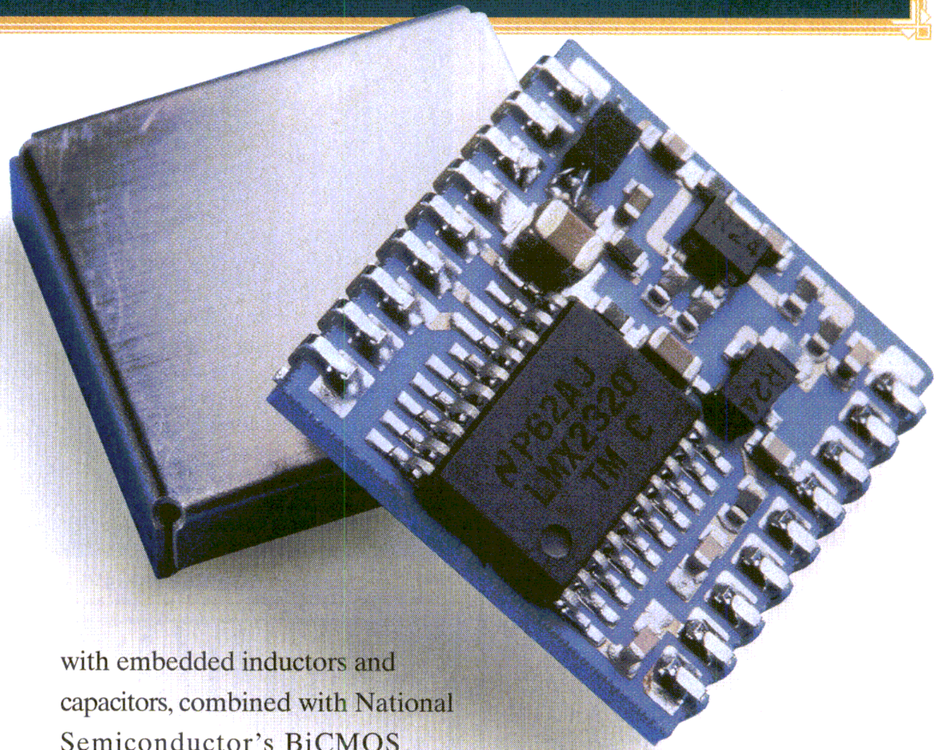
The self-contained modules, called radiation-tolerant module A (RTM-A) and RTM-B, are designed for two different design-rule processes. RTM-A is compatible with complementary-metal-oxide-semiconductor (CMOS) design rules greater than 0.5 μm and RTM-B is for design rules down to 0.35 μm. RTM-A is capable of imparting a total-dose hardness of 200 to 500 krads (SiO_2) while RTM-B can harden from 100 to 200 krads (SiO_2). The authors believe that these modules will provide even greater levels of hardening when the processes are fully optimized.

There are two important results of the module program for commercial satellites. First, advanced CMOS COTS devices can be manufactured with total-dose radiation hardness levels acceptable for most orbits [LEOs, medium earth orbits (MEOs) and geosynchronous earth orbits (GEOs)]. Second, the devices can be produced at a cost that is more in line with commercial satellite system needs and much lower than that of a dedicated rad-hard wafer foundry.

Similar to military satellites before

Application Profile

DuPont Green Tape™ Shrinks Wireless Communications Devices.



National Semiconductor chose DuPont Green Tape™ materials to integrate passive components into circuit substrates used in wireless communications devices, improving performance and reducing circuit size and cost.

The Challenge: Small-Size, High-Performance Circuits for Wireless Telecom

In wireless communications devices such as cellular telephones, wireless LANs and satellite global positioning systems (GPSs), the ratio of passive to active components may be more than 100:1, which increases circuit size, weight and cost. National Semiconductor engineers needed to address the problem of integrating the necessary passive components, while still meeting circuit size and cost-reduction goals.

In addition, the new circuit had to meet complex performance requirements. Portable wireless components typically operate in the 900 MHz–2.4 GHz range, and require high efficiency (or high Qs) to meet stringent power and signal integrity requirements.

The Solution: Dupont 951 Green Tape™ System

The new circuit substrates use the DuPont 951 Green Tape™ system with embedded inductors and capacitors, combined with National Semiconductor's BiCMOS frequency synthesizer ICs.

The combined technology was first used on a VCO/synthesizer designed to embed the resonator and several passive circuit elements of a loop filter in a frequency control module. The embedded technology reduced circuit size by a factor of four and cut costs by half. The circuit has eight conductive layers, 200-micron lines and spaces, and 150-micron vias. It contains two buried inductors and four buried capacitors. Each substrate master produces 64 individual circuits.

The embedded functions of the substrate eliminate many interconnections with poorly controlled impedance, which traditionally require additional components to control the effects of parasitics and impedance mismatches. Concurrent design of active RF ICs with substrates containing embedded passives simplifies IC design and improves circuit efficiency to reduce power consumption, improve battery life, or reduce battery size.

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

DUPONT®

CIRCLE NO. 256

them, commercial versions must carry high-reliability, rad-hard ICs, but the cost/performance trade-offs are more critical for commercial systems. A number of factors affect the selection of COTS electronic components for space, including the height of the satellite's orbit, the required degree of radiation hardness as a function of shielding within the satellite body and electronic modules, ensuring the hardness capability of all devices based on semiconductor materials, and the reliability built into devices by the manufacturer. For a more detailed exploration of the factors about using COTS components in aerospace systems, see "COTS in Space," by Joseph Benedetto, *COTS Journal*, Vol. 1, No. 2, May/June 1998, p. 67.

GPS MOVES ON

Of any technology developed originally for military use, none has greater potential for widespread commercial use than the GPS. Operating from the MEO Navstar satellite system at approximately 12,600 miles from earth, GPS has applications in aviation, marine navigation, land navigation, timing for wireless communications systems, military systems, recreational uses, and some applications not even thought about yet (see "Modern Military Technology Gets A Commercial Look," *Microwaves & RF*, September 1998, p. 31). GPS is the quintessential example of a technology that has made the transition from military use to the civilian sector and back again to the military with commercial enhancements. Although the GPS system is available for civilian use free of charge, it is owned and operated by the US government (US Department of Defense).

The primary function of the GPS is—as originally intended—navigation. A GPS receiver on earth can compute position, velocity, and time. The conventional applications that spring from this are precise positioning and highly accurate time and frequency information. Positioning information is used widely in surveying, geodetic control (for mapping and charting), and plate tectonics (movements of the earth's crust). Time and frequency information is based on precision clocks carried aboard the

satellites. This is used for astronomical observations, telecommunications, and for setting laboratory instruments to precise time and/or accurate frequencies.

In a 1998 report entitled "North American GPS Markets," international marketing consultants Frost & Sullivan (Mountain View, CA) state the total GPS market keeps growing at a rate of approximately 22 percent and shows no sign of letting up until about 2004 when growth will "cool" to just 18.6 percent. The report says that the land market is the biggest, accounting for 50 percent of all revenues through the year 2004. While land-market revenues shot up by 28.1 percent in 1997—to $615 million—the projected compound annual growth rate (CAGR) from 1998 to 2004 is not much smaller at 24.8 percent. Driving the market are uses such as the growth of land-based applications (automobiles, trains, recreational vehicles, handheld units) marine uses (mainly in pleasure craft), telecommunications for nanosecond timing accuracy, wireless communications, and aviation (precision approach and en-route navigation).

The greatest potential for GPS is a so-called enabling technology for consumer mass markets. That is, manufacturers of GPS products and systems join with makers of automobiles, boats, planes, cellular phones, golf carts, etc. to incorporate location, position, tracking, or timing features that were unavailable before the invention of the GPS. These features can create new uses for a product with a resulting increase in demand—automatic vehicle location and mapping directions in automobiles, GPS-enabled wireless communications, enhanced-911 emergency-response capability, even distance measurements from tee to green by specially equipped golf carts. There are a number of applications that could use GPS in a one-way communications system that would be relatively inexpensive. For example, an emergency GPS transmitter could be used to precisely locate a person in trouble when the transmitter was activated with no return signal necessary—in effect, a one-way 911 system. ••

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Boonton's 4530 Series RF power meters—the new peak of power measurement.
The spectacular new 4530 Series RF Power Meter lets you measure the peak and average power of any complex waveform including **UMT, CDMA, W-CDMA, and HDTV**—all at the lightning speed of ATE.
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Graph
Text
Zero CAL
ESC Local
Enter
ON STBY
CDMA
WCDMA
UMT
HDTV
Already A Generation Ahead
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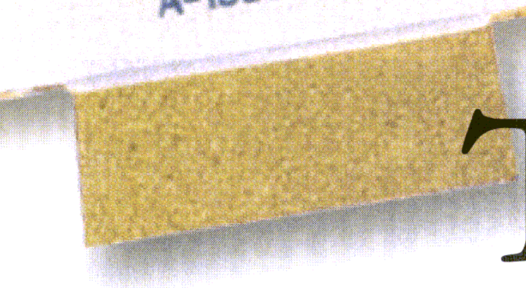

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Ericsson RF Power Products is committed to delivering the best LDMOS RF products in the industry. Your system requires maximum performance and reliability, our devices are ALL GOLD, all the time.

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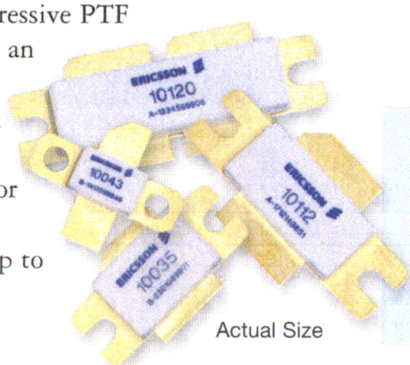
2.0 GHz LDMOS

We offer 2.0GHz LDMOS FETs optimized for PCS. The impressive PTF 10120 displays an intermodulation distortion level of better than -30dBc for peak envelope power levels up to 120 watts.



Actual Size

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900 MHz LDMOS

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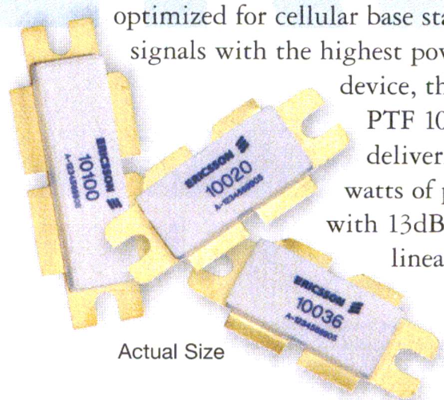

Actual Size

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Richardson Electronics
Engineered Solutions
1-800-RFPOWER (737-6937)
www.rfpowernet.com

AVNET electronics marketing
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www.avnet.com

Spectrum analyzer uses Internet for remote monitoring

Model P9116 spectrum analyzer covers 100 kHz to 1600 MHz and can perform remote monitoring of transmit or receive sites through the Internet. When used with a video-graphics-array (VGA) monitor and the supplied graphical user interface (GUI), it operates as a conventional spectrum analyzer. When used in a remote-monitoring application, it operates in a "server" mode and can be accessed and controlled from any client-designated personal computer (PC) or laptop computer via RS-232, TI line, or the Internet. Data can be processed, stored, displayed, transferred, and printed from virtually anywhere in the world using standard PC techniques. The analyzer has a frequency accuracy of 0.5 PPM and an absolute level accuracy of ±0.5 dB over its –120- to +20-dBm amplitude range. Its internal synthesized local oscillator (LO) yields a frequency resolution of 2 Hz and a step time of less than 150 μs. The analyzer is packaged in an industrial chassis that includes a power supply, Pentium controller, modem, and various serial and parallel communication ports. **Morrow Technologies Corp., 2300 Tall Pines Dr., Largo, FL 33771; (727) 531-4000, FAX: (727) 531-3531.**

CIRCLE NO. 61 or visit www.mwrf.com

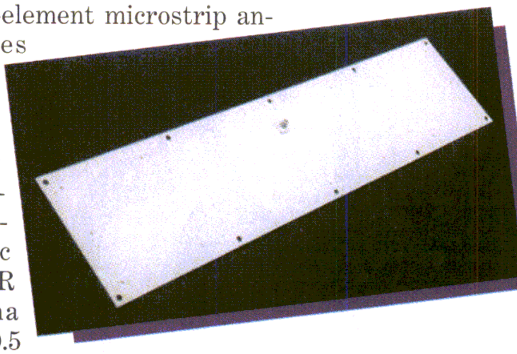
Power/voltage meters conform to VXI standard

The 5730 series of RF power/voltage meters complies with the VME-extensions-for-instrumentation (VXI) hardware-architecture standard, enabling them to integrate and share information with other VXI-compatible devices. They are designed for military and commercial automatic-test-equipment (ATE) applications that require multiple-rack configurations. The meters measure RF power levels from –70 to +44 dBm at frequencies from 10 kHz to 100 GHz with single-sensor dynamic range as wide as 90 dB. They measure RF voltages from 200 μV to 10 V at frequencies from 10 Hz to 2.5 GHz with true root-mean-square (RMS) response below 30 mV. These dual-channel meters can sample up to 200 readings per second and can display two different readings, such as dB and mW, simultaneously on two different channels. **Boonton Electronics Corp., 25 Eastmans Rd., Parsippany, NJ 07054; (973) 386-9696, FAX: (973) 386-9191, Internet: http://www.boonton.com.**

CIRCLE NO. 63 or visit www.mwrf.com

Microstrip antenna covers 5.4 to 5.9 GHz

The model 9859-800 low-profile, 16-element microstrip antenna operates with 45-deg. polarization across the 5.4-to-5.9-GHz frequency range. It is ideally suited for airborne navigational systems. Nominal gain is 16.9 dBic while maximum VSWR is 1.8:1. The antenna measures 21.5 × 5.0 × 0.5 in. (54.61 × 12.70 × 1.27 cm) and includes an SMA connector. **Seavey Engineering Associates, Inc., 28 Riverside Dr., Pembroke, MA 02359; (781) 829-4740, FAX: (781) 829-4590, Internet: http://www.seaveyantenna.com.**



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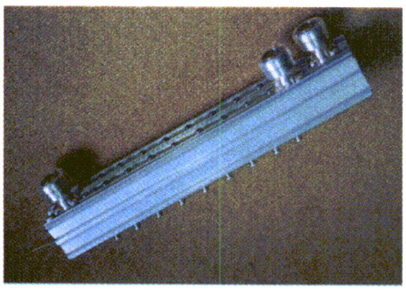
Compact diplexer screens high-power PCS signals

Model W18005D cavity diplexer filters personal-communication-services (PCS) signals at power levels exceeding 250 W. The diplexer passes signals in the PCS low band from 1850 to 1910 MHz and the PCS high band from 1930 to 1990 MHz with less than –120-dBm intermodulation distortion (IMD). Insertion loss is less than 1.2 dB and isolation between receive and transmit channels is better than 90 dB. The diplexer measures 2 × 2 × 10 in. (5.08 × 5.08 × 25.40 cm) and uses 7/16-in. connectors. **Wireless Technologies Corp., 4000 Haile Lane, Springdale, AR 72762; (501) 750-1046, FAX: (501) 750-4657, Internet: http://www.diplexers.com.**



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Female Exec To Head The New HP

Although she would probably balk at being called a "lady" CEO rather than simply a CEO, Carleton (Carly) S. Fiorina was elevated to the position of most powerful female executive in America when she took the reins of the "new" Hewlett-Packard Co. in July. Last March, HP split itself into two independent entities. The Hewlett-Packard that Fiorina will head will retain the company's name as well as its computer and imaging operations. The recently christened Agilent Technologies, now an independent company, comprises the company's core and classic businesses of electronic instruments, test equipment, and semiconductor products.

HP, with Fiorina at the helm, is the world's second-largest computer maker—behind IBM—with $39.5 billion of computer-related revenue in 1998. The Agilent portion of the business trailed far behind at $7.6 billion in the $47.1 billion combined income of the company in 1998. Despite their size, the HP computer operations are feeling the sting of upstarts such as Dell Computer with its direct sales model (over the Internet) which is attracting consumers to the tune of a 38-percent increase in sales last year. HP, by contrast, using the traditional dealers and distributors to sell its machines, posted only a 3-percent sales increase.

Fiorina comes with a mandate to change that, and in addition, to project HP into the global telecommunications market with its potential for enormous growth into the foreseeable future. She is uniquely qualified for the latter task. Before joining HP, Fiorina headed the $20 billion Global Service Provider business of Lucent Technologies and is regarded as very well versed in the telecommunications industry, having launched her career as a sales rep with AT&T in 1980. The business she led sold networking systems and software for telephone, Internet, and wireless services in 43 countries around the world. In fact, some observers expected her to become president of Lucent, but the incumbent, Richard McGinn, seems well entrenched in the position.

If Fiorina has attained the title of the country's most powerful female business leader, this is the second time she has been so anointed. Last year, *Fortune* magazine named her number one in its rankings of the 50 top female executives in America. In her praise, the magazine said, "Fiorina is a star in nothing less than the hottest, most important industry in American business: telecommunications. Without it, or the products her company produces, few of us could do our jobs." It remains to be seen how Fiorina's sales and marketing background will play in her new position, where technical decision making may be the primary factor in HP's future success. ••

Contracts

I-Bus, Inc.—Announced that it has entered into an agreement with I-Link, Inc. with an estimated value of up to $5 million over three years. Under the terms of the agreement, I-Bus will supply fully integrated personal-computer (PC) platforms, including central-processing-unit (CPU) boards, passive backplanes, and system enclosures for I-Link's enhanced Internet Protocol telecommunications-network platform.

Harris Corp.—Has signed a second contract with Telet, a Brazilian mobile-telecommunications operator, for the supply and installation of Harris microwave radios throughout the southern region of Brazil. Telet's recent order brings the total value of Harris' contracts with Telet to $15 million.

M1 and Motorola, Inc.'s Network Solutions Sector—Have signed a contract worth $15 million to upgrade M1's code-division-multiple-access (CDMA) digital cellular-telephone network in Singapore. The enhancement of the CDMA system will include the world's first commercial deployment of the Alcatell 1000 S12 CDMA mobile switching center in an IS-95 network together with the Motorola CDMA radio subsystem.

Berkeley Varitronics Systems—Was awarded a contract by Sprint PCS for 12 wireless telecommunications test systems that are comprised of receivers and transmitters for microcell analysis.

ORBCOMM Global L.P.—Has signed a reseller agreement with Hughes Global Services to deliver ORBCOMM's satellite-based communications services to US government and military customers. Hughes will market ORBCOMM's two-way communications services to government and military customers under a General Services Administration's Federal Technology Service contract, awarded earlier this year to Hughes and valued for all services at approximately $100 million.

Discovery Semiconductors—Was awarded a $65,000 contract from the Ballistic Missile Defense Organization (BMDO) to cover the development of a 100-GHz photodetector/amplifier. Another contract, valued at $736,000 from the US Army, will target the development of multilevel photonic modules. These modules will consist of a monolithic positive-intrinsic-negative (PIN) photodetector/power amplifier (OEIC), providing 44-GHz bandwidth and a power output of 100 mW.

Fresh Starts

Sprint—Has entered into a definitive agreement to purchase American Telecasting, Inc., a Colorado Springs, CO-based company that controls broadcast licenses that can be used to provide high-speed Internet and data services in several major markets including Seattle, WA; Las Vegas, NV; Denver, CO; Portland, OR; as well as Toledo, Columbus, and Cincinnati, OH.

Calibre, Inc.—Announced that it has reached distribution agreements with sales-channel partners in Japan and the Asia/Pacific region.

Andrew Corp.—Has signed a reseller agreement with Divine Tower International Corp. (DTIC), the nation's largest privately held infrastructure provider to wireless carriers. Under the terms of the agreement, DTIC will deploy Andrew-supplied HELIAX® cable, connectors, accessories, and elliptical waveguide; RADIAX® radiating cable; microwave, GridPak®, ValuLine®, base-station, earth-station, and broadcast antennas; pressurization products; towers; and equipment shelters as part of a new partnership between the two companies.

Micro Interconnect, Inc. (MII)—Recently announced business partnerships for the manufacture of expert modules for the Internet and wireless communication markets. Under the terms of the agreements, Apack Technologies, Inc. and Tong Hsing Electronics Industries Ltd., both of Taiwan, will act as manufacturing foundries for MII. Apack will provide a leading-edge interconnect technology, developed by Lucent Technologies, that incorporates embedded passives and uses flip-chip die attach onto a high-density silicon (Si) interconnect substrate. Tong Hsing will offer MII and its customers the most advanced techniques in high-volume ceramic and laminate module manufacturing.

TriQuint Semiconductor, Inc. and RF Integration, Inc.—Announced an agreement to use RF Integration's RF integrated-circuit (RF IC) product-development services with TriQuint's GaAs IC foundry to provide customers with turnkey RF IC product-development capabilities.

Hewlett-Packard Co. and Ando Electric Co.—Announced a three-year contract to develop and market test instruments jointly for the emerging high-bandwidth dense-wavelength-division-multiplexer (DWDM) and Synchronous Optical Network (SONET)/synchronous-digital-hierarchy (SDH) functional test market.

Metawave Communications Corp.—Announced the opening of offices in Taipei, Taiwan and Shanghai, China to support sales, service, and manufacturing operations for its SpotLight® smart-antenna systems.

TeleHubLink Corp.—Signed a letter of intent to acquire all of the shares of wirelessEncryption.com of Burlington, MA, which is developing advanced ultra-secure signal-processing technology for wireless and Internet communications as well as all other data-transmission methods.

SaRonix—Appointed Jacques Dorian as Northeast Regional Sales Manager. Dorian will be responsible for sales and communication with SaRonix's manufacturing representatives within that territory.

RF Micro Devices, Inc.—Opened a new design center in Cedar Rapids, IA. The 5000-sq.-ft. facility will house a team of senior-level RF engineers and technicians designing high-efficiency linear power amplifiers (PAs) and other RF integrated circuits (RF ICs) to be used in cellular and personal-communications-services (PCS) telephones. The RF ICs will be manufactured in the RFMD™ gallium-arsenide heterojunction-bipolar-transistor (GaAs HBT) foundry.

Spike Technologies—Announced the completion of the first phase of a new broadband wireless-local-loop (WLL) installation in Caracas, Venezuela.

Reed Exhibition Companies (REC)—Bob Stewart to senior vice president; formerly industry vice president of NMW and Quality Expo International.

TriPoint Global Communications—William L. (Bill) Shillito to vice president and general manager for technical products; formerly vice president of operations and engineering at Prodelin. Also, Nathan A. Knutson to operations manager for universal antennas; formerly president of Ranger Satellite Communications.

SHILLITO

KNUTSON

TeleHubLink Corp.—Daniel J. Ryan to executive vice president and chief operating officer of the wirelessEncryption.com subsidiary; formerly corporate vice president and division general manager at Science Applications International Corp.

American Technical Ceramics—Sidney Arnow to eastern regional sales manager; formerly president of US Technical Marketing, Inc. Also, Eugene Pinto to western regional sales manager; formerly handled sales of commercial antenna products at Dorne & Margolin, Inc.

LCC International, Inc.—C. Thomas Faulders III to president and chief executive officer; formerly executive vice president, treasurer, and chief financial officer of BDM International, Inc.

Quad Systems Corp.—Roger E. Gower to the board of directors; currently chairman, president, and chief executive officer of Micro Component Technology, Inc.

Narda Microwave-East—Robert Koelzer to vice president of engineering; formerly product-line manager for passive products.

CTS Corp.—William J. Kaska to executive vice president; formerly group vice president.

Storm Products Co., Advanced Technology Group—Lew Backer to western regional sales manager for the Microwave Business Unit; formerly director of marketing and sales for AMP's Precision Cable Division. Also, Colin Scrimgeour to quality assurance manager; formerly quality manager for Telecommunication Devices.

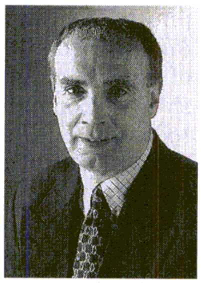
North Light Communications—Tim Ayers to president of North Light Public Affairs; formerly vice president for communications at the Cellular Telecommunications Industry Association.

Superconductor Technologies, Inc.—Richard M. Johnston to the board of directors; currently vice president of investments and a director of The Hillman Co. Also, Joseph C. Manzinger to the board of directors; currently vice president of The Hillman Co.

Piezo Technology, Inc. (PTI)—Carl Sokoloski to instrumentation team manager; formerly worked in instrumentation at CTS Reeves Frequency Products. Also, Robin Sokoloski to production manager; formerly operations manager for CTS Reeves in the area of quartz frequency-control devices.

Trompeter Electronics—Lyn Bresnen to new business development manager; formerly western regional sales manager for Johnson Components.

BRESNEN



ROGERS

Amplifier Research (AR)—Richard R. Rogers to vice president of the AMREP division; formerly worked in human resources at Rohm & Haas.

Goodfellow Corp.—John C. Friday to key accounts manager; formerly sales representative for American Contractors Equipment.

Short Courses

Digital Signal Processing Applications in Wireless Communications
September 29-October 1 (Los Angeles, CA)
UCLA Extension, Department of Engineering
10995 Le Conte Ave., Suite 542
Los Angeles, CA 90024
(310) 825-3344, FAX: (310) 206-2815
e-mail: mhenness@unex.ucla.edu
Internet: http://www.unex.ucla.edu/shortcourses

Mobile Wireless Communications over the Fading Channel: Third-Generation Techniques
October 4-8 (Washington, DC)
The George Washington University
Continuing Engineering Education Program
Washington, DC 20052
(800) 424-9773, FAX: (202) 872-0645
Internet: http://www.gwu.edu/~ceep

Traffic Engineering for Telecommunication
October 11-13 (Madison, WI)
Department of Engineering Professional Development
University of Wisconsin-Madison
432 North Lake St.
Madison, WI 53706
(800) 462-0876, FAX: (608) 262-3160
e-mail: custserv@epd.engr.wisc.edu
Internet: http://epdd.engr.wisc.edu

Introduction to Datacomm and Networks
October 12-15 (Washington, DC)
Learning Tree International
Reston Town Center
Reston, VA 20190-5630
(800) 843-8733, FAX: (800) 709-6405
Internet: http://www.learningtree.com

RF Wireless System Design Fundamentals
October 18-20 (Dallas, TX)
Besser Associates
201 San Antonio Circle, Bldg. E, Suite 280
Mountain View, CA 94040
(650) 949-3300, FAX: (650) 949-4400
e-mail: info@bessercourse.com
Internet: http://www.bessercourse.com

Digital Communications—An Introduction
November 8-9 (University of Oxford, Oxford, UK)
International Electronics and Communications Courses 1999
Oxford University
+44 (0) 1865 288170, FAX: +44 (0) 1865 288163
e-mail: linda.tonge@conted.ox.ac.uk
Internet: http://www.conted.ox.ac.uk/ousep.html

Meetings

The Seventh Annual International Conference on Power Requirements for Mobile Computing and Wireless Communications (Power 99)
October 3-6 (Santa Clara Westin Hotel & Convention Center, Santa Clara, CA)
Power 99
One Longwater Circle
Norwell, MA 02061-1620
(781) 792-2669, FAX: (781) 871-4871
e-mail: power99@gigaweb.com

Internet Device Builder Show and Conference
October 4-8 (Javits Center, New York, NY)
Penton Media, Inc.
611 Route 46 West
Hasbrouck Heights, NJ 07604
(201) 393-6259, FAX: (201) 393-6297
e-mail: tcusumano@penton.com
Internet: http://www.iDeviceBuilder.com

Annual Exhibition and Conference of the Low Power Radio Association
October 20-21 (Birmingham, UK)
Low Power Radio Association
West Yorkshire HX2 6HS, UK
+44 (0) 1422 88 64 63, FAX: +44 (0) 1422 88 69 50
e-mail: info@lpra.org
Internet: http://www.ipra.org.

The 32nd International Symposium on Microelectronics
October 26-28 (Chicago, IL)
International Microelectronics And Packaging Society
1850 Centennial Park Dr., Suite 105
Reston, VA 20191-1517
(888) GO IMAPS, FAX: (703) 758-1066
Internet: http://www.imaps.org

Wireless I.T. 99 (Information Technologies)
November 2-4 (Santa Clara Convention Center, Santa Clara, CA)
Cellular Telecommunications Industry Association
1250 Connecticut Ave., NW
Washington, DC 20036
(800) 334-6147, FAX: (415) 979-2250
Internet: http://www.wow-com.com

Expo Comm Brazil 99
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4th European Symposium on Electromagnetic Compatibility (EMC Europe 2000)
September 11-15, 2000 (Brugge, Belgium)
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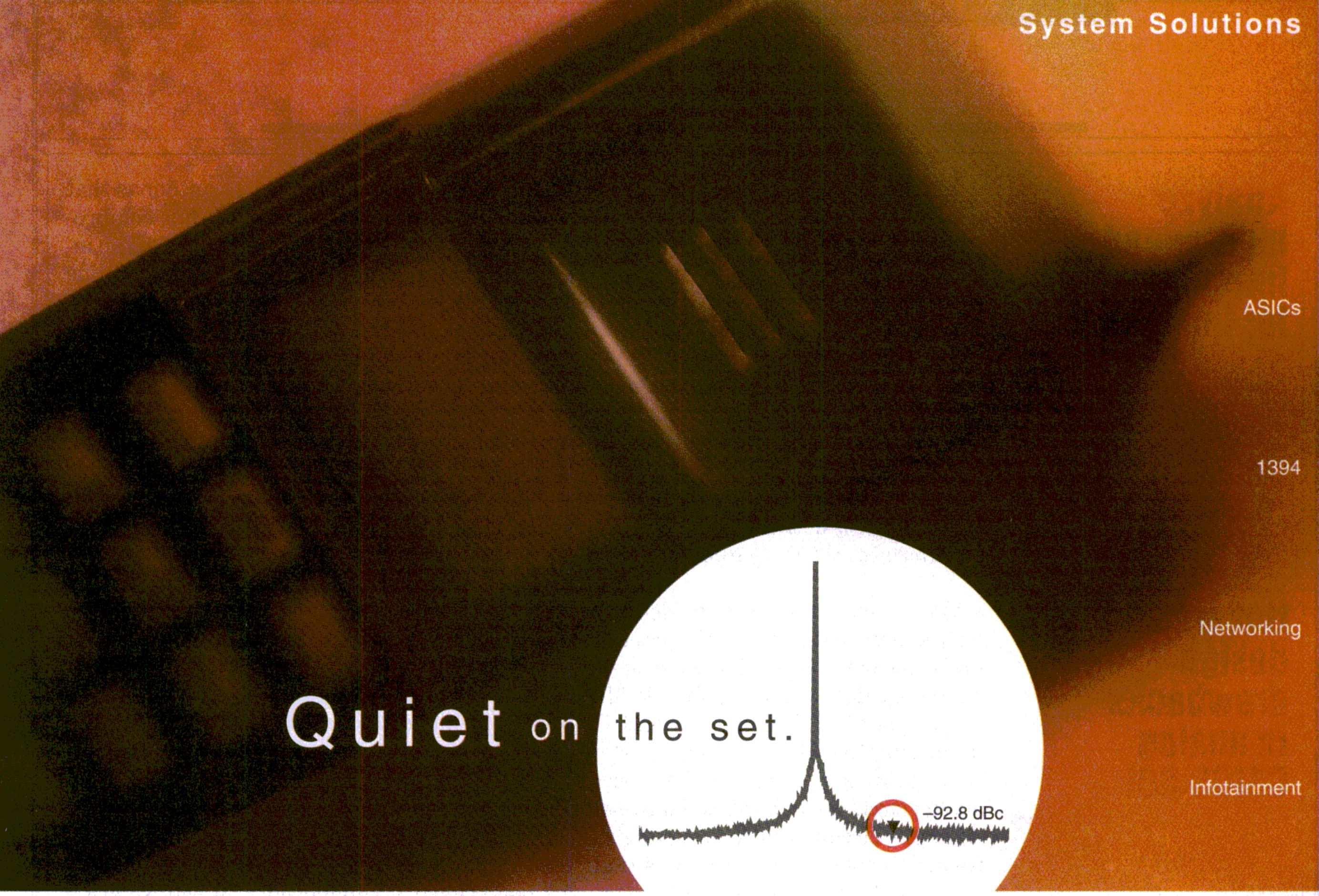
e-mail: johan.catrysse@kh.khbo.be
Internet: http://www.ti.kivi.be/conf/emc2000 htm
Abstract deadline: December 1, 1999



Spurious Noise

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Embedded

Packaging

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Foundry

SL Series PLL Frequency Synthesizers

Single PLLs

Part Number	fIN Max	Icc (mA)	Vcc (V)
MB15E03SL	1.2 GHz	2	2.7
MB15E05SL	2.0 GHz	3	2.7
MB15E07SL	2.5 GHz	3.5	2.7

Dual PLLs

Part Number	fIN Max	Icc (mA)	Vcc (V)
MB15F02SL	1.2 GHz 0.5 GHz	1.8 1.2	2.7 2.7
MB15F03SL	1.75 GHz 0.6 GHz	2.3 1.2	2.7 2.7
MB15F07SL	1.1 GHz 1.1 GHz	2.5 2.5	2.7 2.7
MB15F08SL	2.5 GHz 1.1 GHz	4.4 2.6	2.7 2.7

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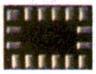
BCC



FUJITSU

FUJITSU MICROELECTRONICS, INC.

CIRCLE NO. 300

Analyze high-speed interconnects for induced EMI

In today's high-speed circuit designs, long interconnects often act as antennas that pick up noise [electromagnetic interference (EMI)] from other systems, thus compromising proper operation and performance. Analyzing the frequency response of the interconnects to incident-field coupling requires solving some complex transmission-line equations, an expensive and time-consuming task in the case of large systems. A new technique for the simulation of transmission lines excited by incident fields was developed by Roni Khazaka and Michel Nakhla, both of the Department of Electronics, Carleton University (Ottawa, Canada). The method involves a model-reduction technique called complex frequency hopping (CFH), which is said to be efficient and reliable for interconnect analysis and 100 to 1000 times faster than conventional simulation methods when run on a computer. It is suitable for VLSI and printed-circuit-board (PCB) designs involving a large number of interconnects. See "Analysis of High-Speed Interconnects in the Presence of Electromagnetic Interference," *IEEE Transactions on Microwave Theory and Techniques*, Vol. 46, No. 7, July 1998, p. 940.

Understand design drawbacks in analog CMOS SOI

Complementary-metal-oxide-semiconductor (CMOS) Silicon-on-Insulator (SOI) is a well-known and widely used process for fabricating digital devices, but it may present problems when applied to the design of VLSI analog circuits. According to studies conducted by Bernard M. Tenbroek, Michael S.L. Lee, and William Redman-White of the Department of Electronics and Computer Science, University of Southampton (Southampton, UK) and John T. Bunyan and Michael J. Uren of the Defense Evaluation and Research Agency (Malvern, UK), the performance of certain types of CMOS analog circuits can be affected by self heating and thermally induced mismatches. For example, circuits that rely strongly on matching, such as current mirrors and digital-to-analog converters (DACs), are affected strongly by self heating and thermal coupling in the SOI process. SOI has very poor thermal conductivity in the isolating buried-oxide layer causing a localized temperature rise that affects a device's channel current. This can result in slowed operation in current mirrors and less accuracy in DACs. See "Impact of Self-Heating and Thermal Coupling on Analog Circuits in SOI CMOS," *IEEE Journal of Solid-State Circuits*, Vol. 33, No.7, July 1998, p. 1037.

Virtual prototyping improves EDA tools

Conventional electronic-design-automation (EDA) tools leave much to be desired when it comes to modeling high-frequency circuits such as those in the RF and mixed-signal domains. Electromagnetic-compatibility (EMC) problems, for example, are modeled with classical numerical EDA tools using algorithms that are computationally intensive and, in some cases, result in incomplete models. A technique called virtual prototyping conceived by Emil M. Petriu and Marius Cordea of the School of Information Technology of the University of Ottawa (Ottawa, Canada) and Dorina C. Petriu of the Department of Systems and Computer Engineering at Carleton University (Ottawa, Canada) uses neural-network (NN) models for EM-field modeling. The virtual prototyping environment using NNs offers a new EDA tool that will enable designers to interactively test complex electronic systems on an enhanced-reality virtual workbench. See "Virtual Prototyping Tools for Electronic Design Automation," *IEEE Instrumentation & Measurement Magazine*, Vol. 2, No. 2, June 1999, p. 28.

Micromachining makes high-quality RF passives

Passive components such as inductors and capacitors can be fabricated directly on silicon (Si) substrates in conventional planar monolithic microwave integrated circuits (MMICs). However, these lumped passive-elements are often plagued by parasitics such as series resistance, stray capacitance, and others. High-quality passives with drastically reduced parasitics can be made through certain Si micromachining techniques according to Yanling Sun of Bell Labs Utrecht, Lucent Technologies (3431 JZ Nieuwegein, The Netherlands), Joseph L. Tauritz of the Delft University of Technology (2600 GB Delft, The Netherlands), and Roel G.F. Baets of the University of Gent (B-9000 Gent, Belgium). The key is a front-side selective etch of the Si substrate, and leads to high-Q reactive passives for high-frequency use (up to approximately 15 GHz). See "Micromachined RF Passive Components and Their Applications in MMICs," *International Journal of RF and Microwave Computer-Aided Engineering*, Vol. 9, No. 4, July 1999, p. 310.

Predict Phase-Noise Effects In Digital Communications Systems

The disruptive effects of phase noise on digital communications systems can be predicted and overcome through these straightforward calculations.

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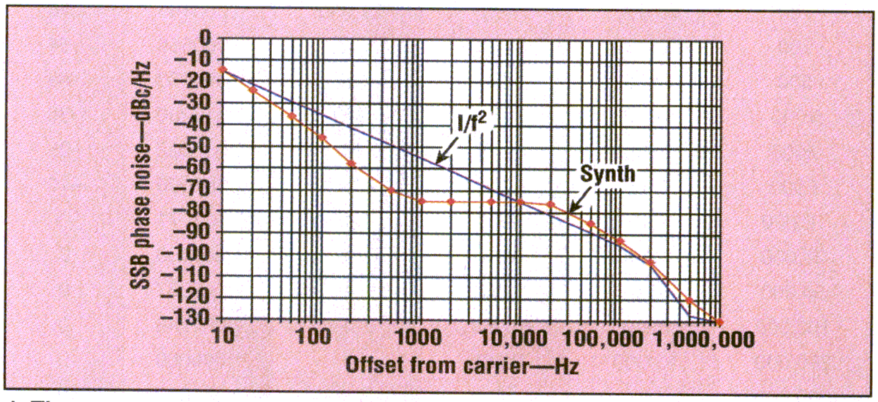
COMMUNICATIONS providers are reaping the benefits of replacing older amplitude-modulation (AM) and frequency-modulation (FM) systems with modern digital communications systems. These newer systems are based on quadrature-amplitude-modulation (QAM) and quadrature-phase-shift-keying (QPSK) methods that encode information in the amplitude and phase of the signal. Unfortunately, frequency-translation stages in most communications systems can still impart unwanted phase noise onto the desired signal. To prevent deleterious effects of phase noise, engineers must plan for it in their link budgets. What follows is an effective procedure for calculating the impairment loss due to phase noise.

Three pieces of information are needed to calculate phase-noise impairment loss:

1. The power-spectral density (PSD) versus frequency of the phase noise {L(f) [in dB]}.
2. The receiver carrier-recovery-loop response versus frequency [H(f) (in dB)].
3. The theoretical SNR (SNR_{theory}) required for the desired bit-error rate (BER).

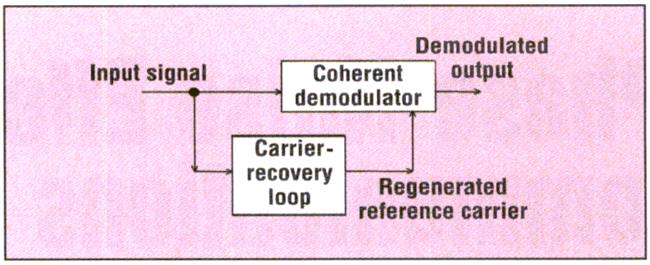
The first step is to calculate the untracked jitter by filtering the PSD of the phase noise by the receiver carrier-recovery-loop highpass error response, H(f). Expressed in dB, the result, L(f) + H(f) = F(f) dB, is the PSD of the filtered phase noise or untracked carrier jitter.

The next step involves finding the total noise power caused by the phase noise that is actually "seen" by the demodulator, by numerically integrating over frequency and multiplying by two (add 3 dB) to make the conversion from single-sideband (SSB) noise to double-sideband (DSB)



1. These two examples of phase noise as a function of offset frequency were generated for a "Synth" source and a white-noise generator with $1/f^2$ response.

noise. The channel additive white Gaussian noise (AWGN) that can be tolerated must be reduced by this amount so that the total noise does not exceed SNR_{theory} for the applicable type of modulation. The amount that AWGN must be reduced is the impairment loss due to the phase noise.



2. This simplified digital demodulator includes a carrier-recovery loop and coherent demodulator.

Phase noise is usually characterized by a mask or plot of the SSB PSD versus offset frequency relative to the total carrier power (in units of dBc/Hz). Due to the presence of a second sideband opposite the carrier, the true phase-noise PSD S(f) = L(f) × 2 or S(f) = L(f) + 3 dB. Figure 1 shows two typical phase-noise curves. The curve labeled "Synth" represents the performance of a typical low-cost local-oscillator (LO) synthesizer. The $1/f^2$ curve, which was generated by feeding white noise into an FM generator, has a –6 dB per octave negative slope over most of its range and is often used for specifying phase-noise performance. Although both curves pass through the same point at –75 dBc/Hz at a 10-kHz offset, they each represent completely different phase-noise characteristics. It can be seen from these two examples that specifying phase noise as only a single value at a particular offset frequency may not provide sufficient information.

In the receiver, the carrier-recovery-loop response of the digital demodulator has a major impact on the impairment loss. The carrier-recovery loop acts as a highpass filter to the incoming phase noise. Figure 2 shows a simplified diagram of a digital receiver coherent demodulator. The carrier-recovery loop regenerates a reference carrier from the average phase of the input signal. Slow phase variations that are within the carrier-recovery-loop bandwidth appear on the recovered reference carrier and on the input signal and are therefore ignored in the demodulation process. Rapid phase variations that are above the carrier-loop bandwidth are not tracked by the carrier-recovery loop and do not appear on the reference carrier. These variations are then indistinguishable from the desired signal-symbol variations and add to the uncertainty in determining the correct symbol. Figure 3 illustrates the effect of phase variations that are not tracked by the carrier-recovery loop on a 64QAM constellation.

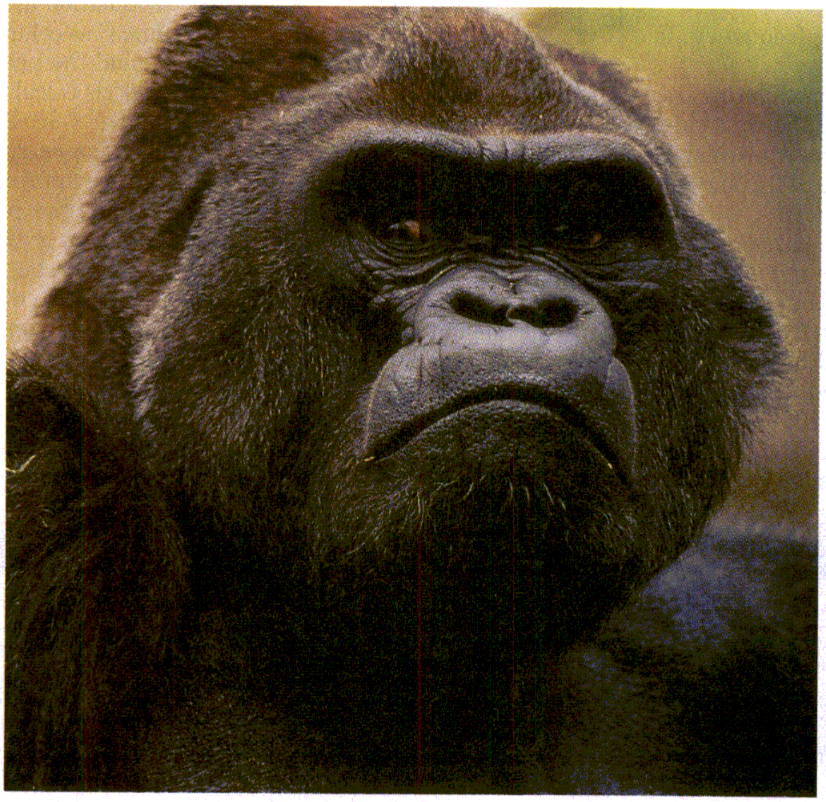
The design of the receiver-carrier-recovery loop requires a compromise in the selection of bandwidth. A wider-loop bandwidth more effectively tracks and removes undesired phase-noise variations. However, a bandwidth that is too wide will cause the channel AWGN and the desired signal-phase variations to appear as additional "self-noise" on the reference carrier, reducing the capability to demodulate the desired symbols. Typically, the carrier-recovery loop is a first- or second-order loop with a bandwidth of approximately 1/100 of the desired symbol rate, yielding a highpass filter with 6 or 12 dB per octave slope.

If the receiver carrier-recovery-loop bandwidth, order, and damping factor are known, it is possible to calculate the loop response [H(f)]. If not, the carrier-recovery-loop response can be measured with the test setup of Fig. 4. The straightforward procedure involves setting the transmitter to an output frequency that is different from that of the receiver input frequency, and using a mixer and LO to translate or convert the transmitter output frequency to the receiver input frequency. The LO is supplied by an FM signal generator which can be used to impart FM and phase noise onto the LO and, therefore, on to the digitally modulated signal.

For narrowband FM (β < 1), the modulation index (β) and the peak deviation (Dev) can be calculated from the first sideband to the carrier ratio seen on the spectrum analyzer as:

Table 1: Demodulation carrier-recovery-loop response with the HP 89441

A	B	C	D	E	F
Offset frequency (Hz)	Measured deviation for a BER of 1×10^{-6} (Hz peak)	β = B/A	Phase modulation (deg. peak) = (C × 360)/2π	Carrier-loop response numerical F(f) = 4/D	Carrier-loop response (dB) = 20logE
10	400,000	40,000	2,292,994	0.000002	–116
20	200,000	10,000	573,248	0.000007	–104
50	80,000	1600	91,720	0.000041	–88
100	40,000	400	22,930	0.000163	–76
200	20,000	100	5732	0.000650	–64
500	8000	16	917	0.004063	–48
1000	4000	4	229	0.016250	–36
2000	2300	1.15	66	0.056522	–25
5000	1000	0.2	11	0.325000	–10
10,000	700	0.07	4	0.928571	–1
20,000	1300	0.065	4	1.000000	0
50,000	3100	0.062	4	1.048387	0
100,000	6300	0.063	4	1.031746	0
200,000	13,000	0.065	4	1.000000	0
500,000				1.000000	0
1×10^6				1.000000	0



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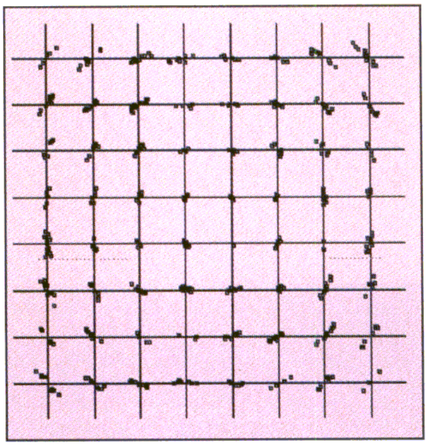
$$\beta = 2\times10^{(R_{dB}/20)} \qquad (1)$$

$$Dev = bf_m \qquad (2)$$

where:
β = the FM modulation index,
R_{dB} = the ratio of the first sideband to the carrier (in dB),
Dev = the peak ±FM deviation (in Hz), and
f_m = the modulating frequency (in Hz).

For example, if the modulating frequency is 10 kHz and the first sideband is seen on the spectrum analyzer at –20 dBc, then β = 0.2 and the peak deviation is ±2 kHz.

To characterize the carrier-recovery loop of the receiver unit under test (UUT), AWGN should be adjusted for a convenient reference BER such as 1×10^{-6} without LO FM. Then, the AWGN should be reduced by 3 dB (which improves the BER), and the LO FM deviation should be increased until the BER is degraded back to 1×10^{-6}. The amount of LO FM deviation present should be measured with a peak deviation meter or spectrum analyzer, and recorded over a range of modulating frequencies from 10 Hz to near the system symbol rate.



3. A constellation diagram can be used to graphically demonstrate phase-noise effects in a system using 64QAM.

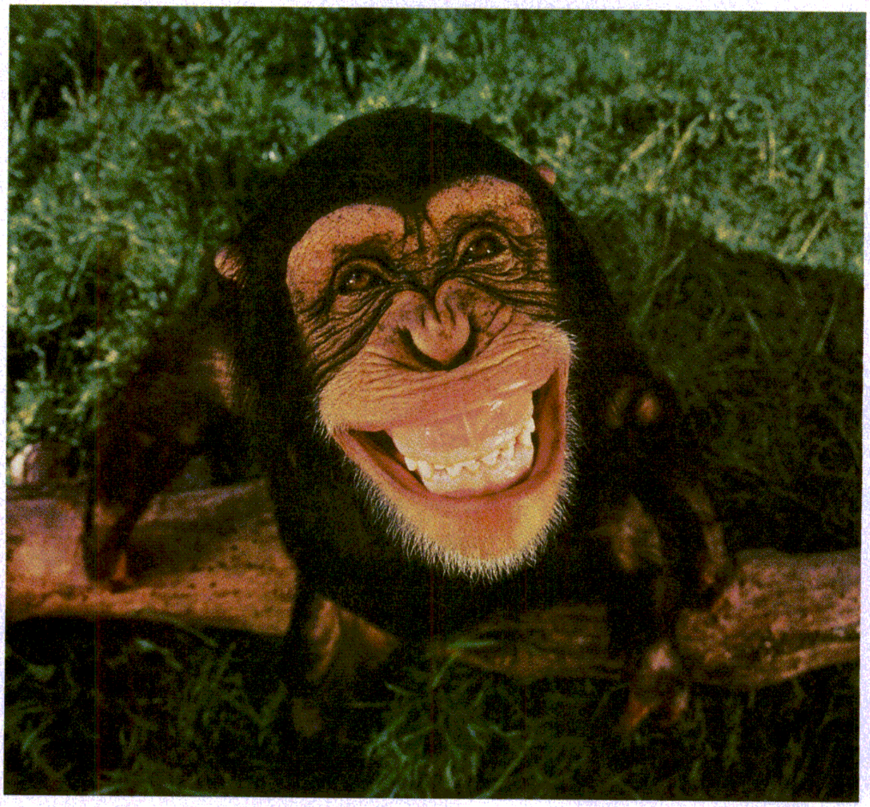
As an example, Table 1 illustrates the 64QAM digital demodulator in the HP 89441 vector signal analyzer analyzed at a symbol rate of approximately 5 Msymbols/s.

Column A lists the modulating or offset frequency, and column B lists the measured peak deviation required to degrade the BER to 1×10^{-6}. The modulation index is calculated in column C as column B divided by column A, and the phase modulation (in deg. peak) is calculated in column D as column C × 360/2π. In this example, the effective carrier-recovery-loop bandwidth is approximately 10 kHz. Above 10 kHz, the phase modulation (PM) needed to create a BER of 1×10^{-6} is relatively constant and equal to 4 deg. peak. Above 10 kHz, the carrier-recovery loop provides no rejection (i.e., the numerical carrier-loop rejection is equal to 1 or 0 dB). At lower frequencies, the carrier-recovery loop rejects the PM, so larger amounts of PM are required to degrade the BER

Table 2: White noise ($1/f^2$) integration to SNR

A Offset frequency (Hz)	B $1/f^2$ PSD (dBc/Hz)	C Carrier-loop response (dB)	D Filtered-noise density (dBc/Hz) = B+C	E Lower-frequency integration limit (Hz)	F Upper-frequency integration limit (Hz)	G Integration bandwidth (Hz) = F-E	H Double-sideband noise power (dBc) = 3+D+10logG	I Power (numerical) = $10^{(H/10)}$
10	–15	–116	–131	0	15	15	–116.0	2.5×10^{-12}
20	–21	–104	–125	15	30	15	–110.0	1.0×10^{-11}
50	–29	–88	–117	30	75	45	–97.3	1.9×10^{-9}
100	–35	–76	–111	75	150	75	–89.0	1.2×10^{-9}
200	–41	–64	–105	150	300	150	–80.0	1.0×10^{-8}
500	–49	–48	–97	300	750	450	–67.3	1.9×10^{-7}
1000	–55	–36	–91	750	1500	750	–59.0	1.3×10^{-6}
2000	–61	–25	–86	1500	3000	1500	–51.2	7.6×10^{-6}
5000	–69	–10	–79	3000	7500	4500	–39.2	1.2×10^{-4}
10,000	–75	–1	–76	7500	15,000	7500	–33.9	4.1×10^{-4}
20,000	–81	0	–81	15,000	30,000	15,000	–36.3	2.4×10^{-4}
50,000	–89	0	–89	30,000	75,000	45,000	–39.0	1.2×10^{-4}
100,000	–95	0	–95	75,000	150,000	75,000	–43.0	5.0×10^{-5}
200,000	–104	0	–104	150,000	300,000	150,000	–49.2	1.2×10^{-5}
500,000	–127	0	–127	3×10^5	7.5×10^5	450,000	–67.5	1.8×10^{-7}
1×10^6	–130	0	–130	7.5×10^5	2.5×10^5	1,750,000	–64.6	3.5×10^{-7}
							Sum = –30.2	0.03 radians RMS
								1.78 deg. RMS

Notes: Calculated equivalent SNR of the impairment = –30.2 dBc
Theoretical threshold SNR = –21.5 dBc
SNR required with the impairment = –22.1 dBc
Impairment loss = 0.6 dB



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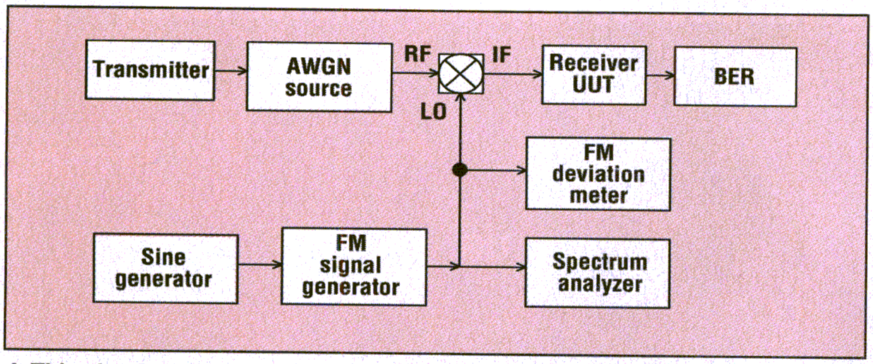


4. This simple test system can be used to measure a communications system's carrier-loop response with a spectrum analyzer or FM deviation meter.

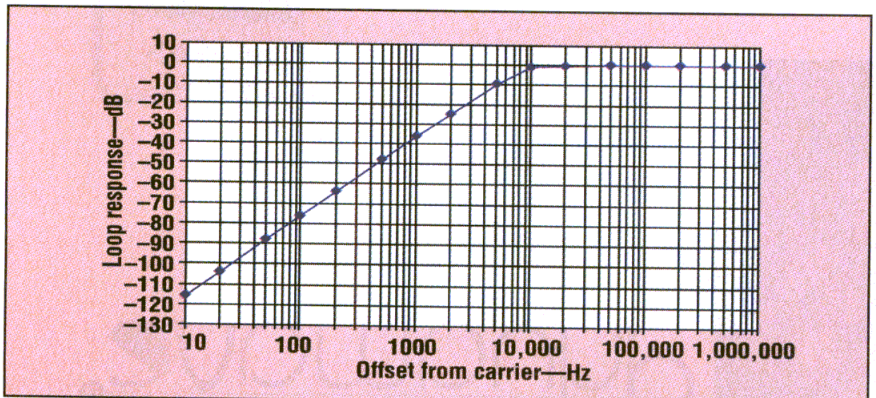
to 1×10^{-6}. The numerical value of the carrier-loop rejection at each frequency is calculated in column E as 4 divided by column D. Last, the carrier-recovery-loop rejection is converted to decibels in column F and is calculated as column F (in dB) = 20 $\log^{10}$ (column E).

The data of Table 1 are graphed in Fig. 5 as the carrier-recovery-loop response versus offset frequency. The results are typical of a well-damped second-order loop. The HP 89441 does not actually use a carrier-recovery loop for demodulation. Instead, it computes the average phase over the duration of the data sample so the "effective" carrier-recovery-loop bandwidth is a function of the data-sample duration.

Table 2 illustrates the filtering of the phase noise by the carrier-recovery-loop response and the numerical integration of L(f) + H(f) for the example $L(f) = 1/f^2$. Column B lists the unfiltered phase noise while column C lists the carrier-recovery-loop response that was calculated in Table 1. Column D lists the filtered phase-noise PSD F(f) calculated as column B + column A. This is the untracked jitter or phase noise that is actually "seen" by the demodulator. Note that the highpass filter response of the carrier-recovery loop has effectively attenuated close-in phase noise.

Columns E and F list the limits of integration around each frequency range. These limits are selected to be contiguous around the center frequencies. The integration bandwidth for each range is then calculated in column G as column F – column E. The filtered FL(f) in each bandwidth is converted to an integrated power by adding a bandwidth factor of $10\log_{10}$ (BW) to each value of FL(f).

This numerical integration technique makes the assumption that the FL(f) is essentially constant within each frequency range. The power in each range is also increased by 3 dB to account for the conversion from SSB power to DSB. Thus, the DSB inte-



5. The carrier-recovery-loop response was measured with an HP 89441 vector signal analyzer from Hewlett-Packard Co. (Palo Alto, CA) for comparison to calculated results.

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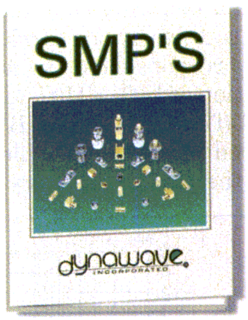

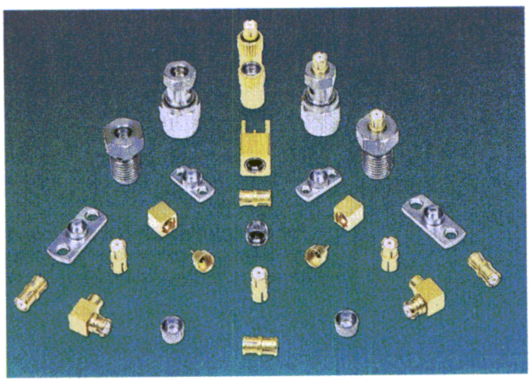

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grated-noise power in each frequency range is calculated in column H as column D + 3 + 1010g^{10} (column G). Integrated noise powers expressed in column H (in dBc) are converted to numerical power values in column I using column I = $10^{(\text{column H}/10)}$. Integration over the entire frequency range is then performed by adding all of the powers together in column I. The total integrated power is converted to root-mean-square (RMS) phase jitter (in radians) and converted to degrees by:

$$J_{RMS\ degrees} = (PWR_{TOT})^{0.5} \qquad (3)$$

$$J_{RMS\ degrees} = J_{RMS\ radians} \times 360/2\pi \qquad (4)$$

where:

J_{RMS} = the RMS phase jitter, and

PWR_{TOT} = the total integrated power (the summation of column I in Table 2).

In this example, the total integrated phase jitter is 1.78-deg. RMS.

The total power is converted back to SNR_{imp} using:

$$SNR_{imp}\ (dB) = 10\log_{10}(PWR_{TOT}) \qquad (5)$$

where:

SNR_{imp} = the SNR caused by the phase noise.

In this example, the resulting SNR_{imp} = 30.2 dB, meaning that the untracked jitter or filtered integrated phase noise creates an impairment with an equivalent SNR of 30.2 dB.

To verify the procedure, a 64QAM signal with phase noise that has a PSD that is the same as the $1/f^2$ curve was applied to the input of the HP 89441. The instrument reported an SNR of 30.5 dB, agreeing closely with the calculated value of 30.2 dB.

To calculate the impairment loss in dB, the theoretical threshold SNR (SNR_{theory}) for the particular type of modulation and FEC must be known. For the International Telecommunications Union specified ITU-T J.83B 64QAM and FEC used in this example, an SNR_{theory} of 21.5 dB is needed to achieve a BER of 1×10^{-6} without impairments that are present. Since the phase-noise impairment adds noise power to the signal, the channel AWGN at the Rx input must also be reduced so that the total noise power remains equivalent to 21.5 dB. Thus:

$$P_{Nin} = P_{Ntheory} - P_{NImp} \quad (6)$$

where:

P_{Nin} = the allowable AWGN input power with the impairment present,

$P_{Ntheory}$ = the allowable AWGN input power without impairment, and

P_{Nimp} = the noise power due to the phase-noise impairment.

When these noise powers are expressed as SNR values in dBc, the equation becomes:

$$SNR_{in} = -10\log_{10}\left[10^{(-SNRtheory/10)}\right] - \left[10^{(-SNRimp/10)}\right] \quad (7)$$

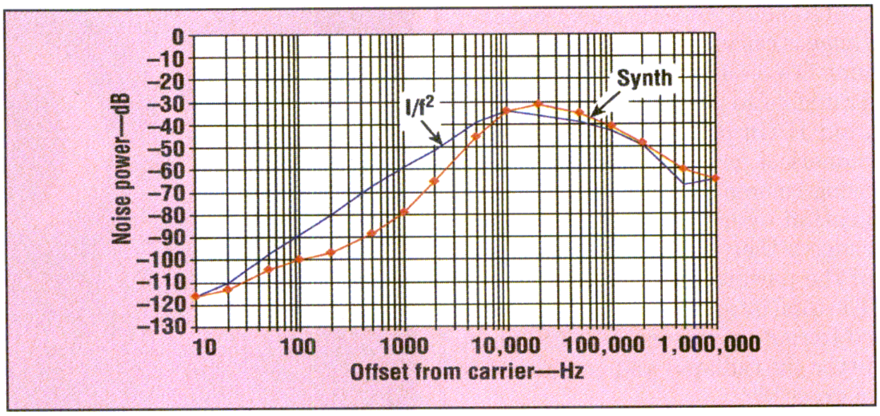
where:

SNR_{in} = the allowable input AWGN SNR dB with the impairment present,

SNR_{theory} = the allowable input AWGN SNR dB without impairment, and

SNR_{imp} = the SNR in decibels due to the phase-noise impairment.

In this example:



6. This plot of noise power as a function of offset frequency was generated for the two sources of Fig. 1.

$$SNR_{in} = -10\log_{10}\left[10^{(-21.5/10)}\right] - \left[10^{(-30.2/10)}\right]$$

$$SNR_{in} = 22.1 \text{ dB}$$

Thus, without phase noise present, the input AWGN SNR must be 21.5 dB for a BER of 1×10^{-6}. When the phase-noise impairment is added, the input AWGN must be reduced so that the SNR is 22.1 dB to maintain the same BER. The difference (0.6 dB) is the impairment loss due to phase noise.

Alternatively, impairment loss can also be calculated directly from the decibel difference between SNR_{theory}

Table 3: "Synth" source integration to SNR

A	B	C	D	E	F	G	H	I
Offset frequency (Hz)	SSB phase-noise spec density (dBc/Hz)	Carrier-loop response (dB)	Filtered-noise spec density (dBc/Hz) = B+C	Lower-frequency integration limit (Hz)	Upper-frequency integration limit (Hz)	Integration bandwidth (Hz) = F-E	Double-sideband noise power (dBc) = 3+D+10logG	Power (numerical) = $10^{(H/10)}$
10	–15	–116	–131	0	15	15	–116.0	2.5×10^{-12}
20	–24	–104	–128	15	30	15	–113.0	5.0×10^{-12}
50	–36	–88	–124	30	75	45	–104.3	3.7×10^{-11}
100	–46	–76	–122	75	150	75	–100.0	9.9×10^{-11}
200	–58	–64	–122	150	300	150	–97.0	2.0×10^{-10}
500	–70	–48	–118	300	750	450	–88.3	1.5×10^{-9}
1000	–75	–36	–111	750	1500	750	–79.0	1.2×10^{-8}
2000	–75	–25	–100	1500	3000	1500	–65.2	3.0×10^{-7}
5000	–75	–10	–85	3000	7500	4500	–45.2	3.0×10^{-5}
10,000	–75	–1	–76	7500	15,000	7500	–33.9	4.1×10^{-4}
20,000	–76	0	–76	15,000	30,000	15,000	–31.2	7.5×10^{-4}
50,000	–85	0	–85	30,000	75,000	45,000	–35.1	3.1×10^{-4}
100,000	–93	0	–93	75,000	150,000	75,000	–41.0	8.0×10^{-5}
200,000	–103	0	–103	150,000	300,000	150,000	–48.2	1.5×10^{-5}
500,000	–120	0	–120	3×10^5	7.5×10^5	450,000	–60.5	9.0×10^{-7}
1×10^6	–130	0	–130	7.5×10^5	2.5×10^6	1,750,000	–64.6	3.5×10^{-7}
							Sum = –28.0	1.6×10^{-3}
								0.0400 radians RMS
								2.2918 deg. RMS

Notes: Calculated equivalent SNR of the impairment = –28.0 dBc
Theoretical threshold SNR = –21.5 dBc
SNR required with the impairment = –22.6 dBc
Impairment loss = 1.1 dB

and SNR_{imp} as follows:

$$L_{imp} = -10\log_{10}[1 - 10^{(-D/10)}] \quad (8)$$

where:

$\Delta = SNR_{theory} - SNR_{imp}$,

$SNR_{imp} = SNR_{theory}$ (in dB),

SNR_{theory} = the allowable input AWGN SNR (in dB) without impairment,

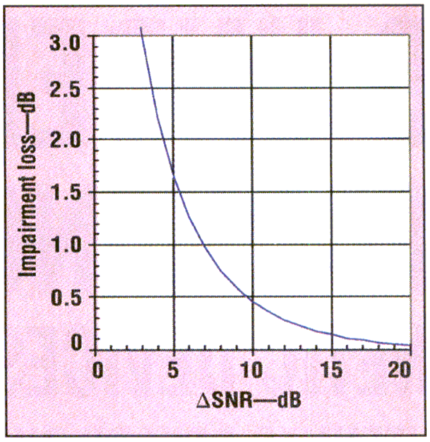
SNR_{imp} = the SNR (in dB) due to the phase-noise impairment, and

L_{imp} = the impairment loss (in dB).

Equation 8 is shown graphically in Fig. 7. The figure shows that SNR_{imp} must be at least 10 dB higher than SNR_{theory} to hold the impairment loss at 0.5 dB or less. Note that the impairment loss increases rapidly as SNR_{imp} approaches SNR_{theory}. In this example, 30.2 dB – 21.5 dB = 8.7 dB, which corresponds to an impairment loss of 0.6 dB.

Note that the modulation used in this example includes forward-error correction (FEC), which provides 6-dB coding gain. The value of SNR_{theory} required for uncoded 64QAM is approximately 27 dB, so per eq. 8, the uncoded impairment loss caused by the phase noise in this example is 2.7 dB. Table 4 shows the SNR_{theory} for BER = 1×10^{-6} and the SNR_{imp} for a 1-dB impairment loss for various types of uncoded modulation. Equation 8 can be used to determine the impairment loss for other values of SNR_{imp} as well.

The same procedure can be used to determine the impairment loss due to phase noise for any arbitrary phase-noise curve. As an example, the curve labeled "Synth" was analyzed in Table 3. The SNR_{imp} due to the phase noise was calculated as 28 dB. The required input channel SNR must be 22.6 dB, representing an impairment loss of 1.1 dB.



7. Impairment loss was plotted as a function of change Δ in signal-to-noise ratio (SNR) to graphically demonstrate eq. 8. Impairment loss increases rapidly as SNR_{imp} approaches SNR_{theory}.

The data shown in column H of Table 3 and graphed in Fig. 6 can provide useful insight into identifying the most-significant frequency regions causing impairment loss. For example, Fig. 6 shows that for L(f) = 1/ f_2, the region around 10 kHz has the most-significant noise-power contribution. For the case of the "Synth" L(f) curve, the entire 1-to-10-kHz region is significant. In order to improve system performance relative to phase noise, the phase noise can be reduced or the rejection provided by the receiver carrier-recovery loop in these regions can be increased by increasing the loop bandwidth. Loop damping should also be investigated since an under-damped loop exhibits peaking, which will actually amplify rather than reject phase noise.

The calculations that are shown rely on numerical integration. Three or more frequency segments per decade should be used in order to minimize inaccuracies.

In addition, the calculation relies on several assumptions. The first assumption is that the amount of phase noise must be sufficiently small so that the uncoded impairment loss does not exceed approximately 2 dB. This is because the phase-noise impairment adds in a linear fashion to the channel AWGN and the carrier-recovery loop does not lose lock or cycle slip. The second assumption is that the phase noise's probability density function (PDF) is assumed to be approximately Gaussian in nature.

Another limitation concerns the different statistical distribution of errors caused by phase-noise versus errors caused by AWGN. AWGN causes errors that are independent of each other (i.e., they are randomly distributed in time). If the phase noise exhibits a concentration of power at frequencies below the system-symbol rate, the errors will no longer be randomly distributed. This may reduce the error-correcting capability of the system FEC, resulting in a coded BER that is higher compared to the coded BER created by an equivalent amount of AWGN. If this type of phase noise is expected in a system, the FEC should be designed to be effective against burst-related errors.

Even with these limitations, this technique provides a useful and simple way to calculate the impairment caused by phase noise in a digital communications system. The result is valuable insight as to what changes may be needed to reduce the impairment, right from the beginning. ••

Table 4: Theoretical SNR for a BER of 10^{-6} and impairment SNR for a 1-dB impairment loss

Uncoded modulation	Approximate SNR_{theory} for BER = 10^{-6} (dB)	SNR_{imp} for 1-dB impairment loss (dB)	Phase jitter for 1-dB impairment loss (RMS deg.)
QPSK	14	21	5.1
16QAM	21	28	2.3
64QAM	27	34	1.1
256QAM	33	40	0.57
1024QAM	39	47	0.26

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Develop A High-Q Oscillator For Millimeter-Wave Applications

The use of an high-Q resonant Invar cylinder helps to stabilize the output frequency of a Gunn diode in this low-cost millimeter-wave oscillator.

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SOLID-STATE oscillators are the signal sources of virtually all modern microwave and millimeter-wave systems. The stability of the oscillator is critical, since it can determine the overall performance of the system. Traditionally, the stability of uncompensated Gunn diode oscillators (approximately 10^{-4}) has been poor compared to crystal-stabilized oscillators and, due to its frequency variations, the Gunn oscillator has been limited in its breadth of applications. Some measures, such as the use of phase-locked loops (PLLs), have improved the stability of Gunn diode oscillators, but with added cost and complexity. In comparison, the use of high-quality-factor (high-Q) resonant elements can help stabilize the output frequency of a Gunn oscillator while maintaining simplicity and low cost.

In particular, a high-Q Invar cylinder was used as a stabilizing resonant element in a millimeter-wave Gunn diode oscillator. The oscillator was designed for use in the TE_{001} resonant mode without additional temperature detection and compensation. Compared to unstabilized Gunn oscillators, the Invar design showed relative good long-term stability of 10^{-6} and improved phase noise at 36.8 GHz.

To explain the design of this millimeter-wave oscillator, it is necessary to review some engineering fundamentals. For example, the stability of a negative-resistance oscillator with one port depends upon:

$$Y_L + Y_D = 0 \qquad (1)$$

where:

Y_L = the load admittance, and

Y_D = the admittance of the active device.

Parameters Y_L and Y_D can be divided into the real and imaginary parts of the conductance (G) and susceptance (B) through:

$$G(\omega, Vac) = G_D(\omega, Vac) + G_L(\omega) = o \qquad (2)$$

$$B(\omega, Vac) = B_D(\omega, Vac) + B_L(\omega) = 0 \qquad (3)$$

where:

G_L = the load conductance,

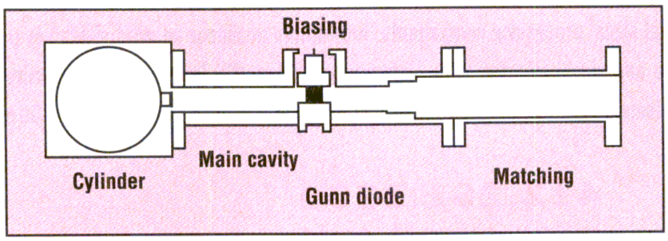
G_D = the device conductance,

B_L = the load susceptance, and

B_D = the device susceptance.

To meet the conditions needed for a stable resonant circuit, the total conductance and susceptance ΔG_s and ΔB_s, respectively, must satisfy eqs. 2 and 3:

$$\Delta Gs + \Delta G = 0 \qquad (4)$$



1. This diagram shows the basic construction of the Gunn diode oscillator, using a high-Q Invar cylinder to achieve good frequency stability with temperature.

$$\Delta Bs + \Delta B = 0 \quad (5)$$

or:

$$\Delta G = |\partial G/\partial\omega|\Delta\omega + |\partial G/\partial Vac|\Delta Vac = -\Delta Gs \quad (6)$$

$$\Delta B = |\partial B/\partial\omega|\Delta\omega + |\partial B/\partial Vac|\Delta Vac = -\Delta Bs \quad (7)$$

because:

$$\partial G/\partial\omega < \partial B/\partial\omega \quad (8)$$

By rewriting eqs. 6 and 7:

$$\Delta Vac \approx -\Delta Gs/(\partial G/\partial Vac) \quad (9)$$

$$\Delta\omega = \{-\Delta Bs + [(\partial B/\partial Vac)/\partial G/\partial Vac]\Delta Gs\}/(\partial B/\partial\omega) \quad (10)$$

From the definition of circuit loaded Q (Q_L):

$$Q_L = \beta/G1 \quad (11)$$

where:

β = the susceptance slope parameter and

G_l = the total loss conductance in the circuit.

Resulting from:

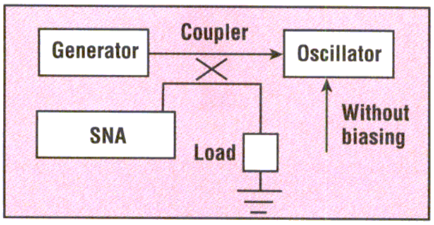
$$\beta = (\omega/2)(\partial B/\partial\omega) \quad (12)$$

As a result, it is possible to derive:

$$\partial B/\partial\omega = (\partial G/\partial\omega)Ql \quad (13)$$

from eqs 11 and 12.

Equation 13 shows that as the value of Q_l increases, the value of ∂B/∂ω also increases, and the frequency (ω) variation in the oscillator is smaller, and the frequency stability is higher. That is, with increasing ∂B/∂ω comes decreasing phase noise.



2. This measurement setup was used to check the resonant peaks of the Gunn oscillator's two resonant sections.

The Gunn oscillator developed to demonstrate these design principles is based on a high-Q Invar cylinder to stabilize frequency. Such a cylinder offers good frequency stability with low loss and ease of tuning. The TE_{011} mode was chose as the operating mode for the cylinder since it offers stable field construction with no degenerate mode and low loss.

Invar has good frequency stability as a function of temperature. By using a material with a roughness value of ∇12, the resulting unloaded Q is more than 20,000. The construction of the millimeter-wave oscillator is shown in Fig. 1. The oscillator's resonant frequency can be tuned by changing the length of the noncontacting piston in the cylinder, with the output frequency continuously variable across a narrow range about the center frequency.

When the cylinder is operated in the TE_{011} mode, its Q is:

$$Q = (\lambda_0/\delta)0.610[1+$$

$$0.168(D/L)^2]^{1.5}/[1+ 0.168(D/L)^3] \qquad (14)$$

where:

λ_0 = the resonant wavelength,

δ = the skin depth of the metal cylinder,

D = the diameter of the cylinder, and

l = the height of the cylinder.

Setting D/L = x and dQ/dx = 0, then x = 1 or:

$$D = l \qquad (15)$$

From:

$$\lambda_0^{TE011} = \frac{l}{\sqrt{(3.832/\pi D)^2 + (1/2L)^2}} = 0.759D \qquad (16)$$

it can be determined that:

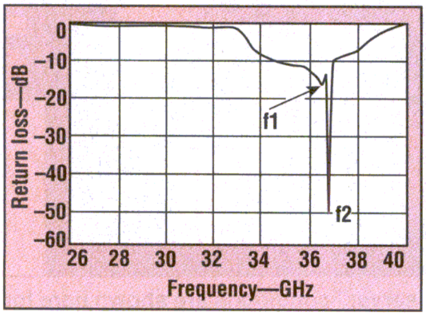
$$D = 1.32\lambda_0 \qquad (17)$$

Once the oscillating frequency is known, from eq. 15 and 17 it is possible to calculate the height and diameter of the cylinder in the TE_{011} mode and obtain the maximum value of Q.

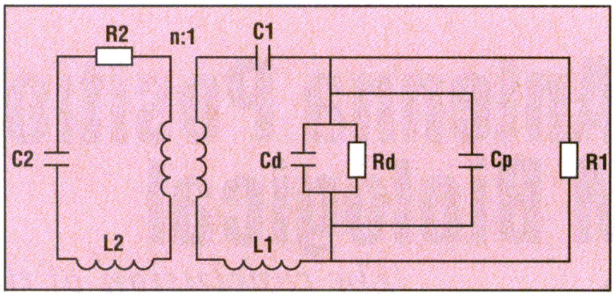
A scalar network analyzer (SNA) was used to determine the resonant frequencies of the main oscillator cavity (low Q), consisting of rectangular waveguide and the Gunn diode, and the high-Q Invar cylinder. Measurements were made without bias applied, using the test setup of Fig. 2. Measured results are shown in Fig. 3.

Figure 4 shows the equivalent circuit of the millimeter-wave oscillator. It is a biresonant circuit with the main resonant cavity formed of the waveguide and the second resonator formed of the Invar cylinder. As a re-

(continued on p. 176)



3. These return-loss results for the Gunn oscillator were made with a scalar network analyzer.



4. This equivalent circuit for the millimeter-wave Gunn oscillator shows the coupling between the two resonantors, one formed with waveguide and the other with the Invar cylinder.

Study Antenna Positioning And EM Biological Effects

The conclusion of this two-part article highlights optimal communications cell placements and the potential health risks of electromagnetic (EM) radiation.

Multiple-Beam Antennas, Part 2

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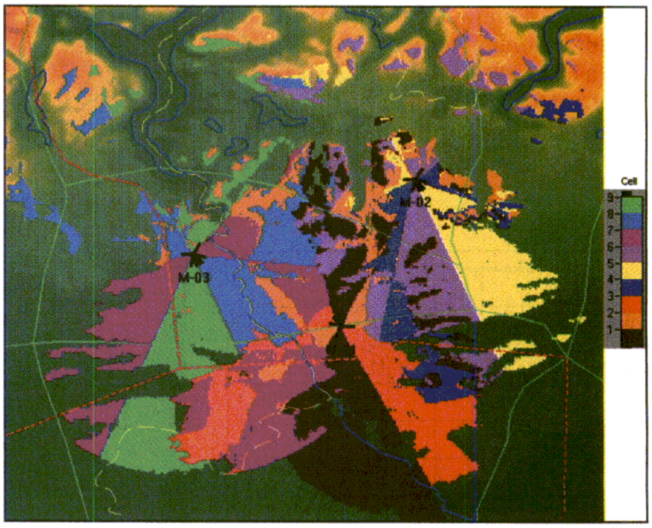
MULTIPLE-BEAM array antennas were presented in Part 1 (see *Microwaves & RF*, August 1999, p. 59) as efficient components in high-capacity cellular-communications systems. Using Quick Plan™ cell planning software from TeS Teleinformatica e Sistemi s.r.l. (Rome, Italy), it was possible to develop cellular-network models, which invited comparison to different antenna approaches. Sectoring with traditional panel-antenna-based clover base-transceiver stations (BTS) was compared to sectoring with BTS using multiple-beam arrays; computer models showed some significant advantages with the arrays. Part 2 continues this comparison, based on an area in northern Italy with a hypothetical arrangement of several existing (clover) BTS. Traffic is assumed to be evenly spread among the cells so that the existing frequency plan is in line with the theoretical frequency-reuse pattern corresponding to the cluster size of the system.

If the central cells in this model become congested, frequency planning can be optimized planning by assigning more frequencies to the congested cells to correct the problem. This solution comes at the expense of somewhat degraded (but still acceptable) system performance due to the general increase of the level of interference. Clearly, frequency optimization is a short-term solution.

Another alternative is to plan the installation of new microcells. This is the intelligent underlay-overlay approach that has already been successfully implemented in several networks. The basic two-layer network structure of the underlay-overlay technique provides seamless



6. The "second-best server" conditions are shown here for the nine-cell area handled by the three existing clover BTS. In this display, different shades or colors show different frequencies or cells.

coverage on one layer (with the macrocells) and high capacity on the other (with the microcells). The approach can improve network capacity through aggressive frequency reuse, although it requires capital investments for new BTS infrastructures which, in turn, contribute to worsen the environmental impact of the mobile network.

Sectoring by a multiple beam (direct radiation) array antenna is a possible alternative. It requires a moderate upgrade investment. Figure 7 shows the "best-server" coverage associated with the replacement (in three of the supposedly congested central cells of Fig. 4) of the existing BTS antenna panels (Fig. 2a) with a multiple-beam array antenna (Fig. 5). The simulated beam patterns of the array antenna are consistent with the assumptions of Fig. 1. It should be noted that the times two frequency reuse among the even and odd beams resulting from the sectoring of the original macrocell (i.e., the original 120-deg. sector) does not require any change of the original frequency plan.

MODERN SAFETY STANDARDS FOR EM RADIATION MAY PROVE TO BE LESS THAN IDEAL IN THE NEAR FUTURE (WHEN MORE COMPLETE DATA WILL BE AVAILABLE).

The sectoring does not even alter the level of interference in adjacent cells. The twofold nominal capacity gain comes directly out from the improved spatial access and not from frequency optimization. In any case, a frequency-optimization process may further improve the effectiveness of sectoring in terms of quality of service.

The "second-best server" picture is shown in Fig. 8. A mobile unit in the assumed congested area has more alternatives to establish a connection with respect to the situation displayed in Fig. 6. The shades in the figure correspond to different cells as well as frequencies, clearly showing the capacity gain.

The European pre-standard deals with exposure to electromagnetic (EM) and other forms of radiation through 300 GHz. The European pre-standard is based on well-established short-term effects which, depending on frequency, include heating and stimulation of electrically excitable cells in nerve and muscle tissue. Basic restrictions are given to prevent any adverse consequences of these effects. They are specified in terms of biologically relevant quantities, typically induced current density and specific absorption rate (or SAR, measured in W/kg) which is the power dissipated in a mass of tissue as a consequence of the exposure to EM fields. These quantities cannot be determined directly, so the standard specifies a set of more readily measurable reference levels, in terms of external electric and magnetic field strengths and power density, derived from the basic restrictions.

Pulsed fields (such as the emissions from radar systems or GSM handsets) may produce other effects, such as the auditory perception of microwave pulses, in addition to those associated with continuous-wave (CW) radiation. In these cases, restrictions in terms of specific energy absorption and energy density are given.

The pre-standard has a two-tier structure with lower levels specified for the general public compared to industrial levels. Some reports state that EM fields at levels lower than the reference levels may have long-term health effects. In Italy, such "preliminary" reports have received significant attention. The Italian limits, based on the principle that "the utmost caution must be used when knowledge is incomplete," are generally much lower than the European pre-standard. Currently available research however has not officially established any adverse effects from low levels of EM radiation and does not yet provide a basis for restricting exposure. For the general population, the SAR limit is 0.08 W/kg for 10 kHz to 300 GHz, averaged over any

7. By partial sectoring, the study area in northern Italy can be broken up into 18 cells. The data are shown overlapping GIS data, with different shades or colors corresponding to different cells.

8. The second-best server conditions are shown for the 18 cells in the study case with 18 cells achieved through partial sectoring of the original three BTS. The different colors refer to the different number of cells.

six-minute interval and over the whole body.

In addition, the SAR averaged over any six-minute interval and any 10 g of "valuable" tissue (such as the brain) must not exceed 2 W/kg. However, when using a telephone handset, most absorption takes place in the user's head, that is, in the brain tissue. It also worth noting that most business users spend long periods of time, much more than six continuous minutes, using a cellular telephone. A typical GSM handset transmits as much as 2-W peak power with a duty cycle of approximately 12 percent (corresponding to one active time slot out of eight slots). The duty cycle is perhaps lower due to the so-called "voice-activated discontinuous transmission" feature of GSM (and DCS-1800) which enables an automatic turn off of the handset transmitter during the pauses of the speaker. Assuming that even 20 percent of the handset transmitted power is evenly absorbed by the brain tissue (approximately 2.5 kg in an adult male), the total power dissipated in the user brain is $2 \text{ W} \times 0.12 \times 0.2 \times 2.5 = 0.12$ W, that is, $0.12/2.5 = 0.05$ W/kg. A SAR of 0.05 W/kg appears well below the limit.

In any case, the SAR cannot be determined directly and hence (for the general population), the limit of ambient root-mean-square (RMS) electric field strength (under continuous exposure) is $1.37 \times f^{0.5}$ (for a frequency range of 400 to 2000 MHz, with f in MHz), whereas the limit for the mean power density is f/200. At the GSM frequency of 900 MHz, this corresponds to an RMS electric-field limit of 41.1 V/m and a power density limit of 4.5 W/m^2. In Italy, the corresponding legal limit is much lower: 20 V/m for open areas accessible to the general population and 6 V/m inside residential buildings. But in Rome, for example, a "quality" target of less than 3 V/m (instead of six) is under discussion. The 6 V/m (RMS) limit inside residential buildings corresponds to approximately 0.1 W/m^2, a much-more prudent limit than the European limit of 4.5 W/m^2.

Assuming an omnidirectional transmission from the GSM handset, at a distance of 0.1 m (10 cm) from the phone, the calculated power density is $(2 \text{ W} \times 0.12)/(4\pi \times 0.1^2) = 2 \text{ W/m}^2$ and the power-density level appears to be very safe. But the calculated power density of 2 W/m^2 generated by the handset surely exceeds the Italian limit of 0.1 W/m^2 and it must also be considered that the transmission from a GSM handset depends on the surrounding environment. For example, when using the handset inside a car, which may be assimilated to a residential area, the power density not only exceeds the Italian limit of 0.1 W/m^2 but it might also easily exceed the limit of 4.5 W/m^2, due to the reflections inside the car.

Modern safety standards for EM radiation may prove to be less than ideal in the near future (when more complete data will likely be available). Focusing a BTS on the energy received from a mobile terminal enables a reduction in transmitted power from the mobile unit (which is close to the user's head), thus alleviating the health risks. Due to the growing number of mobile-handset users worldwide, any reduction in power may lead to a reduction in the probability of developing a disease associated with the use of personal phone handsets.

Analog synthesis of multiple antenna beams is similar to current sector antenna approaches and thus compatible with existing infrastructures. Such compatibility enables service providers to increase their capacity while reducing health risks to their customers, at a moderate cost for improving the existing cellular infrastructure. ••

For Further Reading

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Create A Transceiver IC For Triple-Band GSM Applications

Various requirements must be weighed when developing a low-cost, but flexible GSM transceiver for use at 900, 1800, and 1900 MHz.

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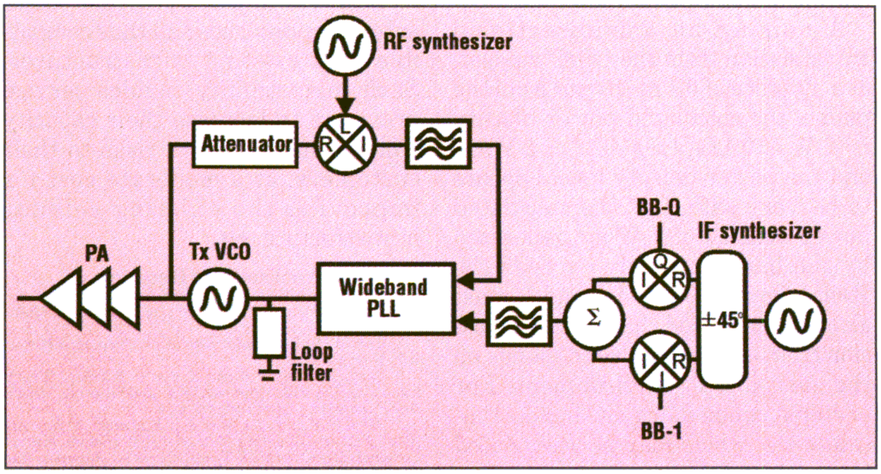
GLOBAL System for Mobile Communications (GSM) is by far the world's most popular cellular standard, with more than 100 million subscribers. Due to its popularity, it is used in at least three different frequency bands throughout the world, challenging designers to develop integrated-circuit (IC) transceivers that offer enough performance at low cost. Responding to the needs of GSM handset designers, a single-chip transceiver IC has been developed that is capable of triple-band operation. The transceiver supports the development of miniaturized, next-generation triple-band GSM telephones.

GSM is a well-established standard, with stringent equipment requirements.[1] GSM radio requirements can be separated into three categories—receiver, transmitter, and frequency synthesizer requirements. A review of these requirements will help to evaluate the performance of the triple-band transceiver IC. A breakdown of the three discrete GSM bands is offered in Table 1.

The key receiver performance parameter relates to the noise performance, the linearity of the receiver, and the performance in the presence of blocking signals. GSM specifies a minimum reference sensitivity of –102 dBm for GSM-900 and GSM-1900, and –100 dBm for GSM-1800. Most handsets, however, offer sensitivity that exceeds the standard, since this leads directly to an increase in coverage. The actual radio noise-figure requirement greatly depends on the handset's digital-signal-processor (DSP) performance. Assuming a required signal-to-noise ratio (SNR) of 9 dB at the input to the DSP (which is the same SNR that is used for the co-channel interference test), a –102-dBm sensitivity implies a noise-figure requirement of 11 dB for the complete radio front end. Not only must the receiver work for a –102-dBm input signal, it also must operate with input signals up to –15 dBm applied to the antenna. Therefore, a dynamic range of approximately 90 dB is necessary.

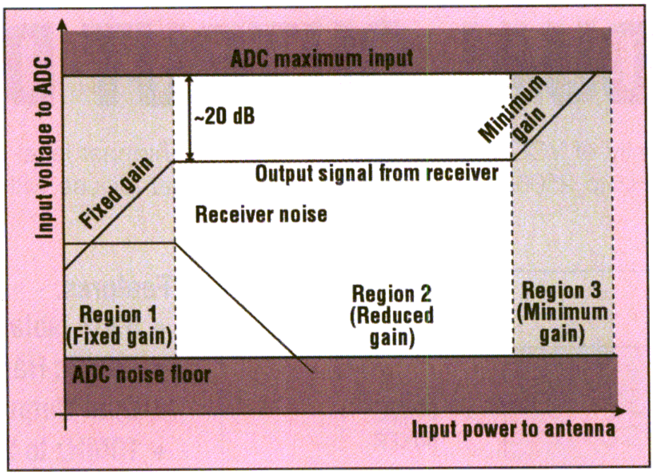
The GSM standard requires the handset to report the received signal strength back to the network with high accuracy in order to support transmit power-management infor-



1. The traditional offset transmit PLL approach to frequency synthesis directly drives the transmit PA with the VCO.

mation. The absolute signal-strength accuracy is provided in Table 2. The relative signal-strength accuracy between two signals at different frequencies must be +2/–4 dB or better when the two signals are in the same frequency band and +4/–6 dB or better when the two signals are in different frequency bands. Slightly higher accuracy is required with stronger signals.

The linearity requirement is defined by a two-tone test with interfering tones of –49 dBm at 800 kHz and 1600 kHz from the desired signal. Given this scenario, the receiver must still be able to receive a signal only 3 dB above the reference sensitivity (–99 dBm). Also, an input third-order-intercept (TOI) requirement for the full receiver of approximately –18 dBm can be derived.



2. This is a typical automatic-gain-control (AGC) algorithm as used in a GSM mobile unit.

A set of blocking scenarios serves as guidelines for evaluating a receiver's channel selectivity. Two fading adjacent-channel signals (200 and 400 kHz from the carrier or desired signal) test the channel selectivity of the receiver and the ability of the receiver's DSP to cope with interfering signals. Strong fixed-power blocking signals that are further away from the desired signal act as desensitizing signals, testing the receiver's ability to suppress strong undesired signals. A summary of the blocking signal definition is presented in Table 3.

Out-of-band blocking scenarios are also defined for the different standards. In GSM-900, the out-of-band blocking signals are defined to be at 0 dBm (–99 dBc) when they are more than 20 MHz away from the receive band (935 to 960 MHz). In E-GSM-900, frequencies lying 10 to 20 MHz from the receive band must support a blocker of –5 dBm (–94 dBc). In GSM-1800 and GSM-1900, the blocking signals are –12 dBm (–87 dBc) up to 100 MHz from the receive band, and 0 dBm (–99 dBc) thereafter (Table 4).

The key transmitter requirements relate to the frequency spectrum of the transmitted signal, the phase accuracy of the produced signal, and the dynamic issues relating to ramping the transmit power of the power amplifier (PA). The latter is out of the scope of this article, as PA ramping is controlled from the baseband or other dedicated PA controller.

The key specification of the required frequency mask of the handset exists at 400 kHz offset from the carrier where the signal must be –60 dBc measured in a 30-kHz bandwidth. Furthermore, GSM has a strict requirement on the amount of power that the transmitter is allowed to leak into the receive band while transmitting. The requirement is –79 dBm measured in a 100-kHz bandwidth for GSM-900 mode (–67 dBm for the E-GSM-900 portion from 925 to 935 MHz) and –71 dBm for GSM-1800 and GSM-1900 while transmitting a maximum output power of +33 dBm for GSM-900 and +30 dBm for GSM-1800 and GSM-1900. Outside the receive bands, the noise requirement is approximately –36 dBm in a 100-kHz bandwidth. These requirements can cause problems for many transmit architectures, often requiring one or more filters in the transmit chain. For multiband applications, this can be very expensive since it will require at least one filter per band.

The other key transmitter requirement is the phase accuracy of the transmitted signal. The requirement is a phase error of less than 5-deg. root-mean square (RMS). There are several contributors to the phase error, but usually the phase-error contribution from the RF phase-locked loop (PLL) in the RF synthesizer dominates.[2]

The GSM 05.10 standard requires the transmit spectrum to be frequency accurate within 0.1 PPM of the desired transmit frequency.[3] The frequency accuracy is provided by a precisely controlled crystal-reference oscillator through the baseband controller. Frequency pushing of the synthesizer or crystal-reference oscillator can occur when the PA and other active circuits are enabled. This can be a challenge when integrating all of the transceiver blocks on a single chip.

Synthesizer lock times are defined by the maximum timing advance (maximum distance to the base station) that the handset must be able to accommodate. GSM specifies a maximum distance to the base station of 35 km. Since the handset's transmit slot follows three slots after the receive slot, subtract the time for the signal to travel from the base station to the handset and back. The maximum lock time can be derived as $[(2 \times 577\ \mu s) - (2 \times 35 \times 10^3\ m)]/(3 \times 10^8\ m/s) = 920\ \mu s$. In the case of dual-slot operation, the maximum lock time requirement is reduced to 343 µs. (Dual-slot operation consists of two consecutive transmit slots followed by two receive slots. It is typically used for data services. The spacing between the first transmit and first receive slots remains as three slots.)

Furthermore, the phase-detector reference spurious content and synthesizer [voltage-controlled-oscillator (VCO)] phase noise can

Table 1: A review of GSM mobile-station bands

Frequency band	E-GSM-900 GSM-900	GSM-1800	GSM-1900
Transmission (MHz)	880 to 915 925 to 960	1710 to 1785	1850 to 1910
Reception (MHz)	890 to 915 935 to 960	1805 to 1880	1930 to 1990

cause violation of the transmit frequency mask or cause blockers to mix on to the desired signal in the receiver. Finally, the in-band phase noise of the synthesizers (PLLs) must meet a certain minimum criteria to meet the 5-deg. RMS phase error for the transmitter.

Many different transceiver architectures have been deployed in GSM, most successfully. Popular transmit architectures range from intermediate-frequency (IF) upconversion and direct upconversion to offset transmit PLL (closed-loop modulation of the transmitter VCO as in Fig. 1). On the receive side, implementations such as direct conversion and receivers with two or more IFs have been used. However, with the event of dual-band telephones, the transmit-architectures seem especially to diverge toward offset PLL-style architectures, since they eliminate the need for band-select filters and duplexers before and after the PA.

Meeting the tight GSM transmit spectral-mask requirements at a 400-kHz offset introduces strict requirements on the noise performance and accuracy of a quadrature modulator. However, the real problem is often the noise that is produced in the receive band during transmissions. GSM-900 requires that the noise in the receive band be better than –79 dBm while transmitting at a level of +33 dBm (with both measured in a 30-kHz bandwidth). Therefore, the noise must be suppressed by 112 dB to approximately –157 dBc/Hz. Single-band architectures using a mixer-based upconversion scheme normally employ one or more filters along with a duplexer to provide the proper attenuation. In dual-band architectures, this becomes impractical since it requires multiple filters and multiple duplexers.

The offset-transmit PLL approach (Fig. 1) offers a flexible solution to this problem. By having a VCO drive the PA, the PLL's loop filter provides low-pass filtering of the noise produced by the quadrature modulator and other blocks inside the transmit loop. Therefore, the noise requirement relates to the phase noise of the transmit VCO. Although these requirements are rather stringent (< –157 dBc/Hz at 20 MHz), high-output VCOs with the required phase noise performance exist. Additionally, by having the VCO drive the PA directly, an inherent constant-envelope signal is produced, minimizing the amplitude-modulation-to-phase-modulation (AM-to-PM) modulation inside the PA. As a result, only a simple harmonic filter is required at the output of the PA.

One problem with the offset PLL transmit architecture is signal leakage from the input to the downconversion mixer (e.g., LO-to-RF leakage) feeding straight into the PA. This can cause spurious-noise violations. To address this issue, an attenuator is usually inserted in the feedback path between the transmit VCO and the downconversion mixer.

The filters shown in Fig. 1 are necessary to prevent harmonic products produced by the downconversion mixer and quadrature modulator from aliasing to the desired transmit spectrum by the wideband transmit PLL.

Adapting the offset transmit PLL scheme to a dual-band architecture is straightforward. Using a band-switched transmit VCO supports a dual-band transmit architecture without external filtering. The RF local oscillator (LO) to the downconversion mixer must be addressed in the frequency planning.

On the receive side, there are no significant changes in going from a single-band receiver to a dual-band receiver. One or more band-select filters will be needed at the input of the dual-band receiver to attenuate out-of-band blocking signals (0 dBm), thereby increasing the number of filters compared to the single-band design.

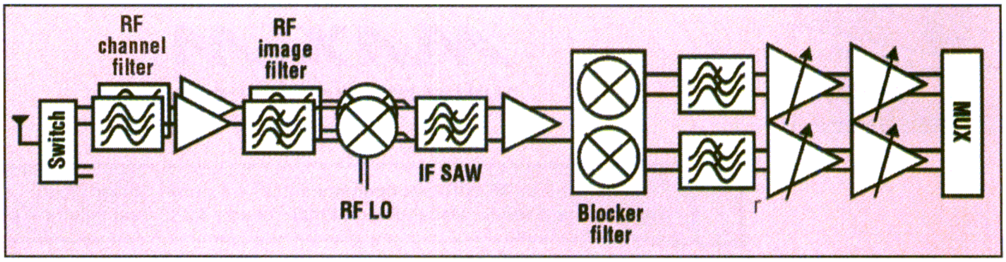
A direct-conversion receiver is a preferred solution since it eliminates the need for additional external filters. This can be a deceptive advantage, however, since few GSM direct-conversion receivers exist. The problem is finding solutions that cope with self-mixing effects (DC offset), automatic-gain-control (AGC) circuits, wide dynamic range, and noise issues.

Traditional dual-conversion and triple-conversion receivers are well-understood, and offer good receive performance. In dual- and triple-conversion designs, it is possible to share the receiver's IF section, thereby sharing (and reducing) the required number of filters between different bands. Of course, any frequency plan must address the sharing of the RF LO to the RF mixer among two or three bands.

It should be apparent that a triple-band receiver for GSM applications can be implemented using an offset PLL transmit architecture and dual-conversion receiver with three sets of band-select filters at the receive input. The problem lies in deriving a frequency plan that supports the sharing of one RF synthesizer between all three bands, and allowing the use of a common surface-acoustic-wave (SAW) IF filter. Normally, a band-switched RF VCO would be used, although its cost is considerably higher than that of a single-band VCO. Many baseband interface options exist, but to achieve compatibility

Table 2: Required RSSI accuracy

Receive level	–110 to –70 dBm	–110 to –48 dBm
Absolute accuracy	±4 dB	±6 dB



3. This receiver architecture supports flexibility in the frequency plan while enabling the use of low-cost SAW filters at IF.

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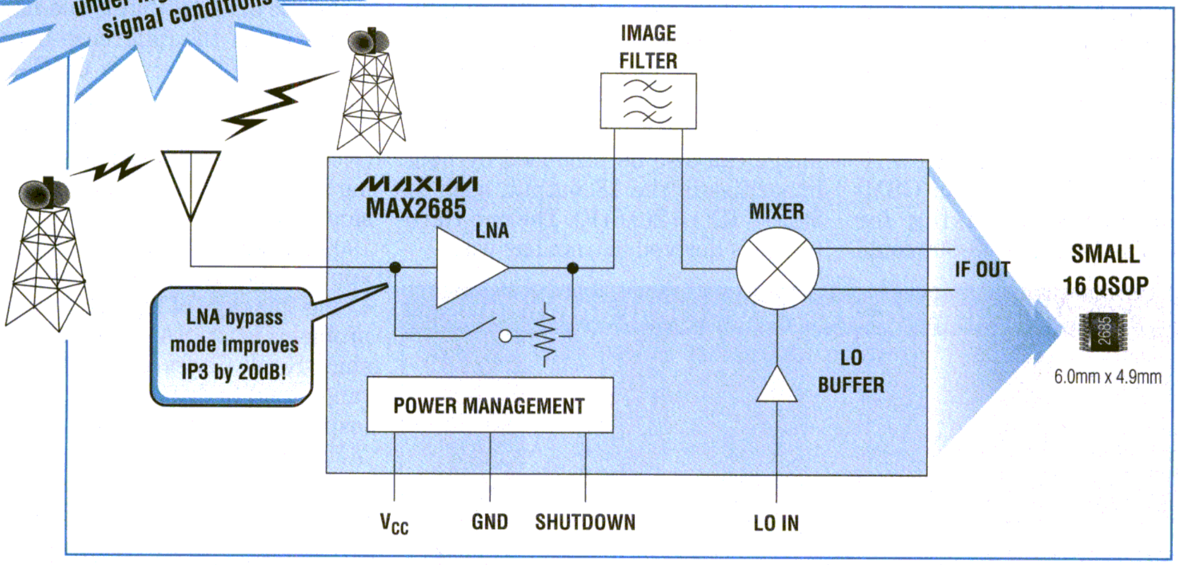

	LNA STAGE	FILTER STAGE	MIXER STAGE	COMBINED FRONT-END PERFORMANCE
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NOISE FIGURE (dB)	1.4	2	13	3.7
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with existing baseband processors, a standard in-phase/quadrature (I/Q) inferface is needed for the transmit and receive functions.

In order to implement a practical triple-band GSM architecture, a superheterodyne receiver with a common intermediate frequency (IF) for all three GSM bands was selected. With a fixed IF, the three different RF LO ranges can be readily derived. Ideally, a single VCO can be used among the three different bands.

One way to share the RF LO frequency range is by choosing a low IF to support low-side mixing for GSM-1800 and high-side mixing for GSM1900. The LO for GSM-900 could then be generated through on-chip frequency division. This would result in a very-limited RF LO frequency range. On the down side, this will also result in a fairly low IF, across the 50-to-100-MHz range. This low IF will introduce problems with the receiver one-half image (IIP2 performance) and the one-third image. The one-half image is that frequency where the second harmonics of the RF and LO signals mix to form the same frequency as the desired IF (e.g., $2LO - 2RF = |IF|$). The one-half image can also be produced by nonlinearities in the IF output, such as $2(LO - RF) = 2(0.5IF)$. The one-third image is derived in a similar way.

Table 3: Summarizing blocking-signal definitions

Blocker offset frequency (kHz)	200	400	600 to 1400	1600 to 2800	> 3000
Levels for					
GSM-900 (dBc)	–9	–41	–56	–66	–76
GSM-1800 (dBc)	–9	–41	–56	–66	–73
GSM-1900 (dBc)	–9	–41	–56	–66	–73

An IF in the 50-to-100-MHz range with a frequency band of 75 MHz (for GSM-1800) has a one-half image located within the receive band for some frequencies. The front-end filters will not reject this blocker when it falls within the operating band, and the receiver will not function unless it has a very high IIP2. An obvious solution is the use of an IF that is one-half the frequency of the difference between the GSM-900 and the GSM-1800/GSM-1900 bands. The RF LO frequency can then fall in between the two bands, providing high-side mixing in the GSM-900 scenario and low-side mixing in the GSM-1800/GSM-1900 scenario. This results in an IF of approximately 440 MHz or higher. Using this type of high IF introduces problems in the SAW filter, however, since SAW filters in this frequency range tend to drift with temperature (especially at frequencies above 300 MHz). Still, the filter can be manufactured with a wider passband to accommodate any center-frequency

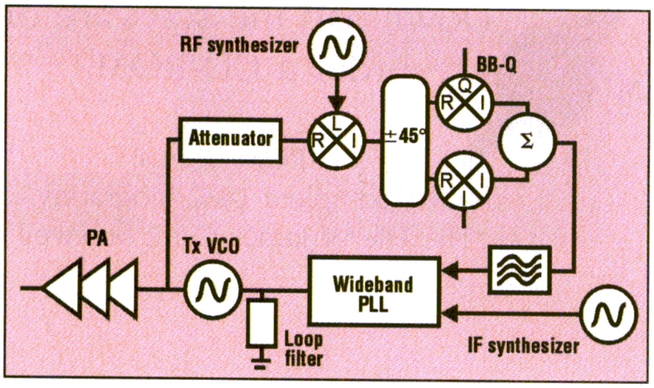
Table 4: Summarizing the triple-band GSM receiver performance

Parameter	GSM-900	GSM-1800/GSM-1900
Noise figure	2.5 dB	3.1/3.3 dB
TOI	–14.9 dBm	–14.2/–14.1 dBm
Compression	–24 dBm	–24 dBm
Gain	92 dB	89/88 dB
Receive current	70 mA	73 mA

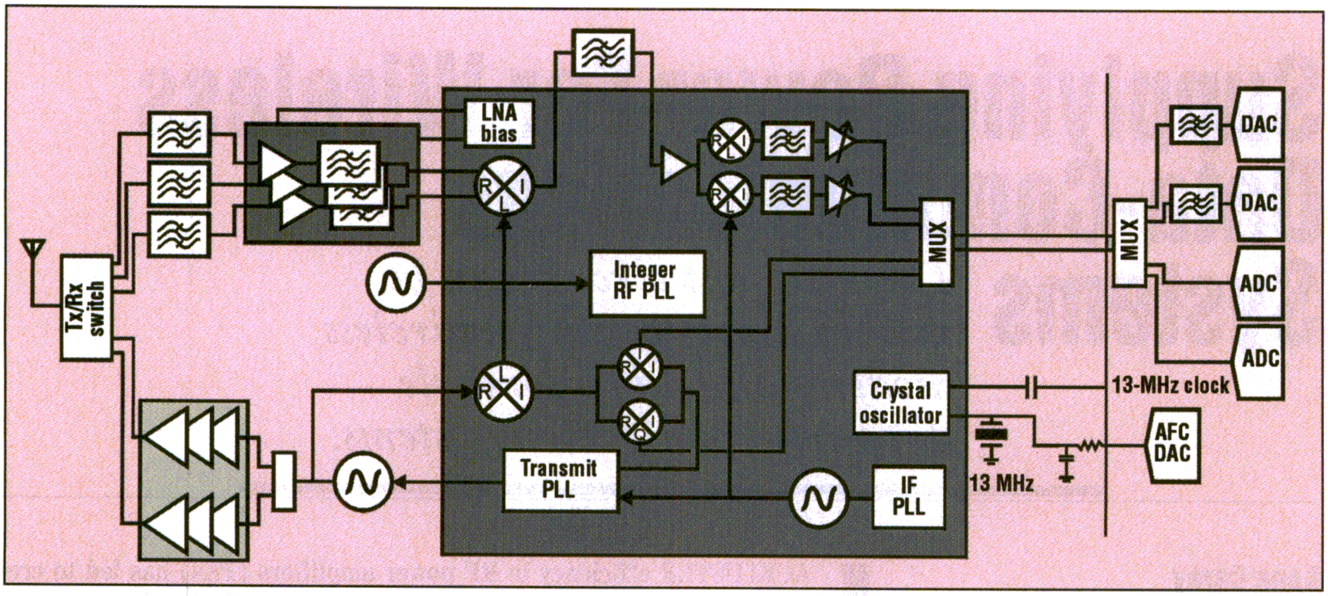
drift, although providing less selectivity. Furthermore, the front end must offer very high isolation between the GSM-900 and the GSM-1800/ GSM-1900 bands using this approach. Assuming the telephone receives a GSM-900 signal, if a blocking signal in the GSM-1800 band at the mixer image (0 dBm) leaks through the GSM-1800 receiver and couples to the GSM-900 path, then this signal will also convert to the desired channel and will most likely dominate a weak desired signal.

To minimize SAW-related problems, an IF was chosen from 200 to 300 MHz. This range will provide significant one-half-image attenuation, and the SAW filter will provide significant channel selectivity. Since the filtering frequencies are below 300 MHz, temperature variations will be minimized.

Given this scenario, the LO range must be at least 185 MHz (1990 to 1805 MHz) to implement a world GSM transceiver using a single-band synthesizer. The LO frequency for the GSM-900 radio can be derived through frequency division. With the choice of an IF in the 200-MHz range, several options exist to complete the receiver. Mixing the signal to a significantly lower second IF before translation to an in-phase/quadrature (I/Q) format could be an option, but this is likely to require extra current for the second mixer. The first mixer should be an image-reject mixer unless exceptional attenuation is obtainable in the IF SAW filter. Approximately 85-dB image rejection is needed for in-band images (65 dB in some cases). For out-of-band images that are close to the



4. This diagram shows the modulation synthesizer architecture that is used to support a flexible frequency plan for multiple-band GSM use.



5. This overview shows how the multiband GSM transceiver was implemented in a single BiCMOS IC.

operating band, the conditions are similar. Clearly, it is desirable to translate IF signals directly to the I/Q baseband representation.

It is important to consider the dynamic range of the receiver. Approximately 90-dB minimum dynamic range is needed with input signals ranging from –102 to –15 dBm. The accuracy of the received signal-strength indicator (RSSI) must be high, which implies the gain of the receiver must be accurate. Besides this overall dynamic range, the receiver must provide enough instantaneous dynamic range to support fading signals during the burst and from burst to burst. The analog-to-digital converter (ADC) in the baseband processor usually defines the instantaneous dynamic range. Typical baseband processors offer 10-b ADCs. To fully utilize the dynamic range of the ADC, the receiver must provide accurate programmable gain that can be used to implement the telephone's AGC algorithm.

Figure 2 shows a typical AGC algorithm for GSM. In region 11, the receiver gain is fixed. The gain remains fixed until a worst-case fading scenario causes the input level to the ADC to rise above a maximum threshold level. This transition point is usually chosen approximately 20 dB below full scale for a static signal (Fig. 2). From this point on (region 2), as the signal increases, the gain of the receiver must be reduced. When the receiver's gain cannot be reduced further (region 3), the signal will increase until the input signal saturates the ADC.

Given the chosen architecture, variable gain can be implemented either directly at IF or at baseband. By implementing the gain at IF, the problems relating to DC offsets can be minimized through AC coupling. However, by introducing the gain at IF, the IF SAW filter requirements become more stringent. The IF SAW filter offers finite attenuation of blocking signals, so if strong blocking signals are not attenuated sufficiently, they may saturate the receiver when the desired signal reaches the IF amplifier.

By moving the majority of the gain to baseband, a simple lowpass filter can be introduced before the baseband gain stage. This lowpass filter will attenuate the blocking signals before the desired signal is fully amplified, thus preventing the blocker from saturating the receiver. Unfortunately, this approach is sensitive to DC offsets. Since the majority of the gain will exist at baseband, any DC offset introduced will be amplified along with the desired signal. Still, simple or advanced DC offset compensation schemes can be implemented to cope with this problem. The latter solution is probably the most desirable, since it supports the use of a lower-cost IF SAW filter, due to the additional filtering before the major gain block. The proposed receive architecture is shown in Fig. 3.

The offset PLL architecture shown in Fig. 1 requires the downconverter mixer's output frequency to be identical to the quadrature modulator's frequency for the spectrum to be produced correctly at the input of the PA. Since the IF synthesizer will only produce a limited frequency range, the tuning range of the RF synthesizer must be wide enough to cover the three desired bands. Using the modulation synthesizer architecture over the traditional offset PLL supports a more flexible frequency plan while maintaining all of the benefits of the offset PLL approach.[5] The modulation synthesizer architecture is shown in Fig. 4.

By inserting the quadrature modulator in the feedback path of the loop, the input to the wideband PLL must be a continuous-wave (CW) signal. Due to this, different divide ratios in the reference (R) counter and the N counter can be used while maintaining the desired modulation spectrum at the input to the PA. This ability greatly improves the flexibility when developing a world GSM transceiver frequency plan. It should be clear that a world GSM transceiver can be implemented with a minimal cost increase of additional band-select fil-

(continued on p. 174)

Supplying Power For Wireless Data-Communications Systems

This efficient circuit can be used to supply power to wireless modems and other portable data-communications systems.

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Application Engineer
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LACKLUSTER efficiency in RF power amplifiers (PAs) has led to creative power-management techniques in many handheld wireless applications. Some communications protocols support burst-transmission techniques and shutdown when they are not transmitting (duty-cycle control), but during a transmission, the typical PA efficiency is no greater than 40 to 60 percent. In contrast, the efficiency for the main power supply in a handheld device is typically 90 to 95 percent. Fortunately, a practical circuit is presented here for powering wireless modems, their associated PA, and wireless data transfers.

Many handheld devices operate from one or more alkaline batteries. Others, which demand the higher levels of instantaneous power associated with backlit displays, achieve longer life (between charges or between battery replacements) with nickel-metal-hydride (NiMH) or even Li^+ batteries. Regardless of the battery type and configuration, however, the modem, PA, and radio circuitry necessary for wireless communications require extra battery capacity if the system is to maintain a reasonable operating life.

Typical of these systems is a PCMCIA wireless modem for transferring cellular digital packet data (CDPD). It might plug into a handheld personal digital assistant (PDA) or a handheld computer running Windows CE, and draw several hundred milliamps from a +3.3-VDC supply. The PCMCIA card usually includes a secondary battery to avoid excessive drain on the host battery. To supply the power surge demanded during transmission, secondary batteries usually exhibit the low-equivalent series resistance (ESR) found in modern rechargeable cells. The actual power for the wireless link depends primarily on the efficiency and transmit power of the PA.

As an example, the power supply (including backup battery) for a wireless data-communications link must connect with a host handheld system operating at +3.3 VDC (see figure). The packaged integrated circuits (ICs) are small and appropriate for a handheld system—IC1 is contained in a 16-pin quarter-sized-outline-package (QSOP) housing while IC2 is contained in an eight-pin micro-MAX package. The secondary battery is a single lithium-ion (Li-ion) cell which exhibits a full-charge voltage of +4.1 to +4.2 VDC and has little residual energy below +2.9 VDC. IC1 converts this voltage to +3.3 VDC, while IC2 forces the resulting backup voltage to track the host supply within 12 mV (0.36 percent).

Voltage tracking is crucial for proper mating of the wireless hardware and host. It ensures that the bidirectional data and control lines achieve valid logic levels. It also prevents excessive current flow from the primary battery to the modem hardware and from the secondary battery to the primary battery and host computer.

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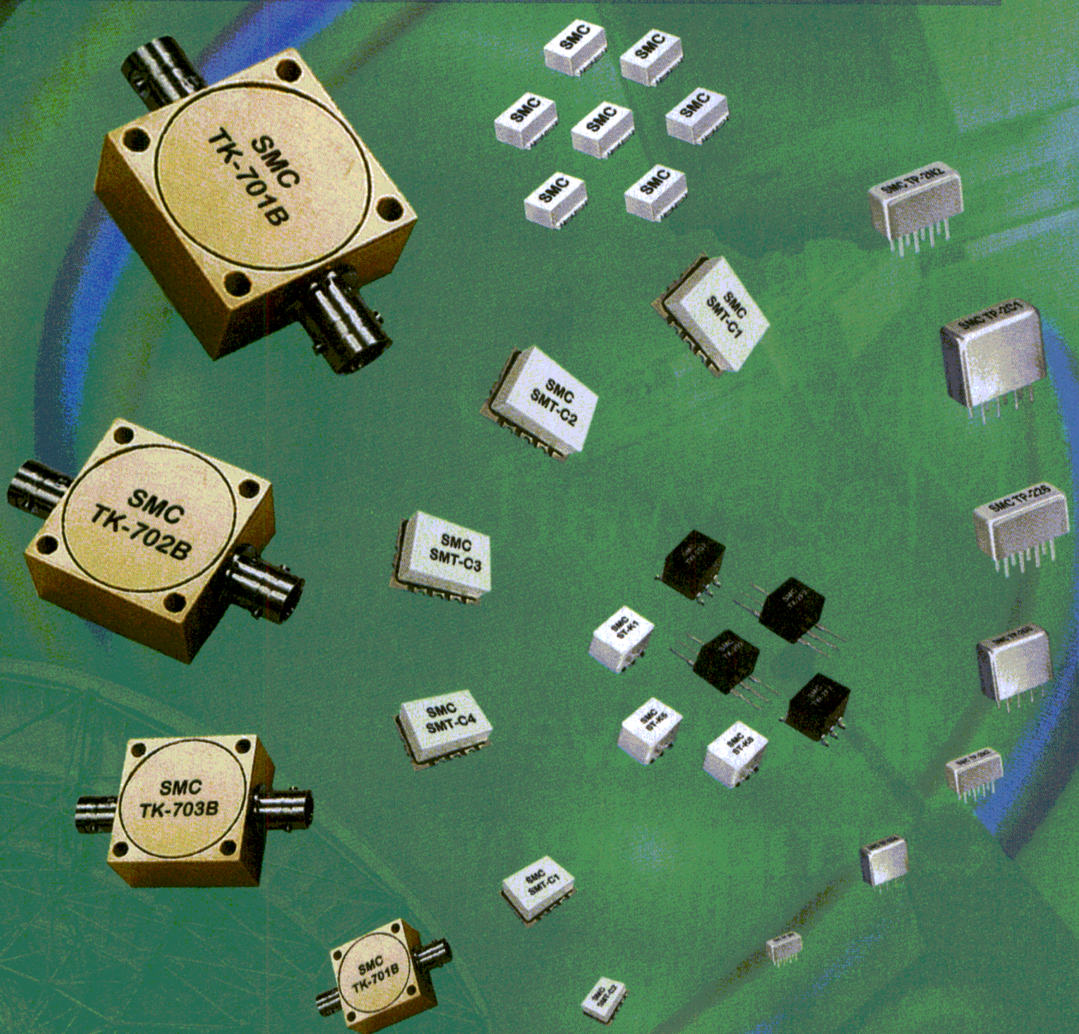

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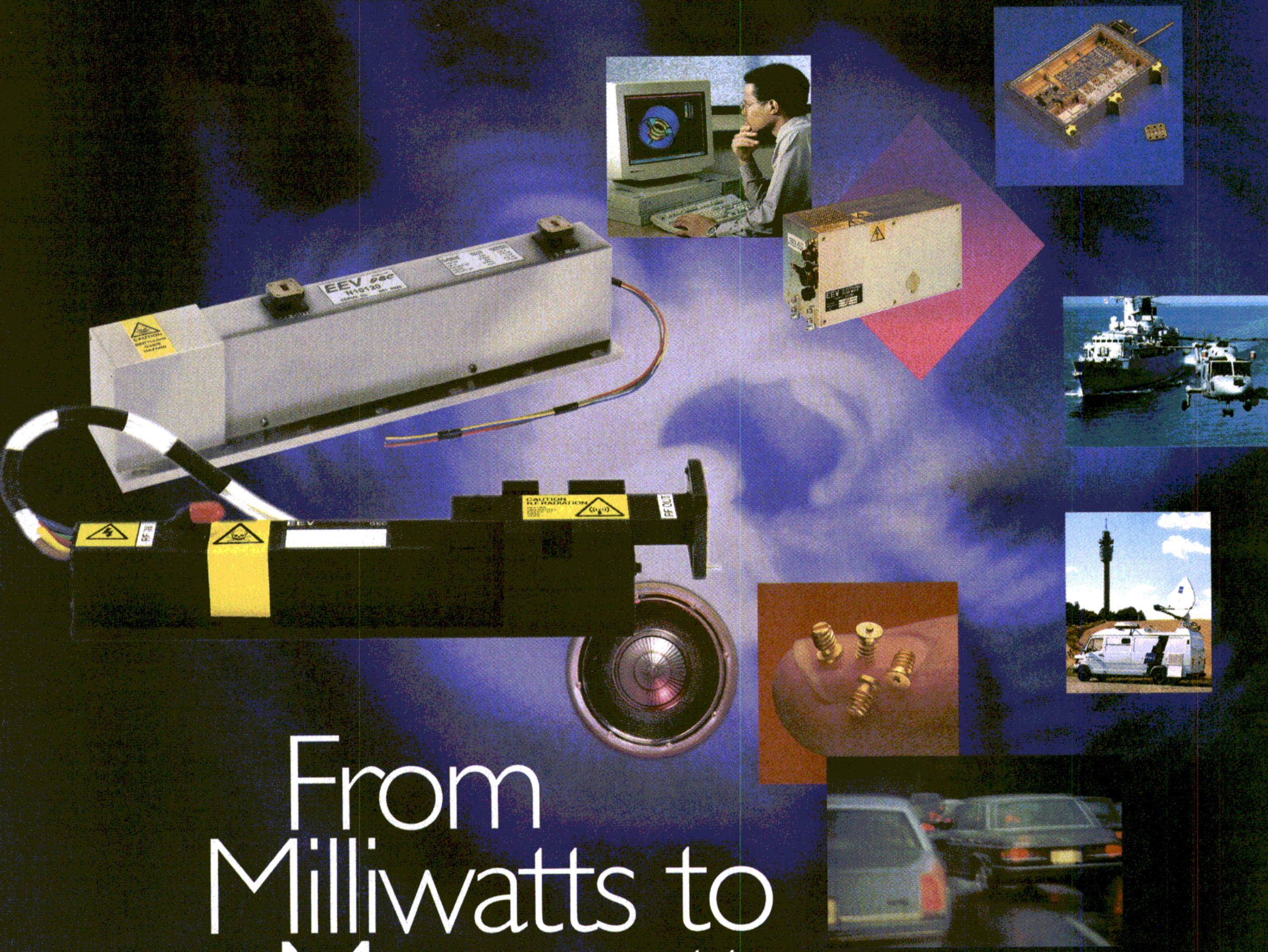


The circuit operates as follows. First, consider the modem's state before it is plugged into the host's PCMCIA slot. Negligible energy should be drawn from the secondary battery, so the modem's power supply must be disabled for that condition. The on/off control line for this secondary supply is the "power-good" (PG) pin of IC2. Since host voltage VHH, which powers IC2, is not present with the modem disconnected, IC2 is powered down.

Since its PG output (an internal open-drain, n-channel MOSFET) offers high impedance with power off, it draws only leakage current. While this PG output is high impedance, the two resistive dividers (the parallel pair R6/R7—which normally monitors the secondary-battery voltage through a comparator internal to IC1, and the parallel pair R3/R4—which sets the boost regulation voltage V_{BOOST} level when power is applied) shut down IC1 by acting as pullups on the ONB line. IC1's switch-mode boost regulator and low-dropout (LDO) regulator are disabled during shutdown. Thus, with 1-μA leakage through the dividers and 1 μA into IC1, the typical battery drain during shutdown is only 2 μA.

Consider the power requirements of this circuit with the supply voltage applied. If the PA must produce 0.6-W output power, and the amplifier has an efficiency of 50 percent, it requires 1.2-W input power. If it operates with a 50-percent duty cycle (equal time for transmitting and receiving), then the root-mean-square (RMS) power into the PA is 0.6 W. At a +3.3-VDC supply voltage, this load draws approximately 180 mA. If the rest of the modem draws 40 mA from +3.3 VDC, the total supply current for the wireless link is approximately 220 mA* at +3.3 VDC.

The IC1 boost regulator can deliver (at V_{BOOST}) 800 mA or so from a +2.7-VDC source and 1 A or more from a virtually depleted Li^+ cell (+2.9 to +3.0 VDC). Even so, the PA and remaining modem hardware are powered by the less-efficient internal LDO, which is rated for 300-mA typical current with approximately 220-mA current as a guaranteed minimum. The reason is noise suppression. The LDO offers approximately 38 dB of PSRR at 300 kHz, proving advantageous when suppressing the pulse-width-modulation (PWM) switching noise at V_{BOOST}. By easing or eliminating the need for subsequent noise suppression on the PA's supply voltage and associated RF emissions, this built-in filter action by the LDO makes it easier to pass the Federal Communications Commission (FCC) emissions requirements. The efficiency hit, on the other hand, is approximately 8.3 percent.

V_{BOOST} tracks V_{HH} in the neighborhood of +3.3 VDC. The secondary battery has a voltage higher than V_{BOOST} when fully charged and lower than V_{BOOST} when near depletion, so the LDO and boost regulator in tandem provide a necessary buck/boost function. SEPIC, flyback, and forward power-supply configurations can also achieve the buck/boost function, but all require bulky and expensive magnetic-storage elements (transformers), and they all lack the noise suppression provided by an LDO. In this regard, the circuit shown in the figure is superior to all of the alternatives.

Next, consider what happens when the modem card is plugged into the host's PCMCIA connector. The action closes electrical connections between the respective circuit commons (GND), and between all bidirectional data and control lines. The host can then enable or disable the modem hardware using the enable (EN) line. If this line is initially low when the hardware is mated, then all of the modem hardware is disabled and presents a high impedance to the LDO node.

IC2 receives power when the host computer's voltage (V_{HH}) supply (nominally +3.3 VDC) charges capacitor C1 through the connector, and IC2's minimum operating voltage supports a proper power-up even when host voltage V_{HH} is at the low end of its range (10 percent below nominal). An internal 15-µs delay enables V_{HH} to settle (at the V^+ terminal) before the PG output goes low, notifying the host that the modem circuitry can now be enabled by the EN line. A low voltage (virtual ground) on PG also grounds the two resistor dividers for proper sensing of the battery and boost-regulation voltages.

When voltage V_{HH} is connected, IC2 pulls ONB low when PG goes low, and IC1 begins shuttling energy through inductor L1, raising V_{BOOST} (through feedback from R3/R4) to approximately +3.7 VDC. Initially, the LDO regulator is held off. It turns on when V_{BOOST} achieves regulation, and IC1 enters its tracking mode when the LDO output is above +2.3 VDC (it should be near +3.3 VDC already, because V_{HH} has charged C2 through R2). Tracking mode, a special feature of IC1 that forces V_{BOOST} to 300 mV above the LDO voltage, is set by the connection between OUT and TRACK. The 300-mV headroom enables the LDO to maintain regulation while providing the required PSRR, up to its maximum output current rating. Since tracking mode supports the boosted voltage to the minimum necessary, the LDO wastes minimum battery power.

The LDO is in regulation when IC1's FB_{LDO} pin assumes the internal reference voltage (nominally 1.23 V). This FB_{LDO} voltage is generated by current through R5, which is proportional to the current through R2. That is, IC2 has the transfer function $V_{OUT} = gm(V_{SENSE})R5$, where V_{OUT} is the voltage across R5, V_{SENSE} is the voltage across the RS+ and RS- terminals (R2), and $gm = 10^{-2}$ S. When in regulation, $V_{OUT} = VFBLDO = 1.23$ V. Therefore:

$$V_{SENSE} = VFBLDO/(gm \times R5) \quad (1)$$

Substituting for V_{SENSE} in the re-

lation $V_{LDO} = V_{HH} + V_{SENSE}$:

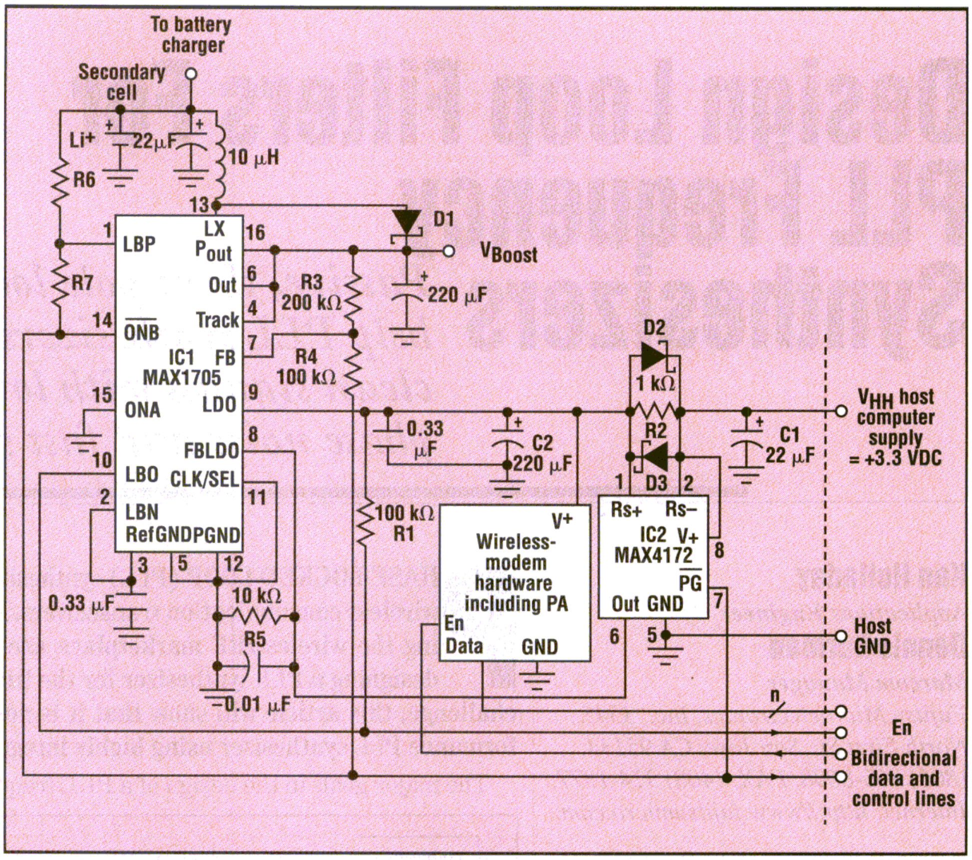
$$V_{LDO} = V_{HH} + V_{FBLDO} / (gm \times R5) \qquad (2)$$

Substituting numerical values from the circuit in the figure:

$$V_{LDO} = V_{HH} + 1.23 / (10^{-2} \times 10^{4}) = V_{HH} + 12.3\ mV \qquad (3)$$

Setting R5 to 10 kΩ forces the sense voltage to 12.3 mV. Knowing this, one can select R2 to program the amount of current flowing from LDO to V_{HH}. At R2 = 1 kΩ, for instance, this R2 current is approximately 12 μA.

The intended use for IC2 (a high-side current-sense amplifier) is in precision measurements of high-side current using low-value, high-wattage, precision current-sense resistors. This application is unusual in that 10-percent-accurate, low-wattage current-sense resistors are fine (1/16-W, surface-mount types, for instance). It is not important how much current flows from LDO to V_{HH}, just that the value remains small.



This circuit adds wireless-modem and PA capability to a handheld instrument, along with appropriate power-management functions.

One advantage of the high-value (1-kΩ) current-sense resistor is that a short circuit or unexpected heavy load at V_{LDO} can draw only 3.3 mA or so through R2 from the host(hardly enough to cause a system crash. Resistor R2 need not be as large as 1 kΩ. IC2 draws approximately 800 μA, so setting R2 = 12 mV/800 μA = 15 Ω enables the LDO node (instead of the host) to power IC2.

In an alternate configuration, IC2's V+ node can be connected directly to LDO instead of V_{HH}. IC2 then receives power from LDO except at power up, when the power comes from V_{HH} through resistor R2. This arrangement requires that the PA and modem hardware be shut down, presenting a high impedance to LDO that forms no significant voltage division with R2, and that R2 be sufficiently small to ensure the minimum voltage enabled at V^{+} for guaranteed operation (+3 VDC). If V_{HH} = +3.6 VDC or more, for example, R2 must be less than 375 Ω. This value ensures that IC2's 0.8-mA operating current drops to no more than 0.3 V at the low end of the V_{HH} range (+3.6 VDC, 10 percent).

The Schottky diodes (D2, D3), in parallel with R2, protect IC2 from excessive voltages between RS+ and RS–. They introduce small-leakage currents, but do not otherwise influence the circuit behavior. The capacitor in parallel with R5 ensures a smooth and stable V_{LDO} voltage by shunting any high-frequency noise at the LDO feedback node to ground. As previously mentioned, IC1 includes a comparator with uncommitted inputs and output. In this circuit, it monitors the secondary battery voltage, and alerts the host when the remaining energy is close to a level critical for maintaining the communications link.

The application circuit accommodates a range of conditions beyond those described. It is compatible with other communications buses suitable for mating a wireless modem to a handheld device, for example. These include Card-Bus and the strongly emerging Universal Serial Bus (USB). It also accepts nominal host supply voltages up to +5 VDC. For greater efficiency, some applications connect the PA directly to the boost-regulation voltage V_{BOOST} instead of V_{LDO}. In that case, voltage V_{BOOST} need not track V_{LDO}. The voltages can be controlled separately with independent sets of feedback resistors.

IC1 starts at +1.1 VDC and operates down to +0.7 VDC, so even a two-cell NiMH secondary battery is acceptable for boost applications that require lower levels of RF output power. Finally, IC1 normally operates as a low-noise PWM regulator under heavy load, with a switching frequency of approximately 300 kHz, which provides the highest efficiency and longest battery life. ••

Note

*If the PA duty cycle is less than 100 percent and the LDO's output current is within the limits rated for IC1, the equation it = CV should be used to size the LDO output capacitor (C2) so that any dips of LDO voltage during transmission are within allowable limits.

Design Loop Filters For PLL Frequency Synthesizers

Passive, three-pole loop filters help PLL synthesizers generate clean signals with low spurious and phase noise and fast switching speeds.

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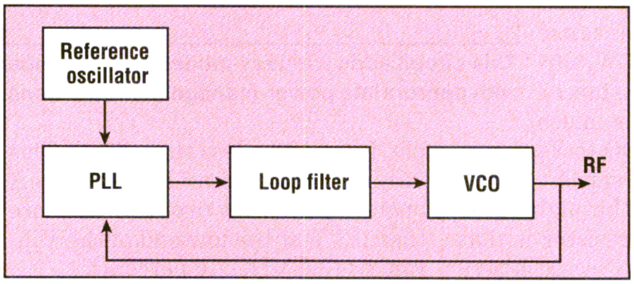
Fujitsu Microelectronics, Inc., 3545 North First St., San Jose, CA 95134; (800) 866-8608, FAX: (408) 324-1377, Internet: http://www.fujitsumicro.com.

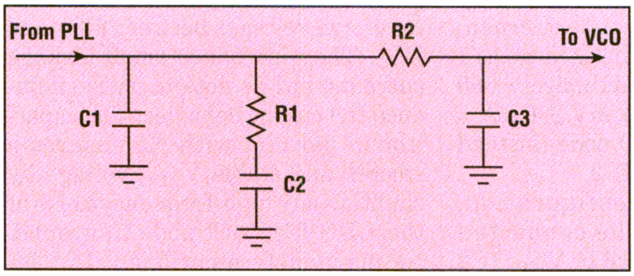
PHASE-LOCKED-LOOP (PLL) synthesizers are integral components in wireless communication transceivers. Many companies that are entering the wireless/RF marketplace are experiencing the challenge of designing a PLL synthesizer for the first time. To help them meet that challenge, this article will show that it is not difficult to design a high-performance PLL synthesizer using highly integrated, off-the-shelf parts.

The major goals in the design of a PLL frequency synthesizer are to achieve low phase noise, low spurious output, and the ability to step, or hop, from one frequency to another in a specified amount of time. These characteristics largely depend on the synthesizer's loop filter. Unfortunately for first-time loop-filter designers, most of the articles and books written on this subject dwell on theory and try to cover all cases of PLL-synthesizer design. This article narrows the focus to the design of a simple, passive three-pole loop filter typically used in low-voltage, low-operating-bandwidth synthesizer applications such as wireless transceivers.

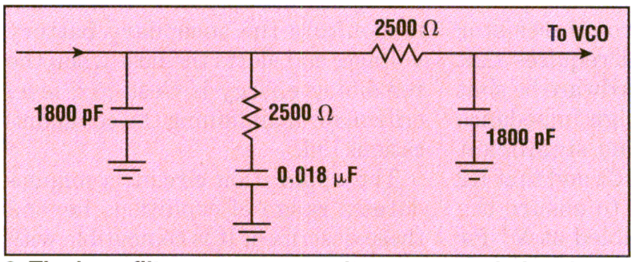
The calculations presented here have been found to



1. This block diagram shows the basic PLL layout.



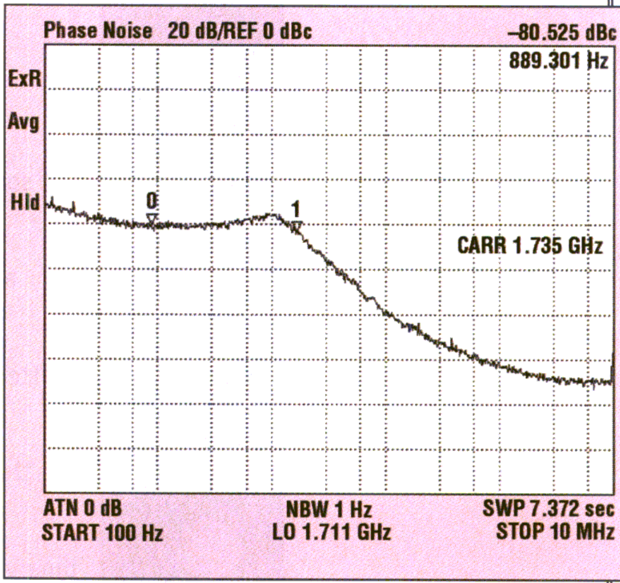
2. This schematic illustrates the loop-filter configuration.



3. The loop-filter component values are rounded to the nearest standard values.

achieve the expected performance goals. If the information used in the calculations is accurate, the PLL synthesizer will perform as designed. Experience has shown that if the PLL synthesizer does not perform as expected, some component part or device specification is in error.

Due to current levels of semiconductor integration, most of the components that make up the synthesizer (Fig. 1)—the PLL, the reference oscillator, and the voltage-controlled oscillator (VCO)—are available as integrated circuits (ICs). The only external components that are needed are the DC-decoupling elements, RF bypass elements, and the passive loop-filter components.



4. The excellent phase-noise performance of the RF PLL of Fujitsu's new MB15F08SL dual 2.5/1.2-GHz PLL is illustrated using the calculated loop-filter values.

The simplified loop-filter design formulas, found in Fujitsu's Super PLL Application Guide, are detailed. The formulas are based on the use of a basic passive two-pole loop filter along with a single-pole spur filter (Fig. 2).

The basic terms and definitions that are related to PLL synthesizers can be defined as follows:

F_{step} = the maximum frequency change during a step, or hop, from one frequency to another,

t_s = the desired time for the carrier to step to a new frequency,

f_a = the frequency of the carrier, within the desired time (t_s), after a step or hop—normally 1000 Hz,

ξ = the damping factor (A value of 0.707 is typical.),

f_n = the natural frequency,

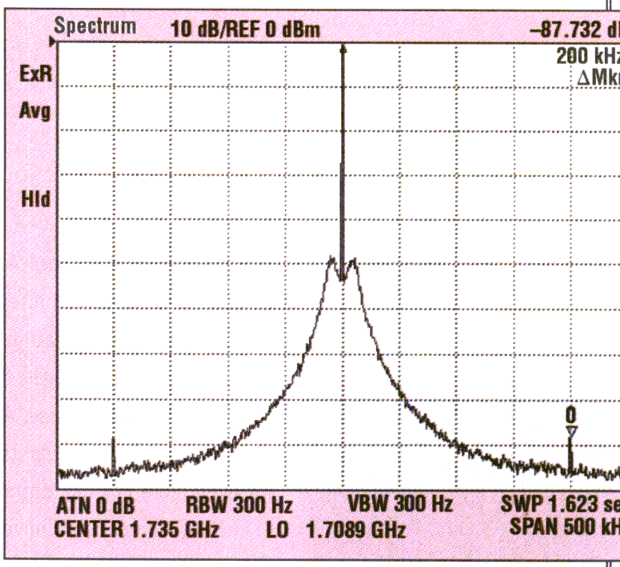
I_{cp} = the charge-pump current, and

K_{vco} = the VCO sensitivity.

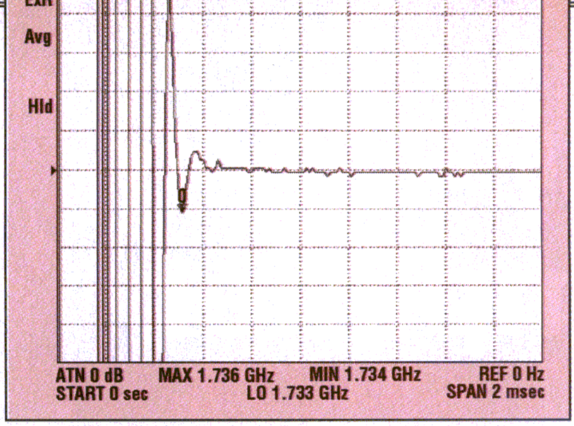
The design of a basic PLL synthesizer loop filter follows a few simple calculations.

First, determine the maximum dividing ratio, N, by:

$$N = \frac{\textit{maximum VCO frequency}}{\textit{channel spacing}} \qquad (1)$$



5. The 200-kHz spurious signals are shown to be an impressive -87.7 dBc, typical of the new Fujitsu SL series of advanced PLL synthesizers, providing optimum performance for the latest digital wireless communications designs.



6. It takes 514 μs to change the frequency from 1675 MHz to 1735 MHz ±1000 Hz.

Then, calculate f_n from eq. 2:

$$f_n = (-1/2\pi t_s \xi)\ ln\,(f_a / f_{step}) \quad (2)$$

The value of capacitor C2 must now be found from eq. 3:

$$C2 = I_{cp} K_{vco} / N(2\pi f_n)^2 \quad (3)$$

Similarly, the value of resistor R1 can be found from eq. 4:

$$R1 = 2\xi [N/(I_{cp} K_{vco} C2)]^{0.5} \quad (4)$$

By knowing the value of C2, it is now possible to find C1.

$$C1 = C2/10 \quad (5)$$

The final step involves the calculations of resistor R2 and capacitor C3, which are used to create the spur filter. R2 and C3 are used to reduce any spurious energy caused by the reference frequency. The product of R2 and C3 should be at least one-tenth the product of C2 and R1.

DESIGN EXAMPLE

As a design exercise, it is first necessary to define the basic synthesizer requirements for a sample application, and then define the specifications for the active components.

For this example, a hypothetical application with arbitrary low-side injection is specified. It uses a model MB15F08SL PLL IC from Fujitsu Electronics (San Jose, CA) and a VCO with a 25-V/MHz sensitivity. The application also requires a frequency range of 1675 to 1735 MHz, channel spacing of 200 kHz, maximum frequency hop of 60 MHz, a frequency hop time of 500 μs, and a frequency accuracy of 1000 Hz after the specified hop time.

The specifications for the active components are defined as follows: the VCO sensitivity is 25 MHz/V and the PLL IC charge-pump current is 6 mA.

Using the loop-filter equations, the synthesizer requirements, and the active-component specifications, the loop-filter component values can now be calculated.

Using eq. 1, N can now be calculated by:

$$N = \frac{1735\ MHz}{200\ kHz} = 8675$$

Using eq. 2, f_n can be found by:

$$f_n = [-1/(6.28)(0.5e^{-3}) \times (0.707)]\,ln(1000/60e^6)$$

$$f_n = 4,955.95\ Hz$$

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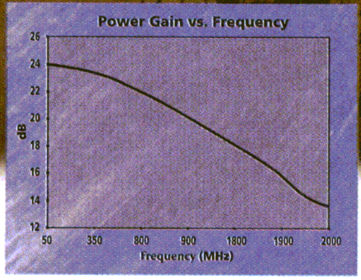

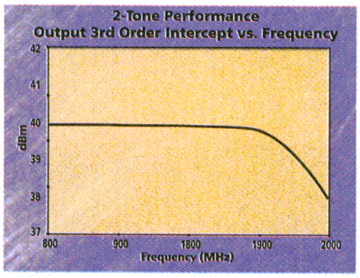

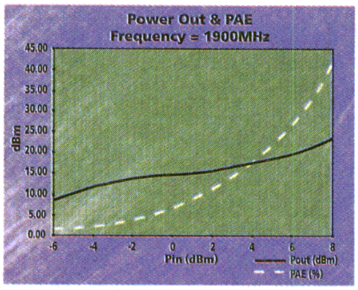

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1800-2000	14.5	1.5:1	1.7:1	22.0	39.0	5.0	95.0

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Using eq. 3, C2 can be determined by:

$$C2 = [(.006)(25e^{6})] / 8675[(6.28)(4955.95)]^{2}$$

$$C2 = 0.01785\ \mu F$$

Using eq. 4, R1 can be calculated by:

$$R1 = (2)(0.707)[(8675)/(0.006) \times (25e^{6})(0.01785e^{-6})]^{0.5}$$

$$R1 = 2545\ \Omega$$

Using eq. 5, C1 can be found by:

$$C1 = 0.01785\ \mu F / 10$$

$$C1 = 0.001785\ \mu F$$

Finally, R2 and C3 (the spur filter) must be determined. The product of R2 and C3 should be approximately one-tenth the product of R1 and C2.

THE GRAPHS CONFIRM THAT THE CALCULATIONS WORK WELL WHEN DESIGNING LOOP FILTERS TO BE USED IN MANY MODERN PLL APPLICATIONS.

$$R2 = 2545\ \Omega$$

$$C3 = 0.01785\ \mu F / 10 = 0.001785\ \mu F$$

After the loop-filter values have been found (Fig. 3), the loop bandwidth can be calculated. Knowing the loop bandwidth will help determine if the PLL is operating correctly when the phase noise is displayed on a spectrum analyzer.

The loop bandwidth is calculated using the following equation:

$$\text{Loop bandwidth} = (2\pi f_n / 2)\ [\xi + (1/4\xi)]\ Hz$$

Using the values determined earlier, the loop bandwidth for this example is:

$$\text{Loop bandwidth} = (6.28)(4955)/2[0.707 + (1/2.828)]$$

$$\text{Loop bandwidth} = 16,500\ Hz$$

The performance of the PLL synthesizer with the calculated values were measured with a spectrum analyzer and plotted (Figs. 4 to 6). The graphs confirm that the calculations work well when designing loop filters to be used in many modern PLL applications.

Marker "0" shows the phase noise inside the loop to be –80.5 dBc/Hz. Marker "1" shows that the loop bandwidth is approximately 1.65 kHz.

An important part of the loop-filter design is using components that will not degrade the performance of the PLL synthesizer. Capacitors must have very-low leakage. Ceramic capacitors should not be used, since the piezoelectric effects may cause noise and even microphonics on the VCO tuning line. Film capacitors are recommended. The resistors should consist of metal or carbon film. Resistors that are composed of carbon are not recommended.

The layout of the printed-circuit board (PCB) can affect the level of VCO spurious signals and noise. When laying out the PCB for minimal noise, two things are important. First, provide the shortest possible ground path between the PLL IC ground pins, loop-filter ground, as well as the ground for the VCO varicap tuning diode. If a packaged VCO is used, it should be mounted close to the PLL IC and loop filter. Second, bypass the V_{CC} lines that feed the

THE LAYOUT OF THE PRINTED-CIRCUIT BOARD (PCB) CAN AFFECT THE LEVEL OF VCO SPURIOUS SIGNALS AND NOISE.

PLL chip with a small-value capacitor (0.1 μF) and a large-value capacitor (10 μF). These capacitors should be placed as close as possible to the V_{CC} pins. Bypassing should also be used for the VCO. If the PLL and the VCO use the same value for V_{CC}, a 22-Ω resistor should be placed in the V_{CC} line between them in order to improve the isolation.

Synthesizers that operate well above 2000 MHz can be manufactured with relatively few problems—as long as good RF techniques are used for board layout and parts placement. ••

For Further Reading

Fujitsu Microelectronics, Inc., *Super PLL Application Guide*, TC-AN-20731-11/98.
Fujitsu Microelectronics, Inc., *MB15FxxSL Series Data booklet*, TC-DS-20788-2/99.
Ulrich L. Rohde, *Microwave and Wireless Synthesizers*, John Wiley & Sons, Inc.
Vadim Manassewitsch, *Frequency Synthesizers*, John Wiley & Sons, Inc.


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Examine The Impact Of Transition-Band Spurs On Downconverters

This analysis will help to explain the effects of transition-band spurious signals on downconverter dynamic range.

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FREQUENCY downconversion is essential to most electronic-warfare (EW) frequency-measurement receivers. While in-band intermodulation distortion (IMD) from the receiver components (such as mixers and amplifiers) can limit downconverter dynamic range, spurious signals produced through filter transition bands can also adversely affect downconverter dynamic-range performance.

In general, spurious signals can be divided into two types. The first type, which occurs within the operating bands of the front-end signal-processing components, mainly consists of IM signals generated by mixer noise as well as harmonic signals generated by intermediate-frequency (IF) amplifiers. Such spurious signals can be suppressed through the careful selection of mixer RF and IF bands and through the use of wide-dynamic-range IF amplifiers. The second type consists of out-of-band responses, mainly image signals generated by the receiver's mixers and caused by mixing of RF and local-oscillator (LO) signals within the stopbands of the RF and IF filters. These out-of-band spurious signals can usually be suppressed through bandpass filters and image-reject mixers.

Unfortunately, RF and IF bandpass filters are not ideal, and some spurious signals will escape. No matter what the specified out-of-band suppression, signals falling within a filter's transition band—that band between the passband and the rejection band—can pass from the input to the output with little attenuation and be presented to the mixer as real signals (or transition-band RF signals). The IFs that result from these RF signals (known as transition-band IF signals) may also fall within the transition bands of the IF filters.

In order to analyze the properties of these transition-band spurious signals, it may help to review the principles of microwave frequency downconverters. Assume that f_R is the frequency of the input signal to the microwave frequency downconverter, with a power level of P_R (in dBm). After the signal passes through a high-frequency RF bandpass filter, which has insertion loss of L_{AR} (in dB), the resulting output power level is P_{Rout} [in dBm] (i.e., $P_{Rout} = P_R - L_{AR}$). For a mixer with conversion loss of L_M, f_R is downconverted to an IF of f_I via mixing with the LO frequency, f_{local} (i.e., $f_I = f_{local} - f_R$ or $f_I = f_R - f_{local}$, with a power level of $P_I = P_R - L_{AR} - L_M$ [in dBm]). The IF signal f_I then passes through the IF bandpass filter with insertion loss (in dB) of L_{AI}, yielding output power of $P_{Iout} = P_I - L_{AI} = P_R - L_{AR} - L_M - L_{AI}$ (in dBm). Following the signal gain through the IF amplifier, the final output of the frequency downconverter is at a frequency of f_{out} and power of P_{out}.

For analysis purposes, the downconverter's RF bandpass filter passband is assumed to lie between a lower bandedge of f_{RL} and an upper bandedge of f_{RH}. The IF range is also assumed to be less than one octave (i.e., $f_{IH} < 2f_{IL}$, where f_{IL} and f_{IH} are the cutoff frequencies of the bandpass filter). If f_{RS} is the RF transition-band signal and f_{IS} is the IF transition-band spurious signal of the microwave frequency downconverter, then:

$$f_{IL} \le 2f_{Is} \le f_{IH} \quad (1)$$

$$f_I = f_{local} - f_R \quad (2)$$

$$f_{Is} = f_{local} - f_{Rs} \quad (3)$$

$$f_{IL} = f_{local} - f_{RH} \quad (4)$$

$$f_{IH} = f_{local} - f_{RL} \quad (5)$$

$$f_I = f_R - f_{local} \quad (6)$$

$$f_{Is} = f_{Rs} - f_{local} \quad (7)$$

$$f_{IL} = f_{RL} - f_{local} \quad (8)$$

$$f_{IH} = f_{RH} - f_{local} \quad (9)$$

where:

f_{IL} = the lower cutoff frequency of the IF bandpass filter, and

f_{IH} = the higher cutoff frequency of the IF bandpass filter.

Solving for eqs. 1 and 2-5 or for eqs. 1 and 6-9 yields:

$$(f_{local} + f_{RL})/2 \le f_{Rs} \le (f_{local} + f_{RH})/2 \quad (10)$$

which are the RF transition-band signals which can give rise to transition-band spurious signals.

If the input power to the microwave frequency downconverter is P_R (in dBm), then the input IF power to the IF amplifier is:

$$P_{Iout}(f_R) = P_R - L_{AR}(Type_R, n_R f_R) - L_M - L_{AI}(Type_I, n_I f_I)\ (dBm) \quad (11)$$

where:

$L_{AR}(Type_R, n_R, f_R)$ = the RF bandpass-filter insertion loss, which is a function of the RF bandpass-filter type, order (n), and IF (f_R),

$L_{AI}(Type_R, n_R, f_R$ = the IF bandpass filter insertion loss, which is a function of the IF bandpass-filter type, order (n), and IF (f_I), and

L_M = the mixer conversion loss.

The relationship between f_R and f_I satisfies eq. 2 or 6. For an in-band signal f_R, $f_{RL} \le f_R \le f_{RH}$, the final output signal of the microwave frequency downconverter is:

$$P_{out}(f_{RL} \le f_R \le f_{RH}) = P_{Iout}(f_{RL} \le f_R \le f_{RH}) + G\ (dBm) \quad (12a)$$

where:

G = the linear gain of the IF amp.

For an RF transition-band signal $f_R = f_{RS}$, satisfying eq. 10, the final output power level of the transition-band spurious signal coming out of the downconverter is:

$$P_{out}(f_{Rs}) = 2[P_{Iout}(f_{Rs}) + G] - Q_2\ (dBm) \quad (13a)$$

where:

Q_2 = the second-order intercept point (IP2) of the IF amplifier.

The transition-band spurious signal power level represented by eq. 13a restricts a downconverter's dynamic range. At the sensitivity limit of a microwave receiver, the minimum detectable input RF signal f_R within the passband of the RF bandpass filter is at a power level of $P_{R\,min}$ (in dBm). By combining eqs. 11 and 12a, the output power level from the microwave frequency downconverter is:

$$P_{out}(f_{RL} \le f_R \le f_{RH}) = P_{R\,min} - L_{AR}(f_{RL} \le f_R \le f_{RH}) - L_M - L_{AI}(f_{IL} \le f_I \le f_{IH}) + G\ (dBm) \qquad (12b)$$

For an RF transition-band signal at the maximum power level of $P_{R\,max}$, the output power of the transition-band spurious signal can be found by combining eqs. 11 and 13a:

$$P_{out}(f_{Rs}) = 2[P_{R\,max} - L_{AR}(f_{Rs}) - L_M - L_{AI}(f_{Is}) + G] - Q_2\ (dBm) \qquad (13b)$$

Provided that eq. 12b is equal to eq. 13b,
$P_{out}(f_{RL} \le f_R \le f_{RH}) = P_{out}(f_{RS})$,
and the dynamic range, D, can be found from eq. 14:

$$D = P_{R\,max} - P_{R\,min} = L_{AR}(f_{Rs}) + L_{AI}(f_{Is}) - 0.5[L_{AR}(f_{RL} \le f_R \le f_{RH}) + L_{AI}(f_{IL} \le f_I \le f_{IH}) + P_{R\,min} + G - L_M - Q_2]\ (dB) \qquad (14)$$

which yields the dynamic range of a downconverter restricted by transition-band spurious signals.

DOWNCONVERTER PERFORMANCE

An example may help to demonstrate the effects of transition-band spurious signals on microwave frequency-downconverter performance. Assume a frequency-downconverter subsystem designed for RF signals from 8 to 10 GHz, with an IF of 2 to 4 GHz and LO or $f_{local} < f_R$, then f_{local} = 12 GHz, f_{RH} = 10 GHz, and f_{LH} = 8 GHz. Substituting these values into eq. 10 yields the RF transition-band frequencies that can cause transition-band spurious products:

$$10\ GHz \le f_{Rs} \le 11\ GHz \qquad (15)$$

Then by using eq. 3, the corresponding IF transition-band frequencies are in the range of:

$$2\ GHz \ge f_{Is} \ge 1\ GHz \qquad (16)$$

Consider the frequency transformation from a bandpass filter to a lowpass filter:

$$f' = (f_0 / B)[(f / f_0) - (f_0 / f)] \qquad (17)$$

where:
f' = the normalized frequency of the prototype lowpass filter,
$B = f_2 - f_1$ which is the bandwidth of the bandpass filter,
$f_0 = (f_2 f_1)^{0.5}$ which is the geometric mean of the cutoff frequencies f_1 and f_2 and is referred to as the midband frequency, and
f = the frequency of the bandpass filter.

A good example for analysis is a prototype lowpass filter with seventh-order Chebyshev response of:

$$L_A = 10 \log[1 + \varepsilon^2 C_7^2(f')] \qquad (18)$$

where:

$$C_7(f') = 64(f')^7 - 112(f')^5 + 56(f')^3 - 7f' \qquad (19)$$

is the seventh-order Chebyshev polynomial of the first kind and $\epsilon = 1$ if the passband ripple is required to not exceed 3 dB.

For the transition-band frequencies of the RF bandpass filter, replacing f in eq. 17 with f_{RS} of eq. 15 and substituting $f_2 = f_{RH} = 10$ GHz and $f_1 = f_{RL} = 8$ GHz in eq. 17 yields:

$$1.0 \leq f'_{Rs} \leq 1.8636 \qquad (20)$$

Replacing f′ of eq. 18 with f_{RS}' of eq. 20, the RF transition-band insertion loss of the filter can be found from:

$$3.0 \leq L_{AR}(f_{Rs}) \leq 69.0 \qquad (21)$$

Similarly, the corresponding IF transition-band insertion loss can be found from:

$$3.0 \leq L_{AI}(f_{Is}) \leq 111.0 \qquad (22)$$

Substituting $L_{AR}(f_{RS} = 10\ \text{GHz}) = 3$ dB, $L_{AI}(f_{IS} = 2\ \text{GHz}) = 3$ dB, $L_{AR}(8\ \text{GHz} \leq f_R \leq 10\ \text{GHz}) \leq 3$ dB, and $L_{AI}(2\ \text{GHz} \leq f_I \leq 4\ \text{GHz}) \leq 3$ dB into eq. 14, the dynamic range restricted by the transition-band spurious signals is:

$$D_1 = 3 + 3 - 0.5\ [3 + 3 + P_{R\ min} + G - L_M - Q_2\ ]\ (dB) \qquad (23)$$

On the other hand, if the IF is changed to a range of 2.5 to 4.5 GHz, and the LO is changed to 12.5 GHz, the RF transition-band frequencies that can cause transition-band spurious responses are:

$$10.25\ GHz \leq f_{Rs} \leq 11.25\ GHz \qquad (24)$$

and the corresponding IF transition-band frequencies would be in the range of:

$$2.25\ GHz \geq f_{Is} \geq 1.25\ GHz \qquad (25)$$

If the same seventh-order prototype lowpass Chebyshev filter is used as in the previous example, then the RF transition-band insertion loss is:

$$33.8 \leq L_{AR}(f_{Rs}) \leq 76.4 \qquad (26)$$

and the corresponding IF transition-band insertion loss is:

$$45.1 \leq L_{AI}(f_{Is}) \leq 117.4 \qquad (27)$$

Substituting $L_{AR}(f_{RS} = 10.25\ \text{GHz}) = 33.8$ dB, $L_{AI}(f_{IS} = 2.25\ \text{GHz}) = 45.1$ dB, $L_{AR}(8\ \text{GHz} \leq f_R \leq 10\ \text{GHz}) \leq 3$ dB, and $L_{AI}(2.5\ \text{GHz} \leq f_I \leq 4.5\ \text{GHz}) \leq 3$ dB into eq. 14, the dynamic range for this modified downconverter is:

$$D_2 = 33.8 + 45.1 - 0.5\ [3 + 3 + P_{R\ min} + G - L_M - Q_2\ ]\ (dB) \qquad (28)$$

The difference, Δ, between D_1 and D_2 is simply:

$$\Delta D = D_2 - D_1 = 33.8 + 45.1 - 3 - 3 = 72.9\ (dB) \qquad (29)$$

Transition-band spurious signals should be considered during the frequency downconverter design process. To limit these signals, the LO frequency should be as far as possible from the RF bandpass-filter cutoff frequencies for IF bands that are not more than one octave. Transition-band spurious signals can also be reduced by careful RF and IF bandpass-filter design and/or selection, choosing filters with the narrowest possible transition bandwidths. ••

Design Approach Yields Broadband Directional Couplers

Application of cosine⁴ coupling-coefficient theory can produce microwave directional couplers with low ripple.

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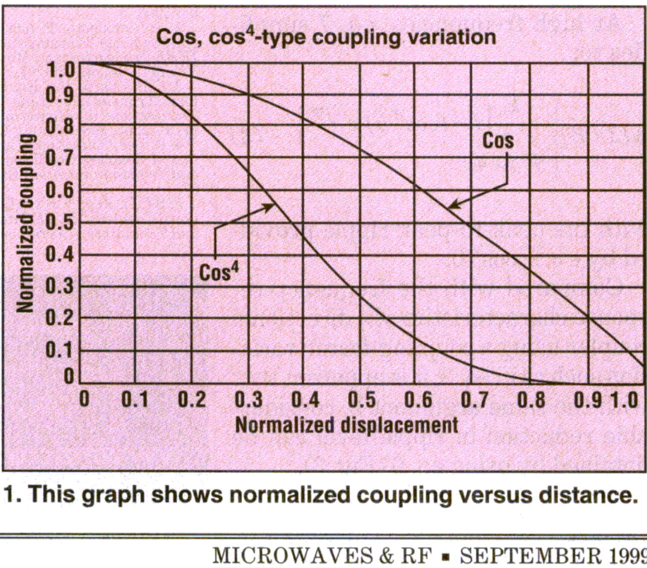
DIRECTIONAL couplers with low ripple over broad bandwidths are vital to certain applications, including instrumentation and electronic countermeasures (ECM). Tapered stripline directional couplers are often used to achieve these wide bandwidths. In the past, a variety of coupling coefficients has been used to design compact directional couplers having low ripple. This article compares two conventional highpass type non-uniformly coupled stripline directional couplers—one with a cosine (cos) coupling variation, the other with a $\cos^4$ coupling variation. The comparison shows that, when the same coupling-factor argument is used and the length of the coupler is fixed, less ripple results with the $\cos^4$-type coupling variation.

Two couplers were designed and built based on cos and $\cos^4$ coupling-coefficient equations. The 2-in. (4.08-cm)-long couplers were designed to have a bandwidth of 6 to 20 GHz and a coupling coefficient of 20 dB using dielectric material with a dielectric constant of 2.2. Empirical data derived from frequency-response tests performed on these couplers show close agreement with the theory.



1. This graph shows normalized coupling versus distance.

Directional-coupler theory states that a transverse-electromagnetic (TEM)-mode directional coupler with continuously tapered, non-zero-coupling physical characteristics should exhibit a highpass type, near-equal-ripple frequency response.[1-3] The coupling factor K(z) along the length of the coupler is a function of the odd- and even-mode impedances of the two coupled transmission lines and is expressed by:

$$K(z) = \frac{Z_{oe}(z) - Z_{oo}(z)}{Z_{oe}(z) + Z_{oo}(z)} \quad (1)$$

For loose coupling, the complex frequency response can be obtained by the following transform:

$$A13\ (\Psi) = j\frac{\Psi}{2}\int_o^l K(x)e^{-j\Psi x}\,dx \quad (2)$$

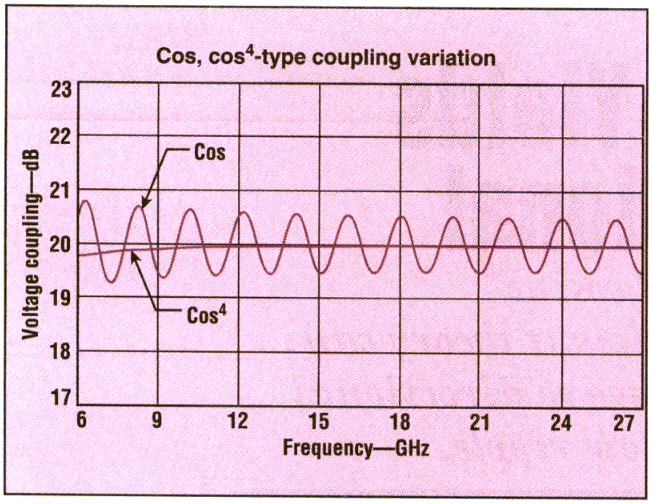
where:

A13 (Ψ) = the complex voltage coupling from port 1 to port 3,

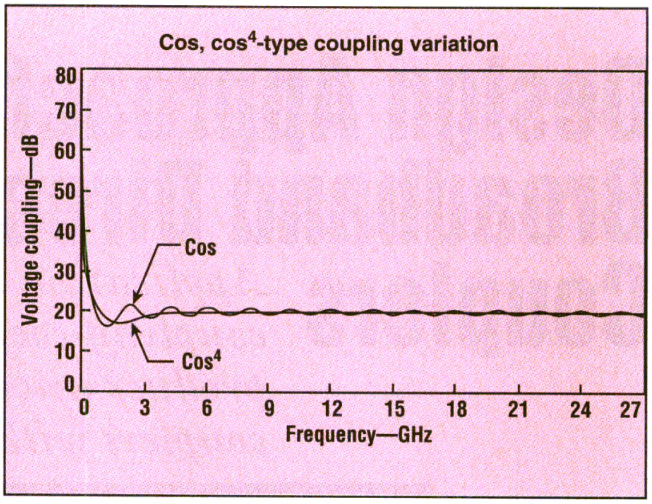
K(x) = the normalized coupling factor,

x = z / 1 = the normalized displacement,

z = the displacement along the



2. Directional coupler frequency response is illustrated in the graph.



3. Similar to Fig. 2, the directional coupler frequency response can also be seen here.

length of the coupler,
1 = the length of the coupler,
$\Psi = 2\beta l$,
$\beta = 2\pi/\lambda$, and
λ = the wavelength.

Extensive analyses have been carried out for exponential and non-exponential cosine types of coupling variations.[4,5] Exponentially varying coupling coefficients have produced large and unequal ripple-voltage coupling responses. To achieve equal ripple responses, tedious and complicated mathematical procedures are required. Reasonable improvement of the large ripples may be achieved with a cosine variation. However, for very-small ripple specification, couplers would need large coupling lengths (Fig. 1). If the coupling variation is chosen to be a cos⁴ type, then the same length can lead to a coupling response with smaller ripple.

The frequency response of a cosine coupling variation such as:

$$K(x) = k_o \cos(bx) \qquad (3)$$

can be found by substituting eq. 3 into eq. 2 and integrating.

$$A13(\Psi) = \frac{k_o}{2}\left\{\frac{\Psi^2}{(\Psi^2 - b^2)}\left[1 - \left(\cos b + j\frac{b}{\Psi}\sin b\right)e^{-j\Psi}\right]\right\} \qquad (4)$$

At high frequencies, eq. 4 simplifies to:

$$A13(\Psi)\Big|_{\Psi \gg b} \rightarrow \frac{k_o}{2}\left[1 - (\cos b)\,e^{-j\Psi}\right] \qquad (5)$$

with the peak-to-peak ripple given by $\epsilon = k_o \cos b$.

If the coupling variation is chosen as:

$$K(x) = k_o \cos^4(bx) \qquad (6)$$

and eq. 6 is substituted into eq. 2 and integrated, the following response (Fig. 2) is obtained:

$$\begin{aligned} A13(\Psi) = \frac{k_o}{2}\Bigg\{ & \frac{3}{8}(1 - e^{-j\Psi}) \\ & + \frac{\Psi^2}{2[\Psi^2 - (2b)^2]}\left[1 - (\cos 2b + j\frac{2b}{\Psi}\sin 2b)\,e^{-j\Psi}\right] \\ & + \frac{\Psi^2}{8[\Psi^2 - (4b)^2]}\left[1 - (\cos 4b + j\frac{4b}{\Psi}\sin 4b)\,e^{-j\Psi}\right]\Bigg\} \end{aligned} \qquad (7)$$

At high frequencies, eq. 7 simplifies to:

$$A13(\Psi)\Big|_{\Psi \gg 4b} \rightarrow \frac{k_o}{2}\left[1 - (\cos^4 b)\,e^{-j\Psi}\right] \qquad (8)$$

with the peak-to-peak ripple provided by $\epsilon = k_o\cos^4 b$.

Compared with the frequency-response characteristics of a directional coupler using a coupling-factor variation such as in eq. 3, it is apparent that with the same argument b, considerable reduction in ripple level can be obtained by using eq. 6 (Fig. 3).

From Figs. 2 and 3, it can be seen that the cos⁴ design is considerably flatter with less overshoot at the low end of the frequency band than the conventional cos design. Within the frequency band of interest, the actual peak-to-peak ripple was 1.6 dB for the cos design and 0.3 dB for the cos⁴ design. In addition to the better ripple response, a very important difference between the two couplers was that, with a finite length of the coupling section, it was not possible to contain the ripple level of a cos-type coupler to within arbitrarily small limits due to the overshoot and the much smaller convergence speed of the design equations. However, with the cos⁴ coupler, a relatively short length would be sufficient for the ripple of the coupler to converge to a negligible level. ••

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Gain Flatness (dB.)	±0.2	±0.2	±0.3	±0.3	±0.2
Pout @ 1 dB. comp. (dBm.)	+27.5	+33.0	+33.0	+33.0	+26.5
IP3 (dBm.)	+45.0	+49.0	+50.0	+49.0	+44.0
Noise Figure (dB.)	0.9	2.5	3.5	3.0	1.0
Vcc/Ic (v/mA.)	+15/200	+15/700	+10/725	+12/725	+15/200
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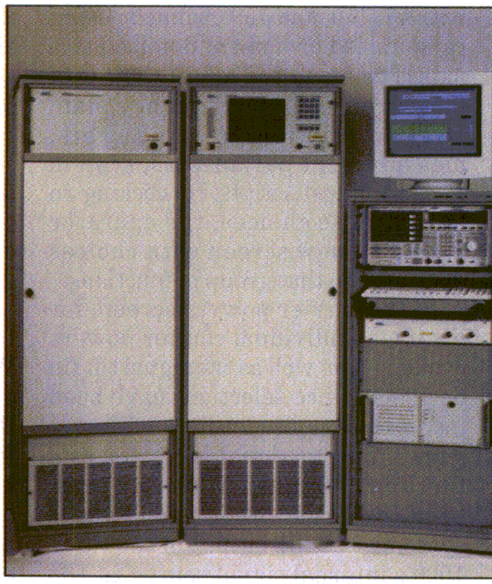
Test System Tackles CATV Components

This turnkey production test system is ideal for checking the CTB and CSO performance of amplifiers and other components in a cable-television system.

JACK BROWNE
Publisher/Editor

CABLE-TELEVISION (CATV) systems grow more complex with time, as operators seek to boost revenues through additional services, such as Internet access and high-speed data transmissions. As a consequence, the characterization of CATV system components also becomes more complicated, especially as these components must cope with newer digital modulation schemes. Fortunately, the CTS-1000 distortion measurement system from RDL, Inc. (Conshohocken, PA) was designed for testing today's and tomorrow's CATV components. This flexible, turnkey system can automatically perform the most-complex CATV linearity tests, such as composite-second-order (CSO), composite-triple-beat (CTB), CROSS-MOD (XMOD), and carrier-to-noise (C/N) measurements, with as many as 159 discrete channels.

The CTS-1000 distortion measurement system (Fig. 1) builds upon the company's model CSG programmable multi-tone signal generator (see *Microwaves & RF*, January 1996, p. 102), with available channel frequencies from 44 to 998 MHz. It adds a modified model ESCS30 receiver from Rohde & Schwarz (Munich, Germany), with intermediate-frequency (IF) capability enhanced to meet the needs of CATV linearity testing. The receiver (a spectrum analyzer) also provides the video-output signals required by the CTS-1000 test system for performing Fast Fourier transforms (FFTs). Additional system components include a computer, monitor, keyboard, interface module, interconnecting cables, and software for making measurements, logging data, and performing analyses. Optional leveling capability is available with the addition of a power meter.

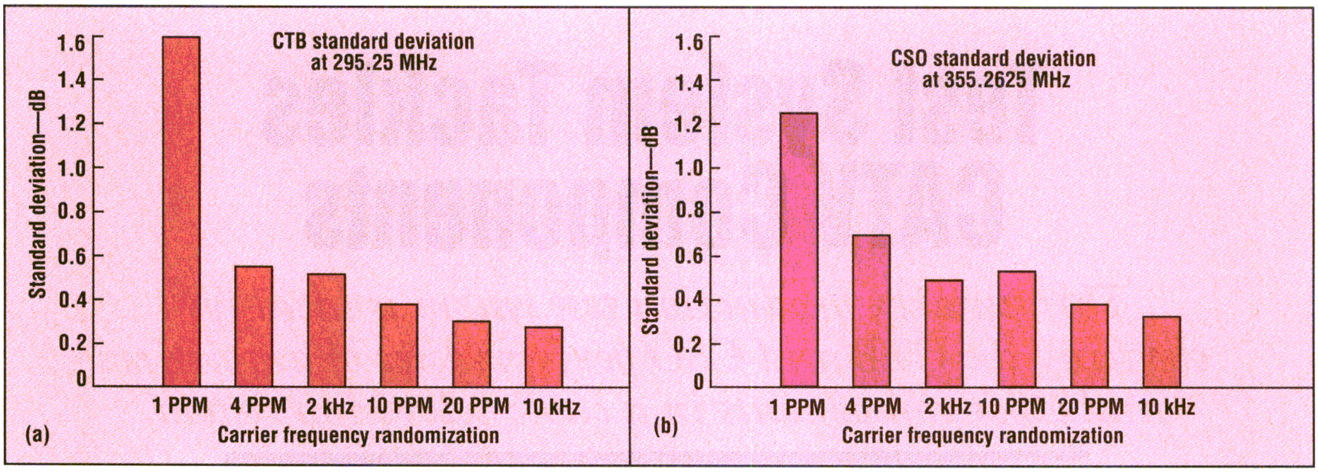


1. The CTS-1000 distortion measurement system flawlessly blends precision test hardware with control and analysis software to speed and simplify complex CATV measurements, such as CSO, CTB, and XMOD.

The CTS-1000 system can be equipped with as many as 159 synthesized frequency sources or channels, although the standard version of the system is supplied with 138 channels. Each source tunes ±3 MHz about a center frequency, accommodating a variety of CATV frequency plans, such as the 6-MHz channels in the US and the 8-MHz channels used throughout Europe. The individual sources tune with 1-Hz frequency resolution. Maximum signal level to the DUT is +45 dBmV with level repeatability of 0.5 dB. In addition, signal levels for each source (or channel) can be adjusted from +10 to +46 dBmV. The system is equipped with as much as 63-dB attenuation, which adjusts all carriers in 1-dB steps. The output level of the individual carriers can be adjusted over a 15-dB range.

ACHIEVING ACCURACY

The CTS-1000 enables operators to improve measurement repeatability by using uniformly distributed test carrier frequencies. Measurement repeatability is affected by a number of factors, including variations in RF test-signal amplitude, frequency drift over time, differences between multiple test-signal sources, and out-of-band signal products. By randomizing the test signals in the CTS-1000, which is a process of using precisely programmed frequency dispersion, the measurement repeatability of CATV testing can be dramatically im-



2. The use of precisely controlled frequency dispersion can help improve the measurement repeatability of CTB (a) and CSO (b) tests.

proved (Fig. 2). In addition to the increase in measurement repeatability and accuracy, the use of randomized test signals can help reduce test time and provide more authentic representations of the signals experienced in an actual CATV system. The randomized function can be programmed in the CTS-1000 in terms of frequency spread (in ±Hz) or frequency deviation (in ± PPM) across a user-selected group of channels.

Along with its impressive hardware, the CTS-1000 is equipped with powerful, but easy-to-use software. Two years in development, and combining several C++ programs with LabView instrument drivers, the CTS-1000 software provides a true Microsoft Windows operating environment for setting levels, adjusting controls, making measurements, gathering and plotting data, and saving results. The software provides seamless control of the receiver and frequency synthesizers, and provides a rational user interface to a series of tests that are by no means simple.

DEMO SOFTWARE

For engineers interested in evaluating the CTS-1000's software interface, the company makes demonstration compact-disc-read-only memories (CD-ROMs) available. This is actually a fully functional version of the software that can also operate offline (without the test equipment). It takes approximately five minutes to install in a personal computer (PC), requiring a machine with Windows 95 or Windows 98, at least 16 Mb or random-access memory (RAM), at least 15 Mb of available hard-disk memory, and a super VGA (SVGA) monitor. The CD-ROM is supplied with a useful booklet with instructions for measuring CSO, CTB, and CNR at each frequency as well as advice on setting up and performing XMOD measurements.

Upon loading the software, the introductory or home screen provides operators with a menu bar and details on the current measurement setup. System configurations include setups with 77, 110, and 159 channels, using randomized or non-randomized signals. The menu bar offers choices of measure, results, frequency plan, DUT description, carrier powers, measurement parameters, system setup, and test scripts. By clicking on any of these choices, the operator calls up a new screen with choices dedicated to that group of functions.

In the carrier powers screen, for example, individual carrier powers can be set, as well as the signal tilt (in decibels). Once selections have been made, the amplitude shape of the resulting output signal is plotted as a function of frequency, using a number of different probe points in the systems—at the output of the CTS-1000's internal signal source, at the DUT input, at the input to the network (such as a filter) preceding the DUT, at the DUT output, and at the output of the network following the DUT. The plot shows frequencies from 0 to 1000 MHz, with amplitude automatically scaled depending upon the choice of probe point.

The other menu choices (and screens) are designed for ease of use. The frequency plan screen offers a spreadsheet-like view of carrier information, including whether the carriers are on or off, the type of modulation, and the percent modulation. A choice of modulation sources includes external, PAL, NTSC, or none.

The DUT description screen includes a simple block diagram of the test fixture. It shows how the forward transmission S_{21} is defined for the connecting networks before and after the DUT. Slopes can be defined across the desired frequency range to account for cable losses.

The measurement parameters screen shows the type of detector (peak or average) used for each measurement (CTB, CSO, XMOD), the options for expressing power (in dBmV, dBm, or dBμV.

The software requires no manual to operate effectively. It has a true Windows feel, and provides ready access to the full operating power of the CTS-1000 system. The system offers a variety of advantages over existing systems, including its ability to quickly modify the frequency and power of the carriers, and its ability to measure nine complete sets of distortion data in under 85 s. **RDL, Inc., 7th Avenue and Freedley St., Conshohocken, PA 19428; (610) 825-3750, FAX: (610) 825-3530, Internet: http://www.rdl-instrumentation.com.** ••

Instrument Makers Adapt To Evolving Applications

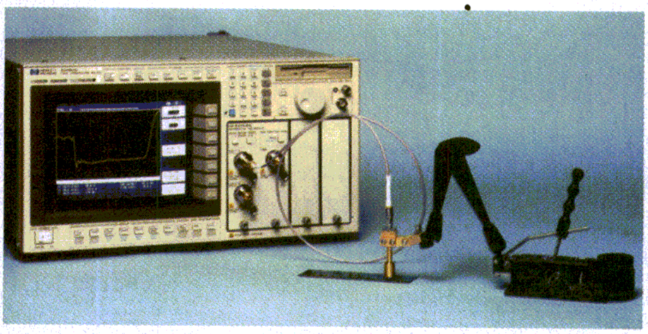
Continual integration of functions and processing power pushes high-frequency test gear to new levels of analog and digital measurement capability.

JACK BROWNE
Publisher/Editor

INTEGRATION is a key to market acceptance not only at the device level, but for test equipment as well. Users of test equipment are increasingly seeking more functions and capability within a single box, driving instrument manufacturers to shrink and compress their architectures to support current requirements and future needs. The end result is more measurement power per dollar then ever before, within instruments that are increasingly similar to self-contained systems.

One of the greatest challenges facing current and future designers of high-speed, high-frequency circuits is the accurate characterization of signals that are digital and microwave in nature. An example of this is the Direct Rambus, where data transfer occurs on the rising and falling edges of 400-MHz clock signals in a controlled-impedance environment of 28 Ω. The bus system achieves peak bandwidths up to 1.6 GB/s, posing a challenge for engineers faced with determining signal integrity. While a traditional logic analyzer can be used to check the data levels, a different solution is needed to analyze the integrity of these high-speed digital signals. One solution offered by Hewlett-Packard Co.'s Lightwave Division (Santa Rosa, CA) is the use of time-domain reflectometry based on the HP 54750A digitizing oscilloscope along with the HP N1020A time-domain-reflectometer (TDR) probe kit (Fig. 1). A special calibration adapter with a 50-Ω air-dielectric termination and SMA short can be attached to the probe tip for calibration.

The oscilloscope enables operators to set the impedance measurement reference plane at the probe tips, in order to accurately determine signal integrity for the fine microstrip circuit traces found in Direct Rambus architectures. Before evaluating Direct Rambus designs, the oscilloscope undergoes a TDR calibration at the reference plane of the probe tip using a two-point calibration with the termination and short. The oscilloscope's internal processing power then performs a normalization routine to correct test-fixture errors in the frequency domain.



1. The HP 54750A digitizing oscilloscope works with the HP N1020A TDR probe kit to make impedance measurements on high-speed digital circuits such as the Direct Rambus digital bus. (Photograph courtesy of Hewlett-Packard Co., Palo Alto, CA.)

The increasing importance in timing measurements is one of the motivations for the development of the Femto-2000 multisite digital time scope from GuideTech (Sunnyvale, CA). It is a time- and frequency-measurement system that combines up to four time-interval analyzers and eight measurement channels in a single enclosure (Fig. 2). The instrument is well-suited for testing the jitter and waveform integrity of semiconductors and high-speed circuits. It offers 1-ps single-shot resolution for signals up to 800 MHz and makes measurements at rates to 2 million/s. The noise floor is 3 ps and the measurement jitter is only 6 ps. The instrument covers a voltage range of ±2.5 V with 1-mV accuracy and accommodates single-ended or differential signals.

The Femto-2000 can characterize phase-locked loops (PLLs) more completely than traditional time-measurement systems—swiftly measuring frequency, step response, loop-filter bandwidth, start-up time, and frequency settling time. For PLLs that use the spread-spectrum technique, the Femto-2000 can directly measure the modulation waveform and the attenuation of the loop filter for modulation and jitter at the

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input, even if the modulation is asynchronous to the clock frequency.

The Femto-2000 can also measure the timing of all pulses in a sequence at frequencies up to 2 MHz, thereby providing a complete test of pulse-width modulated signals in a single pass. To enhance viewing, analysis, and time to actionable data on all traditional and new measurements, the Femto-2000 features a proprietary graphic user interface (GUI) that enables simultaneous viewing of vs-time graphs, histograms, and numerical displays. The unit also comes equipped with 8-Mb memory for each pair of channels.

Much work on communications and video testing with an oscilloscope has been performed by LeCroy Corp. (Chestnut Ridge, NY) which recently published the application note "Using Analog Scopes In Video Applications." The note discusses the frame composition of different video standards, such as NTSC and PAL/SECAM, and explores techniques for successfully triggering on and analyzing video signals with an oscilloscope. The note addresses a variety of television measurements, including subcarrier-horizontal (SCH) phase measurements, jitter measurements, and the use of multiple time bases for zoom measurements on different aspects of complex video signals. Copies of the note are free upon request from LeCroy.

2. The Femto-2000 Multisite Digital Time Scope combines as many as four time-interval analyzers and eight measurement channels in a single enclosure. (Photograph courtesy of GuideTech, Sunnyvale, CA.)

Characterization of television signals is an important aspect of the UFX99CA programmable noise generator from Noise Com, Inc. (Paramus, NJ). Designed to set signal-to-noise ratios (SNRs) for analog

GAINING MEASUREMENT SPEED

Speed often determines the usefulness of a production test system, but can also be invaluable for research and device modeling. In the case of California Eastern Laboratories (Santa Clara, CA), a new noise-parameter test system not only aids the throughput of the company's device testing, but the accuracy as well. The noise-parameter test system makes the characterization of extremely low-noise devices, such as pseudomorphic high-electron-mobility transistors (PHEMTs), easier and more precise. It also simplifies the characterization of hard-to-measure high-gain devices.

The system was initially developed for California Eastern Laboratories by Charlie Woodin while at MicroVue, Inc., based on the firm's Wavevue software modules. Woodin became part of ATN Microwave (North Billerica, MA) following that company's acquisition of MicroVue. Two Wavevue software modules are used in the new noise-parameter test set. The first measures S-parameters and DC characteristics while the second handles the noise measurements. The system eventually became the basis for ATN Microwave's automatic NP5C noise-parameter system.

The differences between the new system and the old ways of measuring transistor noise parameters are staggering, according to Abby O'Connell, manager of California Eastern Laboratories' engineering laboratory. "This new automated system can do in hours what used to take us weeks," according to O'Connell. The old system was largely manual and required approximately two hours of test time to measure the noise figures for 12 different frequencies. The new system can make the same measurements in approximately one minute.

"In addition, the low-frequency performance of the new system is much better than the old system," offers O'Connell. This capability is especially critical for characterizing PHEMTs which typically have very high gain at low frequencies, supporting accurate measurements across the full range of a transistor. Load-pull tuners are used in the system for proper suppression of low-frequency oscillations, which is a potential for many high-gain devices. Finally, O'Connell says, "the new system provides targeted, user-friendly test results, rather than the overwhelming pile of data from the old system."

The new noise-parameter test system was used to evaluate an NE329 PHEMT, which is among the lowest-noise-figure transistors ever developed by California Eastern Laboratories' Asian partner, NEC. With the help of a low-noise test fixture from Inter-Continental Microwave (Santa Clara, CA), the test system was able to measure a noise figure of 0.35 dB at 12 GHz. ●●

cable-television (CATV) systems and signal/noise (E_S/N_O) measurements for) data-over-cable systems, the generator provides white Gaussian noise with a flatness of typically ±0.1 dB in a 10-MHz bandwidth. The instrument, which covers 5 to 1000 MHz with 0-dBm (+49 dBmV) maximum noise output power, includes 127.9-dB attenuation with 0.1-dB attenuation control for precise ratio measurements. The noise generator can be controlled remotely with GPIB. The 75-Ω instrument is equipped with BNC female connectors.

4. The models 57318 and 57518 peak power sensors work with the model 4530 power meter to provide modulation bandwidths of 17 and 6 MHz, respectively. (Photograph courtesy of Boonton Electronics Corp., Parsippany, NJ.)

In preparation for the shift from analog-to-digital broadcasting, Bird Electronic Corp. (Cleveland, OH) has developed the BPM series of broadcast power monitors for use from 54 to 870 MHz (depending on the choice of sensor). These meters and average-reading sensors provide minimum headroom of 10 dB to accommodate the high peak-to-average ratios (crest factors) of digitally modulated signals. The monitors can measure power, VSWR, return loss, and calculate impedance-match efficiency, with ±5-percent full-scale accuracy for power and ±10-percent full-scale accuracy for VSWR.

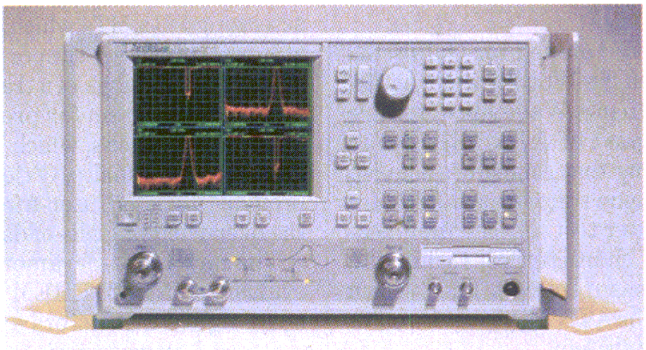
SINGLE SOLUTIONS

As wireless and other high-frequency measurements have evolved, instrument manufacturers have modified their basic offerings to include as many functions as practical for a complete set of measurements. As a result of this, the capabilities and bandwidths of individual instruments have increased drastically in recent years. For example, the MS462x series of vector network analyzers (VNAs) from Anritsu Co. (Morgan Hill, CA) targets the wireless industry's need for a single instrument with multifunction test capability. These are not only measurement receivers and dedicated signal sources, but are actually the equivalent of a large rack of test equipment in a single enclosure. The instruments can measure S-parameters, intermodulation distortion (IMD), harmonics, gain compression, and noise figure at frequencies from 10 MHz to 6 GHz. They are also fast, with measurement speeds to 150 μs/point and as much as 125-dB dynamic range. (For more on accelerated measurements, see sidebar.) They replace up to five separate instruments—a pair of frequency synthesizers, a vector analyzer, noise-figure meter, and a spectrum analyzer.

The company has also reduced the number of instruments that were once needed to cover a bandwidth as wide as 65 GHz. The firm recently announced the expansion of the company's Lightning line of VNAs to include a single system capable of coaxial coverage to 65 GHz. Using the firm's coaxial V connector, the model 37X97 VNA can perform continuous sweeps from 40 MHz to 65 GHz with more than +5-dBm leveled output power and greater than 70-dB dynamic range. The VNA system (Fig. 3) includes a four-sampler architecture and four-channel display for viewing all four S-parameters simultaneously. It can be equipped with doubler or quadrupler modules with GaAs monolithic-microwave-integrated-circuit (MMIC) output amplifier stages for active-device testing.

At even higher frequencies, the HP 8510XF VNA system from Hewlett-Packard Co. offers single-connection coverage to 110 GHz with the company's proprietary 1-mm coaxial connector. Ideal for wideband device characterization, the system operates with high-frequency probes from Cascade Microtech (Portland, OR) for making millimeter-wave S-parameter measurements on MMICs and discrete devices.

With 2.92- or 1.85-mm connectors for applications to 40 and 60 GHz, respectively, the probes include resonant-free power bypass connections for testing mixed-signal devices. Usable for signal rates to 80 Gb/s, the probes offer better than 40-dB crosstalk at 40 GHz.



3. This self-contained model 37X97 VNA system provides continuous sweeps from 40 MHz to 65 GHz using the coaxial V test connector. (Photograph courtesy of Anritsu Co., Morgan Hill, CA.)

NO-HANDS TESTING

In addition to increasing the measurement capability of their oscilloscopes, Hewlett-Packard has also made scopes easier to use. For the ultimate in hands-free control, the firm's Infinium family of oscilloscopes is now available with an optional voice-control interface. The option enables operators to control the instrument through commands spoken into a collar-mounted microphone. With voice control, an

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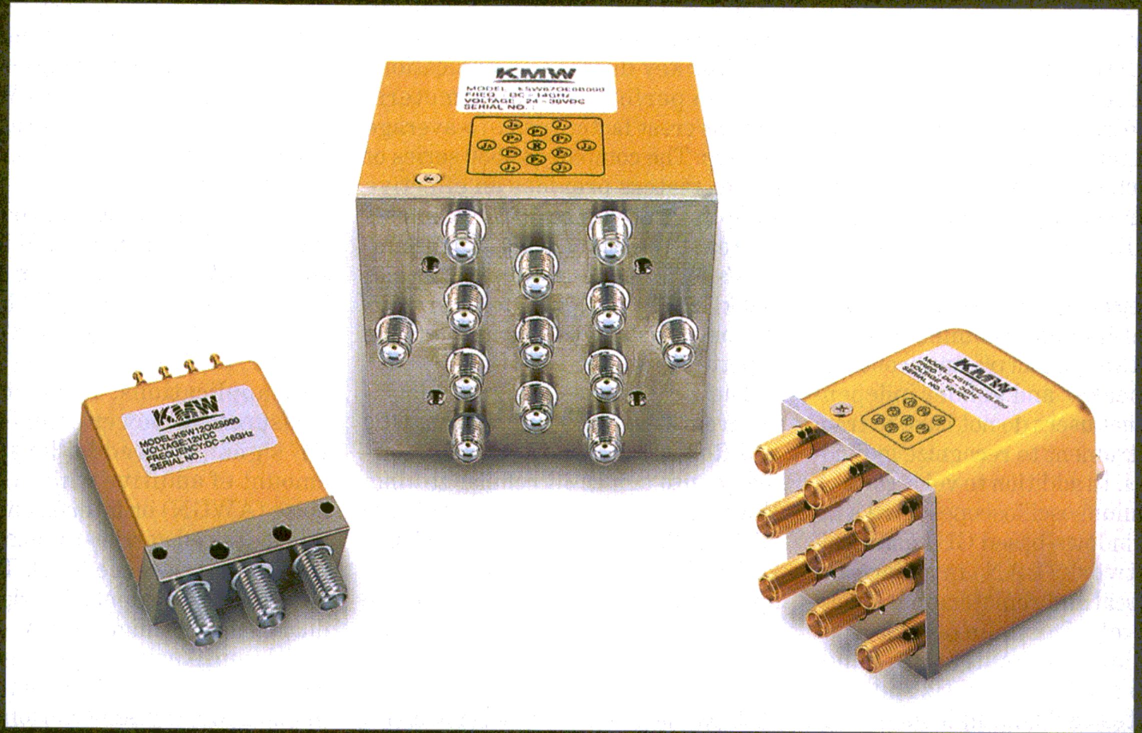

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VSWR (max.)	1.15 : 1	1.15 : 1	1.15 : 1
Isolation (min.)	80dB	80dB	80dB
Operating Mode	Latching	Latching	Latching
Actuating Voltage /Current (max.)	12Vdc ± 10% /240mA (@ 12Vdc, 25°C)	12Vdc ± 10% /165mA (@ 12Vdc, 25°C)	12Vdc ± 10% /165mA (@ 12Vdc, 25°C)
I/O Port Connector	SMA(F) / SMA(F)	SMA(F) / SMA(F)	SMA(F) / SMA(F)
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operator can manipulate waveform views onscreen, make measurements, and print results.

The voice control is expected to greatly simplify the probing of fine-pitch surface-mount circuit boards, which are easily damaged when a probe slips and shorts two pins together.

The voice-control option consists of a natural-language command set in English to adjust all primary controls, and is speaker and gender independent so that users do not have to train their instrument to recognize individual voices. In addition to voice control, the Infiniium oscilloscopes have a familiar Windows-based GUI and local-area-network (LAN) interface for ease of local and remote applications, respectively. Infiniium oscilloscopes are available currently with bandwidths from 0.5 to 1.5 GHz and sampling rates to 8 GSamples/s in two- and four-channel versions.

THIRD-GENERATION NEEDS

With the spread of cellular and personal-communications-services (PCS) communications, many test-equipment suppliers have turned an eye toward the needs of third-generation (3G) digital cellular systems. These systems use advanced digital-modulation formats, such as wideband code-division-multiple-access (WCDMA) techniques and wider channel bandwidths than earlier systems, requiring that test-equipment suppliers keep pace.

One who has is Boonton Electronics Corp. (Parsippany, NJ). The firm recently announced a pair of peak power sensors with the wide modulation bandwidth needed for WCDMA. The models 57318 and 57518 peak power sensors work with the company's model 4530 power meter to provide modulation bandwidths of 17 and 6 MHz, respectively (Fig. 4). Both sensors operate from 0.5 to 18 GHz with a 60-dB dynamic range. The model 57318 offers a typical rise time of better than 15 ns and triggers internally to +20 dBm. Model 57518 offers a typical rise time of less than 50 ns with the same internal triggering capabilities as the model 57318.

By measuring peak and average power levels, the sensors can gauge the percentage of power changes and monitor key digital communications system performance parameters, such as crest factor (peak-to-average power). The company's 4530 series of power meters provides an effective sampling rate up to 50 MSamples/s for repetitive signals and 2.5 MSamples/s for a single-shot bandwidth of 700 kHz. The power meters can trigger on signals with maximum pulse repetition rates of 1.8 MHz.

The FSIQ series of signal analyzers from Tektronix (Beaverton, OR) and Rohde & Schwarz (Munich, Germany) also aim at 3G cellular testing. With three models covering 20 Hz to 3.5 GHz, 20 Hz to 7 GHz, and 20 Hz to 26.5 GHz, the analyzers achieve respectable –152-dBc/Hz phase-noise offset 5 MHz from the carrier. The instruments achieve 75-dB dynamic range for making accurate 4.096-Mchips/s adjacent-channel-power-ratio (ACPR) measurements. Tektronix also offers a wideband real-time spectrum analyzer, model 3086, with a measurement bandwidth of 30 MHz for true 3G signal-analysis research. Unlike a conventional spectrum analyzer, which sweeps across a bandwidth in small segments, the 3086 can capture random bursts and transient signals across the full 30-MHz channel bandwidth instantaneously. It offers modulation analysis capability to 5.3 Msymbols/s.

A twist on spectrum analysis is the Internet-ready testing offered by Morrow Technologies Corp. (Largo, FL). The firm's P9116 spectrum-analyzer system can be controlled through Internet connections for remote measurements from 100 kHz to 1600 MHz. When located in the field, the analyzer acts similar to a network server to provide data access, storage, and display with a connected personal computer (PC). The analyzer itself is a fully synthesized instrument with 2-Hz resolution and better than 150-μs tuning speed. The level accuracy is ±0.5 dB from –120 to +20 dBm.

Emerging wireless standards have driven the development of new models for Hewlett-Packard's ESG line of RF/microwave signal generators (see *Microwaves & RF*, August 1999, p. 157). In addition to new low-phase-noise models, the signal generators can be equipped with personality options for testing according to specific wireless standards, such as WCDMA and cdma2000. The generators are available at frequencies from 0.25 to 4000 MHz.

Similarly, the frequency coverage of the TAS 4600AH noise and interference emulator from Telecom Analysis Systems (Eatontown, NJ) has been extended to 2.5 GHz for coverage of cellular, PCS, and emerging wireless data applications. The instrument measures the power of an incoming RF signal, then adds the specified amount of additive white Gaussian noise (AWGN) or interference to obtain a desired carrier-to-noise or carrier-to-interference level. The TAS 4500AH can be supplied in signal- and dual-channel configurations.

Telecom Analysis Systems was recently selected by the Cellular Telephone Industry Association (CTIA) as the vendor of choice for CDMA mobile-handset evaluation. The firm's CDMA-ATS automatic mobile-phone test system is the CTIA's specified mobile-telephone test bench for the CTIA certification program, which is designed to ensure user satisfaction with carrier networks. The CDMA-ATS system combines the company's TAS 4500 RF channel emulator (which emulates RF propagation conditions) as well as TAS 4600 noise and interference emulator (which emulates carrier-to-noise and carrier-to-interference conditions) with the firm's TASKIT/CDMA for Windows software.

When electric-field levels are a concern, the model 2001 RAHAM radiation-hazard measuring instrument from General Microwave (Amityville, NY) can check a dynamic range of 10 to 1000 V/m (equivalent power-density levels of 0.0275 to 265 mW/cm^2). With a frequency range of 3 MHz to 1 GHz, the instrument features a three-element diode detector which provides an isotropic (omnidirectional) response. The alarm-level threshold in standard units is set at the factory to 200 V/m (which is a density of 10.6 mW/cm^2), but can be factory reprogrammed to meet specific customer needs. Visual- and audio-alarm indicators are activated when the alarm threshold is exceeded. ••

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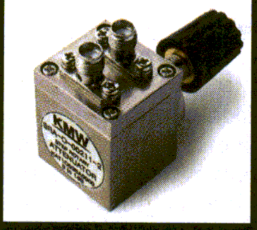

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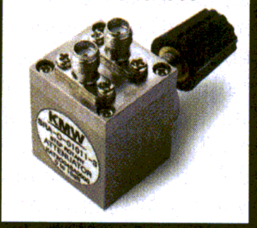

KAT13O4CA000

KAT13O4CA001

■ Step-Rotary Attenuators

Product Code No.	KAT1O04SA000	KAT2O04SA000	KAT1SO4SA000	KAT2SO4SA000
Operating Type	Break-Before-Make		Make-Before-Break	
Frequency Range	DC ~ 3GHz	DC ~ 3GHz	DC ~ 3GHz	DC ~ 3GHz
Insertion Loss (max.)	0.2dB	0.2dB	0.2dB	0.2dB
VSWR (max.)	1.15:1	1.15:1	1.15:1	1.15:1
Incremental Attenuation Range (dB)	0 ~ 1	0 ~ 10	0 ~ 1	0 ~ 10
Attenuation Step (dB)	0.2	1	0.2	1
Nominal Impedance	50 ohm		50 ohm	
I/O Port Connector	SMA(F) / SMA(F)		SMA(F) / SMA(F)	
Average Power Handling	2W @ 2GHz		2W @ 2GHz	
Temperature Range	-55°C ~ +85°C		-55°C ~ +85°C	
Dimension (inch)	1.93*1.56*1.51		1.93*1.56*1.51	

■ Continuously Variable Attenuators

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Product Code No.	A type : KAT13O4CA000 B type : KAT13O4CA001		
Frequency Range	DC ~ 1GHz	1 ~ 2GHz	2 ~ 3GHz
Insertion Loss (max.)	0.15dB	0.3dB	0.35dB
VSWR (max.)	1.25 : 1	1.25 : 1	1.25 : 1
Attenuation Range (max.)	4dB @ 1GHz	13dB @ 2GHz	25dB @ 3GHz
Nominal Impedance	50ohm		
I/O Port Connector	SMA(F) / SMA(F)		
Average Power Handling	2W @ 2GHz & 25°C, without Heat-Sink		
Temperature Range	-55°C ~ +85°C		
Dimension (inch)	A type : 1.496*1.102*0.470, B type : 1.225*1.102*0.470		

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Wide-Range Sensor Gauges Power Of Complex Signals

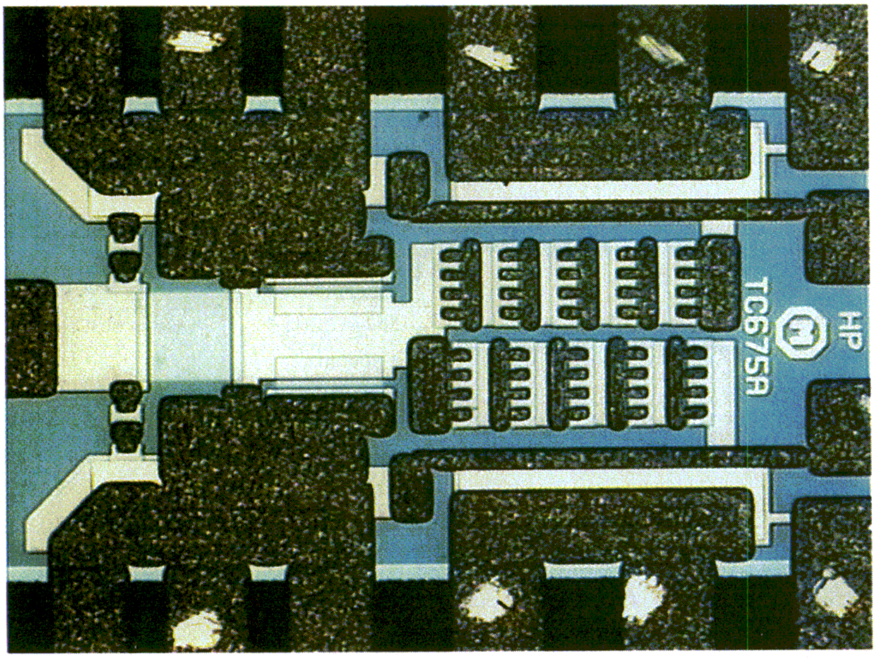
This average-reading power sensor provides the dynamic range needed for handling the latest digital modulation formats.

Ron Hogan

Development Engineer

Hewlett-Packard Co., Microwave Technology Div., Santa Rosa, CA 95409; (707) 577-1400, Internet: http://www.hp.com.

POWER sensors have traditionally relied on thermocouples for high-power (–30- to +20-dBm) measurements or diodes for low-power (–70- to –20-dBm) measurements. Both approaches offer a 50-dB dynamic range for a single sensor making average power measurements, even though many modern complex modulation formats require an even greater dynamic range. However, a new diode sensor design, based on a combination of diodes and attenuators, provides the accuracy associated with diodes operating in their "square-law" region while also achieving the wide dynamic range needed for measuring the power of complex modern modulation formats.



1. The modified-barrier-integrated-diode (MBID) sensor uses a diode pair/attenuator/diode pair topology to make wide-dynamic-range measurements from +20 to –60 dB.

Traditional diode sensors operate at power levels from –70 to –20 dBm, which is commonly referred to as the diode square-law region. In the square-law region, a diode's detected output voltage is proportional to the (logarithm of the) input power and so measures power directly. A common approach to extend the dynamic range of diode power sensors above their square law region has been the use of correction factors. The correction factors are derived from measurements of a continuous-wave (CW) source and stored in each sensor's electronically erasable programmable read-only memory (EEPROM). While this approach is fine for measuring CW signals over a wide dynamic range, it fails to provide an accurate average power reading for modulated signals, such as code-division-multiple-access (CDMA) signals, when the signal level is above the diode power sensor's square-law region.

NOVEL NEW DESIGN

The new wide-dynamic-range sensor approach is based on a dual-sensor, diode pair/attenuator/diode pair topology as proposed by Szente *et. al.* in 1990.[1] The topology has the advantage of always maintaining the sensing diodes within their square-law region and therefore will respond properly to complex modulation formats as long as the sensor's correct range is selected.[2]

This approach was further refined by incorporating diode stacks in place of single diodes to extend square-law operation to higher power levels at

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KPH90OSCL001

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VSWR (max.)	1.25 : 1	1.25 : 1	1.25 : 1
Incremental Phase Shift	90 degree min. @ 2GHz		
Electrical Delay	125 psec min.		
Nominal Impedance	50 ohm		
I/O Port Connector	SMA(F) / SMA(F)		
Average Power Handling	20W @ 2GHz		
Temperature Range	-30°C ~ +60°C		
Dimension (inch)	A type : 1.496*1.102*0.470, B type : 1.225*1.102*0.470		

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the expense of sensitivity, as proposed by Ehlers *et al.*[3] As described in the reference, a series connection of m diodes results in a sensitivity degradation of 10log(m) dB and an extension upward in power of the square-law-region maximum power of 20log(m) dB, yielding a net improvement in square-law dynamic range of 10log(m) dB compared to a single diode detector.

The approach has been implemented in the HP E-series E9300 power sensors from Hewlett-Packard Co. (Palo Alto, CA), as a modified barrier integrated diode (MBID)[4] with a two-diode-stack pair for the power-power path from –60 to –10 dBm and a resistive divider attenuator and five-diode-stack pair for the high-power path from –10 to +20 dBm (Figs. 1 and 2). Additionally, series field-effect-transistor (FET) switches were used off-chip to enable the low-path diodes to self-bias to an off condition when not in use.

Figure 3 shows the typical deviation from square-law behavior for the low-power path (–60 to –10 dBm) and the high-power path (–10 to +20 dBm) for the new HP E-series E9300 power sensors. The linearity at the –10 dBm switching point is specified as being typically ≤ ±0.02 dB.

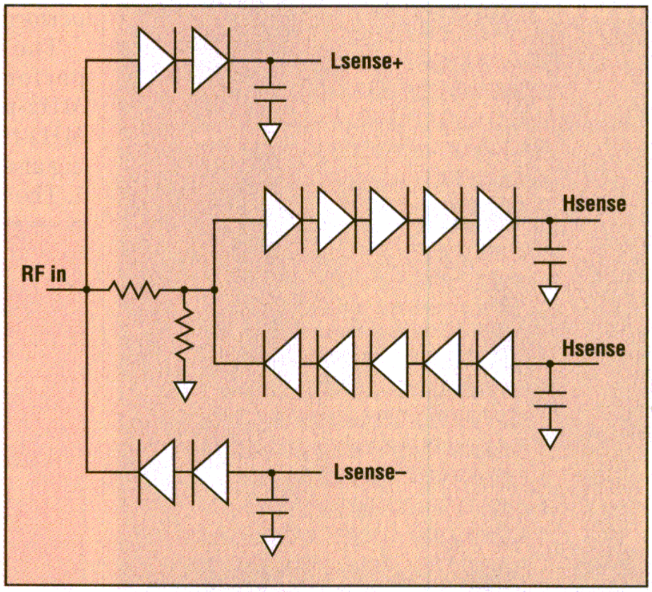
Table 1: Checking the linearity of the HP E93000 sensors

Power	Linearity (25 ±10°C)	Linearity (0 to 55°C)
–60 to –10 dBm	±3.0 percent	±3.5 percent
–10 to 0 dBm	±2.5 percent	±3.0 percent
0 to +20 dBm	±2.0 percent	±2.5 percent

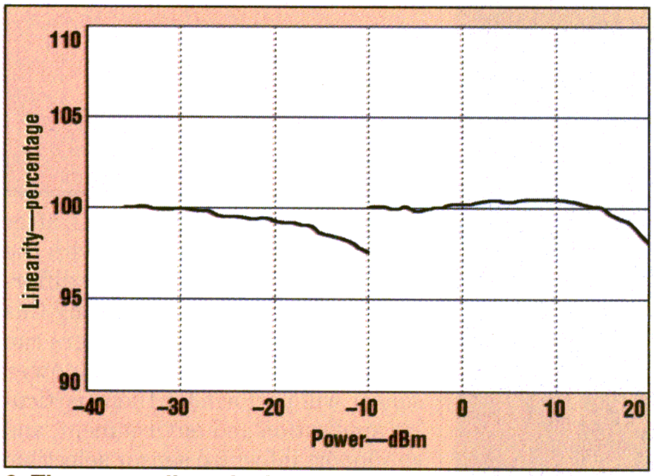
Switching between the low-power and high-power paths is seamless, providing a transparent 80-dB dynamic range.

The decision for switching between the low- and high-power paths is made on the basis of the average power detected by the power meter. For example, a signal with –12-dBm average power and a high crest factor (peak-to-average ratio) would be measured through the sensor's low-power path. For most CDMA signals, such as those meeting the IS-95A standard, the sensor will give better than ±0.05-dB accuracy for power levels to –10-dBm average power using the low-power path. However, for some CDMA signals used in third-generation (3G) digital cellular systems, with aligned symbols that represent an almost 20-dB maximum peak-to-average ratio, the accuracy of the measurement during the high-power crests would be compromised. This is due to the sensor being in the low-power path and the diodes being operated well outside the diode square-law region of the low-power path during the crests. Fortunately, a "Range Hold" function in the power meter for the high-power path deals with this situation as it enables the peak and average powers to be measured in the square-law region of the high-power path.

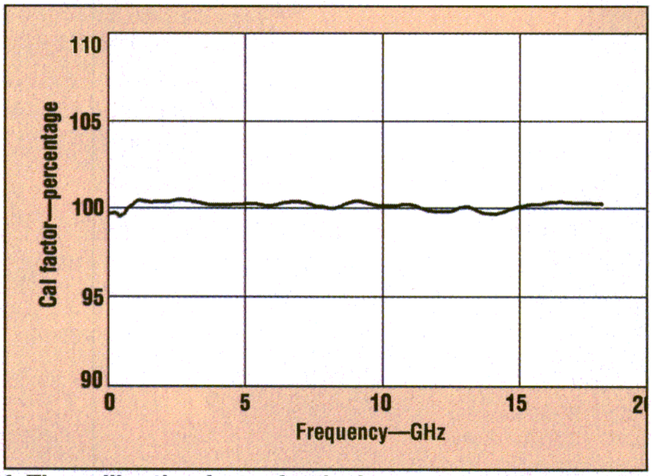
To avoid unnecessary switching when the power level is near –10 dBm,



2. The MBID sensor consists of a two-diode-stack pair for the low-power path and a resistive divider attenuator and five-diode-stack pair for the high-power path.



3. The sensor linearity is a result of separating a wide dynamic range into two measurement power levels.



4. The calibration factor for the low-power path was determined for a wide range of operating frequencies.

Table 2: Reviewing the measurement uncertainty of the HP E93000 sensors

Power range	–30 to –20 dBm	–20 to –10 dBm	–10 to 0 dBm	0 to +10 dBm	+10 to +20 dBm
Measurement uncertainty	±0.90 percent	±0.80 percent	±0.65 percent	±0.55 percent	±0.45 percent

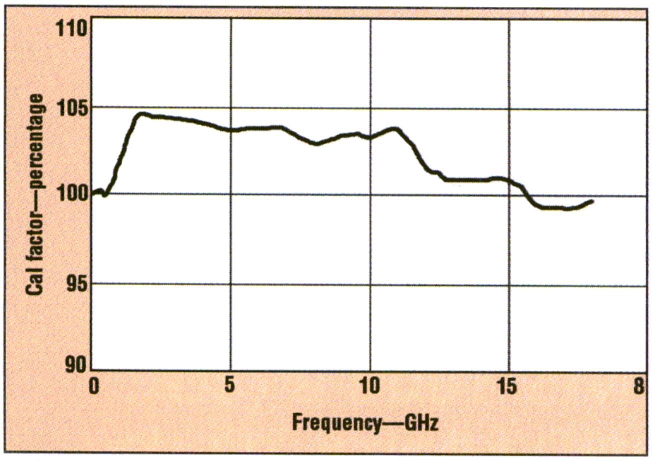
switching-point hysteresis has been added. This hysteresis causes the low-power path to remain selected as signal power levels increase until approximately –9.5 dBm. Above this power level, the high-power path is selected. The high-power path remains selected until approximately –10.5 dBm is reached for decreasing power levels. Below this level, the low-power path is selected.

To provide sensor information that is relevant to the numerous power-measurement scenarios and meaningful to a user, from a temperature-controlled manufacturing or research and development (R&D) environment to field installation and maintenance applications, warranted specifications for HP E9300 sensors are provided over the temperature ranges of +25 ±10°C and 0 to +55°C. In addition, supplemental information is provided at +25°C to illustrate the typical performance.

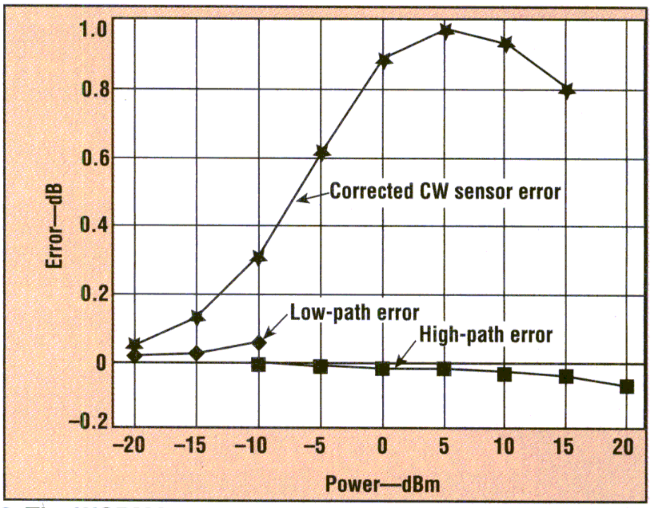
The warranted power-linearity performance for the E9300A (10 MHz-to-18-GHz) and E9301A (10-MHz-to-6-GHz) sensors, after zeroing and calibration, is shown in Table 1. The table shows the typical power linearity performance at +25°C, after a zero and calibration, along with the measurement uncertainty for different power ranges. By providing this type of specification data, users can better understand how the power sensor will perform for their particular application and environment, and so support informed sensor selection.

THE FLAT CALIBRATION FACTOR VERSUS FREQUENCY RESPONSE MAKES THE SENSOR IDEAL FOR AMPLIFIER INTERMODULATION DISTORTION (IMD) MEASUREMENTS, WHERE THE TWO-TONE OR MULTI-TONE TEST SIGNAL CAN BE SEPARATED BY HUNDREDS OF MHz.

The HP E9300 power sensors also have an exceptionally flat calibration



5. The calibration factor for the high-power path was determined for a wide range of operating frequencies.



6. The WCDMA measurement errors for the HP E9300 power sensor were determined relative to corrected CW measurements.

factor versus frequency response across the entire range, as shown in Figs. 4 and 5. This makes the sensor ideal for amplifier intermodulation-distortion (IMD) measurements, where the two-tone or multi-tone test signal can be separated by hundreds of MHz. All users have to do is select an appropriate, single calibration factor frequency for the measurement.

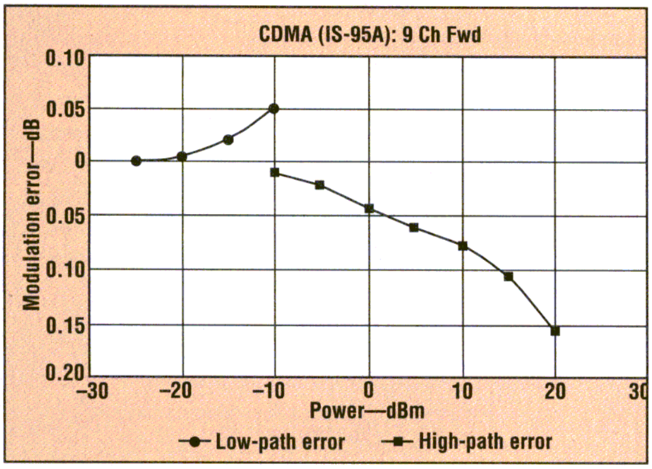
The error in measuring single-channel wideband CDMA (WCDMA) was characterized at 50 MHz and is compared to the error measured for a corrected CW diode sensor (Fig. 6). A negative WCDMA error means the sensor reads power levels lower than it should, and a positive error means the sensor reads higher than it should. The uncertainty in these numbers is estimated to be on the order of approximately ±0.01 dB.

The method used to measure WCDMA error was to compare power-meter readings for CW versus WCDMA modulation with an HP ESG series signal generator at the level of interest. Attenuation was then added between the source and the sensor 10 dB at a time until the difference read between the CW and WCDMA measurements stopped changing. This is the "actual" difference between the two assuming the ESG source sees a good 50-Ω load during this procedure (which was not measured or included in the numbers above). The difference between CW and WCDMA at the power level of interest minus the difference between CW and WCDMA at the lower level is referred to as the WCDMA error.

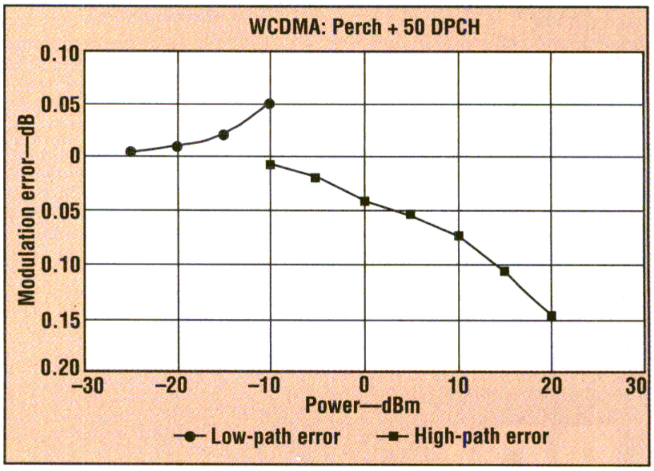
The advantages of the technique are:

1. This isolates only the WCDMA component of the error. The total error is the WCDMA error plus any other errors, for example, linearity, calibration factor, and errors in temperature correction.

2. The method is independent of source-leveling errors for WCDMA. This was verified using a 2-W amplifier to shift the source level down by approximately 15 dB for a particular sensor power level, to get the source-leveling loop to a more accurate re-



7. The new measurement technique and MBID sensor were used to test an IS-95A nine-channel CDMA signal with maximum crest factor of 10 dB.



8. The new measurement technique and MBID sensor were used to evaluate a WCDMA Perch + 50 DPCH signal with a maximum crest factor of approximately 11.5 dB.

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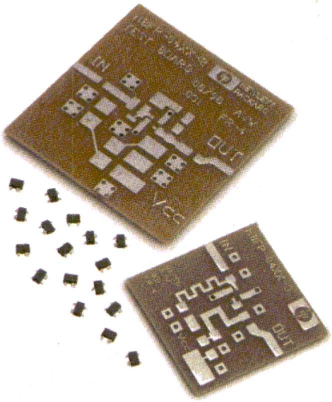
HBFP-0420	2V, 5mA	1.1	17	12 @ 2V, 20mA
HBFP-0450*	2V, 20mA	1.3	14	16 @ 3V, 50mA

*Coming spring 1999



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gion of operation. This resulted in virtually the same WCDMA sensor error numbers and also enabled high-power measurements to be made to +20 dBm. The peak-to-average ratio of the modulation had previously been compromised by source compression at this level.

3. The exact attenuation of the 10-dB step attenuator does not contribute significant error since the difference between modulation on versus modulation off is being examined.

4. Using the sensor in its square-law region to compare difference readings at the power level of interest eliminates transfer and connection repeatability errors (versus comparison to a thermoelectric or another attenuated diode sensor).

RF AND MICROWAVE ENGINEERS INVOLVED IN DESIGN, MANUFACTURE, OR SERVICE CAN NOW SIMPLIFY THEIR MEASUREMENT EQUIPMENT REQUIREMENT WITH A SINGLE POWER SENSOR.

Additional measurements of CDMA signals using this technique and the new sensor are shown in Figs. 7 and 8 for an IS-95A nine-channel signal with approximately 10-dB maximum peak-to-average ratio and a WCDMA Perch + 50 DPCH signal with approximately 11.5-dB maximum peak-to-average ratio, respectively.

Used with the HP EPM series of power meters, the HP E9300 wide-dynamic-range average-reading power sensors have been shown to provide accurate power-level measurements for digitally modulated signals over a wider range of power levels than is currently achieved with traditional thermocouple or diode-based power sensors (Table 2). RF and microwave engineers involved in design, manufacture, or service can now simplify their measurement equipment requirement using a single HP E9300 power sensor to measure complex digital-modulation formats as well as multi-tone and CW signals. ••

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1. United States Patent No. 4943764, assigned to Hewlett-Packard Co.(Palo Alto, CA).
2. Hewlett-Packard Application Note 64-1A, pages 27 and 34.
3. Ehlers *et al.*, private communication to the author.
4. *Hewlett-Packard Journal*, November 1986, pp. 14-21.

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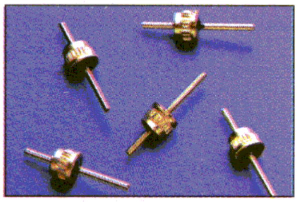

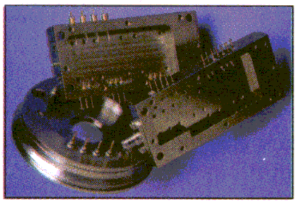

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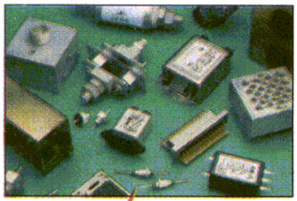

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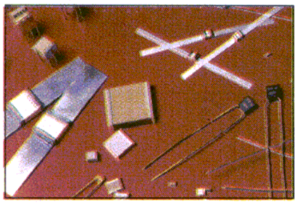

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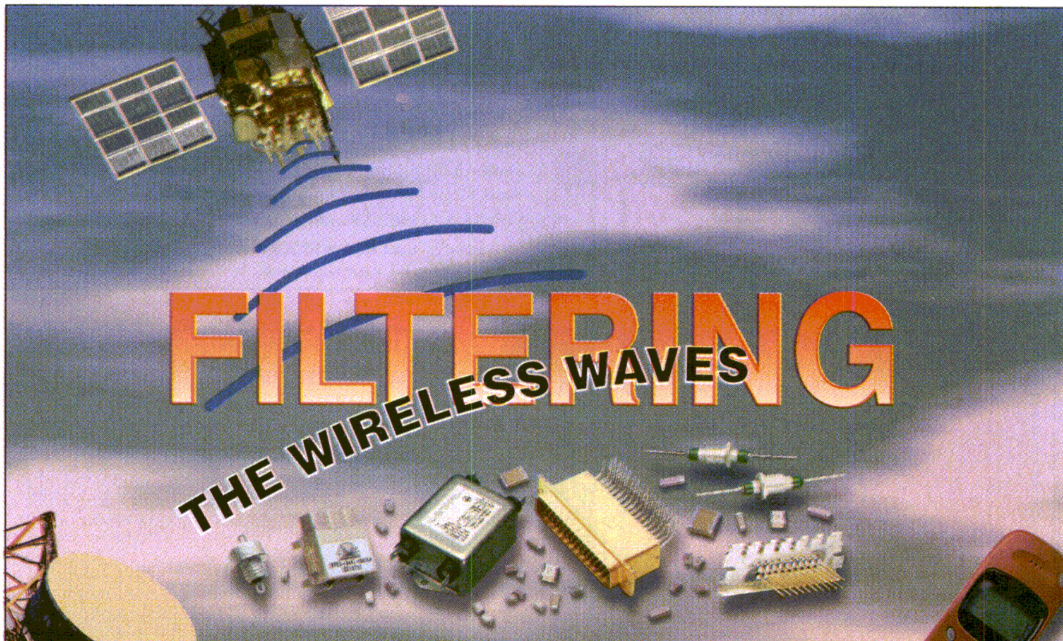

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CIRCLE NO. 268

Designing antennas for a low-cost UHF receiver

Wireless systems are composed of basic building blocks that include data encoders and decoders, baseband-to-RF electronics, RF-to-baseband electronics, and an antenna system. Availability of multifunction integrated circuits (IC), such as the MICRF001 ultra-high-frequency (UHF) receiver from Micrel, Inc. (San Jose, CA), helps simplify many portions of a wireless design, leaving only the antenna. An application note from Micrel, "MICRF001 Antenna Design Tutorial," helps specifiers of this IC with this final system detail.

The MICRF001 UHF receiver IC has been designed for fairly simply implementation with basic wire antennas. It can be directly connected to the antenna, without impedance-matching components. The application note discusses the various antenna options for the IC, with a comparison of three simple wire antennas—straight-wire monopole antennas, helical coil antennas, and loop antennas.

Written by Tom Yestrebsky, the application note reviews antenna characteristics, radiation patterns for aligned and misaligned monopole and dipole antennas, ground-plane effects, and tuning. Details are provided for constructing all three types of wire antennas for use with the MICRF001, with a comparison of the results with each type.

For readers wishing more in-depth treatment of antennas, the application note offers a brief bibliography of reference books covering general antenna design as well as specific topics, such as antenna impedance matching. Copies of the application note are available as part of the company's *QwikRadio RF Receiver/Demodulator Handbook*, from: **Micrel, Inc., 1849 Fortune Dr., San Jose, CA 95131; (408) 944-0800, FAX: (408) 944-0970, Internet: http://www.micrel.com.**

CIRCLE NO. 194 or visit www.mwrf.com

Questioning the price of bandwidth efficiency

Bandwidth efficiency is the basis for all new cellular-communications designs. But designing for high bandwidth efficiency can be too expensive for some applications, notes Andrew Bateman of Wireless Systems International Ltd. (http://www.wsil.com) in the July 1999 edition of the *LPRA News*. At issue is the recent availability of limited bandwidth at 868 MHz in the UK, ostensibly for low-power telemetry applications. Designers of low-cost systems argue that wide channel spacing is necessary to support the loose tolerances of inexpensive components, while proponents of spectral efficiency say that the little remaining bandwidth must be used sparingly.

Bateman points to the use of digital filters as a low-cost solution for spectrally efficient, low-power wireless telemetry systems at 868 MHz. He provides a table with the components and their specifications for implementing the necessary digital filters for telemetry, and feels that digital filters can lead to efficient use of bandwidth for low-cost wireless modems. Copies of the *LPRA News* are available from the **Low Power Radio Association, Brearley Hall, Luddenden Foot, Halifax HX2 6HS, United Kingdom; (44) 0-1422-886463, FAX: (44) 0-1422-886950, e-mail: info@lpra.org, Internet: http://www.lpra.org.**

CIRCLE NO. 195 or visit www.mwrf.com

Applying wavelet filters to real-time applications

Wavelets represent a powerful method of building selective filters in digital form. In digital-signal-processing (DSP) applications, wavelets are already used for data compression, active equalization, and in transmultiplexers. Wavelets enable various levels of resolution to be implemented, and can be used for multirate signal analysis in complex filter banks. An article in the Summer 1999 edition of *MATLAB News & Notes* from The MathWorks, Inc. (Natick, MA), "Wavelets in Real-Time Applications: The Design of Multirate DSP Systems," provides some insights into the use of wavelet filters in real-time signal-processing systems, using the MATLAB software.

Several forms of multirate systems are presented, including the use of a sample-by-sample dead-zone transfer function where all coefficients with values less than a specified level are removed (for efficient filtering of noise). Several different frame-based implementations are also highlighted. To learn more about wavelets and how they can be implemented in math-based software, ask for a copy of the Summer 1999 *MATLAB News & Notes.* **The MathWorks, Inc., 3 Apple Hill Dr., Natick, MA 01760-2098; (508) 647-7000, Internet: http://www.mathworks.com.**

CIRCLE NO. 196 or visit www.mwrf.com

Revolutionary RF IC Performs RMS-to-DC Conversion

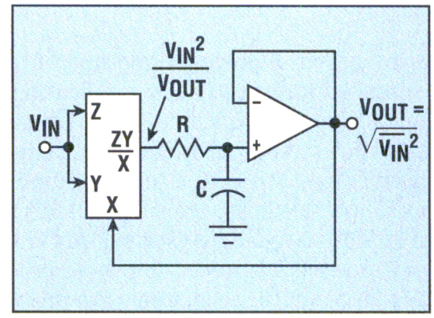
This innovative power-measurement circuit is designed for accuracy evaluation of complex-modulation signals with high crest factors.

Eamon Nash
Applications Engineer
John Greichen
Product Marketing Manager, RF/IF Products
Analog Devices, Inc., 804 Woburn St., Wilmington, MA 01887; (781) 937-1292, FAX: (781) 937-1024, Internet: http://www.analog.com.

POWER measurements serve almost all communications-system designs. Receiver dynamic range is influenced by detection accuracy, while transmitters rely on accurate level detection to ensure compliance with regulatory emissions limits, control loop stability, and to protect output devices. Until recently, power detection had been accomplished either by discrete diode detectors or slower-response thermal conversion for true root-mean-square (RMS) power detection. But a better way is available by using a patent-pending silicon (Si) integrated circuit (IC) from Analog Devices, Inc. (Wilmington, MA)—the AD8361. Part of the company's new line of True Power Detection™ products, the IC supports true RMS power measurements on complex waveforms, such as code-division-multiple-access (CDMA) signals, through 2.5 GHz.

Recent innovations in integrated logarithmic detectors have overcome many of the limitations of diode detectors (see "Approach Performs RF Detection Directly," *Microwaves & RF*, September 1998, p. 84). Logarithmic amplifiers solve the power-detection problem for waveforms which have a fairly constant peak-to-average ratio such as quadrature-phase-shift-keying (QPSK) and Gaussian minimum-shift-keying (GMSK) signals and are unique in their ability to measure signal strength over dynamic ranges of up to 100 dB.

Yet with changing communications standards, RF circuit and system designers face many challenges, including the need to support multiple standards and frequency bands, the capability to handle higher-order modulation schemes, the flexibility to detect signals with multiple modulation formats (especially in test equipment), and the capability to support emerging modulation formats. The newer modulation formats, such as CDMA, and the desire to measure multiple modulated waveforms, reveal the limitation of current detection approaches. CDMA signals exhibit peak-to-average signal levels (crest factors) of 14 dB and higher, pushing them well beyond the constant-envelope capability of existing power-detection schemes.



1. The implicit RMS computation method uses feedback to perform the square-root function indirectly at the input. The average signal levels are divided by the average of the output and vary linearly with the RMS level of the input.

The True Power Detection product line was developed to overcome the limitations of existing power-detection schemes. The ICs address the difficulty of measuring the signal levels of spread-spectrum CDMA signals and other digital modulation formats. The first RMS-to-DC converter ICs were patented by Barrie Gilbert of Analog Devices

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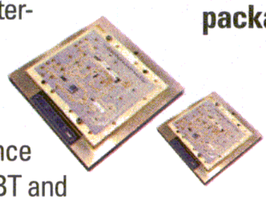

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CIRCLE NO. 217

more than 20 years ago. While accurate for a variety of waveforms, these devices were limited to frequencies up to approximately 10 MHz. The devices are based on "implicit RMS computation," where feedback is used to perform the square-root computation function indirectly at the input of the power-detection circuit (Fig. 1). In this circuit, average signal levels are divided by the average of the output and vary linearly (instead of as the square) with the input RMS level.

The implicit computation method has advantages of lower complexity and greater dynamic range than the explicit RMS method. A critical performance parameter of RMS-to-DC converters is accuracy as a function of an input-signal crest factor. A waveform's crest factor refers to the ratio of its peak value to its RMS value. Simple sine-wave signals have a crest factor of 1.41 (or $\sqrt{2}$). Single-channel (i.e., reverse-link) CDMA signals have a crest factor of approximately 2, while multichannel, forward-link CDMA signals, with behavior similar to pseudorandom noise, can have a crest factor up to 6.

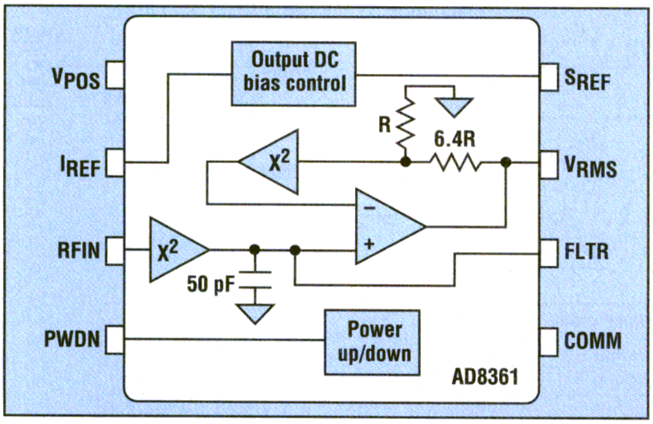
Barrie Gilbert has built a rich career innovation by innovation, and his latest development is an RMS-to-DC converter that overcomes the bandwidth limitations of his earlier designs. The model AD8361 IC is the first member of the resulting True Power Detection™ family of RMS-to-DC products. The low-power IC is designed for use in high-frequency receiver and transmitter chains through 2.5 GHz (Fig. 2). The device is simple to apply, requiring only a single supply between +2.7 VDC and +5.5 VDC, a power-supply decoupling capacitor, and input coupling capacitors for many applications. The AD8361 provides a linear-responding DC signal at its output port, with conversion gain of 7.4 V/V_{RMS}.

Table 1: The AD8361's specifications at a glance

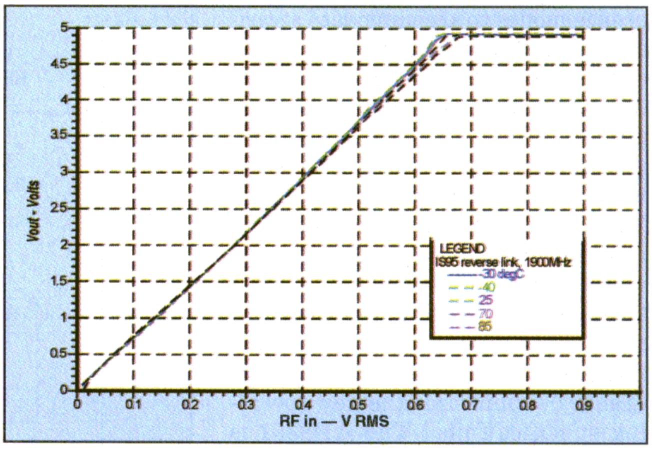
Parameter	Condition	Specification
Frequency range		0.1 to 2.5 GHz
Maximum input voltage	V_s = +3 VDC V_s = +5 VDC	390-mV RMS 660-mV RMS
Input impedance	900 MHz	200/1 Ω/pF
Temperature range		–30 to +85°C
Positive supply voltage		+2.7 to +5.5 VDC
Output offset	0 at S_{REF}, V_s at I_{REF} 0 at S_{REF}, I_{REF} open V_s at S_{REF}, V_s at I_{REF}	0 V 345 mV 400 mV
Current drain (enabled)	V_{RFIN} = 200 mV	2 mA
Current drain (standby)		1 µA

Table 2: The AD8361's conversion specifications at a glance

Parameter	Condition	Specification
Conversion gain		7.4 V/V RMS
0.25-dB error dynamic range	CW input	14 dB
1-dB error dynamic range	CW input	23 dB
2-dB error dynamic range	CW input, V_{pos} = +5 VDC	30 dB
Deviation from CW	For a crest factor of 5.5 dB (IS-95 reverse link) For a crest factor of 12 dB (four-channel WCDMA)	0.2 dB 1.0 dB



2. The AD8361 performs an implicit root-mean-square (RMS) computation, with the filtered output of a squaring circuit being applied to a square-rooting circuit with a conversion gain of 7.4.



3. This plot shows that the 1900-MHz CDMA (IS-95 reverse link) response of the AD8461 is stable and linear over the standard industrial temperature range.

The AD8361 performs explicit RMS computations. In operation, RF input signals are applied to a wideband squaring circuit. An onboard 50-pF capacitor filters the output of the squaring circuit. Additional filtering can be supplied by connecting an external capacitor to the device's FLTR pin. The filtered output of the squaring circuit is applied to a square-root circuit, which consists of an operational-amplifier-type circuit with a second squaring circuit in its feedback loop. This results in the inverse function (i.e., the square root) in the forward direction. In addition, a resistive attenuator (1/6.4) in the feedback loop adds a gain of 7.4 in the forward direction. As a result, the AD8361's overall transfer function can be given by:

$$V_{OUT} = 7.4\ V_{IN(RMS)}$$

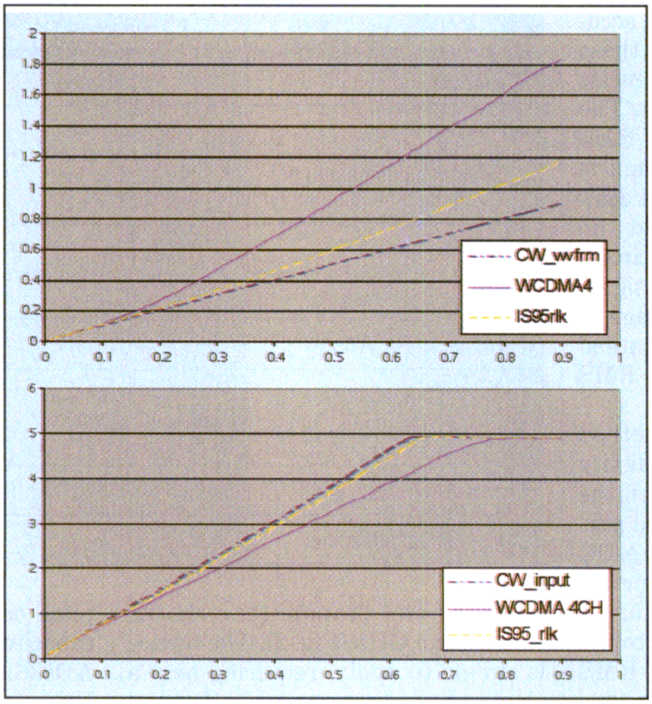
The AD8361 is intended for true power measurement of simple and complex waveforms. The device is particularly useful for measuring high-crest-factor signals, such as CDMA and wideband-CDMA (WCDMA) signals. It is also accurate for measurement of most other complex modulation waveforms. This first product offers up to 30-dB dynamic range.

The AD8361 has three output reference modes to accommodate a variety of analog-to-digital-converter (ADC) requirements—ground referenced; internally referenced, with the output offset by 345 mV above ground; and V_{POS} as the reference, with the output offset to V_{POS}/7.4. A single external filter capacitor modifies the response time of the device. The detector consumes 2-mA typical current during normal operation, while a power-down function reduces current drain to only 1 µA. (Table 1 summarizes the key operating specifications.) The AD8361 is specified for operation from –30 to +85°C and is available in an 8-pin micro-SO package. It is fabricated on a proprietary Si bipolar process.

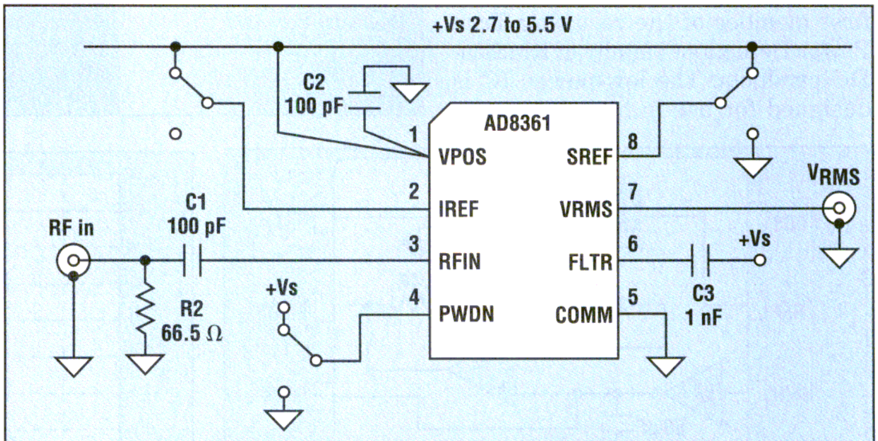
An important benefit of the RMS approach is its accuracy for different modulation waveforms. Table 2 provides conversion specifications for continuous-wave (CW) sine-wave and high-crest-factor (CDMA/WCDMA) waveforms. As the table shows, the response of the AD8361 varies only 0.2 dB for IS-95 reverse-link (single-channel) conditions and 1 dB for IS-95 forward-link (multi-channel) conditions, relative to the CW sine-wave response. Very-accurate (0.25-dB) conversion is possible over a 14-dB dynamic range. Reasonable accuracy is possible to a dynamic range of 30 dB with a +5-VDC supply.

The AD8361 offers outstanding temperature stability, minimizing the need for calibration during manufacturing or normal operation and improving system performance. Figure 3 illustrates the device's highly stable 1900-MHz CDMA (IS-95 reverse-link) response over a standard industrial temperature range.

The value of the True Power Detection™ approach is clear when compared with the current alternatives (Table 3). Note that the diode detector, a model 51015 from Boonton Electronics (Parsippany, NJ), and the thermal bridge, a model 51101 from Boonton Electronics, are instrument modules costing



4. The accuracy of a traditional diode detector was compared with a thermal bridge (top) for CW sine wave, CDMA (IS-95), and WCDMA signals. The AD8361 True Power Detection™ IC was then compared with the thermal bridge (bottom), with favorable results.



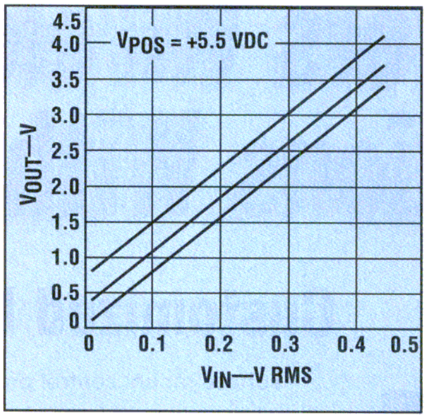
Reference mode	I_{REF}	S_{REF}	V_{RMS} range
Ground	VPOS	COMM	7.4-V RMS
Internal	OPEN	COMM	7.4-V RMS +0.345 V
Supply	VPOS	VPOS	7.4-V RMS + VPOS/7.4

5. The AD8361 requires a single supply voltage between +2.7 and +5.5 VDC. It consumes only 2-mA supply current in normal operation. The voltages on pins I_{REF} and S_{REF} set the output offset voltage.

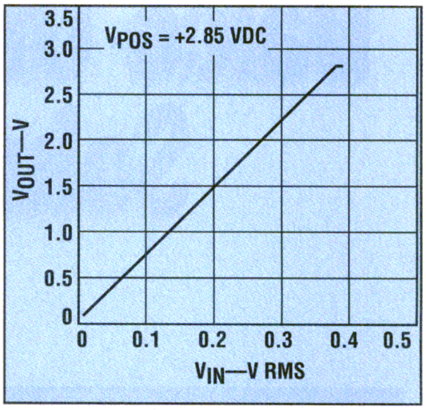
approximately $1000. The logamp and True Power Detection™ ICs are less than $10 each in large volumes. In this comparison, the thermal bridge was treated as the reference, since it can measure true power and offers waveform-independent operation. The accuracy of the AD8361 compares favorably with that of the thermal bridge, with the AD8361 enjoying technical and cost advantages over the thermal bridge. The logamp IC and diode response is repeatable despite the effective offset due to crest factor. The poor temperature stability of the practical diode circuits (which are not $1000 instruments) is a major disadvantage. Further comparisons show the differences between a thermal bridge and diode detector for sine-wave, single-channel CDMA signals; WCDMA signals (Fig. 4a); and for the new true-RMS approach versus the thermal bridge (Fig. 4b) for the same waveforms.

The operation of the AD8361 is very straightforward, with some basic connections (Fig. 5). Supply voltage, in the range of +2.7 to +5.5 VDC, is applied to the V_{POS} pin which is decoupled by a single 100-pF capacitor. Supply current of 2 mA in the operating mode can be reduced to 1 μA by pulling the PWDN pin to a voltage of V_{POS}. The RF input, which is AC coupled, has an input impedance of 200 Ω. A 66-Ω external shunt impedance combines with the AC-coupled input to provide an input impedance of 50 Ω.

The output voltage is nominally 7.4 times the input RMS voltage. The three different modes of operation are set by the pins S_{REF} and I_{REF}. In addition to the ground-referenced mode (where $I_{REF} = V_{POS}$ and S_{REF} = GND), where the output voltage swings from approximately 100 mV to 3.4 V on a +5.5-VDC supply (Fig. 3), two additional modes enable an offset voltage to be added to the output. This is useful in applications where the AD8361's output is being digitized by an ADC, where the input voltage range is not ground referenced (a typical input range on these devices is 0.5 to +4.5 VDC). In the first of these modes, the output voltage swing can be shifted upward by an internal reference voltage of 345 mV. Alternatively, a supply-referenced offset voltage, where the effective offset voltage is equal to $V_{POS}/7.4$, can be added to the output voltage. The output swings for all three of these modes were evaluated for a sine-wave input signal (Fig. 6). The response for lower supply voltages will look similar (in the supply referenced mode, the offset will obviously be smaller), but the dynamic range will be reduced as headroom decreases (Fig. 7).



6. The output of the AD8361 swings from 100 mV to 3.4 V when using a +5.5-VDC supply.

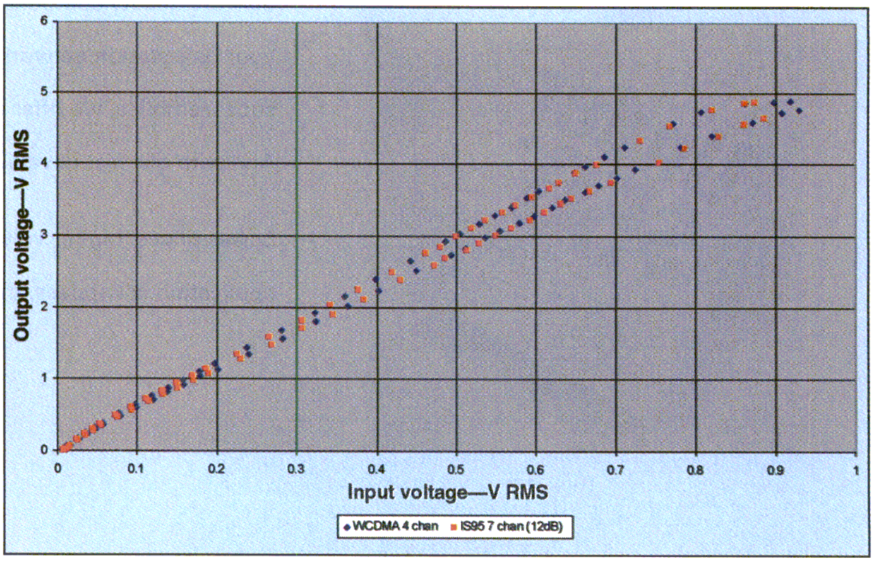


7. The response of the AD8361 for a supply voltage of 2.85 V is limited by the output-voltage headroom.

A key quality criterion of an RMS measurement circuit is its response to different signal types, such as CDMA and WCDMA signals at cellular (900 MHz) and personal-communications-services (PCS) [1800-MHz] frequencies (Fig. 8). While there is some shift in response as a function of frequency, measurements show that the overall response is fairly independent of modulation scheme.

Obvious uses of True Power Detection™ products include terminal and base-station transmit power control, overload protection, receive-signal-strength-indicator (RSSI) measurements, receiver signal leveling, and gain monitoring. Creative radio designers will undoubtedly find numerous other uses.

The AD8361 finds use in any application where power measurement is required, especially in applications where the peak-to-average ratio of the signal is continually changing. In



8. The AD8361 IC can effectively measure signals with high crest factors, such as pseudorandom-noise-like CDMA (IS-95) and WCDMA signals at cellular (900-MHz) and PCS (1000-MHz) frequencies.

transmit applications, the need to accurately control power stems from two requirements. In many spread-spectrum systems, particularly those using CDMA and WCDMA signals, the amount of power being transmitted from each terminal in a particular cell is critical to system capacity.

In high-power transmitters, thermal management is another critical issue. In addition to complying with emission regulations, a base station must not transmit more power than that for which it is thermally dimensioned. In multicarrier base-station transmitters, therefore, the total power (the composite RMS energy in all of the channels) is a measure of a transmit power amplifier's (PA's) safe operating region.

The need to measure RMS energy also crops up in single-channel applications where the modulation scheme can change. A radio-link platform, for example, might be designed to transmit and receive multiple phase-modulation schemes. Simple QPSK might be used to transmit at a low data rate while a more complex scheme such as 16-state quadrature amplitude modulation (16QAM) or 64QAM would be used for higher data rates. These different modulation schemes have different crest factors, necessitating the need for a detector to measure the RMS signal strength.

Table 3: Comparing true power-measurement methods

Parameter	Diode detector	Logamp IC	Thermal detector	True Power Detection™ IC
Circuit-board volume	Moderate	Low	High	Low
Power consumption	Low	Moderate	High	Low
Single-channel CDMA response delta (crest factor of 5.5 dB)	~ +2 dB	+3.55 dB	0 dB (reference)	+0.2 dB
Multichannel CDMA response delta (crest factor of 12 dB)	> +5 dB	> +5 dB	0 dB (reference)	+0.8 dB
Temperature stability	Low	Moderate	High	High
Response time	Microseconds	< 10 µs	Milliseconds	< 5 µs
Linear dynamic range	20	Up to 100	20	25 to 30
Cost	Low	Moderate	Moderate	Moderate

In a typical handset power-control application (Fig. 9), the AD8361 is used in conjunction with a directional coupler. In this application, maximum power transmitted by the PA is +30 dBm. The coupler samples a portion of this RF power. With most off-the-shelf directional couplers for handset applications that have a coupling factor of approximately 20 dB, additional attenuation of the signal is required to translate the power into the range of the AD8361, with the IC running on a +2.7-VDC supply. Also, some bandpass filtering will generally be performed before the signal is applied to the true power detector. The output voltage from the AD8361 may then be used in an analog automatic-gain-control (AGC) loop to regulate the transmitted power. Alternatively, the output voltage may be digitized (as shown). The regulation is then performed in the digital domain and any adjustments are made using a digital-to-analog converter (DAC). While discrete ADCs and DACs can be used in this type of application, a more integrated device such as the ADuC812 can also be used. The ADuC812 incorporates an eight-channel 12-b ADC and two 12-b DACs along with an 8051 microcontroller core. Once digitized, the RMS voltage can be converted to a value that is proportional to the power in watts.

The AD8361 offers a new approach to accurate, cost-effective measurement of RMS signal strength at RF. The IC supports a wide range of applications. With inherent waveform independence and temperature stability, it eases the RF designer's task of measuring complex waveforms. **Analog Devices, Inc., 804 Woburn St., Wilmington, MA 01887; (781) 937-1292, FAX: (781) 937-1024, Internet: http://www.analog.com.**

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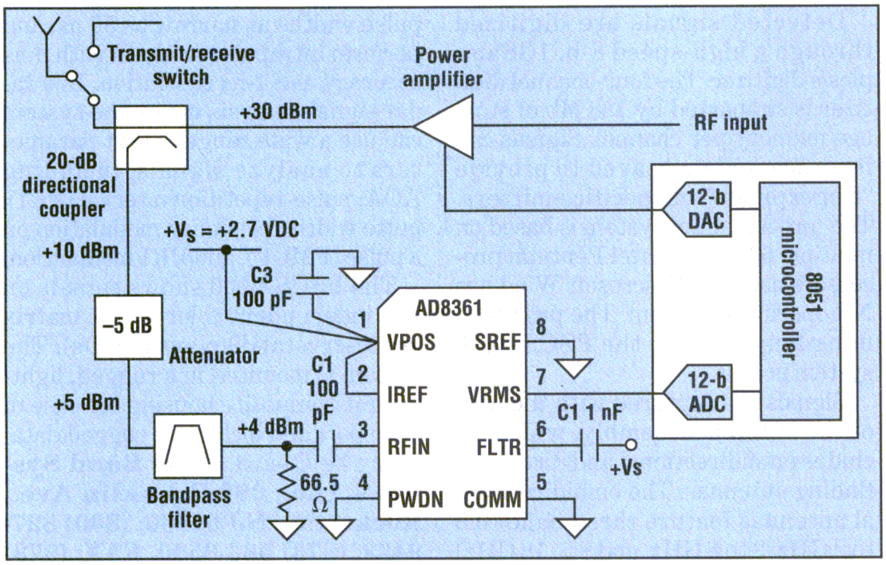


9. In a typical power-control application, the transmitted power is sampled using a directional coupler and fed into the AD8361 true power detector. An ADuC812 microcontroller provides ADC, digital-to-analog-converter (DAC), and control functions for the control loop.

Portable ELINT System Captures Data From 0.5 to 18 GHz

This portable system provides sophisticated signal-capture-and-display capabilities through advanced RF and digital signal processing.

JACK BROWNE
Publisher/Editor

ELECTRONIC intelligence (ELINT) has gained in importance since the end of the Cold War. Capturing information can be vital to survival, and depends upon the agility and flexibility of the signal-collection system. The PSCS-2000 Portable Signal Collection System from Wide Band Systems (Rockaway, NJ) has been designed for reliability and accuracy in demanding modern ELINT applications. The system, which can be configured with a variety of display modes, for signal angle-of-arrival (AOA) measurements from 0.5 to 18 GHz.

The PSCS-2000 system consists of the RF/microwave-processing hardware, digital-signal-processing (DSP) hardware, an antenna assembly with omnidirectional and direction-finding (DF) antennas, multiple display screens, and dedicated software. The receiver assembly actually incorporates multiple receivers. Two wideband instantaneous-frequency-measurement (IFM) receivers and a digitally synthesized tuner are used to cover the band from 2 to 18 GHz with –60-dBm sensitivity, 1.25-GHz frequency resolution, and 2.5-MHz root-mean-square (RMS) frequency accuracy. The amplitude resolution is 0.5 dB and the RMS amplitude accuracy is 1 dB. The IFM receivers can resolve a signal pulse-width time of arrival (TOA) within 25 ns.

A narrowband receiver subassembly switches between intermediate frequencies (IFs) of 1 GHz and 160 MHz to locate an emitter of interest, under control of the main receiver assembly and RS-422 bus. The selection of IF is based upon selected emitter parameters. Either IF provides a dynamic range of 70 dB with better than 0.5-MHz nominal frequency resolution and 1-MHz RMS frequency accuracy. The amplitude resolution and RMS amplitude accuracy are 0.5 dB and 1 dB, respectively. Both narrowband channels can resolve a pulse width TOA to 25 ns.

Detected signals are digitized through a high-speed 8-b, 1GSamples/s digitizer. The four-channel digitizer is supported by 128 Mb of storage memory per channel. Signals can be processed and saved to provide "fingerprints" of specific emitters. The processor subsystem is based on multiple 600-MHz Intel Pentium processors using the Microsoft Windows NT operating system. The programming language for the PSCS-2000 system is C++.

Signals are captured with the help of the antenna assembly, which includes omnidirectional and direction-finding antennas. The omnidirectional antennas feature three-band (0.5 to 2 GHz, 2 to 8 GHz, and 8 to 18 GHz) slant linear polarization while the DF antenna uses two-band (0.5 to 1 GHz and 1 to 18 GHz) slant linear polarization. The DF antenna achieves better than 10.5-dBi gain at 1 GHz and 20-dBi gain at 18 GHz.

The PSCS-2000 clock is referenced to the Global Positioning System (GPS), which is also used for platform location. The system, with built-in demodulation capability, can operate in a variety of modes, including surveillance mode and communications signal-analysis mode. In the latter mode, the system can decipher a wide range of modulation formats, including amplitude modulation (AM), frequency modulation (FM), pulse modulation, quadrature amplitude modulation (QAM), quadrature-phase-shift-keying (QPSK) modulation, and pulse-width modulation (PWM). In the surveillance mode, the system can handle pulse widths as narrow as 50 ns, and perform intrapulse analysis with 3-ns accuracy and 1-ns resolution. In a radar signal-analysis mode, the system can use a wide range of sort parameters to analyze signals, including AOA, pulse-repetition interval (PRI), pulse width, frequency modulation on a pulse (FMOP), and PRI modulation.

The PSCS-2000 shows signals on two independent color-active matrix liquid-crystal displays (LCDs). The system is mounted in a rugged, lightweight-composite housing for ease of transport and includes a rugged data-entry keyboard. **Wide Band Systems, Inc., 389 Franklin Ave., Rockaway, NJ 07866; (800) 827-9433, (973) 586-6500, FAX: (973) 627-9190, e-mail: bilsul@widebandsystems.com, Internet: http://widebandsystems.com.**

CIRCLE NO. 53 or visit www.mwrf.com

Power Modules Reduce Amplifier Design Time

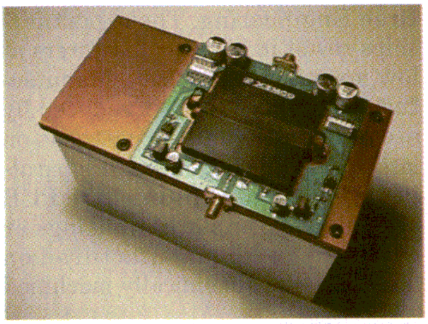
Targeting base-station power amplifiers, these new 200-W power modules reduce the number of individual stages that must be combined to achieve power levels of more than 400 W.

PETER STAVENICK
Managing Editor

FAST turnaround is a quality that can set a company and its product apart from the competition. In today's world, the company that can get products to market faster wins the most customers. The QPP-017 200-W QuikPAC™ power module from XEMOD (Sunnyvale, CA) provides cellular-amplifier designers with a building-block approach to rapidly turn a design into working hardware compared to conventional laterally-diffused-metal-oxide-semiconductor (LDMOS) transistor amplifier designs. QuikPAC modules improve return-on investment by reducing design time and lowering material and manufacturing costs.

The QPP-017 module is 869-to-894-MHz Class AB amplifier stage for use in the output stage of high-power linear RF power amplifiers (HPAs). Together with other surface-mount, flat-pack modules from XEMOD, it is designed to be used in a 50-Ω stripline assembly that connects the power modules with couplers, attenuators, as well as other passive and active components of the designer's choice to create a complete PA (see figure). A complete system includes a set of RF power stages that has the capability to amplify low levels of input power up to hundreds of watts of output.

The modules truly simplify the design of cellular HPAs with a power level that more than matches the highest-output discrete devices currently available. In the past, it was difficult to design these types of high-power stages due to the difficulty of developing matching networks. The new QuikPAC modules, however, reduce the stages that must be combined to achieve power levels above 400 W. The combination of a new and proprietary XeMOS metal-oxide-semiconductor-field-effect transistor (MOSFET) and internal impedance matching circuits solves the problem of matching to the very low impedances of high-power transistors.



The 200-W QuikPAC module with fixture can be used in sets to create a complete amplifier or portions of an amplifier.

The QPP-017 module is designed for the output stage of a cellular base-station linear PA over the range of 869 to 894 MHz. Features include 50-Ω input and output impedances, high gain, low intermodulation distortion (IMD), and a load impedance that results in 10:1 maximum VSWR with a +24-to-+32-VDC supply (drain) voltage range. Maximum control (gate) voltage is +4 VDC.

At +28 VDC, minimum peak-envelope power (PEP) at 1-dB two-tone compression is 200 W. Minimum gain at 200-W PEP is 11.5 dB with a nominal gain of 12 dB. Nominal gain flatness over frequency at two-tone PEP is ±0.05 dB. Nominal quiescent current is 2.1 A with +3.1-VDC bias voltage. Nominal operating current is 10.5 at 200-W PEP.

The advantages in using the QuikPAC modules in amplifier design are significant. For one, the modules' gain linearity over a wide dynamic range enables integration into feedforward amplifier designs. Polarity biasing is used on all RF modules and the use of the company's scaled devices improves stage-to-stage tracking. As a thermally efficient design, a feedforward correction-compatible architecture is used.

The major elements in the QuikPAC system consist of power modules to 200-W PEP, RF driver modules that are scaled to output modules, and gain blocks for input and intermediate stages. Precision-error amplifiers with a high intercept point and flat response for high cancellation ratios are included, along with phase and amplitude modulators for adaptive correction. **XEMOD, Inc., 333 Soquel Way, Sunnyvale, CA 94086; (408) 733-7229, FAX: (408) 733-7327, e-mail: info@xemod.com, Internet: http://www.xemod.com.**

CIRCLE NO. 54 or visit www.mwrf.com

Firm Fosters Co-Military Filter-Design Approach

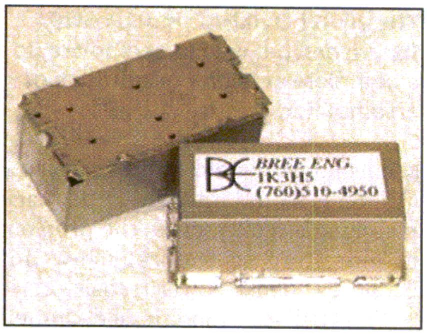
Although a startup, this company brings more than 25 years of experience to the design of RF/microwave filters and filter assemblies.

JACK BROWNE
Publisher/Editor

FILTER design involves a blend of engineering experience, imagination, and perseverance. All three of these ingredients can be found at Bree Engineering (San Marcos, CA), a new firm in Southern California that was founded by longtime filter developer Dan Bree. Bree uses the term "co-military" on the firm's website (at www.breeeng.com) to summarize the product philosophy: "the best of both worlds, military quality, commercially marketed." Bree Engineering offers lumped-element and cavity filters at frequencies from kilohertz through gigahertz ranges with everything from bandpass, lowpass, and highpass filters to complex multiplexer assemblies.

The firm features a line of surface-mount lumped-element filters at prices that are competitive with dielectric and surface-acoustic-wave (SAW) filters. Frequencies can range from approximately 100 kHz to about 6 GHz, restricted by the bandwidth limitations of the surface-mount packaging. Filter bandwidths are available from 1 percent to multi-octave designs. Filter functions need not be restricted to standard equal-ripple Chebyshev responses, but can also be Butterworth, elliptical, and time-domain responses.

The Southern California facility accommodates fast-turnaround designs and prototypes, although higher-volume production-line products are manufactured at additional local and offshore facilities. Less process-sensitive circuit elements (such as coils and toroid windings) are produced offshore. More-sensitive processes, such as component placement on the printed-circuit board (PCB), are performed by local pick-and-place PCB manufacturing facilities that serve as strategic partners with Bree Engineering. These partnerships result in high-quality filters at low manufacturing costs. Manufacturing costs are further reduced by etching multiple filter patterns on large pre-scribed PCBs, and separating the individual filters once PCB assembly is complete. Housings or covers are also manufactured on large sheets of chemically machined nickel-silver (NiAg) alloys that do not require plating.



A line of surface-mount RF/microwave filters is available at frequencies through 6 GHz. The filters can be incorporated into larger designs using standard solder-reflow and pick-and-place manufacturing processes.

The surface-mount filter product line (see figure) is assembled with high-temperature solder, allowing each filter to withstand a standard 62-percent tin (Sn) solder-reflow process. As results, users can handle these filters as they would with a standard surface-mount package, using pick-and-place assembly equipment and standard solder-reflow manufacturing techniques. Although the surface-mount filters are commercial products, all of the process steps are inspected for the highest quality. Complete incoming inspection processes as well as on-site source inspections (where practical) are performed on all of the products.

In addition to the surface-mount filters, the company also offers a complete line of lumped-element and cavity filters with connectors for commercial and military applications. A visit to the website provides specifiers with an on-line quote request form for bandpass as well as highpass/lowpass filters. Additionally, specifiers are asked for mechanical requirements and additional electrical specifications. Request forms can be automatically e-mailed to Bree Engineering for a quick response. All requests are modeled at the quote stage to ensure compliance with a customer's requirements. **Bree Engineering, 1269 Linda Vista Dr., San Marcos, CA 92069; (760) 510-4950, FAX: (760) 510-4959, Internet: http://www.breeeng.com.**

CIRCLE NO. 55 or visit www.mwrf.com

Engineering Kit Tames Pesky RF Problems

This free Engineer's Survival Kit contains six samples of EM absorber that can be applied to troublesome designs through 18 GHz.

JACK BROWNE
Publisher/Editor

CONQUERING electromagnetic interference (EMI) can challenge the best design engineers. It may help to have the Engineer's Survival Kit from ARC Technologies, Inc. (Amesbury, MA). It contains six samples of the company's standard EM absorbing materials, each backed with a peel-and-stick adhesive for attachment to circuits and housings. The materials can be used to control EM radiation at frequencies through 18 GHz.

The Engineer's Survival Kit offers six 6 × 6-in. (15.24 × 15.24-cm) panels of absorbing materials of different types and thicknesses. The kit submitted for this review includes ARC-UD-11554, a urethane-based 0.175-in. (0.4445-cm)-thick absorber material with a frequency range of 0.7 to 18 GHz, ARC-DD-10214, a silicone-based 0.030-in. (0.0762-cm)-thick absorber material with a frequency range of 4 to 18 GHz, ARC-DD-10017, also silicone based, with in a 0.087-in. (0.22098-cm) thickness for applications from 2 to 18 GHz, ARC-LS-10055, a lossy 0.125-in. (0.3175-cm)-thick foam material with an acrylic protective coating for use from 4 to 18 GHz, the similar but thicker [0.25-in. (0.635-cm)] ARC-LS-10211 foam material for applications from 2 to 18 GHz, and ARC-ML-10049, a 0.375-in. (0.9525-cm)-thick broadband absorber with acrylic protective coating for applications from 8 to 18 GHz.

The "DD" and "UD" products are resonant-tuned microwave absorber materials available with a choice of four binders: silicone (which is the version supplied in the kit, for maximum flexibility and good environmental characteristics), urethane (for improved abrasion resistance and tear strength), neoprene (for good economy, durability, and weather resistance), and Nitrile (for resistance to water and fuel). These materials are offered for frequencies from 2 to 18 GHz, with an EM absorption of 20 dB or more. Depending upon the choice of binder, these materials can handle temperature ranges as wide as –60 to +375°C (for silicone). Standard sheet sizes for the "DD" and "UD" absorber materials are 12 × 12 in. (30.48 × 30.48 cm) and 24 × 24 in. (60.96 × 60.96 cm).

LOW-DENSITY FOAMS

The "LS" materials are low-density, high-loss flexible foams that are precisely and uniformly impregnated with carbon for good EMI suppression. They can be used in waveguide applications as a dissipative material or within resonant cavities to reduce unwanted reflections. The LS materials are rated for temperatures from –70 to +270°C and are supplied in a standard sheet size of 24 × 24 in. (60.96 × 60.96 cm). Standard thicknesses range from 0.125 to 1.00 in. (0.3175 to 2.54 cm) [with attenuation at a particular frequency increasing as a function of thickness], although thicknesses to 6.00 in. (15.24 cm) are available. Non-corrosive versions are also available. These versions can be used in hostile environments.

Finally, the "ML" materials are comprised of multiple layers of foam. Each layer is loaded with a different loss tangent to produce a gradient design. The layers are laminated for maximum absorption. For weather-resistant requirements, the multilayer foams can be wrapped with a neoprene nylon fabric or sprayed with a flexible polyurethane coating. The ML materials, which can be specified in thicknesses from 0.25 to 4.5 in. (0.635 to 11.43 cm), are supplied in standard sheets of 24 × 24 in. (60.96 × 60.96 cm). The thickest materials provide the lowest-frequency coverage, down to 0.5 GHz for the 4.5-in. (11.43-cm)-thick ML materials. The ML materials are rated for temperatures from –70 to +270°C.

Each Engineer's Survival Kit includes a capabilities and cross reference guide to help engineers with the selection process. The small sheets, which are suitable for commercial and military applications, are ideal for prototyping as well as experimentation. In most cases, attachment to a metal surface is recommended for optimum EM absorption. P&A: free; stock. **ARC Technologies, Inc., 11 Chestnut St., Amesbury, MA 01913; (978) 388-2993, FAX: (978) 388-6866, e-mail: sales@arc-tech.com, Internet: http://www.arc-tech.com.**

CIRCLE NO. 56 or visit www.mwrf.com

HP's Second Generation PHEMT process FETs and amplifiers are so consistent, you'd think we cloned them!

Introducing the first products in HP's Second Generation PHEMT process – a process designed specifically for repeatability and high performance. These new parts allow predictable designs, better yields and fewer surprises in your base station, handset or other wireless application.

In fact, we're so confident of their performance that we include distribution charts of key specifications right on the data sheets. An industry first.

Already designed into several major CDMA handsets, the MGA-72543 LNA's integral switch allows you to bypass the amplifier, reducing system current needs. The ATF-34143 FET with its low noise figure and excellent linearity at 4V, is perfect for base station LNA applications. The ATF-35143 FET offers economical low noise performance for portable applications.

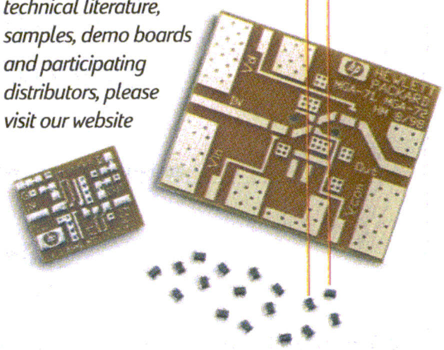
Add it all up and you'd have a hard time finding a more consistent bunch of parts anywhere.

Typical performance @ 2 GHz

Part Number	Bias	NF (dB)	Gain (dB)	IP3 (dBm)
MGA-72543*	3V, 5-60 mA	1.5	14.4	3.5-14.8 (input)
ATF-34143	4V, 60 mA	0.5	17.5	31.5 (output)
ATF-35143	2V, 15 mA	0.4	18.0	21.0 (output)
ATF-38143 *coming soon*	2V, 10 mA	0.5	16.0	22.0 (output)

** as a switch (amp bypassed): insertion loss = 2.5 dB, IIP3 = 35 dBm*

For more product information, technical literature, samples, demo boards and participating distributors, please visit our website



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Coaxial Contacts Offer Low Loss And High Repeatability Past 6 GHz

This innovative coaxial connector system provides reliable electrical connections with minimal insertion force.

JACK BROWNE

Publisher/Editor

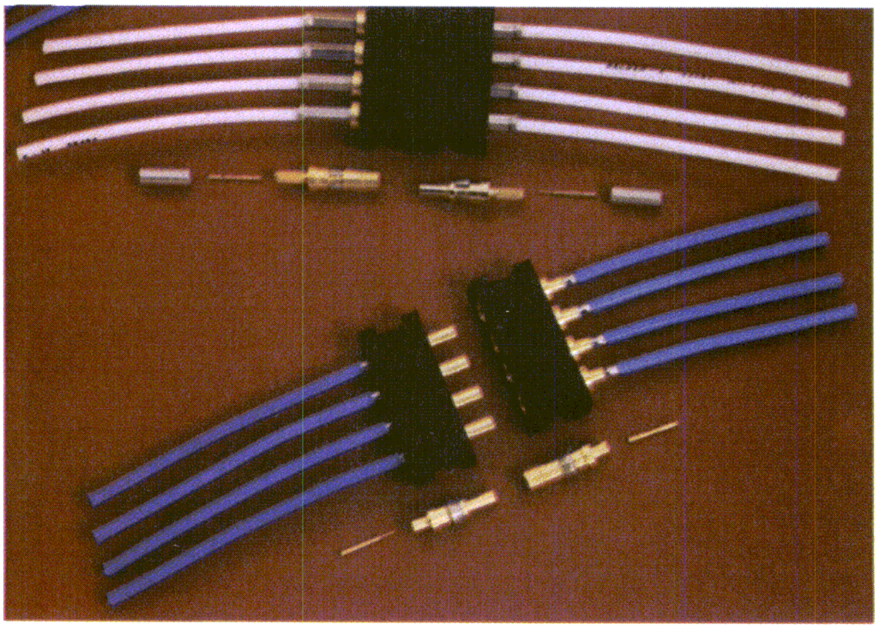
CONNECTORS are vital to system design, but often overlooked. Poor connectors can add excessive loss and VSWR to a design, and decrease reliability. Good connectors are invisible to the system, forming electrical links without adding their own electrical characteristics. The Hypertac Coaxtac coaxial connectors from Hypertronics Corp. (Hudson, MA) are an example of connectors that perform almost invisibly, adding minimal loss and reflections while maintaining high performance levels over more than 25,000 mating cycles.

The Hypertac Coaxtac components (see figure) have been designed for reliable, low-insertion-force connections with good electrical performance through 6 GHz (depending upon the choice of associated coaxial cable). The mating mechanism consists of a pin contact and a sleeve formed by multiple wires strung at an angle to its socket's axis. As the pin is inserted into the wire sleeve, the wires stretch to accommodate the shape of the pin. Since the sleeve wires wrap around the pin at multiple contact points, the electrical resistance is low, with good power-handling capability. By maintaining the sleeve wires at a controlled angle, the engagement and disengagement forces are kept low. In addition, the smooth, light wiping action of the wires on the contact pin results in negligible wear on the contact surfaces.

WITHSTANDING VIBRATION

The Hypertac Coaxtac contact system features a low-mass design. The contact wires exhibit low insertion relative to their resilience, enabling them to follow the most abrupt or extreme excursions of the pin without loss of contact. Tests performed across the full 6-GHz operating range revealed no discontinuities, even when checked for events as small as 3 ns. The contact area in the Hypertac Coaxtac connectors is not limited to a single plane, but extends 360 deg. around the pin and is uniform over its entire length. This ensures good electrical continuity regardless of the direction or intensity of external or inertial forces.

The Hypertac Coaxtac connectors exhibit maximum VSWR of 1.20:1 at 3 GHz and 1.40:1 through 6 GHz. Even with slight misalignment, the VSWR performance remains acceptable for most applications (see table). Only 8-oz. (maximum) force is required to engage the connectors, and only 6-oz. force (maximum) to disengage them. The connectors are designed to withstand more than 25,000 connect/dis-



The Coaxtac low-insertion-force coaxial contacts provide low-loss interconnections with good repeatability at frequencies through 6 GHz.

connect cycles with only negligible degradation in electrical performance. The maximum insertion loss is 0.30 dB at 3 GHz. The connectors suffer only –70-dB RF leakage at the connector interface when fully mated. The connectors feature contact resistance of 8 mΩ maximum at the inner contact and 2 mΩ maximum at the outer contact.

The contacts are available as discrete components or in rack-and-panel housings. Optional signal, 15-to-200-A power, 50-Ω coaxial connectors, and 8-kV modules are available for a wide range of applications. Standard rack-and-panel and quick-release cable-to-chassis housings are available to mount three-position coaxial modules as well as optional signal and power contacts. Up to 39 coaxial contacts can be mounted in standard rack-and-panel frames with a misalignment capability of 0.049 in. (0.12446 cm).

The innovative coaxial connectors are ideal for applications requiring blind-mate connectors. They are rated for operating temperatures from –55 to +125°C and root-mean-square (RMS) dielectric withstanding voltage of 500 V.

The coaxial connectors meet the appropriate provisions as defined within Group C inspection of MIL-C-39012 in conjunction with MIL-STD-202 and its test methods. P&A: $20 (mated coaxial pair) (1000 qty); 4 to 6 wks. **Hypertronics Corp., 16 Brent Dr., Hudson, MA 01749-2978; (800) 225-9228, (888) HYPERTAC, (978) 568-0451, FAX: (978) 568-0680, e-mail: info@hypertronics.com, Internet: http://www.hypertronics.com.**

CIRCLE NO. 57 or visit www.mwrf.com

Checking performance with various amounts of axial disengagement

Gap (mm)	VSWR at 3 GHz	VSWR at 6 GHz	Insertion loss (dB)
No gap	1.13:1	1.23:1	0.09
0.51	1.17:1	1.33:1	0.09
0.76	1.18:1	1.43:1	0.11
1.01	1.23:1	1.57:1	0.13
1.27	1.26:1	1.67:1	0.15
1.52	1.28:1	1.69:1	0.19

THE CONNECTORS ARE DESIGNED TO WITHSTAND MORE THAN 25,000 CONNECT/DISCONNECT CYCLES WITH ONLY NEGLIGIBLE DEGRADATION IN ELECTRICAL PERFORMANCE.

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1999 ANALOG WORKSHOPS

Choose from 8 workshops to suit your design needs.

Special Guest, Keynote Speaker
ROBERT DOBKIN
Join us at the Luncheon and hear one of the world's leading experts on analog design from Linear Technology set the tone for this extraordinary workshop.
Tuesday, Sept. 21, 12:30 p.m.

Electronic Design Continuing Education (EDCE) has instituted an ongoing educational program for design engineers and engineering managers. EDCE's 1999 Analog Workshop is the first in a series of events to help designers stay in tune with state-of-the-art and emerging technology approaches to problem solving.

This is your opportunity to meet, mix, and mingle with your peers and proven analog experts from your industry. Explore the inside track and hear why and how the experts make their design solution decisions. See for yourself how emerging technologies help deliver better products to market faster. Experience the excitement of learning from those who have found solutions to new and more complicated analog design technical problems. This is a not-to-be-missed workshop experience.

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Ashok Bindra (left) will chair this important two-day Analog Workshop program. Ashok is *Electronic Design's* Analog, Power Supplies & DSP Editor, and has gained considerable experience organizing numerous technical conferences, workshops, panel discussions, and forums. His latest accomplishment was co-chairing the World Spring Design Conference in 1998.

Turn page for workshop information, schedules, and registration forms.

For workshop information, call Fred Sklenar at 201.393.6082

Analog Workshops at a glance

Tuesday, Sept. 21, 8:30 a.m. - 12:30 p.m.; Room: Stevens Creek

1. Applying High-Speed Operational Amplifiers in Communication Systems

Instructor: Michael Steffes, Senior Strategic Engineer, Burr-Brown Corp.

Topics to be covered:

I. Relative merits of voltage-feedback and current-feedback type amplifiers: a.) Frequency response; b.) Noise and distortion; c.) External techniques to improve performance.

II. Amplifiers in wireless systems: a.) Receiver channel applications; b.) Transmit channel applications.

III. Amplifiers in wired systems: a.) DSL line interface options; b.) Cable modem requirements.

Tuesday, Sept. 21, 8:30 a.m. -12:30 p.m.; Room: Winchester

2. Designing Class D Amplifiers for Audio Systems

Instructor: Pat Begley, Senior Manager of Audio Programs, Harris Semiconductor

The workshop agenda includes an introduction and overview of amplifier techniques, switching amplifier approaches versus Class D, benefits of Class D in terms of efficiency, size and weight, future directions for audio, and demonstration of hi-fidelity Class D amplifier solutions.

Tuesday, Sept. 21, 1:30 p.m. - 5:00 p.m.; Room: Stevens Creek

3. Understanding the Concepts of Electro-Magnetic Compatibility (EMC) and PCB Suppression Techniques

Instructor: Mark Montrose, Consultant, Montrose Compliance Services

Besides learning EMC fundamental and implementation techniques, this course will provide an overview of European EMC directives. It will discuss in a simplified manner how and why electro-magnetic interference (EMI) exists within a printed-circuit board (PCB) without the use of sophisticated math. It will disclose proper design and layout techniques to attain first-time compliance with International EMC requirements.

Tuesday, Sept. 21, 1:30 p.m. - 5:00 p.m.; Room: Winchester

4. Designing With High-Speed Data Converters

Instructor: Paul Hendriks, Senior Applications Engineer, High-Speed Data Converter Group, Analog Devices Inc.

Starting with an overview of analog-to-digital and digital-to-analog (ADC and DAC) architectures, this workshop will show how to drive high-speed high-resolution ADCs for direct IF sampling in a variety of applications, including digital receivers. Understanding DAC specifications also will be covered, as it simplifies the workings of direct digital synthesis. Optimizing DAC performance also will be discussed. Finally, this workshop will demonstrate the use of ADCs and DACs in high-speed communications systems.

Wednesday, Sept. 22, 8:30 a.m. - 12:30 p.m.; Room: Stevens Creek

5. Performance Verification For High-Resolution Data Converters: Getting All The Bits You Paid For

Instructor: Jim Williams, Staff Scientist, Linear Technology

Instrumentation, waveform generation, data acquisition, feedback control systems and other applications are utilizing high-resolution ADCs and DACs. 16-,18-, and even 20-bit resolution measurements are becoming increasingly common. This lecture describes hardware-based methods for verifying high-resolution converter performance. In particular, settling time measurements of 16-bit DACs is covered. Additionally, techniques for testing ADC linearity beyond 20 bits are also presented.

Wednesday, Sept. 22, 8:30 a.m. - 12:30 p.m.; Room: Winchester

6. Designing Switching Power Supplies

Instructor: Robert A. Mammano, Vice President of Advanced Technologies, Unitrode Corp.

This workshop will start with an introduction to switching techniques and topologies, and then go directly into practical topologies for off-line applications, DC/DC converters, and AC/DC converters. This discussion will also describe specialized topologies such as compound converter, resonant, and softswitching topologies. Additionally, it will focus on component considerations in the design of switch-mode power supplies, as well as magnetic fundamentals including inductor and transformer designs. Finally, the workshop will address myriad control issues and solutions, as it presents design examples to highlight the control methodology.

Wednesday, Sept. 22, 1:30 p.m. - 5:00 p.m.; Room: Stevens Creek

7. Design Of AC/DC Motor Control Circuits

Instructor: Dal Y. Ohm, Principal Consultant, Drivetech Research

The pupose of this workshop is to provide basic concepts and technical skills necessary to design various types of AC and DC motor drives. Types of motors to be discussed will include brushed and brushless DC, stepper, reluctance, and induction motors, with special emphasis on low to medium power drives. Practical and useful procedures in selecting components and control methods, design rules, and performance versus cost tradeoffs will be discussed.

Wednesday, Sept. 22, 1:30 p.m. - 5:00 p.m.; Room: Winchester

8. Building the RF Front-End for a Software Radio

Instructor: Clive Winkler, Vice President, Engineering, Cubic Communications

There are powerful reasons for providing an analog RF downconverter prior to digitizing the signal for a software radio. Amongst them, dynamic range and the signal environment are a couple of the most important. We start with the need for dynamic range, phase noise and aperture jitter and derive the theoretical equations that relate them. Then we develop a clear perspective of the physical processes and requirements that need to be met for a modern software radio which is also able to cope with the high accuracy (amplitude and phase response) necessary for more advanced digital modualtions (high order m-ary PSK and QAM).

The newer evolving architecures for both the receive and transmit RF elements are considered along with ways of correcting for the amplitude and phase errors encountered in practical components.

Register Today and Save!

Advanced registration with payment must be postmarked or faxed by September 13, 1999. After September 13, please register on site.

STEP 1.

Name ____________________

Company ____________________

Title ____________________

Address ____________________

Mailstop/Suite ________ City ____________ State ______ Zip ________

Phone ____________________ Fax ____________________

E-mail ____________________

STEP 2.

Workshop Selection(s):

Tues., Sept. 21

- ❑ 8:30 a.m. Workshop 1
- ❑ 8:30 a.m. Workshop 2
- ❑ 1:30 p.m. Workshop 3
- ❑ 1:30 p.m. Workshop 4

Wed., Sept. 22

- ❑ 8:30 a.m. Workshop 5
- ❑ 8:30 a.m. Workshop 6
- ❑ 1:30 p.m. Workshop 7
- ❑ 1:30 p.m. Workshop 8

❑ Yes, I will attend the Keynote Luncheon on Tuesday, September 21. (Complimentary for Workshop attendees.)

STEP 3.

Payment:

One Workshop Fee:
Advanced Registration: **$250**
On-site Registration: **$300**

Two Workshops Fee:
Advanced Registration: **$400**
On-site Registration: **$500**

Three Workshops Fee:
Advanced Registration: **$500**
On-site Registration: **$650**

Four Workshops Fee:
Advanced Registration: **$600**
On-site Registration: **$800**

TOTAL: $ ____________

❑ Check enclosed (Please make payable to "Penton - EDCE") ❑ MasterCard ❑ VISA ❑ American Express

Account Name ____________________
(Please Print)

Card Number ____________________

Expiration Date ____________________

Member's Signature ____________________

Refund Policy:
For full refund, written request must be postmarked by September 13, and mailed to EDCE Analog Workshops, 611 Route 46 West, Hasbrouck Heights, NJ 07604.
Please allow 6-8 weeks for refund to be processed.

Analog Workshop Schedule

TUESDAY, SEPTEMBER 21, 1999

	TIME	WORKSHOP DESCRIPTION	LOCATION
REGISTRATION	7:30 a.m. - 8:30 a.m.	Continental Breakfast	Outside Stevens Creek
WORKSHOP 1	8:30 a.m. - 10:15 a.m.	Applying High-Speed Amplifiers in Communication Systems	Stevens Creek
WORKSHOP 2	8:30 a.m. - 10:15 a.m.	Designing Class D Amplifiers for Audio Systems	Winchester
BREAK	10:15 a.m. - 10:30 a.m.	Refreshments	In rooms
WORKSHOP 1	10:30 a.m. - 12:30 p.m.	(Continued)	Stevens Creek
WORKSHOP 2	10:30 a.m. - 12:30 p.m.	(Continued)	Winchester
LUNCHEON	12:30 p.m. - 1:30 p.m.	Keynote Speaker, Bob Dobkin	
WORKSHOP 3	1:30 p.m. - 3:15 p.m.	Understanding concepts of EMC and Suppression Techniques	Stevens Creek
WORKSHOP 4	1:30 p.m. - 3:15 p.m.	Designing with High-Speed Data Converters	Winchester
BREAK	3:15 p.m. - 3:30 p.m.	Refreshments	In rooms
WORKSHOP 3	3:30 p.m. - 5:00 p.m.	(Continued)	Stevens Creek
WORKSHOP 4	3:30 p.m. - 5:00 p.m.	(Continued)	Winchester

WEDNESDAY, SEPTEMBER 22, 1999

	TIME	WORKSHOP DESCRIPTION	LOCATION
REGISTRATION	7:30 a.m. - 8:30 a.m.	Continental Breakfast	Outside Stevens Creek
WORKSHOP 5	8:30 a.m. - 10:15 a.m.	Performance Verification for High-Resolution Data Converters	Stevens Creek
WORKSHOP 6	8:30 a.m. - 10:15 a.m.	Designing Efficient Switching Power Stations	Winchester
BREAK	10:15 a.m. - 10:30 a.m.	Refreshments	In rooms
WORKSHOP 5	10:30 a.m. - 12:30 p.m.	(Continued)	Stevens Creek
WORKSHOP 6	10:30 a.m. - 12:30 p.m.	(Continued)	Winchester
LUNCH	12:30 p.m. - 1:30 p.m.		Mezzanine
WORKSHOP 7	1:30 p.m. - 3:15 p.m.	Design of AC/DC Motor Control Circuits	Stevens Creek
WORKSHOP 8	1:30 p.m. - 3:15 p.m.	Building the RF Front-End for a Software Radio	Winchester
BREAK	3:15 p.m. - 3:30 p.m.	Refreshments	In rooms
WORKSHOP 7	3:30 p.m. - 5:00 p.m.	(Continued)	Stevens Creek
WORKSHOP 8	3:30 p.m. - 5:00 p.m.	(Continued)	Winchester

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Hasbrouck Hghts., NJ 07604

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201.393.6073

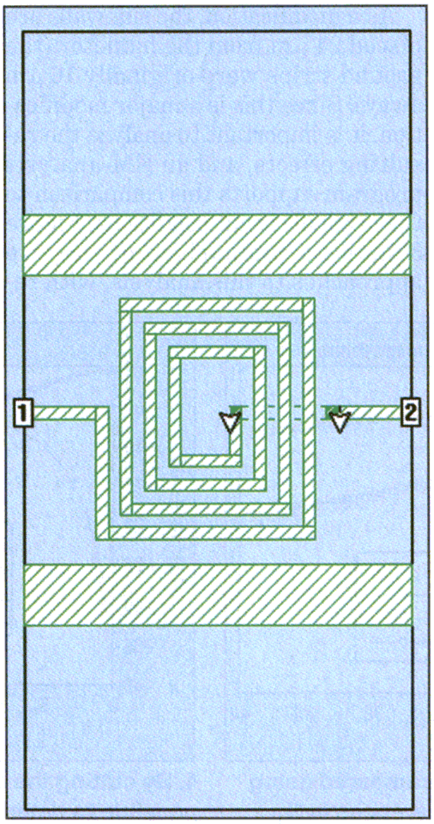
Free EM Software Analyzes Spiral Inductor On Silicon

One of the more difficult-to-model components, the spiral inductor, can be flawlessly analyzed with this free and powerful electromagnetic simulator.

James C. Rautio
President
Sonnet Software, Inc., 1020 Seventh North St., Suite 210, Liverpool, NY 13088; (315) 453-3096, FAX: (315) 451-1694, e-mail: info@sonnetusa.com, Internet: http://www.sonnetusa.com.

SPIRAL inductors appear simple but are difficult to analyze. With the complexity of a conductive silicon (Si) substrate, the problem challenges the most advanced modern analysis tools. Amazingly, a free-of-charge electromagnetic (EM)-analysis program called Sonnet Lite can be used to provide a high-accuracy solution to this problem.

Sonnet Lite performs analysis of the three-dimensional (3D) EM fields around planar circuits. It is based on the industry-standard suite of EM software tools from Sonnet Software (Liverpool, NY). Sonnet Lite can be supplied on a compact-disc read-only memory (CD-ROM) from Sonnet® (see phone number at the end of article) for only the cost of shipping, or it can be downloaded free of charge from the company's website (also listed at the end of the article) [the download requires approximately 14-MB available hard-disk memory]. This is not a demonstration program but a fully operating (albeit scaled-down) version of Sonnet's em® program (see *Microwaves & RF*, August 1999, p. 152 for a review of Sonnet Lite). Sonnet Lite can handle all small-to-medium-size problems using up to 16-MB random-access memory (RAM) for a full matrix solution, up to two metal levels (and three dielectric layers), and as many as four ports. With care, a large number of problems can be solved within these constraints. The spiral inductor on Si is one of these problems.



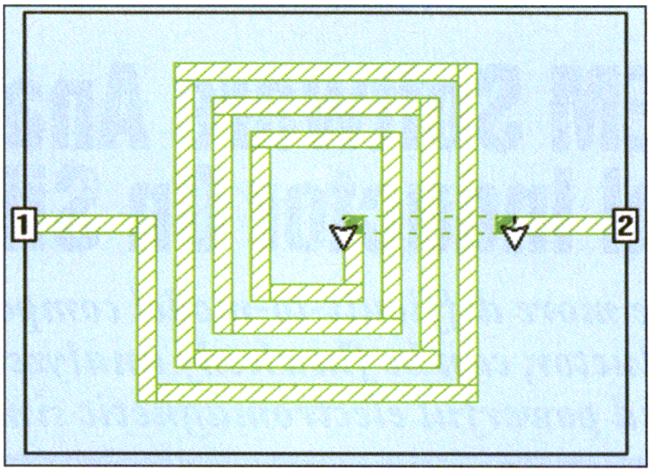
1. The baseline inductor has coplanar ground-return lines for low loss.

The basic characteristic of Sonnet Lite (and its full-fledged counterpart) that makes it ideal for analyzing a spiral inductor on Si is the accuracy when conducting dielectrics are analyzed. The software is based on Fast Fourier transform (FFT) analysis. Currently, there is no accuracy-degrading numerical integration. Also, all fields in each layer are represented as a (large) sum of simple rectangular waveguide modes (the sidewalls of the conducting box containing the circuit form the rectangular waveguide). When loss is present, only the characteristic impedance and velocity of propagation of each waveguide mode need to be changed. Since these characteristics are known exactly for rectangular waveguide, substrate conductivity is included with precisely the same accuracy as seen in an equivalent lossless analysis. This is difficult to perform in analyses based upon direct numerical integration of an underlying Green's function. Thus, the only problem to overcome in analyzing an inductor on Si using Sonnet Lite is to cast the problem into a form that satisfies the analysis constraints.

Figure 1 shows the baseline inductor. To reduce substrate loss, a coplanar inductor is being used. The wide conductors in the figure represent the coplanar ground strips. Ground-return current flows in these strips. Thus, the electric field from the signal line to ground need not pass through the entire Si substrate. This results in lower loss.

The inductor's transmission lines are 8 μm wide with 8-μm separation.

The ground strips are 16 μm from the edge of the inductor. The substrate is 1000 μm thick with a relative dielectric constant of 12 and a conductivity of 20 S/m. In addition, there is a 1-μm layer of silicon dioxide (SiO_2) with a relative dielectric constant of 4.0 on top of the Si. Most of the inductor is on top of the SiO_2 layer. The connection from the center of the inductor out is below the SiO_2 on top of the Si. A metal loss of 0.04 Ω/square (plus appropriate skin effects) is included.



2. The inductor of Fig. 1 was modified so that the box sidewalls take the ground-return current. Removing the ground strips results in a faster analysis.

The task at hand is to simplify the model of the inductor without affecting analysis accuracy. The Sonnet software meshes only the metal surface and problem size increases rapidly with the number of subsections. Thus, the number of subsections should be reduced as much as possible.

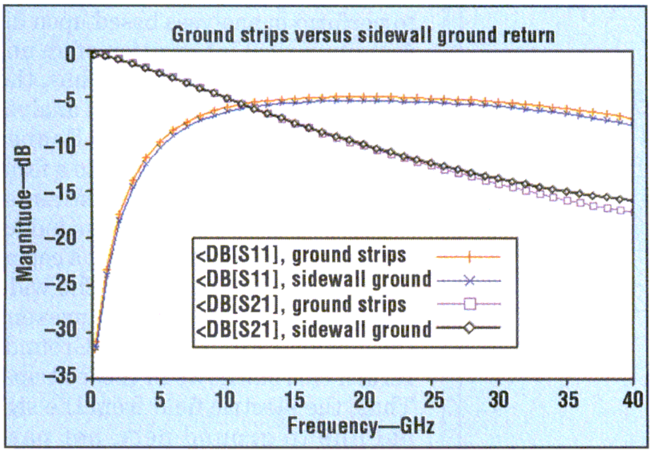
The first way this can be performed is by making the subsection size as large as possible. With 8-μm-wide lines on 16-μm centers, this is simple enough using an 8-μm cell size. Now, all 8-μm lines are meshed one cell wide. The impact of this subsection size on analysis error is quantitatively evaluated later.

Another way that the subsection count can be reduced is by reducing the metal area. Since the coplanar ground-return lines are wide, it would be helpful if they could be eliminated from the analysis. Since Sonnet Lite places a perfectly conducting box sidewall at the edge of the substrate, the coplanar ground-return lines can be eliminated. This is performed by removing each ground strip and substituting a box sidewall in its place. Now, the ground current flows in the box sidewalls rather than in the ground strips.

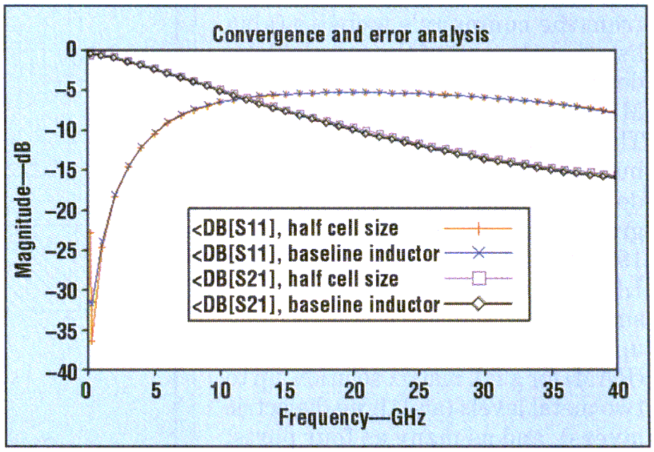
As a modification, the sidewalls are placed 24 μm from the inductor (the ground strips were originally 16 μm away). Since this is a major modification, it is important to analyze the resulting effects, and an EM-analysis program supports this comparison to the original circuit conditions. Figure 3 shows a comparison of the two approaches to this analysis, with reflection parameter S_{11} differing by approximately 0.5 dB in the two approaches while forward transmission, S_{21}, differs up to 2 dB, but only at the higher frequencies where S_{21} is down to approximately 20 dB. If the differences are assumed to be small compared to the design requirements, it is possible to adopt the second approach and proceed with the evaluation of the inductor of Fig. 2.

Using a typical Pentium-based personal computer (PC) running at a 450-MHz clock speed, the modified inductor of Fig. 2 can be analyzed in 1 s/frequency using only 1 MB of random-access memory (RAM). The original coplanar inductor requires 2-MB RAM and 2-s/frequency-analysis time. This may seem like a small difference at this point, but it could prove to be significant later. For those who are concerned with the 0.5-dB difference in analysis results, trade-offs can be performed using the basic inductor of Fig. 2. When the trade-offs are completed, one final analysis can be performed with all of the changes incorporated into the inductor of Fig. 1 if desired.

At this point, it is appropriate to determine the accuracy of the analysis. While suppliers of EM-analysis tools claim their programs to be accurate, the analysis errors rather than analysis accuracy are generally of in-



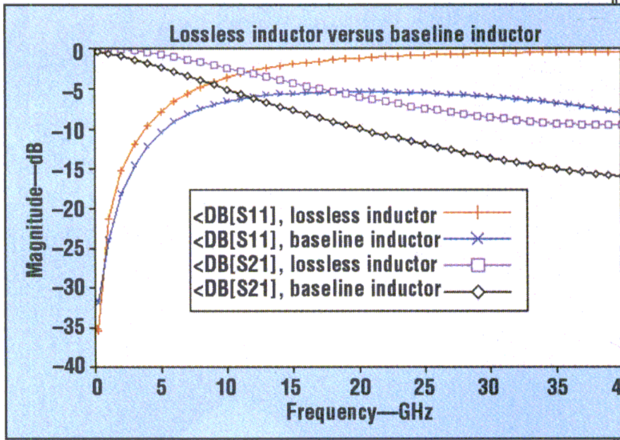
3. The inductors of Figs. 1 and 2 were compared using Sonnet Lite software. Note that the differences in S_{11} are approximately 0.5 dB while the differences in S_{21} approach 2 dB, but only at higher frequencies.



4. By cutting the cell size in half (from 8 to 4 μm), it is possible to determine error bounds. With at most 0.15-dB difference between the two results, model data should be within ±0.3 dB of the correct answer.

terest to high-frequency engineers. It is also desirable to have a quantitative value for the error. By comparing the estimated error with the project requirements, it is possible to tell if one's trust in an analysis approach is justified.

The error mechanisms for the full version of Sonnet have been extensively investigated. In nearly all of the cases, the principal error source is error that is due to the size of the cell. A unique characteristic of this error source is the fact that the error can be cut in half by shrinking the cell size in half (with a few well-understood exceptions). Error that is due to cell size is easily evaluated. If the cell size is cut in half and the circuit is then analyzed, then the error is also cut in half.



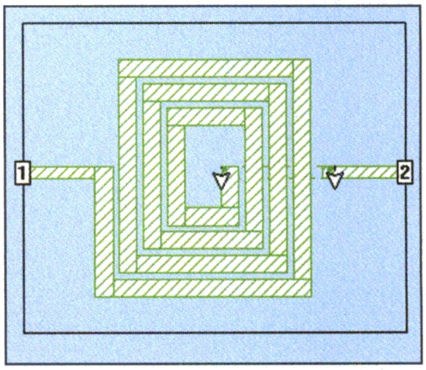
5. Comparing the loss of the baseline inductor (from Fig. 2) with the lossless version of the inductor shows that there is room for approximately 7-dB improvement at 40 GHz.

The results of this type of a convergence analysis are shown in Fig. 4. The original cell size is 8 μm. The second analysis uses a cell size of 4 μm with an analysis time of 6 s/frequency. The difference between the two curves is not easily seen, but is less than 0.15 dB nearly everywhere. This means that the results obtained using the 8-μm cell size should be good to approximately ±0.3 dB.

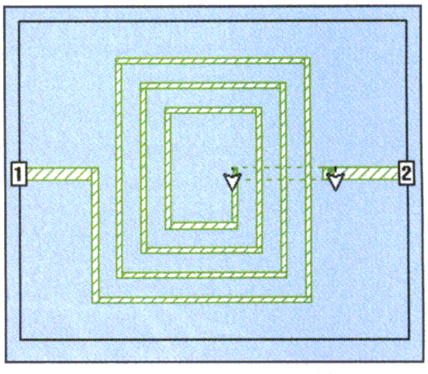
If this level of accuracy (or error) is sufficient for the needs of a particular analysis, then further modification is not necessary. But if more accurate results are needed, then an approach known as the "Richardson extrapolation" can be tried. In this technique, the total error for each data point is first determined (i.e., double the difference between the 8- and 4-μm cell-size results), and then the error is subtracted from the 8-μm answer. A spreadsheet is useful for performing a Richardson extrapolation. The results should now be accurate to approximately ±0.05 dB.

To ensure that a Richardson extrapolation is indeed providing increased levels of accuracy, a third analysis should be performed with a one-quarter-cell size. While this problem is slightly too large for the memory constraints of Sonnet Lite, it was performed in the full-featured version of Sonnet and confirmed expectations that the results are converging.

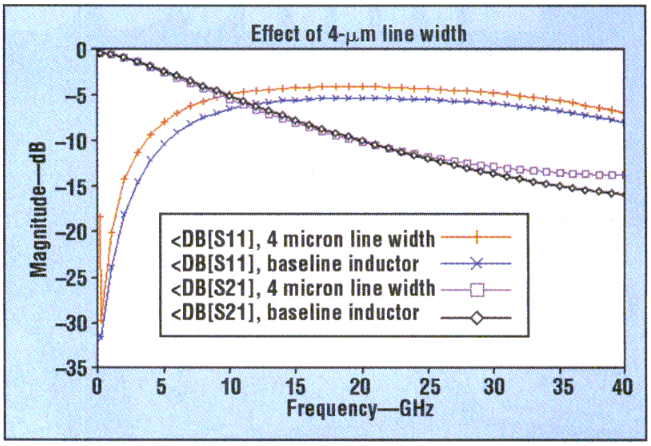
There is one exception to the ±0.3-dB error bound—the lowest-frequency data point for the forward trans-



6. The line width was increased from 8 to 12 μm to evaluate the effect on inductor loss.



8. Decreasing the line width from 8 to 4 μm results in this geometry.



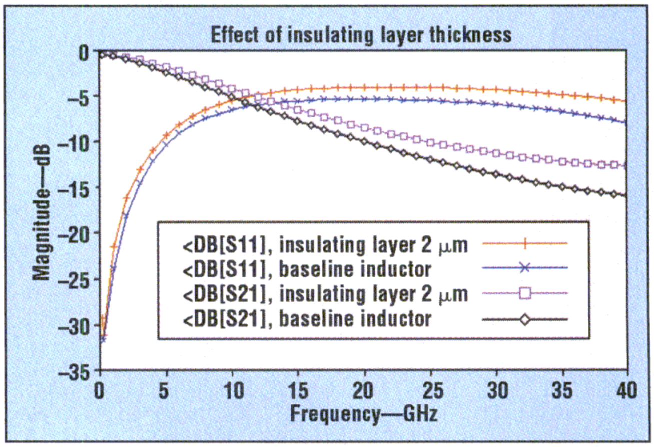
9. Decreasing line width from 8 to 4 μm results in the desired decrease in loss.

mission (S_{21}) at 100 MHz appears to be incorrect.

This can happen in any EM analysis when the cell size is small compared to the wavelength. The 200-MHz point is much better than the 100-MHz point, but still appears to have some error. Due to their smaller wavelengths, frequencies that are above 200 MHz do not pose a problem. The low-frequency data are included here to show that no EM analysis can be trusted completely. Some form of convergence test should be applied in order to check the accuracy of the software.

Before attempting to reduce loss, it is helpful to perform an analysis with loss that is completely removed. A comparison of this lossless analysis with the baseline lossy inductor should provide an upper limit on how much improvement remains for the lossy inductor. In the lossless analysis, all losses (metal and substrate) were removed from the inductor model of Fig. 2. Figure 5 compares this lossless analysis with the original lossy results.

This comparison reveals approximately 7-dB loss in S_{21} and S_{11} at 40 GHz. Certainly, there is room for improvement. By comparing S-parameter data for the lossless case, the baseline case, and any case under consideration, the merits of each alternative can be quickly determined.



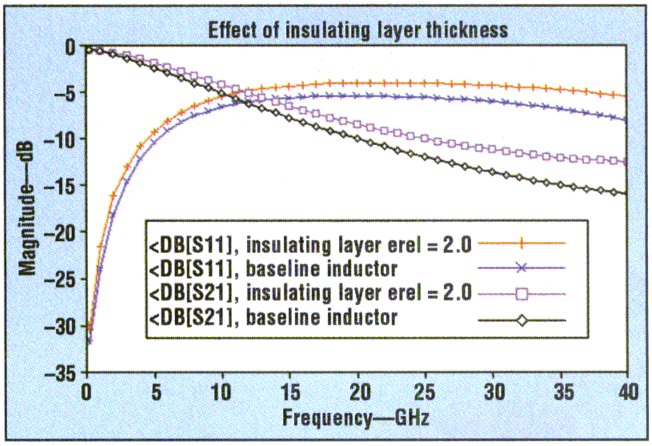
10. By increasing the thickness of the insulating layer on top of the Si substrate from 1 to 2 μm, it is possible to reduce the inductor loss substantially.

Note that there is no need to calculate inductor quality factor (Q) which, for the complicated equivalent circuits possible with planar spiral inductors, is neither uniquely defined nor simple to calculate.

Reasoning that wider line widths should yield lower loss, modifications to the baseline inductor of Fig. 2 begin by increasing the line width from the original 8 μm to a width of 12 μm, with a 4-μm cell size for analysis (Fig. 6). Figure 7 shows the results of analyzing this modified inductor model. The analysis now requires 11 s/frequency. Note that the loss has generally increased. While the effect of conductor loss has undoubtedly gone down, the effect of substrate conductivity has increased. The loss drops unexpectedly around 40 GHz. The reason for this was not investigated, but may be due to a resonance above 40 GHz.

Since it did not help to cut losses by increasing line widths, perhaps it will help to decrease loss by decreasing the line widths from the nominal 8 μm to a new value of 4 μm (Fig. 8). Figure 9 shows the analysis results (at 5 s/frequency). In these results, it is apparent that the loss has decreased substantially. If desired, this process of reducing the line width and analyzing the results can continue until the optimum line width is found.

A spiral inductor on Si usually has a thin insulating layer (for example, SiO_2) deposited underneath it. This reduces the effect of substrate conductivity, especially at low frequency. To reduce the loss, the insulating layer can be made thicker and/or formed with a material having a lower dielectric constant. To investigate the effects of changing the insulator thickness, it can be increased from its nominal thickness of 1 μm to a new thickness of 2 μm, then analyzed (Fig. 10). Following this, the dielectric constant of the original 1-μm-thick layer can be changed from 4.0 to 2.0 and then analyzed (Fig. 11). Both actions substantially reduce inductor loss at all frequencies. Now, armed with the knowledge of the relative importance of each parameter with respect to loss, and with knowledge of manufacturing and design constraints, a designer has a variety of options for realizing the lowest possible inductor losses.

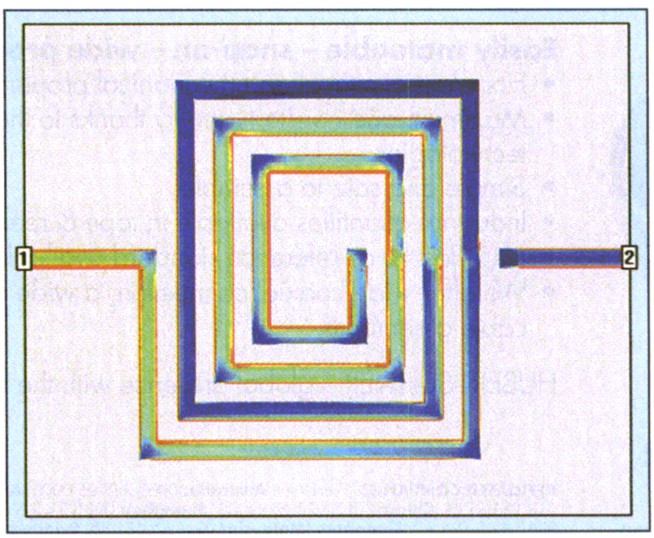


11. By decreasing the dielectric constant of the thin insulating substrate on top of the Si substrate from 4.0 to 2.0, the inductor loss is reduced substantially.

To gain some further insight into this particular spiral inductor design, the inductor of Fig. 2 was analyzed with a small cell size of 0.5 μm. As a result, each line of the inductor is now 16 cells wide. Due to the size of this problem, the full-featured Sonnet software suite was used in the analysis, rather than Sonnet Lite. The resulting current distribution is shown in Fig. 12. Note that there is some disruption of the current distribution at the location of the underpass connection to port 2. This is common in underpass and overpass situations and is real rather than a numerical artifact.

In Fig. 12, the color red represents a current density of approximately 1200 A/m. Port 1 is excited with a 1-V source at 40 GHz connected in series with a 50-Ω resistor. Port 2 is terminated in an impedance of 50 Ω. The analysis requires more than 12,000 subsections and 305-MB RAM. Since the PC used in this investigation has only 256-MB RAM, substantial memory swapping increased the analysis time to approximately 8 h. Normally, with adequate RAM for the problem, about one hour of analysis time would be expected.

Although free of charge, Sonnet Lite software is quite effective for solving a difficult and troublesome problem—the analysis of a spiral inductor on a conductive Si substrate. Techniques were shown for modifying the inductor in order to reduce inductor loss and characterizing the analysis error of an EM investigation quantitatively. Richardson extrapolation was shown to be an effective method for reducing analysis error even further without resorting to the full Sonnet suite of programs. **Sonnet Software, Inc., 1020 Seventh North St., Suite 210, Liverpool, NY 13088; (315) 453-3096, FAX: (315) 451-1694, e-mail: info@sonnetusa.com, Internet: http://www.sonnetusa.com.**



12. The current distribution for the baseline spiral inductor was analyzed with a very-fine 0.5-μm cell size. The disruption near the underpass connection to port 2 is not a numerical artifact but a real phenomenon.

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Acknowledgment
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(continued from p. 89)
ters and a few extra components over a traditional dual-band architecture.

The goal of the frequency plan is to define a transceiver that will work for all frequencies in the three targeted frequency bands of a world GSM radio, preferably using a single-band RF VCO. By default, the radio will also be able to operate as a traditional dual-band transceiver.

A proposed solution for a triple-band GSM radio incorporates dual conversion with an IF in the 250-MHz range. The IF LO frequency was chosen at two times the desired IF. A divide-by-two brings the LO IF into the 500-MHz range and prevents the IF synthesizer from leaking to the IF amplifier input, avoiding a large DC offset. The RF synthesizer frequency range was chosen for high-side mixing in the cases of GSM-1800 and GSM-1900, resulting in an RF LO frequency range of approximately 2050 to 2250 MHz. A selectable divide-by-three circuit at the RF VCO produces an RF LO signal usable for GSM-900 low-side mixing. Using the same RF LO frequency range for the transmit downconversion mixer results in transmit IFs from 340 to 465 MHz for GSM-1800 and 200 to 340 MHz for GSM-1900.

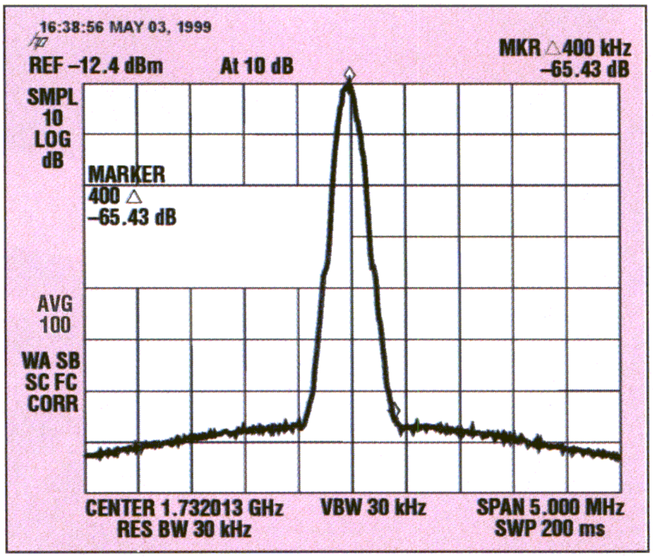
It is desirable to run the wideband PLL at a high phase-comparison frequency. This allows the loop to track the modulation easily, and ensures a low-divide ratio in the PLL, thus reducing in-band PLL phase noise (which must be better than –104 dBc inside the loop). By using a high comparison frequency, reference spurious products will be attenuated significantly by the loop filter, so the harmonics of the comparison frequency will be wider space and will not cause noise problems in the receive band. The loop filter must be wider than 600 kHz to ensure low-transmit phase error.

Given an IF LO frequency in the 500-MHz range, the possible comparison frequencies will be in the 250-MHz range (with a divide-by-two), in the 167-MHz range (with a divide-by-three), and in the 125-MHz range (with a divide-by-four). From the identified transmit IFs, it follows that the divide-by-three and the divide-by-four options are possible alternatives.

The GSM-900 transmit operation can be constructed either by introducing a divide-by-two, or by introducing a divide-by-four, resulting in IF ranges of 145 to 210 MHz or 353 to 367 MHz, respectively. The divide by two for GSM-900 forces a comparison frequency in the 167-MHz range, while a divide-by-four forces a comparison frequency of approximately 125 MHz. It should be noted that it is possible to change the IF between transmit and receive functions, leaving some flexibility in the transmit frequency plan. A swapping of the I/Q inputs/outputs (I/O) is required in the receive and transmit modes when changing from high-side to low-side mixing. Usually, this is orchestrated by the baseband processor, although it can be implemented directly in the transceiver at the baseband ports.

A single chip implementing the previously mentioned transceiver architecture has been implemented with a 0.5-μm BiCMOS process (Fig. 5). It is housed in an 81-pin CABGA package. The receiver includes a low-noise-amplifier (LNA) biasing circuit that can deliver optimum bias for three external LNA transistors. Desired input signals are converted to IF through an RF mixer, which actually consists of two mixers in parallel. One of these mixers processes the GSM-900 LNA output and the other handles the GSM-1800/GSM-1900 LNA output. This approach was chosen to ensure proper isolation between the GSM-900 and the GSM-1800/GSM-1900 mixers, and to ease the matching of the different bands.

A 13-MHz crystal oscillator is included on the chip. The crystal oscillator supplies a reference signal for the two frequency synthesizers as well as a sinusoidal 13-MHz output as the clock signal to the baseband processor. A programmable capacitor array is included for production-line tuning of the 13-MHz crystal oscillator.



6. This typical GSM-1800 transmit spectrum was measured for the GSM transceiver IC when used in a system application.

The IC has been characterized on a stand-alone basis and as a transceiver in a World GSM radio. A single-band external RF VCO was used with the RF PLL for the measurements, and a dual-band transmit VCO (GSM-900 and GSM-1800/GSM-1900) was used for the transmitter. Both the RF PLL's and the transmit PLL's charge pumps operate at +4.6 VDC while the rest of the circuits operate at +2.8 VDC. Figure 6 shows a typical transmit spectrum at the output of the transmit VCO. The measured phase error is approximately 3-deg. RMS. The RF PLL's phase noise dominates the RMS phase error, so the GSM-1800/GSM-1900 transmit-operation is the worst-case scenario. This is because the RF synthesizer phase noise is reduced when divided in GSM-900 mode. Measured receiver performance is summarized in Table 4. ••

Acknowledgments

The authors would like to acknowledge Paul D. Boyer, Bill Burdette, Yuko Kanagy, and David J. Green of National Semiconductor for their contributions toward this project.

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Table 1: Charting output frequency and power versus time

Time (min.)	Δf (kHz)	ΔP (dB)
0	0	0
5	–60	–0.50
15	–62	–0.50
30	–5	–0.60
45	–20	–0.65
60	–72	–0.67
75	–132	–0.60
90	–110	–0.67
103	–192	–0.60
130	–165	–0.67
140	–100	–0.60
160	–120	–0.60
200	–110	–0.60

Table 2: Checking output frequency and power versus temperature

Temperature (°C)	Δf (kHz)	ΔP (dB)
0	+120	+1.0
+5	+110	+1.0
+10	+100	+0.7
+15	+50	+0.4
+20	+20	+0.2
+25	0	0
+30	–10	–0.20
+35	–20	–0.20
+40	–130	–0.40
+45	–150	–0.50
+50	–165	–0.55
+55	–180	–0.60

(continued from p. 75)
sult, there are two resonant peaks, (f_1 and f_2) on the measured results of Fig. 3, where the resonant frequency of the high-Q cylinder is f_2 and the resonant frequency of the main resonant cavity is f_1. The peak value of f_2 is better than 50 dB. The Gunn diode was not biased during testing with the SNA; its junction capacitance will decrease with biasing ($V_d > 0$), although the resonance frequency of the main cavity will rise upon being biased.

The junction capacitance (C_d) of a Gunn diode can be found from:

$$C_d = (N_{oe}\varepsilon/2Vd)^2 A \qquad (18)$$

where:

$$V_d = 0.5(E_d - E_r)_d \qquad (19)$$

The total capacitance (C_t) can be found from:

$$1/C_t = 1/C_d + 1/C_p \qquad (20)$$

where:

C_p = the seal capacitance of the Gunn diode.

Without biasing, $V_d = 0$ and $C_d = \infty$. That is, $C_{t1} = C_p$. Without biasing, $V_d > 0$, and C_d is limited; that is:

$$C_{t2} = (C_d C_p)/(C_d + C_p) < C_p = C_{t2} \qquad (21)$$

The resonant frequency is:

$$f_0 = 1/[2\pi(LC)^{0.5}] \qquad (22)$$

and

$$C_{t2} < C_{t1} \qquad (23)$$

Thus:

$$f_{lv} = 0 < f_{lv} > 0 \qquad (24)$$

The EM fields within the main cavity are complex, with the Gunn diode and the biasing wire within the cavity. As a result, it is difficult to calculate $f_{lv>0}$ from the above equations. But it can be found by testing, and the difference of $f_{lv>0} - f_{lv=0}$ should be between 30 to 50 MHz. Once the output frequency of the Gunn oscillator is known, the resonant frequency of the high-Q cylinder, f_2, should be the same as f_0, provided that the height and diameter of the oscillator have been guided by eqs. 15 and 17. The resonant frequency of the main cavity ($f_{lv>0}$ is by 340 to 50 MHz lower than that of f_2, due to the positioning of the Gunn diode and the biasing wire, as well as the moved position of the tuning screw when testing with the scalar network analyzer without biasing. With bias applied to the Gunn diode, $f_{lv>0}$ will rise and coincide with f_2, and the output frequency stability and output power of the oscillator will be at their optimum (maximum) levels.

An HP 8562A spectrum analyzer and HP 11970A harmonic mixer from Hewlett-Packard Co. (Palo Alto, CA) were used to evaluate the Gunn oscillator's performance. By slightly shifting the positions of the Gunn diode and the biasing wire, the difference between f_1 and f_2 as a result of biasing can be changed slightly, to achieve higher oscillator stability at the required output frequency.

When the Gunn diode is biased with +4.5 VDC and 0.8-mA current at room temperature (+25°C), the output power and frequency of the Gunn oscillator are P_{out} = +16 dBm and F_0 = 36.800088 GHz. The phase noise of the oscillator is equal to or better than –100 dBc/Hz offset 100 kHz from the carrier and equal to or better than –90 dBc/Hz offset 50 kHz from the carrier.

Table 1 shows the relationships of the output frequency and the power as functions of time while Table 2 shows the relationships of the output frequency and output power as functions of temperature. When evaluated after operating for three hours over a temperature range of 0 to +55°C, the maximum frequency shift (Δf_{max}) was 300 kHz, with a resulting frequency stability of $\Delta f_{max}/f_0 = 8 \times 10^{-6}$ and output-power ripple, ΔP_{max}, of 1.6 dB.

In summary, it is possible to build a highly stable millimeter-wave Gunn oscillator using a properly dimensioned high-Q Invar resonant cylinder. The frequency stability of the Gunn oscillator can be improved further through preheating for several hours. The basic design can be applied at a wide range of millimeter-wave frequencies with maximum frequency variations of 300 kHz or less. ••

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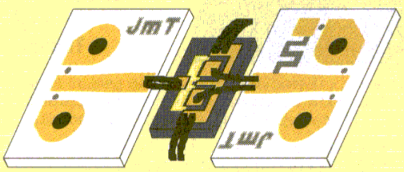

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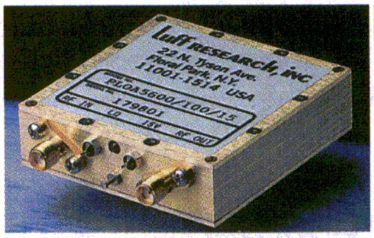

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Size:	2.25" x 2.25" x 0.65"

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FAX: (516) 358-2757
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LUFF RESEARCH — CIRCLE 577

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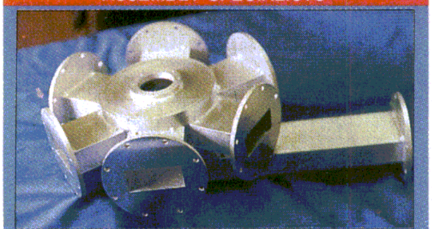

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Attenuator covers DC to 2 GHz

The model AD-280-25 gallium-arsenide (GaAs) integrated-circuit (IC), 5-b digital attenuator covers DC to 2 GHz with a total attenuation of 15.5 dB. Its bit values are 0.5, 1.0, 2.0, 4.0, and 8.0 dB, and its typical insertion-loss ranges from 1.1 to 2.2 dB. For frequencies above 0.5 GHz, its third-order intercept point (IP3) is +45 dBm. It is well-suited for applications requiring high attenuation accuracy, low insertion loss, and low intermodulation (IM) such as cellular radio, wireless data, and wireless local loop (WLL). **Alpha Industries, 20 Sylvan Rd., Woburn, MA 01801; (508) 894-1904, FAX: (617) 824-4579, Internet: http://www.alphaind.com.**

CIRCLE NO. 65 or visit www.mwrf.com

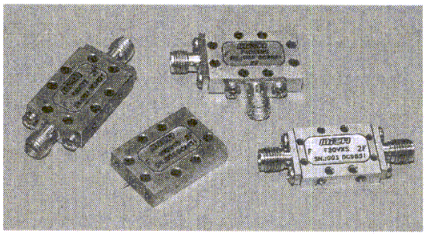
Dual-band diplexer simplifies installation

The model W9180 diplexer serves the 860-to-980-MHz band for Global System for Mobile Communications (GSM) and the 1690-to-1920-MHz band for distributed communication systems (DCS). It was designed to eliminate the need for two tower feed cables at the base station. The diplexer can handle a maximum continuous-wave (CW) input power of 40 W. It provides more than 60-dB isolation between the two bands. Insertion loss is less than 0.5 dB and return loss is better than –18 dB at all ports. The diplexer measures 3.0 × 4.0 × 1.5 in. (7.62 × 10.16 × 3.81 cm) and is available with SMA or N connectors. **Wireless Technologies Corp., 4000 Haile Lane, Springdale, AZ 72762; (501) 750-1046, FAX: (501) 750-4657, Internet: http://www.diplexers.com.**

CIRCLE NO. 66 or visit www.mwrf.com

Frequency doubler reaches 26 GHz

Model F90K accepts frequencies from 9 to 13 GHz and produces signals from 18 to 26 GHz with a typical conversion loss of 10 dB. Fundamental isolation is 25 dB. The frequency doubler works with input-power levels from +10 to +13 dBm. The doubler is available with in-line or right-angle port configuration, and hermetic or



non-hermetic topography. Packages include pin-pack, open-carrier-plate drop-in, surface-mount, or housings with SMA coaxial connectors. **Mica Microwave Corp., 7017 Realm Dr., San Jose, CA 95119-1312; (408) 363-9200, FAX: (408) 363-9220, Internet: http://www.mica-mw.com.**

CIRCLE NO. 67 or visit www.mwrf.com

Digitizer enhances ATE

The Catalyst and A5 family of automatic test equipment (ATE) is now available with a 1-GHz digitizer for multisite testing of high-frequency, system-on-chip (SOC) or mixed-signal devices such as 1000BaseT local-area networks (LANs) and 1.2-Gb/s, partial-response, maximum-likelihood (PRML) disk drives. The digitizer features 1-GHz differential bandwidth and 12-b resolution. It is designed with four differential or eight single-ended inputs for multisite testing. It also features a programmable termination voltage that enables users to test 100BaseT and 1000BaseT devices without transformers and capacitors on the device-interface board (DIB). **Teradyne, Inc., 321 Harrison Ave., Boston, MA 02118-2238; (617) 482-2700, Internet: http://www.teradyne.com.**

CIRCLE NO. 68 or visit www.mwrf.com

Tiny amps span to 26.6 GHz

The AGM-series of compact, self-contained amplifiers spans frequencies from 1.8 to 26.5 GHz. The smallest of the series is approximately the size of a postage stamp, measuring 0.50 × 0.78 × 0.22 in. (12.7 × 19.8 × 5.6 mm). They can be used as stand-alone amplifiers, either with the supplied SMA connectors or without the connectors in drop-in applications. Features include internal voltage regulation, reverse-voltage protection, and open/short-circuit protection. They are available in all of the popular communications and satellite bands as well as in octave bandwidths for other applications. Each band is available with single-, double-, or triple-gain stages providing up to 40-dB gain and output power to 1 W. For example, the two-stage, 13.75-to-14.5-GHz unit has 18-dB gain with ±0.75-dB flatness. Its output power is +20 dBm and it draws 160 mA from a +9- to +15-VDC power supply. Typical VSWR for these amplifiers is 2.0:1, although some models achieve 1.5:1. Operating temperature ranges from –55 to +85°C. **Atlantic Microwave Ltd., 40A Springwood Dr., Braintree, Essex CM7 2YN, England; 01376 550220, FAX: 01376 552145, Internet: http://www.atlanticmicrowave.co.uk.**

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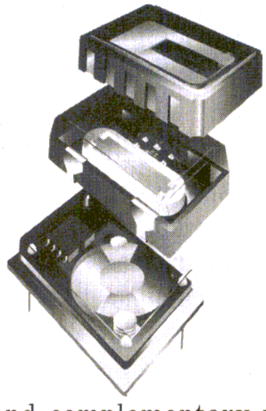
Oscillators sport new package

The EP1400SJ and EP1500SJ families of programmable oscillators are now available in a plastic surface-mount, J-leaded package measuring 9.8 × 14 × 4.7 mm. The oscillators are available in frequencies ranging from 1 to 125 MHz with a frequency stability of ±50 or ±100 PPM. Options include tri-state or power-down func-



tion and complementary-metal-oxide-semiconductor (CMOS) or transistor-transistor-logic (TTL) output. The EP1400SJ is a +5-VDC device while the EP1500SJ operates at +3.3 VDC for low-voltage applications. **Ecliptek Corp., 3545 Cadillac Ave., Costa Mesa, CA 92626; (714) 433-1200, FAX: (714) 433-1234, Internet: http://www.ecliptek.com.**

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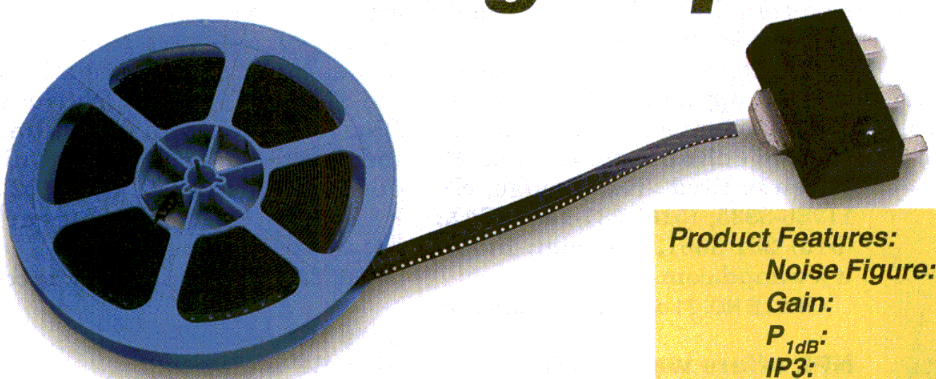
Now in Production Packaged pHEMTs!



Product Features:

Noise Figure:	*0.5 dB @ 1800 MHz**
Gain:	*15 dB @ 1800 MHz*
P_{1dB}:	*29 dBm @ 1800 MHz*
IP3:	*46 dBm*

Filtronic Solid State is now in production with Low Noise and High Dynamic Range *low-cost* SOT-89 packaged pHEMTs. Filtronic Solid State is offering Millennium-leading technology for the year 2000 and beyond for Base Station, Wireless Local Loop, and W-CDMA applications from 500 MHz to 4 GHz.

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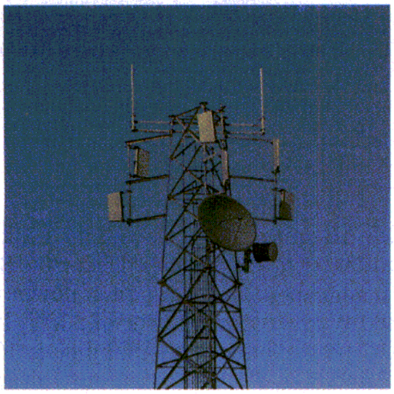
Model Number	Noise Figure	Gain	P_{1dB}	IP3
LP750SOT89	0.7 dB*	14 dB	24 dBm	40 dBm
LP1500SOT89	0.5 dB*	16 dB	27 dBm	44 dBm
LP3000SOT89	0.5 dB*	15 dB	29 dBm	46 dBm

*with optimum Noise Figure biasing

Base Stations



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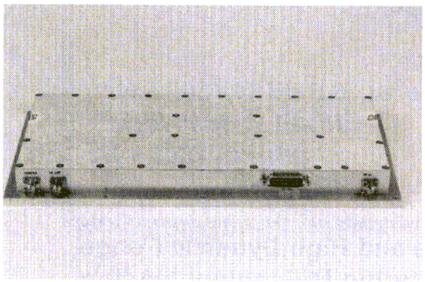
CIRCLE NO. 334

Amplifier covers PCS band

Model HPA1920-13 power amplifier (PA) covers the entire personal-communications-services (PCS) transmit band of 1930 to 1990 MHz with an average power of 13 W. The amplifier is designed to boost code-division-multiple-access (CDMA) carriers.



It has a gain of 40 dB and gain flatness of 1.5 dB peak-to-peak or 0.07 dB across a 1.23-MHz bandwidth. Input/output (I/O) VSWR is 1.5:1. The amplifier operates on a power-supply voltage of +26 to +29 VDC and draws a nominal current of 5 A at +28 VDC. Baseplate operating temperature range is 0 to +85°C. The amplifier provides circulator protection for transmission into any load mismatch, and monitors lines for major fault status. It is housed in a 12.0 × 5.8 × 1.1-in. (30.48 × 14.732 × 2.794-cm) package. Applications include single-channel tower tops and base stations. **MPD Technologies, Inc., 49 Wireless Blvd., Hauppauge, NY 11788-3935; (516) 231-1400, FAX: (516) 231-8081, Internet: http://www.mpd.com.**

CIRCLE NO. 71 or visit www.mwrf.com

NPL offers waveguide noise-measurement service

The United Kingdom's National Physical Laboratory (NPL) now offers microwave noise measurements in waveguide type WR22 (WG23) between 33 and 50 GHz. The test lab can measure noise temperatures from 77 to 105 K ±1.1 to 1.5 percent (excess noise ratio to 25 dB ±0.05 to 0.07 dB). This service complements a well-established range of RF and microwave calibration services including noise in other waveguide sizes, complex noise parameters of amplifiers, and phase noise. The new service will be of particular interest to systems integrators operating in the 33-to-50-GHz band. **National Physical Laboratory, Teddington, Middlesex TW11 0LW, United Kingdom; 0181 977 3222, FAX: 0181 943 6458, Internet: http://www.npl.co.uk.**

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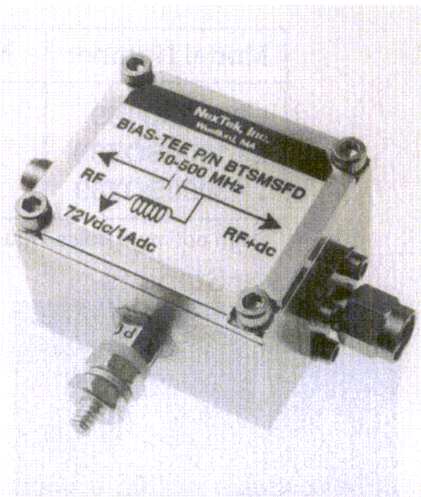
Bias tee features high RF isolation

The model BTSMSFD bias tee operates from 10 to 500 MHz with a minimum RF/DC isolation of 50 dB. It is specifically designed to enable DC injection or retrieval on a coaxial line, but can also be used for DC blocking. Maximum insertion loss is



0.2 dB, maximum VSWR is 1.15:1, and intermodulation (IM) is –80 dBc. It can handle an average RF power of 300 W and a peak power of 2 kW. Its DC power-handling capability is 100 V and 3 A, with a maximum voltage drop of 300 mV at 3.75 A. The weatherproof tee measures 1.5 × 1.25 × 1.0 in. (3.81 × 3.175 × 2.54 cm) and can operate at temperatures from –40 to +100°C. Applications include biasing of amplifiers and analog/digital circuits or general lab use. **NexTek, Inc., 439 Littleton Rd., Westford, MA 01886; (978) 486-0582, FAX: (978) 486-0583, Internet: http://www.ultranet.com/~nextek.**

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If you're a wireless or portable, engineering or design professional and have something to share with the professional community, you're eligible to submit an abstract for consideration!

How Speakers Are Selected

Every year we survey the industry to identify our audience's most current issues. Papers, topics and speakers are then selected by the conference advisory committee based on content, originality, and timeliness. Once selected, we work closely with speakers to make sure our sessions are right on target. We will not duplicate programs given at other shows. Each presentation must be original, and intended to inform, not sell.

Types of Sessions Offered

Wireless/Portable offers three types of sessions:

Paper Presentation Sessions: These sessions are led by a "session chair" and usually include 4-5 research papers that focus on a general topic. The authors make approximately 30-minute presentations based on their papers.

Mini-Tutorial "Expert" Sessions: These sessions are presented by one expert instructor on a concise topic, a case study, a narrow discipline, or "tips and tricks". Sessions are 1 hr. to 1.5 hrs. in length.

Workshop Tutorial Sessions: 1 or 2-day sessions featuring in-depth, discussions, presented by an expert instructor, and require separate or VIP registration for attendance.

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The Wireless/Portable Conference is divided into tracks. Prospective speakers should indicate the most appropriate track for their paper or topic.

- Cellular, Cordless & PCS Design
- Digital Communications Systems
- Modulation Techniques
- Wireless Data Transmission
- Wireless Local Area Networks
- Point-to-Point and Base Station Design
- High Power Design
- Wireless IC Solutions
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- Satellite Systems
- Test & Measurement Solutions
- Embedded Internet Systems
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- Software Radio/Advanced Wireless Systems
- Portable-System Software Issues
- Systems, Buses and Architectural Issues
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- Mobile and RF System Packaging Technologies
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1. Name, title, company, address, phone/fax, and e-mail address.
2. A short professional biography (50 words max.)
3. Your proposed paper/session title, and a 50-word abstract.
This material must be included or your submission will not be considered.
4. Indicate Session Type (Paper Presentation, Mini-Tutorial, or Workshop and appropriate Conference Track.)
5. If and where you are speaking in the next 18 months.
All submissions will receive a response (please allow four weeks.)
Conference schedule will be completed in November.

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Penton Media, Inc. must approve all co-presenters. The co-presenter must meet the same criteria as the presenter.

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No further expenses will be reimbursed.

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Frequency multipliers

A 52-page catalog features an assortment of passive and active multipliers. Doublers, triplers, and higher-order products are listed. Complete specifications are provided. **MITEQ, Inc.;** *(516) 436-7400, FAX: (516) 436-7430, Internet: http://www.miteq.com.*

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RF design

An RF-design handbook focuses on high-voltage, high-power, and metal-oxide-semiconductor-field-effect-transistor (MOSFET) technology. Data sheets for all of the newest high-voltage RF MOSFETs, Simulation Program with Integrated Circuit Emphasis (SPICE) models, and application notes for uses up to 81.36 MHz are included. **Advanced Power Technology;** *(800) 522-0809, FAX: (541) 388-0364, e-mail: custserv@advancedpower.com, Internet: http://www.advancedpower.com.*

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Microwave connectors

A four-page brochure describes a line of custom and standard RF/microwave coaxial-connector products. Adapters, receptacles, high-frequency connectors, cable connectors, and interface gauges are described. **SRI Connector Gage Co.;** *(407) 259-9688, FAX: (407) 259-9681, e-mail: info@sriconnectorgage.com.*

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Capacitor assemblies

Power-capacitor assemblies are overviewed in a 12-page brochure. Performance advantages, typical applications, capabilities, configurations, special test options, and component specifications are offered. Performance data are provided. **American Technical Ceramics;** *(516) 622-4700, FAX: (516) 622-4748, Internet: http://www.atceramics.com.*

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Low-noise VCOs

A catalog features low-noise voltage-controlled oscillators (VCOs) and low-phase-noise synthesizers for wireless communications applications. Specifications and outline drawings are provided. **Princeton Electronic Systems, Inc.;** *(609) 275-6500, FAX: (609) 799-7743, Internet: http://www.pesinc.com/~pesinc/.*

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Power-supply test

AC-to-DC and DC-to-DC power-supply test systems, test-system software, power-supply test systems for unipolar DC-to-DC converters, a voltage-regulator-mode (VRM) adapter card, power-supply measurement boards, electronic-load controller boards, and test manifolds are described in a 24-page catalog. Interface boards and multichannel analog-to-digital and a digital input/output (I/O) board are also featured. A glossary of power-supply tests and terms is provided. **ELTEST;** *(800) 701-9347, (508) 339-8210, FAX: (508) 337-4789, e-mail: sales@eltest.com, Internet: http://www.eltest.com.*

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Frequency synthesizer

A 16-page catalog examines products for RF/frequency synthesizer solutions. Direct-digital-synthesizer (DDS) and phase-locked-loop (PLL) products, including a digital chirp synthesizer, low-cost personal-communications-services (PCS), synthesizers, a fractional-N evaluation board, and static microwave prescalers are offered. Specifications and features are provided. **Osicom Technologies, Inc.;** *(888) OSICOM-8, (619) 558-3960 (outside the US), FAX: (619) 558-3980, e-mail: info@osicom.com, Internet: http://www.osicom.com/products/rf.html.*

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Synthesis modules

An 18-page brochure overviews high-speed circuit simulation, schematic entry, PWB layout, inductive-capacitive (LC) filter synthesis, microwave-filter synthesis, and active-filter synthesis. Oscillator synthesis, match and amplifier synthesis, phase-locked-loop (PLL) synthesis and simulation, as well as export and interfaces are discussed. **Eagleware Corp.;** *(770) 939-0156, FAX: (770) 939-0157, e-mail: eagleware@eagleware.com, Internet: http://www.eagleware.com.*

CIRCLE NO. 81 or visit www.mwrf.com

Integrated circuits

A 12-page brochure covers integrated circuits (ICs). Application-specific ICs (ASICs), application-specific standard products (ASSPs), standard ICs, and power-management ICs are described. Specifications are included. **Ricoh Corp., Electronic Devices Division;** *(408) 432-8800, FAX: (408) 432-8375.*

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Crystal oscillators

Fixed-frequency and voltage-controlled crystal oscillators (VCXOs) are presented in a 122-page catalog. New generations of extended temperature-range oscillators, low-jitter oscillators, low-height surface-mount-device (SMD) units, and oscillator-building application-specific integrated circuits (ASICs) are covered. Application information is included. **MF Electronics;** *(914) 576-6570, FAX: (914) 576-6204.*

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Test equipment

AC voltage sources; amplifiers; analyzers; calibrators; component testers; data-communications equipment; frequency counters/timers; generators; as well as light, laser, and optical equipment are provided in a 52-page catalog. Directional couplers, multimeters, voltmeters, oscilloscopes, and temperature-measuring devices are also listed. Pricing information is included. **Test Equipment, Inc.;** *(800) 336-7723, FAX: (707) 995-7151, e-mail: sales@naptech.com, Internet: http://www.naptech.com.*

CIRCLE NO. 84 or visit www.mwrf.com

RF switching

A 126-page catalog offers programmable attenuators, fixed attenuators and terminations, manual variable attenuators, RF coaxial switches, power dividers, and test accessories. Outline drawings are included. A model-number index is provided. **JFW Industries, Inc.;** *(317) 887-1340, FAX: (317) 881-6790, e-mail: sales@jfwindustries.com, Internet: http://www.jfwindustries.com.*

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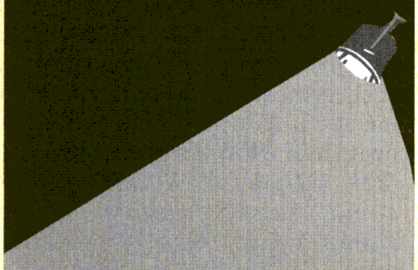

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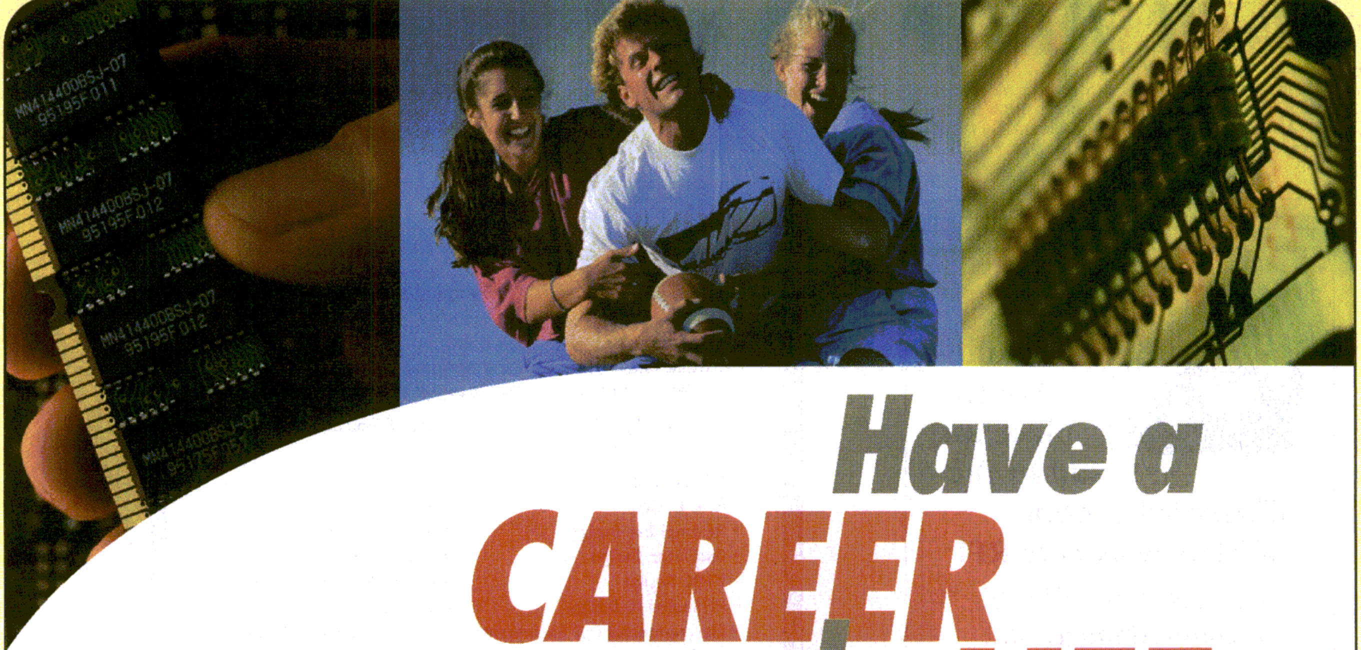

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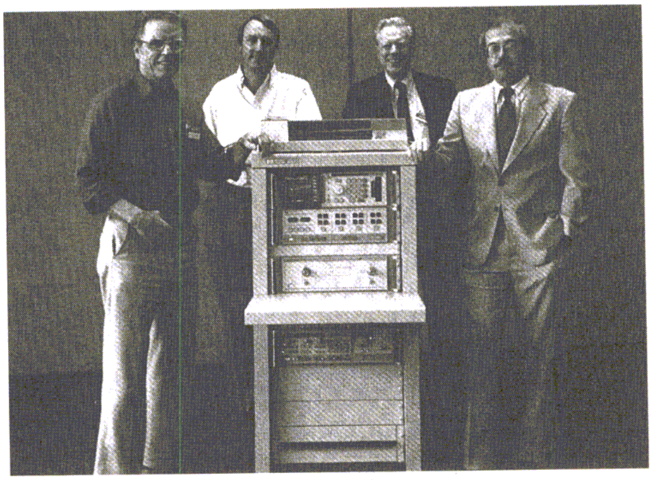

Approximately 14 years ago, the lead news story and cover detailed the shortcomings of this country's educational system to prepare students for a career in microwave engineering. In special cases, such as Hewlett-Packard's grant of an HP 8510 vector analyzer to Cornell University (Ithaca, NY), industry helped pave the way for future educational successes.

Microwaves & RF October Editorial Preview

Issue Theme: Integrated Circuits

News

Canada is home to majestic mountain ranges and untamed wilderness. It is also home to a growing microwave industry with technology centers in Montreal, Toronto, Ottawa, and Vancouver. The lead News Report in October will explore the capabilities of Canada's high-frequency industry, and highlight some of the key individuals who are helping the high-frequency industry to grow north of the border.

Design Features

Design Features in October support the issue theme of integrated circuits (ICs) from software and hardware standpoints. Software-based articles describe techniques for developing small-signal silicon-germanium (SiGe) device models and methods for extracting large-signal device parameters. Hardware-based articles show how to measure and set signal levels in high-speed digital systems, and how to design GaAs ICs.

Product Technology

In keeping with the silicon-germanium (SiGe) and Canadian themes, a Product Feature will highlight a SiGe power amplifier (PA) for Global System for Mobile Communications (GSM) applications. Developed in Ottawa, the 2.4-GHz device offers 45-percent efficiency. Additional product articles will examine a new line of phase-locked loops (PLLs) and a choke capable of supporting circuits to 8 GHz.