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Microwaves & RF

Integrated Circuits Issue

NEWS
Microwaves on the Canadian Plains

DESIGN FEATURE
Create small-signal SiGe HBT models

PRODUCT TECHNOLOGY
SiC MESFETs deliver clean RF power

Wideband choke biases amplifier circuits to 8 GHz

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ULTRA BROADBAND MIXERS

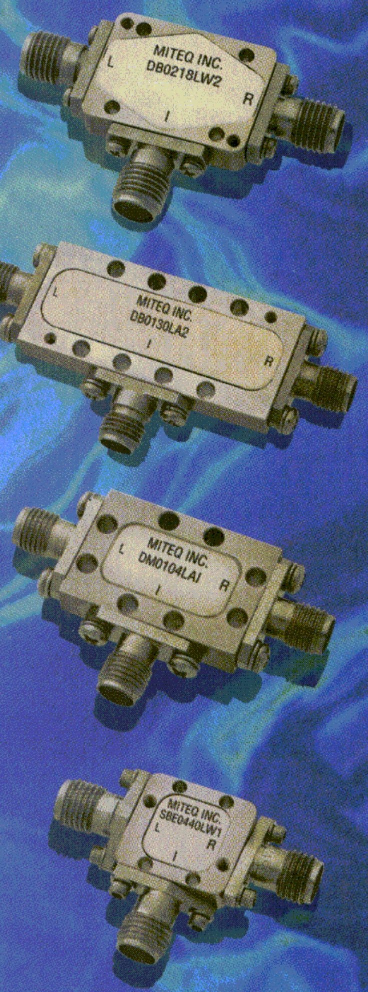
MODEL NUMBER	FREQUENCY RANGE RF/LO (GHz)	IF (GHz)	CONVERSION LOSS (dB, Typ.)
DB0130LA2	1 to 30	DC to .5	8.5
DB0218LW2	2 to 18	DC to .75	6.5
DB0218LA1	2 to 18	DC to .75	6.5
DB0226LA1	2 to 26	DC to .5	6.5
DB0426LW1	4 to 26	DC to 2	7.5
DB0440LW1	4 to 40	DC to 2	9
TB0218LW2	2 to 18	0.5 to 8	7.5
TB0218LA1	2 to 18	0.5 to 8	7.5
TB0226LW2	2 to 26	0.5 to 8	10
TB0440LW1	4 to 40	0.5 to 20	10

HIGH ISOLATION MIXERS

MODEL NUMBER	FREQUENCY RANGE RF/LO (GHz)	IF (GHz)	LO-RF ISOLATION (dB, Typ.)
DM0052LA2	0.5 to 2	DC to 0.5	40
DM0104LA1	1 to 4	DC to 1	40
DM0208LW2	2 to 8	DC to 2	40
DM0408LW2	4 to 8	DC to 2	40
DM0412LW2	4 to 12	DC to 4	40
DM0812LW2	8 to 12	DC to 4	35
DM0520LW1	5 to 20	DC to 8	35

SUB-HARMONIC MIXER

MODEL NUMBER	FREQUENCY RANGE RF (GHz)	LO (GHz)	IF (GHz)
SBE0440LW1	4 to 40	2 to 20	DC to 1.5



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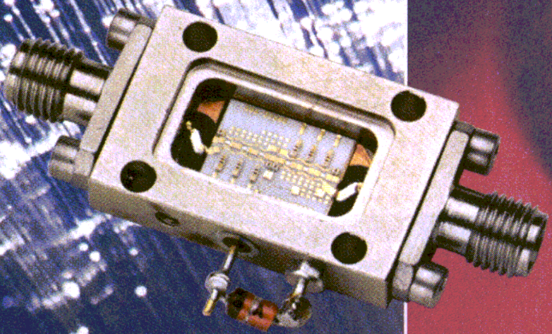

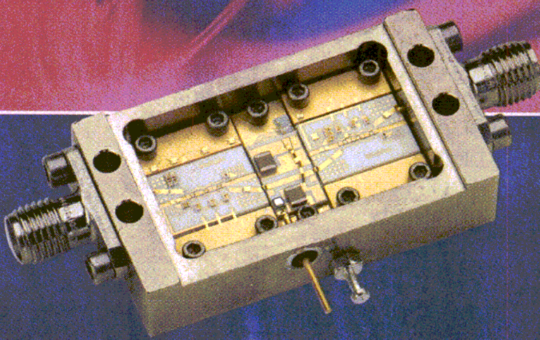

MODEL NUMBER	FREQ. RANGE GHZ	GAIN dB MIN	NOISE FIG. dB MAX	GAIN FLATNESS +/-dB	1dB COMP. PT. dB MIN	3RD ORDER ICP TYP.	VSWR IN/OUT MAX	DC CURRENT MA
JCA008-201	.01-8.0	25	*5	2.0	0	10	2.0:1	175
JCA008-202	.01-8.0	24	*5	2.0	5	15	2.0:1	200
JCA008-203	.01-8.0	22	*5	2.0	10	20	2.0:1	225
JCA008-301	.01-8.0	35	*5	2.5	0	10	2.0:1	300
JCA008-302	.01-8.0	34	*5	2.5	5	15	2.0:1	325
JCA008-303	.01-8.0	32	*5	2.5	10	20	2.0:1	350
JCA0010-201	.01-10.0	24	*5	2.0	0	10	2.0:1	175
JCA0010-202	.01-10.0	22	*5	2.0	5	15	2.0:1	200
JCA0010-203	.01-10.0	20	*5	2.0	10	20	2.0:1	225
JCA0010-301	.01-10.0	34	*5	2.5	0	10	2.0:1	300
JCA0010-302	.01-10.0	32	*5	2.5	5	15	2.0:1	325
JCA0010-303	.01-10.0	30	*5	2.5	10	20	2.0:1	350
JCA0012-201	.01-12.0	23	*5	2.0	0	10	2.0:1	175
JCA0012-202	.01-12.0	21	*5	2.0	5	15	2.0:1	200
JCA0012-203	.01-12.0	20	*5	2.0	10	20	2.0:1	225
JCA0012-301	.01-12.0	33	*5	2.5	0	10	2.0:1	300
JCA0012-302	.01-12.0	31	*5	2.5	5	15	2.0:1	325
JCA0012-303	.01-12.0	30	*5	2.5	10	20	2.0:1	350
JCA0118-201	.1-18.0	22	**5	2.5	3	13	2.0:1	200
JCA0118-202	.1-18.0	20	**5	2.5	5	15	2.0:1	250
JCA0118-203	.1-18.0	20	**5	2.5	7	17	2.0:1	300
JCA0118-301	.1-18.0	31	**5	2.5	3	13	2.0:1	250
JCA0118-302	.1-18.0	29	**5	2.5	5	15	2.0:1	300
JCA0118-303	.1-18.0	29	**5	2.5	7	17	2.0:1	350

* Noise Figure is specified above 300 Mhz
** Noise Figure is specified above 500 Mhz

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Microwaves & RF

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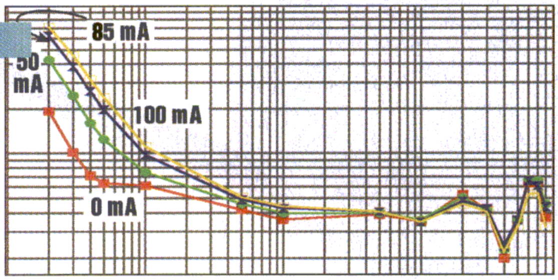


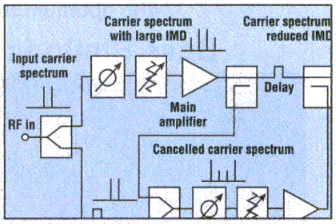
DEPARTMENTS

NEWS

SPECIAL SECTION

DESIGN FEATURES

PRODUCT TECHNOLOGY



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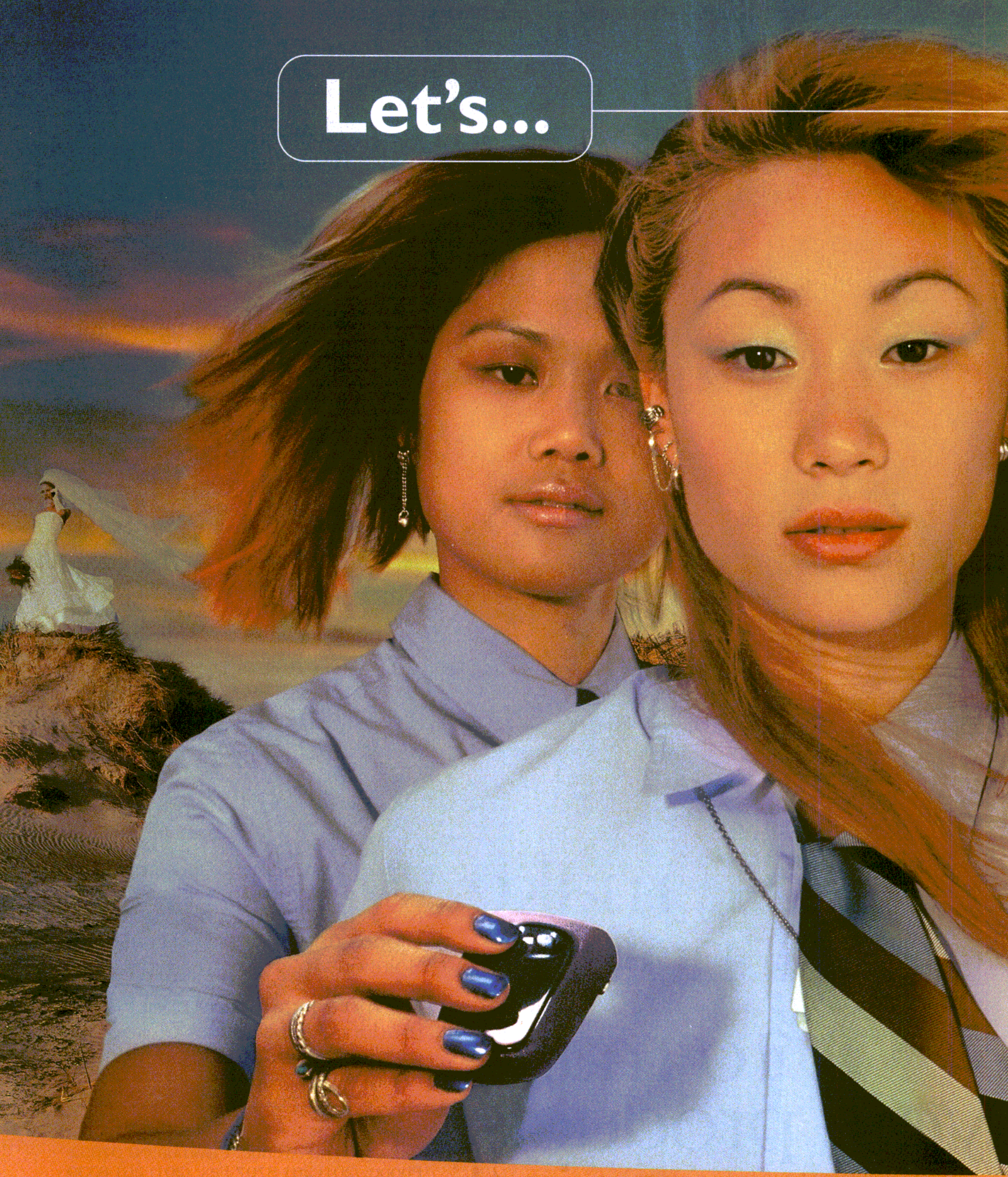

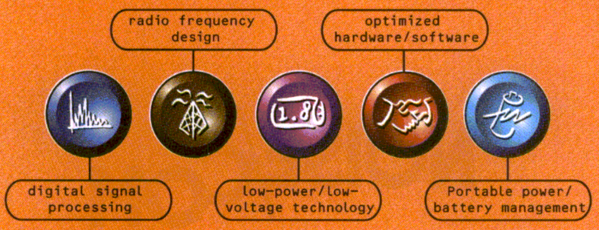

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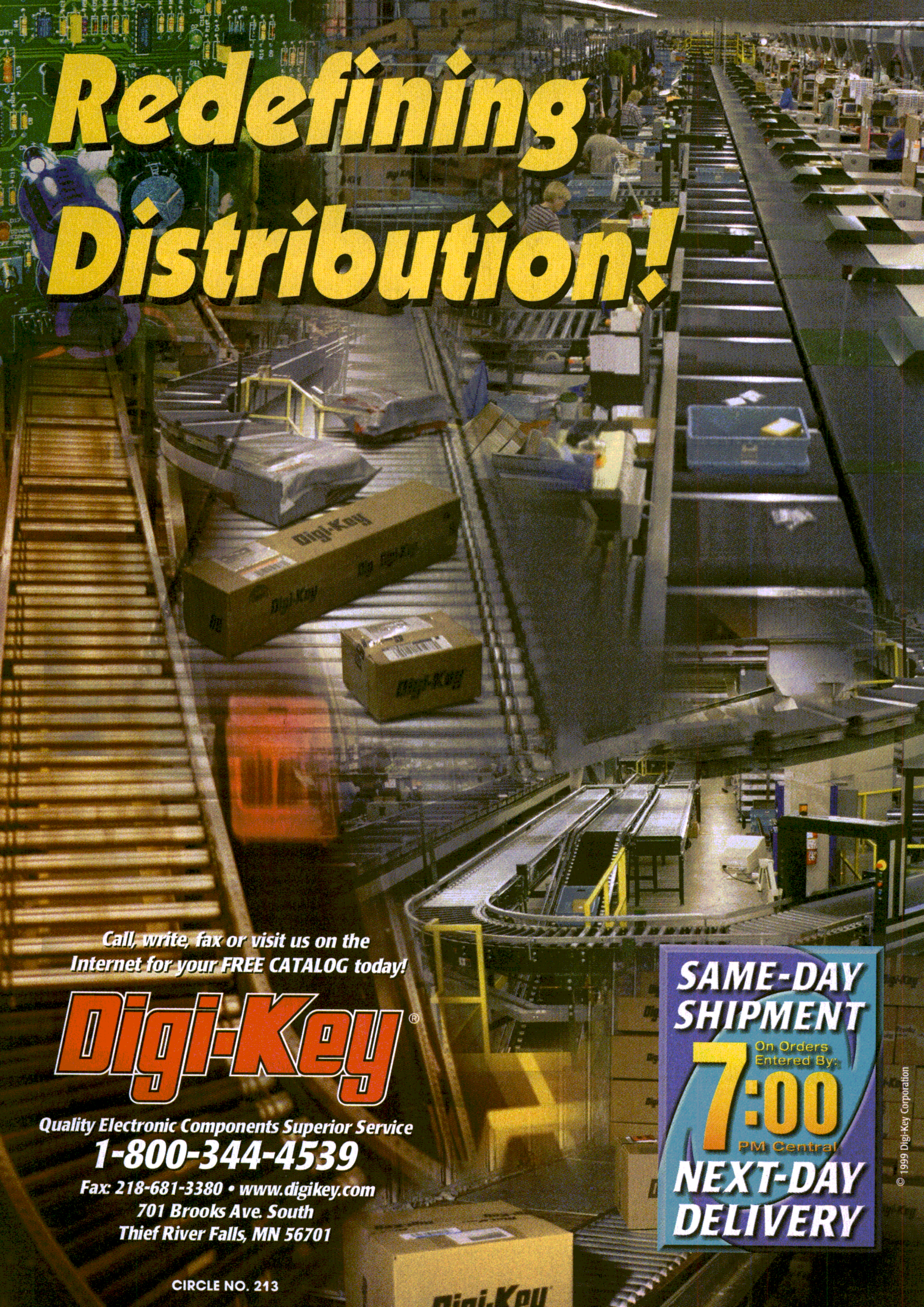
in communications



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CIRCLE NO. 246

READER CLARIFICATION

To the editor:

We have revised material pertaining to my article, "Fractional-N Synthesizers Meet HSCSD And GPRS Demands" which appeared in the August issue (p. 105). We would like the following information to be added for reader clarification:

1. Equation 7, which appears in the second column on page 109, is incomplete. The correct equation should read:

$$L(\omega) = (\Phi_{PLL\_RMS} / 2)^2 / BW$$

with

$$\Phi_{PLL\_RMS} = \Phi_{PLL} / \sqrt{2}$$

Please note that the brackets for ω were missing and that the divide-by-2 was also written incorrectly.

2. Next, in the first column of page 119, the incomplete paragraph that was continued from the previous page includes sentences that are out of order. The second-to-last sentence, "As with parameter T_{tb}, this minimum requirement will only be used when adjacent-cell power measurements are not required by the service selected," should be moved to the end of the paragraph as it is an explanation of, "Term T_{rb} related to the necessary time to the MS to get ready to receive."

3. The last change involves the third and fourth equations, which appear in the first column on page 121. The equations should be revised to:

$$t_{ra} = (T_{ra} - T_{monitor}) / 2$$

$$t_{rb} = T_{rb}.$$

Thank you for addressing these points.

Christian Wuensch
RF System Engineer
Philips Semiconductors

DSL TECHNOLOGY

To the editor:

Last month, one of your readers wrote to ask about *Microwaves & RF*'s plans to write articles on up and coming 21st century technologies such as high-definition television (HDTV). Well, here is another subject that could have an impact on the way we communicate in the coming years—digital subscriber line (DSL). This technology has many variations such as asymmetric DSL (ADSL), G.Lite (a subset of ADSL), high-speed DSL (HDSL), and splitterless ADSL, known as UDSL. As these technologies proliferate and become complex technically and economically, it becomes difficult for those of us who are not directly involved to keep abreast of developments in the industry. For example, ADSL has had only limited success in terms of home installations, yet it seems to have significant potential for high-speed Internet access, something many home users desire.

Hal L. Cassidy
TEXCOM
Fort Worth, TX

Silicon Germanium (SiGe), Stanford Microdevices' latest RF semiconductor process, offers benefits not attainable by conventional silicon-bipolar technologies: lower noise figures, lower power consumption, high output power at high efficiency, and high integration level.

As part of a series of Stanford Microdevices' initiatives to apply silicon germanium technology to a broad spectrum of communications applications, we are now offering the wireless market devices built using a patented silicon germanium manufacturing process. As these devices enter the marketplace, consumers will benefit from cellular phones, pagers, and other wireless communications devices that have extended battery life, carry out multiple functions, and are smaller, lighter, and less expensive.

PRODUCT SELECTION GUIDE

General Purpose Amplifiers

Part Number	Vd (V)	Id (mA)	3dB BW	P1dB (dBm)	IP3 (dBm)	Gain@ 1 GHz	Gain@ 2 GHz	NF 50 Ohm
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

SGA 2000 through 4000 series are also available in SOT-363
SGA 5000 and 6000 series are also available in SOT-89 & SOT23-5

BLUETOOTH MAY BE NEXT BIG IC CONSUMER

Internet access was the big buzz at last month's PCS '99 show in New Orleans, LA. Almost every major service provider and equipment supplier—including AT&T Wireless Services, Nortel Networks, Sprint PCS, Ericsson, Nokia, Lucent Technologies, and Qualcomm—heralded the availability of wireless Internet access and the continued displacement of wired communications by its wireless counterparts. Yet, for all the loud cheering surrounding wireless Internet connections, a quiet but steadily increasing roar could also be heard for Bluetooth.

Admittedly, wireless Internet access (and wireless data in general) was the major single focus of many booth displays and technical presentations last month (September 22-24) at the Ernest N. Morial Convention Center. For example, Ted Leonsis, President of the America-On-Line (AOL) Interactive Properties Group, appeared as a keynote speaker at PCS '99. He referred to wireless as a "sleeping giant" of the Internet and that the combination of wireless technology with the Internet may be one of the biggest opportunities in this or any lifetime.

While the technical session rooms were awash in projections of millions of subscribers and billion-dollar wireless Internet markets, the exhibit floor at PCS '99 was alive with demonstrations of software that would enable cellular telephones to reach directly into the World Wide Web. PIXO (Cupertino, CA) provided a convincing display of software that converts a cellular telephone into a mobile Internet device. Using clever display strategies to limit graphics content and selectively organize text from a website, the software represents a good first step toward wireless Internet access. The true mobile Internet-access solution, however, will not be a cellular telephone that has been adapted to the application but one that has been specifically designed for it.

To that end, the wireless Internet may offer a huge opportunity for some integrated-circuit (IC) manufacturers, such as makers of digital and baseband ICs, but not for others, such as RF ICs, since these are devices that would have been needed for the cellular-communications portion of the telephone anyway.

Bluetooth may just be that one new application area that drives RF semiconductor manufacturers for the next few years. Designed for personal-area connectivity and short-range communications in the home, office, and home office, Bluetooth (http://www.bluetooth.com) is meant to replace wire between electronic devices. Given that the specification's founding members Ericsson, IBM, Intel, Nokia, and Toshiba are now joined by over 1000 companies interested in manufacturing or supporting Bluetooth products, it may be difficult for the market not to succeed. And the potential for RF (and other) ICs in Bluetooth products is staggering. For a rough estimate, just think of how many electronic devices might one day connect to a home's main personal computer (PC) and multiply by the number of global households. ••

Jack Browne

Publisher/Editor

Microwaves & RF

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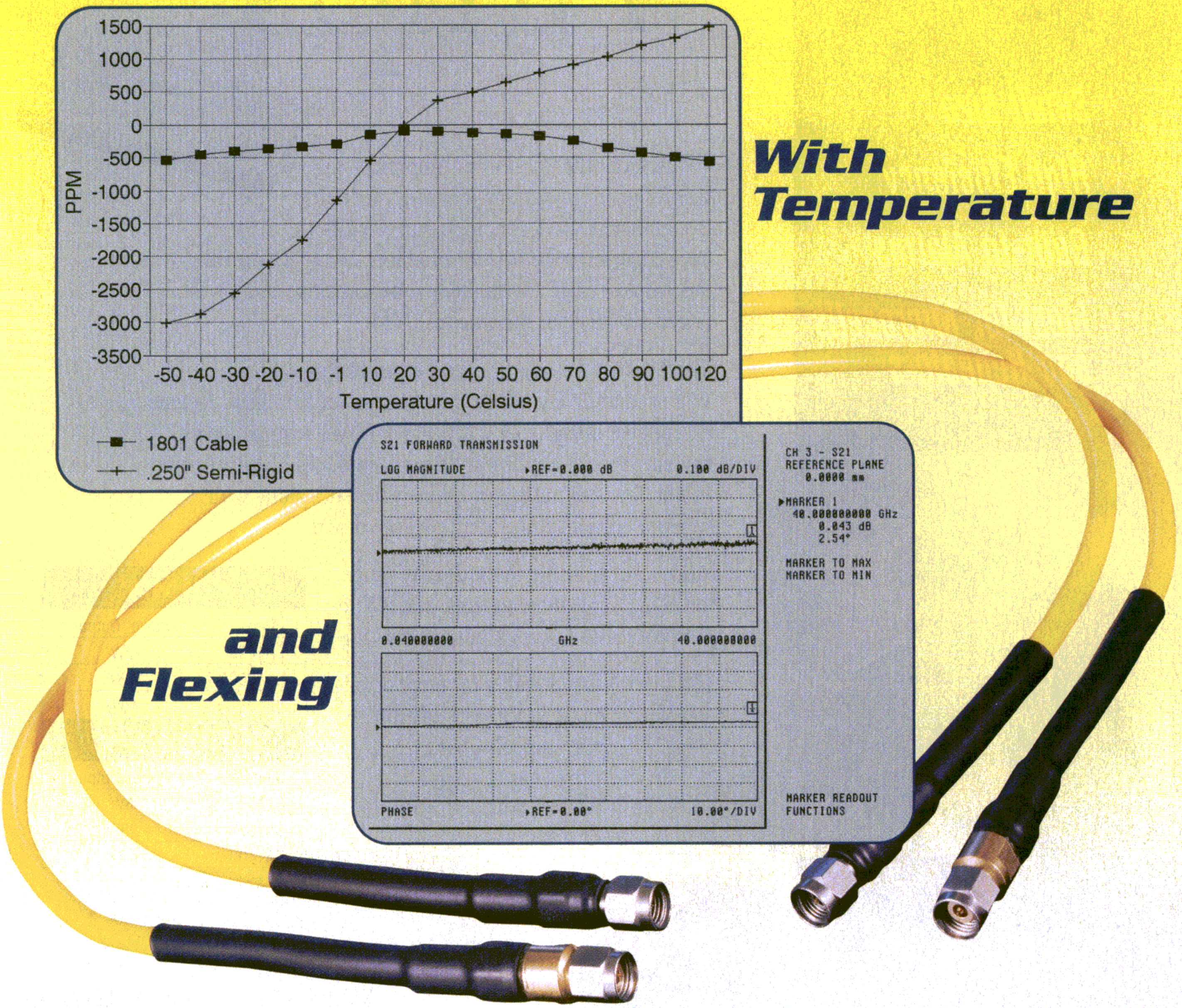

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CIRCLE NO. 228

Quartz-Crystal Industry Mourns Passing Of Juergen H. Staudte

CEDAR CITY, UT—The quartz-crystal industry is mourning the loss of Juergen H. Staudte, a pioneer in the quartz-crystal industry, who died on May 19, 1999 in New Mexico. He perished in the crash of a twin-engine plane that he was piloting.

Staudte revolutionized the quartz-crystal industry in the late 1960s when he combined the technologies of the quartz-crystal and the semiconductor industries to fabricate quartz crystals by photolithographic batch processing. By inventing this process, Staudte changed the way crystals were built, supporting production of low-cost, miniature tuning-fork crystals. His advances laid the foundation for the quartz-crystal watch business.

"The quartz-crystal industry—and science itself—has suffered a great loss in Mr. Staudte's passing," says E.L. Fox, president of Fox Electronics, a supplier of frequency-control products. "The processes he developed revolutionized and advanced the industry."

North Dakota Implements Data-Communications Plan

BISMARCK, ND—North Dakota Highway Patrol troopers are speeding toward the new millennium with computers on wheels.

Sixty-five patrol vehicles have been equipped with high-tech mobile data-communication systems. This system uses a laptop in a wireless environment with the aid of 10 tower sites that are located statewide. The installation of the system is the culmination of a detailed, four-year planning process involving leadership from the governor's office, approval of the Legislative Assembly, and citizen support. The project was accomplished with the assistance of Motorola as well as an inter-agency steering committee.

With this system, North Dakota becomes one of the first states in the country to implement a statewide mobile data-communications plan. Motorola has provided a complete data-communications solution for the first phase of the statewide mobile data-communications plan including hardware, software, and system installation.

"The substantial cost of approximately $10,600 per vehicle will be more than offset through the efficiency this system provides," says Colonel James M. Hughes. "This system will allow troopers to spend more time on the highway being visible and assisting the motoring public. Increasing the efficiency of service to the citizens of North Dakota is extremely important to us. I believe this system does just that."

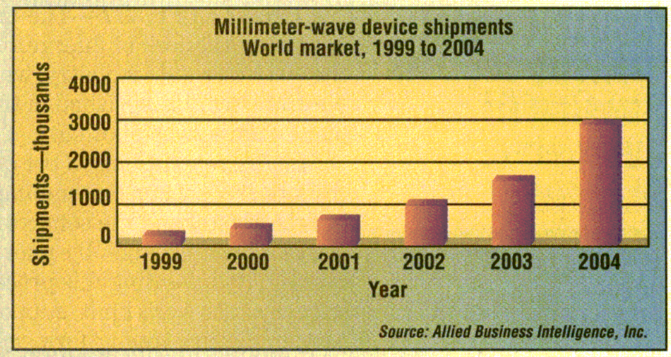
Microcosm Leads CAD Portion Of $14 Million Optical MEMS Project

RALEIGH, NC—Microcosm Technologies, Inc. recently announced that it will lead the computer-aided-design (CAD) portion of a $14 million, three-year optical micro-electro-mechanical-systems (MEMS) project. The project, titled "Micro-Opto-Electro-Mechanical Systems Manufacturing," has been funded by the US National Institute for Standards and Technology (NIST). Its goal is to develop and demonstrate technologies for efficient manufacturing of optical MEMS for telecommunications, imaging, medicine, entertainment, and information technology. Microcosm shares the NIST Advanced Technology Program (ATP) project with Xerox, Maxim Integrated Products, Optical Micro-Machines, Standard Microsystems Corp., and Microscan Systems, Inc. Xerox is the prime contractor on the project.

In the NIST project, Microcosm will extend its CAD software to include the modeling of optical components which are typically incorporated on optical MEMS devices. The modeling of optical components is inherently different from ray-tracing optical design at the macroscopic level. In microdevices, the structures that are fabricated are often not any larger than the wavelength of the light being manipulated. Thus, near-field and diffraction effects are important. Also, due to the high mechanical accelerations inherent in these devices, optical elements can warp dynamically, and this needs to be taken into consideration in a combined dynamic optical-electro-mechanical analysis. Microcosm is among the leaders in the MEMS CAD field in the development of software which can co-solve electromechanical problems.

LMDS Network Build Out To Spur Millimeter-Wave Device Market's Growth

OYSTER BAY, NY— The millimeter-wave device market will grow 10-fold by 2005, with the satellite end-user segment surpassing the local-multichannel-distribution-system (LMDS) customer premise equipment market by that year, according to a report from Allied Business Intelligence, Inc. (ABI). Total shipments of millimeter-wave devices are expected to grow from 314,000 in 1999 to slightly under 3 million in 2004 (see figure), according to the findings of "Millimeter Wave '99—Broadband Wireless and Automotive Radar Markets, Opportunities and Forecasts."



Traditional, point-to-point, and back-haul applications are expected to constitute the majority of the millimeter-wave shipments in 1999 with greater than 90 percent of the market. Less than 5 percent of the point-to-point millimeter radios sold are expected to be used to provide access to end users this year. This ratio will change dramatically during network build outs. Despite the connotations of its name, early LMDS deployments are expected to be primarily point to point.

Electronics' Factory Sales Reach $244 Billion In First Half Of 1999

*ARLINGTON, VA—*According to data recently released by the Electronic Industries Alliance (EIA), US factory sales of electronics equipment reached $244 billion, up 9 percent over the same period last year.

Highlighting the strength of this year's sales figures to date, EIA President Dave McCurdy says, "Virtually all sectors of the US electronics industry are doing well. Strong, consistent growth in electronics and related goods and services continues to drive the American economy forward."

McCurdy adds, "Our impressive growth reflects the fact that the American electronics industry is robust enough to weather the recent effects of the economic slump in Asia. The most impressive gains were in telecommunications with an outstanding 18-percent increase in sales over last year, and components, which showed strong increases on already high volume, reaching over $72.5 billion in sales."

Largest Nationwide Internet Survey Of Wireless Subscribers Is Launched

*SAN FRANCISCO, CA—*Wireless carriers have traditionally used small and expensive telephone or mail surveys to try to understand wireless subscribers' changing expectations, experiences, and the success of carriers' programs. Telephia has harnessed the power of the Internet by partnering with Harris Interactive to provide carriers with a far more powerful way to gather critical competitive information.

Telephia and Harris Interactive, an Internet market-research company and the sponsor of the renowned Harris Poll, have launched the Telephia/Harris Interactive Wireless E-Trac. This nationwide survey of more than 50,000 wireless subscribers will measure brand awareness, buying and usage habits, customer satisfaction, interest in new product offerings, and key factors affecting switching and loyalty for the 10 largest US markets.

"Harris Interactive's Internet Database, the largest data base of cooperative respondents in the world, and the company's 40-year history conducting market research, make Harris a valuable partner in helping Telephia deliver unparalled market research for wireless carriers," remarks Bill Bondurant, Telephia's Vice President of Client Services. "Our joint Internet research will provide multiclient, timely market-level research never before available."

Wireless E-Trac provides an all-in-one information solution. Today, carriers commission market research on multiple topics, sometimes only being able to afford surveying their own customers and not their competition. Through offering these products, Telephia will now be able to deliver a complete, integrated, cost-effective solution to all carriers in the market.

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Production Is Commenced At World's First 6-in. GaAs Analog Fab

WARREN, NJ— ANADIGICS recently announced that the company's top two customers have accepted and will receive production shipments of key high-volume products from its newly completed 6-in. (15.24-cm) gallium-arsenide (GaAs) fab. ANADIGICS announced plans for the world's first 6-in. (15.24-cm) GaAs analog fab (Fab 2) for integrated-circuit (IC) production earlier this year. ANADIGICS accelerated the development of the 6-in. (15.24-cm) fab in order to lower cost structures, improve efficiency, and increase capacity.

"This world-class facility enables ANADIGICS to better serve the needs of our customers by delivering high-quality products with shorter time to market," comments Dr. Bami Bastani, ANADIGICS' president and CEO. "Now, in addition to industry-leading RF IC design expertise, we offer true world-class operations that are ahead of the competition."

The new facility is currently being used for GaAs metal-semiconductor-field-effect-transistor (MESFET) and pseudomorphic-high-electron-mobility-transistor (PHEMT) processing. In addition, ANADIGICS is developing in-house heterojunction-bipolar-transistor (HBT) manufacturing capabilities, which they will implement in Fab 2. "We brought Fab 2 online in record time and ahead of original plans," remarks Bruce Diamond, senior vice president of manufacturing operations. "After passing all industry-standard reliability qualifications, we are very pleased to begin delivering production shipments from our new state-of-the-art 6-in. GaAs fab."

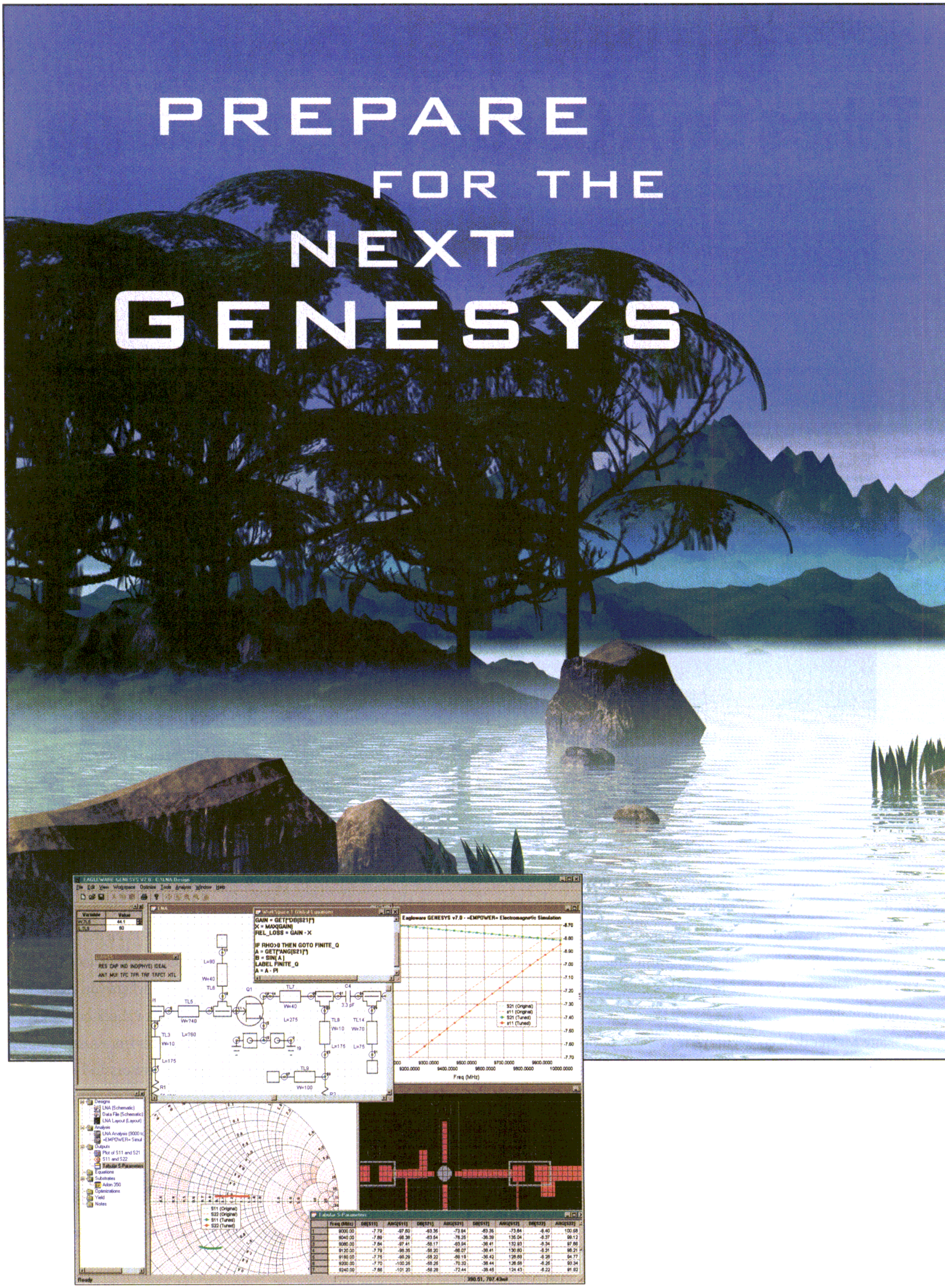
Fiber Deployment To Push Revenues To $25 Billion In 2003

*NEWPORT, RI—*The competitive-local-exchange-carrier (CLEC) industry in the US will increase from $4.4 billion in 1998 to $25 billion in 2003, according to a KMI Corp. report, *CLECs in the United States and Canada: Market Developments and Fiberoptic Systems Deployment*. This represents a 42-percent compound annual growth rate (CAGR) in CLEC revenue for the next five years.

CLECs fall into three broad categories—fiber based, non-fiber based, and wireless. Fiber-based CLECs—37 in the US at the end of 1998—deploy their own fiber to provide service. This category includes carriers' carriers that provide fiber to other CLECs. Non-fiber based CLECs may use digital-subscriber-line (DSL) technology to provide service over the incumbent local-exchange carriers' (ILECs) copper (Cu) loops, or lease ILEC and CLEC fiber to provide service. The third type of CLEC, the wireless CLEC, provides local service through wireless technology. Distinctions aside, all CLECs drive fiber development. Most non-fiber-based carriers must lease fiber, while those who use DSL technology drive bandwidth demand, increasing fiber deployment beyond the local loop.

Kudos

Motorola, AirTouch Communications, and CommNet Cellular, in partnership with the Wireless Foundation CALL TO PROTECT: Wireless Phones for Domestic Safety program, will donate 500 wireless phones and matching airtime to victims of domestic violence in Colorado. CALL TO PROTECT is a national educational and philanthropic program aimed at combating domestic violence nationwide...Tellium's Aurora 32 optical cross-connect has been chosen to receive the prestigious 1999 InfoVision award from the International Engineering Consortium (IEC). The award recognizes products and services based on their level of innovation, uniqueness, market impact, customer benefit, and value to society...Interstate Electronics Corp. (IEC), a Southern California manufacturer of cutting-edge Global Positioning System (GPS) receivers and display systems and an employer of more than 550 Orange County employees, recently announced that it will donate over 30 computers to two nonprofit schools in Corona, CA. The selected schools, based on recommendations from IEC employees, are Susan B. Anthony and Grace Lutheran School. The computers are configured to provide the students with easy and fast access to the Internet. Each school will receive 15 personal computers (PCs) to be delivered and installed at a later date at no cost...AT&T Corp. recently announced a $1.42 million donation to further technology education to underserved communities. More than half of the grant amount is allocated to organizations in Southern California. The contribution from AT&T is part of a larger program announced by President Clinton in a speech at the National Academy Foundation in Anaheim, CA.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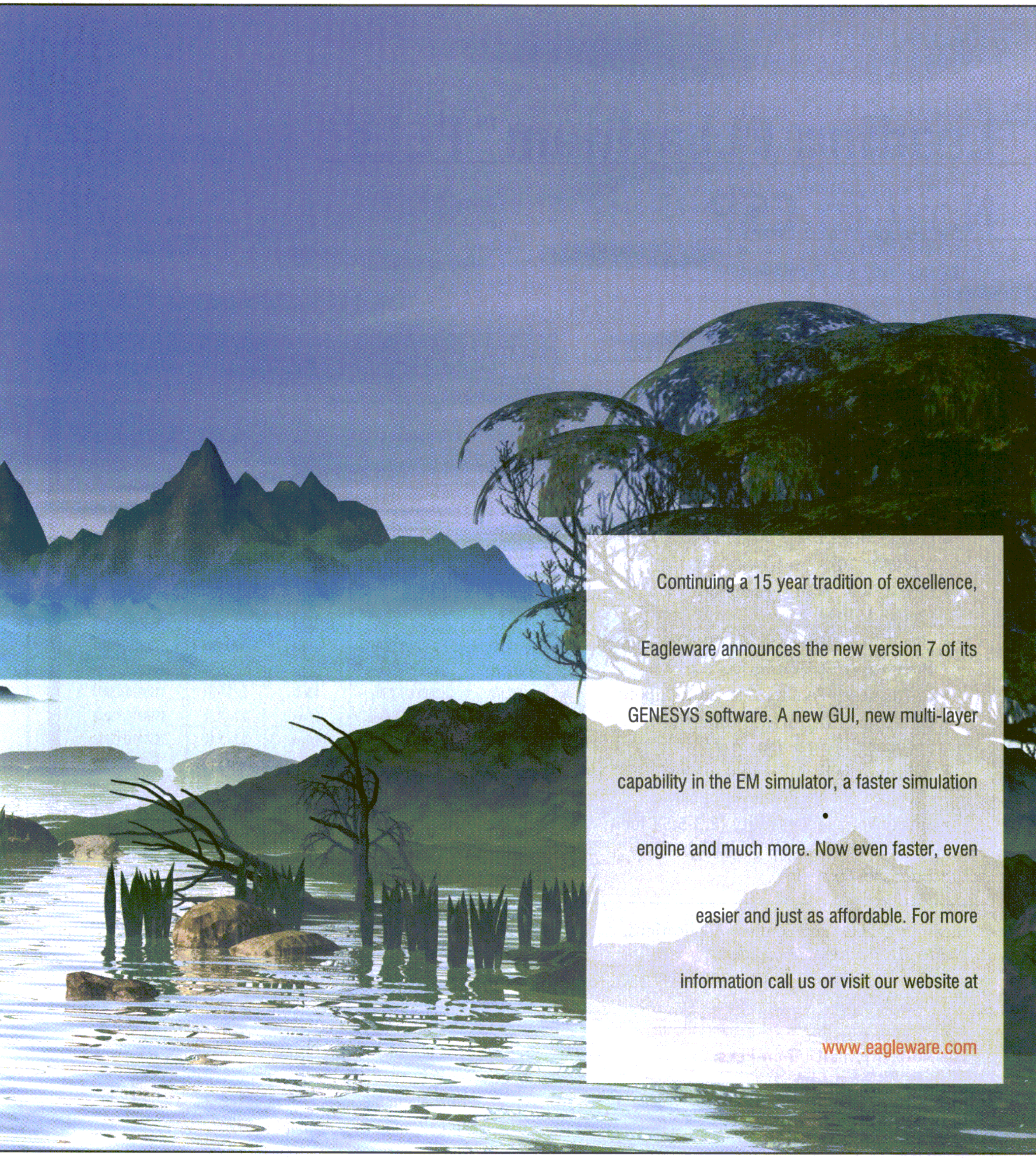

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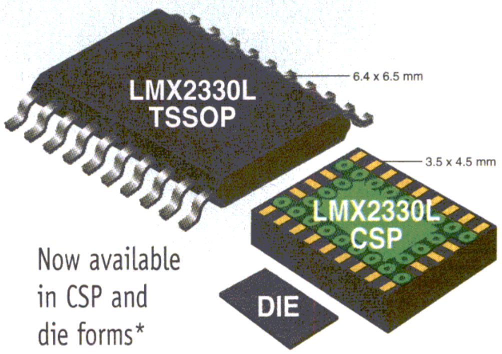
	LMX2331L	2.0GHz & 510MHz	4.0mA	2.7 - 5.5V	TSSOP20/CSP24
	LMX2332L	1.2GHz & 510MHz	3.0mA	2.7 - 5.5V	TSSOP20/CSP24
	LMX2335L	1.1GHz & 1.1GHz	4.0mA	2.7 - 5.5V	TSSOP16/SO16/CSP16
	LMX2336L	2.0GHz & 1.1GHz	5.0mA	2.7 - 5.5V	TSSOP20/CSP24
NEW	LMX2370	2.5GHz & 1.2GHz	6.0mA	2.7 - 5.5V	TSSOP20/CSP24
NEW	LMX2371	2.0GHz & 1.2GHz	5.0mA	2.7 - 5.5V	TSSOP20/CSP24
NEW	LMX2372	1.2GHz & 1.2GHz	4.0mA	2.7 - 5.5V	TSSOP20/CSP24
	LMX1600	2.0GHz & 500MHz	5.0mA	2.7 - 3.6V	TSSOP16/CSP16
	LMX1601	1.1GHz & 500MHz	4.0mA	2.7 - 3.6V	TSSOP16/CSP16
	LMX1602	1.1GHz & 1.1GHz	5.0mA	2.7 - 3.6V	TSSOP16/CSP16
	Fractional N PLLs				
	LMX2350	2.5GHz/550MHz	7mA	2.7 - 5.5V	TSSOP24/CSP24
	LMX2352	1.2GHz/500MHz	5.5mA	2.7 - 5.5V	TSSOP24/CSP24
NEW	LMX2353	2.5GHz	4.5mA	2.7 - 5.5V	TSSOP16/CSP16



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Canada Extends Borders Of RF Technology

JACK BROWNE
Publisher/Editor

CANADA is often mistaken for a northern extension to the US. But Canada is a country with its own strong identity, and with proud people. It is also a country with noteworthy technology from coast to coast.

Montreal is certainly Canada's most culturally diverse city, with residents claiming roots from around the globe. It is also one of the nation's most technologically diverse cities, with more than a dozen closely located firms, including Focus Microwaves, in the city's St.-Laurent section. The company's precision mechanical load-pull and noise-parameter impedance tuners are the basis for measurement systems capable of characterizing devices and components through 100 GHz. Founded in 1987, the firm offers more than 100 product and application notes on its website (http://www.focus-microwaves.com).

The firm's latest products include the minimum-loss test fixture (MLTF). It provides a mount for packaged high-power RF transistors, and an interface for connection between impedance tuners and vector network analyzers (VNAs). Usable to 5 GHz, the test fixture consists of a main body with coaxial connectors and connector flanges and a set of through-line-reflect (TRL) calibration standards.

In addition, the PMT line of prematching tuners permit high-VSWR load-pull testing. They use two independent RF probes to generate VSWRs up to 100.0:1 for load-pull testing of high-power transistors without need of quarter-wave transformers. Model PMT-1816-N, for example, is a PMT unit with type-N connectors. It is capable of operation from 1.6 to 18 GHz.

Also, the GPTS general-purpose RF and DC parameter test software is a Windows-based program designed for measuring the de-embedding RF and DC parameters of RF and microwave transistors and components under constant or swept input-power levels, frequencies, and bias levels. The test software helps obtain measured results that can be used in design and optimization computer-aided-engineering (CAE) packages, such as the Advanced Design System (ADS) from Hewlett-Packard Co. (Palo Alto, CA).

Software is also the strong suit of Infolytica Corp., located in downtown Montreal. Infolytica, founded in 1978, was actually the first company dedicated to the creation of affordable electromagnetic (EM)-analysis software. In contrast to its earliest two-dimensional modeling tools, the firm's latest iteration (Version 6.0) of the MagNet finite-element program is a powerful three-dimensional (3D) modeling tool that operates seamlessly within Windows 95/NT environments. It offers flexible scripting facilities and powerful parameterization capabilities, and enables mechanical as well as electrical-design engineers to visualize and analyze the most-complex 3D structures and circuits.

Also in St.-Laurent, GHz Technologies, Inc. offers extensive lines of coaxial couplers, circulators, isolators, and power dividers from 0.5 to 18 GHz, as well as low-loss waveguide filters, diplexers, and waveguide-to-coax adapters through 50 GHz for commercial and military customers. Many of the waveguide components are manufactured from Invar to achieve good frequency stability with temperature. Founded in 1985 by David Geller, the firm now numbers 105 employees with high-frequency products shipping into many wireless markets. Having achieved ISO9001 certification in 1994, GHz Technologies specializes in custom designs developed with quick turnaround times, due in part to the effectiveness of proprietary design software as well as a large bank of networked computer-numerically-controlled (CNC) milling and drilling machines for prototype of production fabrication.

TQF Technologies is next door to GHz Technologies, specializing in selective and multiplayer plating with a cumulative mechanical tolerance of less than 0.001 in. (0.0025 cm).

The company provides plating services for a number of Canadian and US electronics manufacturers, offering silver (Ag) plating, copper (Cu) plating per MIL-C-14550, electroless nickel (Ni) plating per MIL-C-26074, cadmium (Cd) plating, passivation, and baking to +220°C.

Future Electronics in Montreal is a major global distributor of electronic components and products, and is involved in more than 100 major markets. With 5500 employees worldwide, the company offers 24-hour service on its extensive product lines, including products from Motorola and Hewlett-Packard Co.

Montreal is also home to GaGe Applied Sciences, a company specializing in the marriage of computers and test equipment. Founded in 1987,

Canada's high-frequency firms at a glance

ADVANTECH Advanced Microwave Technologies, Inc., 657 Orly Ave., Dorval, Quebec H9P 1G1, Canada; (514) 420-0045, FAX: (514) 420-0073, Internet: http://www.advantech.ca.
CIRCLE NO. 94

Apollo Microwaves Ltd., 275 Hymus Blvd., Pointe-Claire, Quebec H9R 1G6, Canada; (514) 426-5959, FAX: (514) 426-5788, Internet: http://www.apollomw.com.
CIRCLE NO. 95

Avtech Electrosystems, Box 5120, Station F, Ottawa, Ontario K2C 3H4, Canada; (613) 226-5772, FAX: (613) 226-2802, e-mail: info@avtechpulse.com, Internet: http://www.avtechpulse.com.
CIRCLE NO. 96

CAL Corp., an Electromagnetic Sciences, Inc. Co., 1725 Woodward Dr., Ottawa, Ontario K2C 0P9, Canada; (613) 727-1771, FAX: (613) 727-1200, Internet: http://www.calcorp.com.
CIRCLE NO. 97

Com Dev Canada, (519) 622-2300, FAX: (519) 622-1691.
CIRCLE NO. 98

Communications & Power Industries (CPI) Canada, Inc., 45 River Dr., Georgetown, Ontario L7G 2J4, Canada; (905) 877-0161, FAX: (905) 873-7416, e-mail: marketing@cmp.cpii.com, Internet: http://www.cpii.com/cmp.
CIRCLE NO. 99

EEV Canada Ltd., (905) 678-9811, FAX: (905) 678-7726, e-mail: info@eevinc.com, Internet: http://www.eev.com.
CIRCLE NO. 100

Filtran Microcircuits, Inc., 2475 Don Reid Dr., Ottawa, Ontario K1H 1E2, Canada; (613) 737-0706, FAX: (613) 737-0495, e-mail: fmi@filtranmicro.com, Internet: http://www.filtranmicro.com.
CIRCLE NO. 101

Focus Microwaves, Inc., 970 Montee-de-Liesse, Suite 308, Ville St.-Laurent, Quebec H4T 1W7, Canada; (514) 335-6227, FAX: (514) 335-6287, e-mail: focusmw@compuserve.com, Internet: http://www.focus-microwaves.com.
CIRCLE NO. 102

Future Electronics, 237 Hymus Blvd., Pointe-Claire, Quebec, Canada H9R 5C7; (514) 694-7710, FAX: (514) 695-3707, Internet: http://www.future.ca.
CIRCLE NO. 103

GaGe Applied Sciences, Inc., 2000 32nd Ave. Lachine, Montreal, Quebec H8T 3H7, Canada; (514) 633-7447, FAX: (514) 633-0770, Internet: http://www.gage-applied.com.
CIRCLE NO. 104

GHz Technologies, Inc., 6602 Van Den Abeele, St.-Laurent, Quebec H4S 1Y3, Canada; (514) 335-6090, FAX: (514) 335-9297,
CIRCLE NO. 105

Infolytica Corp., P.O. Box 1144, Station Place du Parc, 300 Leo Pariseau, Suite 2222, Montreal, Quebec H2W 2P4, Canada; (514) 849-8752, FAX: (514) 849-4239, e-mail: info@infolytica.com, Internet: http://wwsw.infolytica.com.
CIRCLE NO. 106

ITS Electronics, Inc., 200 Edgeley Blvd., Units 24-26, Concord, Ontario L4K 3Y8, Canada; (905) 660-0405, FAX: (905) 660-0406, Internet: http://itselectronics.com.
CIRCLE NO. 107

Lap-Tech, Inc., 230 Simpson Ave. South, Bowmanville, Ontario L1C 2J3, Canada; (905) 623-4101, FAX: (905) 623-3886, Internet: http://www.laptech.com.
CIRCLE NO. 108

Mitec Telecom, Inc., 104 Gun Ave., Pointe-Claire, Quebec H9R 3X3, Canada; (514) 694-6666, FAX: (514) 694-3933, Internet: http://www.mitectelecom.com.
CIRCLE NO. 109

Optotek Ltd., 62 Steacie Dr., Kanata, Ontario K2K 2A9, Canada; (613) 591-0336, FAX: (613) 591-0584, e-mail: optotek@opteotek.com, Internet: http://www.optotek.com.
CIRCLE NO. 110

Philsar Electronics, Inc., 146 Colonnade Rd. South, Nepean, Ontario K2E 7Y1, Canada; (613) 274-0922, FAX: (613) 274-0915, Internet: http://www.philsar.com
CIRCLE NO. 111

Scientific Microwave Corp., 707 Lajoie Ave., Dorval, Quebec H9P 1G7, Canada; (514) 828-9212, FAX: (514) 828-9227, e-mail: Smc@smcq.com, Internet: http://www.smcq.com.
CIRCLE NO. 112

SDP Components, Inc., 222 Brunswick Blvd., Pointe Claire, Quebec H9R 1A6, Canada; (514) 428-8749, FAX: (514) 428-8757, e-mail: info@sdp.ca, Internet: http://www.sdp.ca.
CIRCLE NO. 113

SiGe Microsystems, Inc., 1500 Montreal Rd., M50 IPF, Ottawa, Ontario KIA OR6, Canada; (613) 748-1334, FAX: (613) 748-1635, Internet: http://www.sige.com.
CIRCLE NO. 114

TQF Technologies, Inc., 6640 Van Den Abeele, St.-Laurent, Quebec H4S 1Y3, Canada; (514) 333-3895, FAX: (514) 333-8576.
CIRCLE NO. 115

Wavesat Telecom, Inc., 4600 Rue Cousens, St.-Laurent, Quebec H4S 1X3, Canada; (514) 956-6300, FAX: (514) 956-8587, Internet: http://www.wavesat.com.
CIRCLE NO. 116

the firm's extensive product lines include deep-memory-oscilloscopes (DMOs)-analysis instruments capable of capturing 1 billion samples of data. With as many as eight channels, DMOs offer a sample rate of 10 MSamples/s with the capacity to grab 10 s of waveform data at that sampling rate. The DMOs work with 12-b analog-to-digital converters (ADCs) capable of 70-dB spurious-free dynamic range, although 8-b versions are also available. The instruments are ideal for hard-to-analyze signal problems in telecommunications, medical systems, radar systems, optical systems, and video equipment. The DMOs can also operate in a digital-sampling-oscilloscope (DSO) mode to make measurements using sampling rates faster than those available in the DMO mode. Models are available with sample rates from 5 to 100 MSamples/s and bandwidths to 250 MHz.

Apollo Microwaves Ltd., located in the Montreal suburb of Pointe-Claire, designs and manufactures standard and custom components and subsystems for telecommunications, wireless, and satellite-communications applications. The 25,000-sq.-ft. ISO9001 facility was established in 1981 and now has more than 115 employees. Product lines cover frequencies from S-band through Ku-band and include filters, duplexers, multiplexers, circulators, isolators, terminations, power combiners, switching networks, as well as monitor and control systems.

Also located in Pointe-Claire, Mitec Telecom operates two facilities with a total of 43,000 square feet. The firm, with more than 215 employees, also owns facilities in the United Kingdom and the US and has a controlling interest in a manufacturing facility in Thailand. Mitec Telecom's Canadian products include amplifiers, diplexers, filter subsystems, flexible waveguide, and waveguide components. Established in 1972, the company supports applications in terrestrial microwave-communications systems to 38 GHz, satellite-communications applications from C-band through Ku-band, as well as personal communications services (PCS) and cellular communications systems to 2 GHz.

SDP Components was formed in 1994 to provide high-quality coaxial connectors. The connectors are designed through CAE tools in the Pointe-Claire facility and built in China to minimize manufacturing costs. With approximately 15 employees in Quebec and another 55 workers in China, the firm can quickly make special custom connector designs on request. It offers all industry-standard connectors, such as type-N, TNC, BNC, and SMA connectors at a fraction of the cost of most leading suppliers.

Scientific Microwave Corp., located near Dorval Airport on the outskirts of Montreal, was founded in 1986 to design and manufacture microwave and millimeter-wave passive components. The company applies innovative design techniques to various evanescent-mode rigid waveguide and planar structures, for narrowband, wideband, low-power, and high-power applications. Product lines include RF-interference (RFI) filters, lowpass, highpass, bandpass, and band-reject microwave filters, equalizers, couplers, terminations, attenuators, circulators, and isolators.

ADVANTECH Advanced Microwave Technologies, Inc. (Dorval, Quebec, Canada) numbered only five employees less than eight years ago. Today the firm, which specializes in high-power solid-state amplifies, boasts 160 employees in a 60,000-sq.-ft. facility. The company supports a variety of fast-growing markets, including wireless-local-loop (WLL) and multichannel-multipoint-distribution-system (MMDS) applications. Digital MMDS is typically used for television-broadcast ("wireless cable") applications. The MMDS products include solid-state power amplifiers (SSPAs), frequency upconverters, transmitters, and a frequency-reference system.

ADVANTECH's satellite-communications amplifiers include rack-mount and hub-mount single and multiple amplifier systems for L-, S-, C-, X-, and Ku-band systems. Single-amplifier systems with power levels of 250-, 600-, or 700-W output power, and multiple-amplifier systems with 100-, 2400-, and 2800-W output power are offered. The firm's repeater systems are designed to extend the coverage of a CDMA PCS cell in areas where the rough terrain otherwise obstructs the wireless signals.

ONTARIO'S OFFERINGS

One of Ontario's better known high-frequency firms has earned a strong following in the US through their CAE software. Optotek, in Kanata, offers the powerful MMICAD suite of software tools for designing and analyzing high-frequency active and passive circuits. The low-cost software tools can accurately model passive components such as filters and couplers, or active components, such as amplifiers, as well as a complex assemblies consisting of several components. The firm recently announced Version 2 of its MMICAD SCHEMATIC/LAYOUT software which simplifies the development of schematic diagrams and supports a wide range of layout data files, including HPGL, GERBER, and AUTOCAD layout formats. The company's LASIMO software is a parameter-extraction package that aids in the creation of device modeling (see p.75).

One of the better (and lesser-known) success stories in Ontario is that of EMS Technologies (formerly CAL Corp.) in Ottawa. Part of the US firm, Electromagnetic Sciences, CAL Corp. has a wide range of expertise in analog, digital, microwave, power, and shielding design for terrestrial and satellite-communications applications. The company recently announced the 300th installation of its AMT-50 multichannel Inmarsat satellite-communications antenna. The AMT-50 satcom antenna operates as part of either the Honeywell MCS 3000/6000/7000 or Rockwell Collins SAT-906 satellite-communications avionics system. The antenna system, which provides 12-dBic gain from 1525 to 1660.5 MHz, was installed on board a Falcon 2000 jet aircraft for Dassault Falcon Jet Corp. (Little Rock, AK).

Lap-Tech, Inc. of Bowmanville, Ontario, manufactures precision quartz resonators and clock oscillators for military and commercial applications. Founded in 1972, the

11,500-sq.-ft. facility can provide AT-cut or SC-cut crystal blanks, crystals, or complete oscillators, depending on a customer's needs.

CPI Canada (formerly known as Varian Canada) has long been associated with satellite communications and high-power electron tubes. Located in Georgetown, Ontario, the firm's VKA 2400 series of extended interaction klystrons (EIKs) deliver power levels to 1 kW at Ka-band, with instantaneous bandwidths as wide as 1.2 GHz.

Recently, the company developed an innovative digital tuning system for its satellite-communications uplink klystrons. The system incorporates optical encoders and DC stepper motors, and is a noncontacting positioning system without discernible passband shift for the klyston during lifetime tests of more than 100,000 channel changes. The tuning system can be operated locally, through a high-power amplifier's (HPA's) controls, or remotely, with RS-232 or RS-422 connections. The digital tuning system provides 50 preset channels, full backup of settings in electronically erasable programmable read-only memory (EEPROM), and capability for in-field mechanical tuning as a backup.

Avtech Electrosystems (Ottawa, Ontario), in business since 1975, is a leading supplier of high-speed pulse generators and laser-diode drivers. The firm offers a line of high-voltage pulse generators, such as the AVL-2 series generators with output voltages up to 350 V for pulse rise times as fast as 2 ns and pulse widths from 5 to 100 ns. In addition, the AVRH series, with 1000-, 2000-, and 3000-V models provides pulse widths from 0.2 to 5 μs.

The model AVP-AV-1-C pulse source supports 100-ps rise/fall time and as much as 5-V output into a 50-Ω load. The pulse generator offers repetition rates from 0 to 1 MHz with pulse widths variable from 0.2 to 4.0 ns. Also, the model AVN-C pulse source features 100-ps rise/fall times, with pulse widths variable from 0.2 to 1.0 ns and repetition rates from 50 to 250 MHz.

Filtran Microcircuits, Inc. in Ottawa brings the precision of semiconductor processing to the world of circuit-board manufacturing. The company's 20,000-sq.-ft. facility enables the company to achieve 1-mil lines and spacings (held to tolerances of $\pm$0.2 mils) on soft-substrate circuit boards clad with Cu and other metals. Rather than using chemical etching for fabricating conductive through holes in the substrates, the firm employs a proprietary technique to form sputtered blind holes with extremely tight tolerances and very repeatable electrical characteristics.

The firm, recently acquired by Merrimac Industries (West Caldwell, NJ) for its ability to fabricate Merrimac's Multi-Mix multilayer high-frequency circuit technology, supports its circuit-board fabrication capabilities with a wide range of machining capabilities. Using combinations of milling, laser machining, and computer-numerically-controlled (CNC) routing machines, the firm can achieve machined tolerances as tight as $\pm$0.001 in. ($\pm$0.0025 cm)

Some of the more established Canadian companies have already given birth to several generations of additional companies. Wavesat Telecom, which started in 1993, grew out of ADVANTECH. Now employing more than 100 people, Wavesat Telecom is on the verge of ISO9001 certification. In addition to extensive development of feedforward amplifiers for high-linearity cellular and PCS applications, including next-generation cellular systems based on wideband CDMA (WCDMA) modulation techniques, the company has made investments in creating an offset-frequency-division-multiplex (OFDM) wireless modem for point-to-point and point-to-multipoint data-communications systems. The wireless modem is expected to support data rates to 25 Mb/s.

The company produces large volumes of HPA modules for satellite communications and point-to-point communications systems. Available at frequencies from L-band through K-band, the modules are supplied with overvoltage and output VSWR protection.

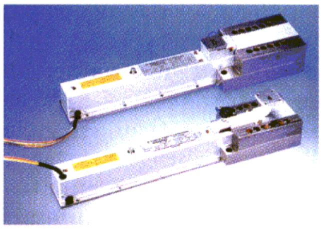
Philsar Electronics (Nepean, Ontario) is an innovative developer of low-power radio and frequency-synthesizer designs. The firm's PS-XX00 series of fractional-N frequency synthesizers is the first products of their kind to provide step resolution of finer than 100 Hz. The PS-XX00 family employs a high internal reference and delta-sigma technology to achieve excellent phase-noise performance –100 dBc/Hz for carriers to 2.5 GHz) and small step sizes. The PS-XX00 family is among the most highly integrated devices ever built on IBM's silicon germanium (SiGe) BiCMOS process (see "SiGe Technology Makes Practical Advances," p. 121).

At least one manufacturer in Canada has responded to the growing interest in wireless Internet access—ITS Electronics (Concord, Ontario). Founded in 1987 to design and manufacture microwave and millimeter-wave solid-state amplifiers, frequency converters, and subsystems in commercial and military markets, the company's Inter Wave 99 lines of products are aimed at high-speed Internet and wireless communications applications. The family of products includes a base-station transmitter unit, an Internet subscriber unit, and a dual-input base-station receiver unit. The base-station transmitter unit is designed to operate in any 200-MHz segment of the 2100-to-2700-MHz MMDS band. It provides 40-dB input-to-output gain with less than $\pm$1.5-dB gain variation and +36-dBm output power at 1-dB compression. Harmonics are less than –40 dBc at the output, while output spurious products are a miniscule –55 dBc. Output monitoring and control are provided with an RS-422 interface.

The Internet subscriber unit is compatible with DOCSIS cable modems and is factory set to operate in any 200-MHz segment of the 2100-to-2700-MHz MMDS band. It features 50-dB transmitter gain and 40-dB receiver gain, with a receiver noise figure of 5 dB.

Regrettably, not all of Canada's high-frequency companies could be covered in this brief report, and apologies to those who were omitted. But even this sampling (see "Canada's high-frequency firms at a glance") should provide some indication of the formidable capabilities that lie north of the US border. ••

High-efficiency TWTs power DBS uplinks

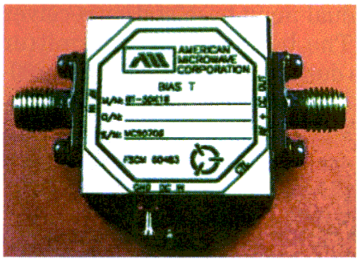


Two additions have been made to a new generation of traveling-wave tubes (TWTs)—models TH 3976D and TH 3977D. Designed specifically for uplink applications in direct-broadcast-satellite (DBS) systems, the TH 3976D delivers 270-W RF output power while the TH 3977D generates 500-W RF output power. Both TWTs are designed for use in the 17.3-to-18.4-GHz satellite band. The tubes feature dual-collector, conduction-cooled technology, making them well-suited for rack-mount and hub-mount high-power amplifiers (HPAs). The high-efficiency dual-stage collector makes it possible to place the tube closer to the antenna in a hub-mount configuration, increasing overall system efficiency and providing significant operating-cost savings. Models TH 3976D and TH 3977D are mechanically and electrically compatible with the company's 400-W, 13.75-to-14.5-GHz model TH 3976 and 750-W, 13.75-to-14.5-GHz model TH 3977 TWTs. **Thomson Tubes Electroniques, 18 avenue du Marechal Juin, 92366 Meudon la Foret Cedex, France; (33) 130703500, FAX: (33) 130703535, Internet: http://www.tte.thomson-csf.com.**

CIRCLE NO. 59 or visit www.mwrf.com

Broadband bias tee covers 50 kHz to 18 GHz

Sporting an impressive bandwidth of 50 kHz to 18 GHz (and usable to 15 kHz), model BT-50K18 is a wideband bias tee that simplifies the connection of DC power to wideband amplifiers and other active devices. The bias tee, which exhibits typical VSWR of 1.75:1, suffers an insertion loss of only 0.75 dB across its full bandwidth. The unit offers 60-dB typical RF-to-DC isolation from 10 MHz to 18 GHz. The bias port has a series resistance of 2.2 Ω and will handle DC voltages up to 15 V and maximum current of 200 mA. The component measures only 1.0 × 1.0 × 0.5 in. (2.54 × 2.54 × 1.27 cm) with SMA connectors, although a miniature version measuring just 0.80 × 0.62 × 0.40 in. (2.032 × 1.5748 × 1.016 cm) with removable SMA connectors is also available. **American Microwave Corp., 7311-G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, e-mail: amcpmi@aol.com, Internet: http://www.amwave.com.**



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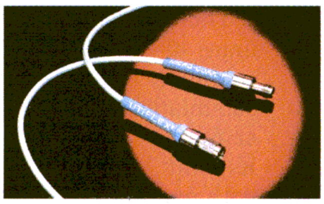
Cable assemblies minimize losses

The UFA series of UTiFLEX low-loss cable assemblies has been designed for low electrical loss in commercial wireless as well as military applications. The UFA series features a maximum insertion loss of 0.11 dB/ft at 1 GHz to 1.21 dB/ft at 50 GHz, depending on the choice of cable assembly. Phase stability as low as 2 deg. can be specified at 10 and 18 GHz. The five cable assemblies in the series include the DC-to-26.5-GHz UFA147B, UFA210A, and UFA210B, the DC-to-40-GHz UFA147A, and the DC-to-50-GHz UFA125A. The cables are integrated with a patented connector attachment that provides high reliability. The connector body, dielectric material, and center contact are completely captivated for low VSWR and high reliability. **MICRO-COAX, 206 Jones Blvd., Pottstown, PA 19464-3465; (610) 495-0110, FAX: (610) 495-6656, Internet: http://www.micro-coax.com.**



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Duplexer/receive filter screen PCS signals

Model WSDA-00019 is a full-band duplexer diversity receive-filter assembly for personal-communications-services (PCS) applications. It consists of a base-transceiver-station (BTS) fullband duplexer and a BTS fullband diversity receive filter. The duplexer has a receive passband of 1850 to 1910 MHz and a transmit passband of 1930 to 1990 MHz. The maximum insertion loss within the passband is 1 dB for the receiver portion and 0.8 dB for the transmit portion, while the passband return loss is at least 16 dB. The duplexer offers minimum passband-to-passband isolation of at least 65 dB with antenna-to-receive rejection of at least 50 dB from DC to 1790 MHz and at least 65 dB from 1930 to 1990 MHz and 50 dB from 1990 to 5000 MHz. The diversity receive filter has a passband of 1850 to 1910 MHz with a maximum insertion loss of 1 dB and passband return loss of at least 16 dB. The second-harmonic rejection is at least 50 dB from DC to 1790 MHz, 65 dB minimum from 1930 to 1990 MHz, and 50 dB minimum from 2090 to 5000 MHz. The full-band duplexer receive filter is rated for 20-W maximum CW power and 400-W peak instantaneous power. **K&L Microwave, Inc., 2250 Northwood Dr., Salisbury, MD 21801; (410) 749-2424, FAX: (410) 749-2788, e-mail: wireless@klmicrowave.com, Internet: http://www.klmicrowave.com.**

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AT&T Gives Ericsson The Boot

Plagued by problems that include a network outage, customer difficulties in completing calls, and slow equipment deliveries, AT&T Corp., the largest US cellular company, has ousted Ericsson AB as one of its network equipment suppliers. Stepping in to replace Ericsson will be Nortel Networks, which together with current supplier Lucent Technologies, will be the two top providers of the company. This is the first time that Nortel has sold wireless equipment to AT&T. Financial terms of the contract with Nortel were not announced.

For Ericsson, currently AT&T's major supplier of wireless network equipment, the loss means that instead of having approximately two-thirds of AT&T's business, the company will retain only one-third—a loss of 50 percent. AT&T spent approximately $385 million with Ericsson on infrastructure equipment in 1998 according to financial sources. The positive spin for Ericsson on the turn of events is that the lost business makes up only 1 percent of the company's total revenue. Not only that, but company executives in the network operations group feel that retaining one-third of AT&T's overall infrastructure spending still represents a substantial business.

AT&T believes that the new equipment it will purchase from Nortel and Lucent will increase the capacity of its network, improve reliability, and lower its costs by $900 million over the next four years. The equipment will enable AT&T to provide high-speed data services and permits customers to use their wireless phones while overseas.

AT&T said it will add equipment manufactured by Lucent in New York City, parts of Los Angeles, and in 11 states where Lucent equipment is installed. The Nortel equipment will go to Arizona, California, Nevada, Oregon, and Washington.

The trouble began for AT&T when its One Rate plan attracted many more customers than anticipated, clogging networks with telephone traffic volume that the networks were not meant to handle. Wireless revenues rose 42 percent in the second quarter, but the company experienced a major outage and had problems keeping up with demand. Apparently, the Ericsson-designed equipment could not keep up with the increased load, nor could Ericsson stay abreast of AT&T's equipment demands for expanding its network.

An unusual aspect of the AT&T/Nortel deal is that it is based on the amount of capacity added to the network, not the equipment it takes to add that capacity. Thus, if Nortel is able to build more-efficient networks—greater capacity for a particular equipment size—it will increase its profit margin in the arrangement. ••

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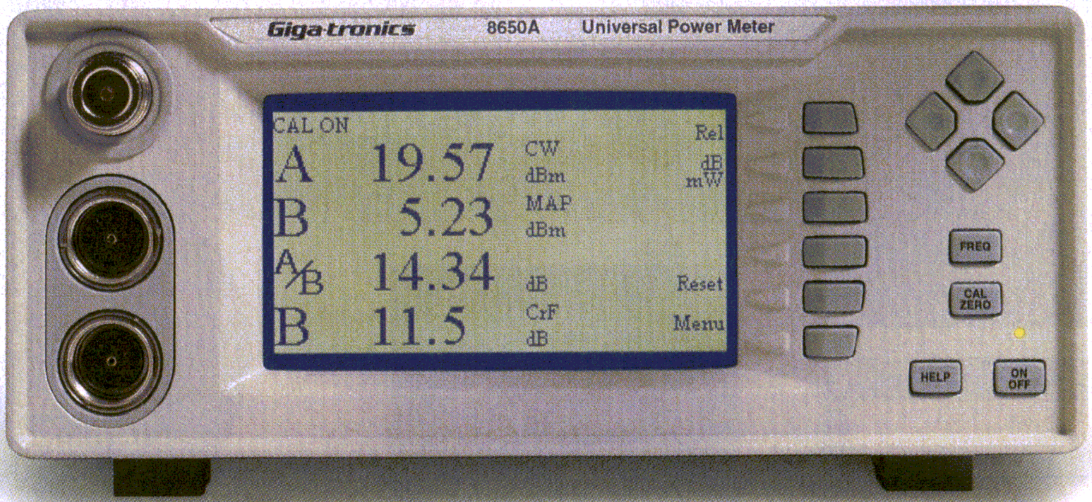

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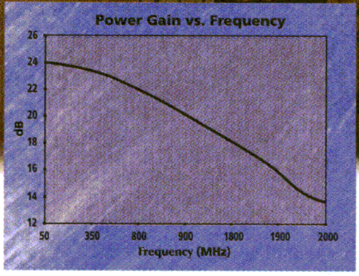

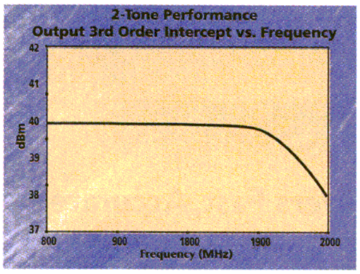

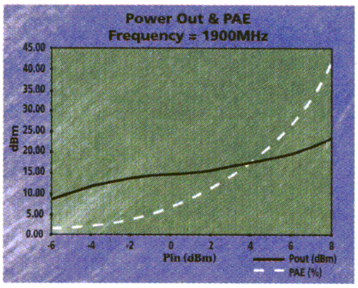

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This GaAsHBT amplifier is fabricated using molecular beam epitaxial growth technology, which produces reliable and consistent performance from wafer to wafer and lot to lot. The SXH-1 was specifically designed for use as drivers stages for infrastructure equipment in the 50-2000MHz cellular, ISM and narrowband PCS bands. Operating at a stingy 95ma of current, the SXH-1 is an ideal choice for multi-carrier as well as digital applications.

Performance Matrix
SXH-1

Freq. (MHz)	Gain (dB) Typ.	S11 Typ.	S22 Typ.	P1dB (dBm)	TOIP (dBm)	Voltage	Current (mA) Typ.
800-960	20.0	1.5:1	1.9:1	22.0	39.0	5.0	95.0
1800-2000	14.5	1.5:1	1.7:1	22.0	39.0	5.0	95.0

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Contracts

Berkeley Varitronics Systems, Inc.—Has been awarded a contract by Procelbras of Brazil to produce code-division-multiple-access (CDMA) test transmitters and Rhino rubidium frequency sources to help with rapid build out in Brazil's underground rail system.

Srico, Inc.—Has won a $100,000 contract from the US Air Force to develop a unique optical chip-based modulator component that reduces the noise figure of high-frequency communications systems to approximately 0 dB. The component may be used for cable-television (CATV) transmission, remote antennas, satellite communications, wireless and cellular communications, as well as other data-transmission systems.

Harris Corp.—Announced the extension of its contract with US Cellular, a wireless telecommunications provider. The three-year agreement is for the supply and installation of Harris radio links and is valued at $20 million.

TRW, Inc.—Has agreed to produce a set of indium-phosphide (InP) chips designed by the Australian government-owned Commonwealth Scientific and Industrial Research Organisation (CSIRO).

Decibel Products—Has been awarded a $6.6 million contract to supply the largest cellular operator in China with key components to expand their existing Global System for Mobile Communications (GSM) wireless communications systems. The contract requires Decibel products to manufacture and deliver more than 5000 base-station antennas, including directional panel antennas; omni antennas; along with state-of-the-art dual-polarized, variable down-tilt antennas.

SatCom Systems, Inc.—Has been awarded contracts from the North American Electric Reliability Council (NERC) and member organizations to provide mobile satellite equipment and services for a nationwide Y2K backup communications network.

LCC International, Inc.—Announced that its contract with XM Satellite Radio, Inc. ("XM") has been extended to include, among other things, the deployment of XM's terrestrial repeater network that is used to augment the company's coast-to-coast satellite signal coverage. The extension, which may reach $5 million in value, enables LCC to increase its level of activity from defining the technical requirements for XM's satellite terrestrial repeater network in an initial 30 markets to now include preliminary services in the remaining 40 markets of XM's coast-to-coast network.

Tellabs—Announced that it will provide multiservice networking solutions for the Indian State Government through Andhra Pradesh Technology Services Ltd. (APTS). The contract was awarded to United Telecom Ltd. (UTL), a Tellabs' strategic partner in India. Tellabs' MartisDXX managed access and transport solution will provide data, voice, and video-transmission services to the APTS organizations. These include the Government of Andhra Pradesh, state public-sector undertakings, local bodies in Andhra Pradesh, and other organizations for government interactions.

Wireless, Inc.—Has received an initial order valued at $2.5 million from Quest Wireless, Inc., a wholly owned subsidiary of Quest Net Corp. for wireless-access equipment. Quest Net Corp. is using the equipment to initially expand their network from the South of Miami up to West Palm Beach, FL.

ADTRAN, Inc.—Announced a contract with Bell Atlantic to provide the industry's first NEBS level 3-compliant dual T1, license-free radio—ADTRAN's TRACER®. Bell Atlantic has chosen ADTRAN for wireless unlicensed T1 connectivity for voice and data transmission.

Fresh Starts

Technology Futures, Inc. (TFI)—Announced the opening of an office in Washington, DC. Their services include consulting, research, and education.

Raytheon Co.—Announced that it will change the name of its Raytheon Microelectronics business unit to Raytheon RF components. The name change reflects a move to better align the business within its market segment and enhance customer awareness of the business.

Optical Cable Corp.—Announced that it has been informed by the US Army, Communications Electronics Command, Fort Monmouth, NJ, that it is qualified to supply the military tactical fiber-optic cable used in the manufacture of the Army's Tactical Fiber Optic Cable Assembly "TFOCA."

Cardinal Components, Inc.—Has added Coakley, Boyd, & Abbett as its newest representative in the New York/New Jersey metropolitan area.

Reactel, Inc.—Has moved to a new 10,000-sq.-ft. stand-alone building located at 8031 Cessna Ave., Gaithersburg, MD 20879. The facility houses manufacturing, assembly, testing, and offices.

Bowthorpe, plc.—Completed the acquisition of the xDSL simulator division of Consultronics Ltd. The $25 million purchase enhances Bowthorpe's position as a supplier of advanced test instruments and systems for telecommunications testing applications.

ANADIGICS—Has begun developing internal heterojunction-bipolar-transistor (HBT) capabilities while enabling immediate capacity through an agreement with Global Communication Semiconductors, Inc. (GCS). GCS will provide ANADIGICS with open foundry service for HBT wafers as part of a non-exclusive long-term agreement between the two companies.

RF Micro Devices, Inc.—Approved plans to build a second facility to fabricate gallium-arsenide (GaAs) heterojunction-bipolar-transistor (HBT) semiconductor wafers. The new plant will be located near its existing GaAs HBT facility in Greensboro, NC.

Tektronix, Inc.—Announced a worldwide strategic program for its customer-service organization in the Measurement Business Division. Named Services 2000, this program for the new millennium focuses on customer demand for decreased downtime and increased quality of service. To manage customer service in the US better, Tektronix has launched a National Call Center with a single, toll-free number and has consolidated the company's 12 service depots into two automated service centers.

DSP Communications, Inc.—Davidi Gilo to chief executive officer; remains chairman of the board.

Abpac, Inc.—Gil Olachea to marketing vice president; formerly worked in sales and marketing in the integrated-circuit (IC) packaging business.

Methode Electronics, Inc.—Gregory J. Gartner to senior executive officer of the Methode AEC business unit; formerly general manager with General Electric Co.


GARTNER


WILSON

Uniphase Corp.—David A. Wilson to director of administration for the Broadband Products division; formerly director of finance and accounting for Harris Corp.'s Electronic Systems Sector (ESS).

Trompeter Electronics, Inc.—Gayland Fisher to regional sales manager; formerly national distribution manager at Power-One.

Narda Microwave—Michael J. Sanator to vice president of operations; formerly director of operations.

Quad Systems Corp.—Jay DiGiovanni to director of software engineering; formerly director of systems engineering at CFM Technologies, Inc.

CTS Corp.—Jeannine M. Davis to executive vice president of administration, general counsel, and secretary; formerly senior vice president, general counsel, and secretary.

The Personal Communications Industry Association (PCIA)—Harold Salters to director of government relations; formerly director of regulatory affairs for ANMEX, Inc.

Kepco, Inc.—Martin Kupferberg to president; remains financial vice president. Also, Saul Kupferberg to secretary/treasurer; remains vice president of sales and marketing.

Microwave Instrumentation Technologies—Michael W. Murphy to director of customer support; formerly programs manager for advanced systems and phantom works site director at the Boeing Co.'s military aircraft and missiles site.

STMicroelectronics—Georges Auguste to corporate vice president; formerly director of total quality and environmental management.

Stanford Microdevices, Inc.—Robert Van Buskirk to chief executive officer; formerly executive vice president for business development and operations at Multilink Technology Corp.

Signal Technology, Olektron Operation—Arthur Butts to vice president of sales and marketing; formerly sales manager.

Taconic—Dr. Thomas F. McCarthy to product development manager; formerly headed the development of new fluoropolymer dispersions with Allied Signal.

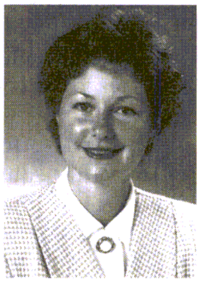
MYDATA Automation—Ed Ruppert to northwest regional sales manager; formerly western regional sales manager at Sanyo.

National Dispatch Center, Inc. (NDC)—Melinda Ann Krupcyznski to manager of customer support; formerly assistant manager of customer service.

IPC—Holly Evans to vice president of government relations.

TriPoint Global Communications—James Griscavage to vice president of quality assurance; formerly quality assurance director.


GRISCAVAGE


DUNN

Kanthal Globar—Sharon A. Dunn to components sales engineer; formerly held sales and marketing positions in the ceramic process equipment and materials industry.

EDUCATIONAL MEETINGS

Short Courses

Engineering and Planning Telecommunication Local Loop Facilities
November 2-5 (Madison, WI)
Department of Engineering Professional Development
University of Wisconsin-Madison
Madison, WI 53706
Katie Peterson
(800) 462-0876, FAX: (608) 262-3160
e-mail: custserv@epd.engr.wisc.edu
Internet: http://epd.engr.wisc.edu

Global Positioning System: Principles and Applications
November 2-5 (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

Digital Communications—An Introduction
November 8-9 (University of Oxford, Oxford, UK)
International Electronics and Communications Courses 1999
Oxford University
Linda Tonge
+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

Using the IS-136 TDMA Wireless Air Interface
November 9-12 (Dallas, TX)
Department of Engineering Professional Development
University of Wisconsin-Madison
Madison, WI 53706
Katie Peterson
(800) 462-0876, FAX: (608) 262-3160
e-mail: custserv@epd.engr.wisc.edu
Internet: http://epd.engr.wisc.edu

Low-Power Circuits and Systems for Digital Wireless Communications
November 11-12 (Redwood City, CA)
Continuing Education in Engineering
UC Berkeley Extension
1995 University Ave.
Berkeley, CA 94720-7010
(510) 642-4151, FAX: (510) 642-6027
e-mail: course@unx.berkeley.edu

Tutorial Colloquium on Design of RF ICs and MMICs
November 24 (Savoy Place, London, England)
Claire Ewings
+44 (0) 171 344 5423
Internet: http://www.iee.org.uk/Events/e24nov99.htm

Short Course on Measurements and Characterization of Broadband Access Technologies
November 30-December 1 (Atlanta, GA)
National Institute of Standards and Technology (NIST)
David K. Walker
(303) 497-5490, FAX: (303) 497-3970
e-mail: dwalker@boulder.nist.gov
Internet: http://www.arftg.org

Meetings

Wireless I.T. 99 Convention and Exposition (Mobile Wireless Solutions)
November 2-4 (Santa Clara Convention Center, Santa Clara, CA)
Cellular Telecommunications Industry Association (CTIA)
Registration and Housing
(800) 334-6147, FAX: (415) 979-2250
Internet: http://www.wow-com.com

Low Temperature Co-fired Ceramic (LTCC) Technology for Wireless, RF, and Microwave: Measurement and Testing Issues
November 4-5 (Gaithersburg, MD)
Green Tape Applications Group and the National Institute of Standards and Technology (NIST)
Steve Freiman/Carolyn Sladic (NIST)
(301) 975-6119
e-mail: stephen.freiman@nist.gov

1999 Tower Summit & Trade Show
November 8-9 (Paris Hotel, Las Vegas, NV)
Shorecliff Communications, Inc.
27127 Calle Arroyo, Suite 1909
San Juan Capistrano, CA 92675
(800) 608-9641, FAX: (949) 443-9206
e-mail: registration@scievents.com

Expo Comm Brazil 99
November 9-12 (Centre Norte Facility, Sao Paulo, Brazil)
E.J. Krause & Associates
6550 Rock Spring Dr.
Bethesda, MD 20817
(301) 493-5500
e-mail: windom@ejkrause.com

7th International Symposium on Recent Advances in Microwave Technology
December 13-17 (Malaga, Spain)
University of Malaga/University of Nevada (Reno, NV)
Professor Banmali Rawat
University of Nevada
(702) 784-6927, FAX: (702) 784-6627
e-mail: rawat@ee.unr.edu
Internet: http://www.isramt99.ic.uma.es

Call For Papers

2000 International Conference on Gallium-Arsenide Manufacturing Technology (GaAs MANTECH)
May 1-4, 2000 (Washington, DC)
Bruce Bernhardt
Motorola, MD EL-720
2100 East Elliot Rd.
Tempe, AZ 85284
(480) 413-4427
Internet: http://www.gaasmantech.org
Abstract deadline: November 1, 1999

4th European Symposium on Electromagnetic Compatibility (EMC Europe 2000)
September 11-15, 2000 (Brugge, Belgium)
TI-KVIV, Mrs. R. Peys
B 2018 Antwerpen, Belgium
+32/(0)3 216 09 96, FAX: +32/(0)3 216 06 89
e-mail: johan.catrysse@kh.khbo.be
Internet: http://www.ti.kivi.be/conf/emc2000 htm
Abstract deadline: December 1, 1999

Micromachined resonators offer high-Q characteristics

Diplexers, filters, and other components that run at millimeter-wave (Ka-band) frequencies are often implemented using high-Q structures such as resonant waveguide cavity filters or dielectric resonator filters. These components must exhibit low insertion loss, high out-of-band rejection, and high channel-to-channel isolation. A micromachining technique that suspends microstrip lines on thin dielectric membranes was found to produce resonators that have quality factors greater than 450, much higher than is possible with traditional manufacturing techniques. In addition, according to Andrew R. Brown and Gabriel M. Rebeiz of the Radiation Laboratory, Department of Electrical Engineering at the University of Michigan (Ann Arbor, MI) and Pierre Blondy of the University of Limoges (Limoges, France), the resonators support planar integration in complex filter designs such as millimeter-wave low-loss filters. See "Microwave And Millimeter-Wave High-Q Micromachined Resonators," *International Journal of RF and Microwave Computer-Aided Engineering*, Vol. 9, No. 4, July 1999, p. 326.

Network improves direction-finding antenna capability

Radio direction-finding systems used in mobile communications require accurate calibration and may be sensitive to noise and external interference. It is possible to avoid the hardware calibration by implementing an artificial neural network (NN) at the output of an antenna array's combining circuit based on research performed by ...Èric Charpentier and Jean Jacques Laurin of the Department of Electrical and Computer Engineering, Ecole Polytechnique de Montrèal (Montreal, Canada). The topology of system is a hybrid, combining a simple RF beamformer with an NN. The system uses a very simple RF architecture including patch antennas, low-noise amplifiers (LNAs), standard hybrid junctions, and diode detectors. See "An Implementation Of A Direction-Finding Antenna for Mobile Communications Using A Neural Network," *IEEE Transactions on Antennas and Propagation*, Vol. 47, No. 7, July 1999, p. 1152.

Hardware and software team for better telecom testing

Built-in self test (BIST) is a hardware technique incorporated in the integrated circuits (ICs) of communication systems to detect and diagnose possible device failures. Traditional diagnostic solutions rely on software to detect and diagnose failures, ignoring the hardware capabilities built into modules or components. This approach, however, delivers less than adequate fault coverage besides requiring a considerable investment in product-specific software generation. A new approach to telecom system design developed by Saman M. I. Adham, Ken Brough, Bruce Ecroyd, Stephen Foster, and Paul Soong of Nortel Networks (Ottawa, Canada) links diagnostic software to hardware BIST. The linking reduces software complexity while increasing diagnostic accuracy, and permits subcritical faults to be detected and recorded in non-volatile memory for subsequent failure analysis. The key to a successful test strategy is the use of hardware BIST which provides improved flexibility and versatility to the creation of test, programming, and system-maintenance capability. See "Linking Diagnostic Software To Hardware Self-Test In Telecom Systems," *IEEE Communications Magazine*, Vol. 37, No. 6, June 1999, p. 79.

Finding the best way to measure ADC linearity

Linearity measurements represent a substantial portion of the production cost of an analog-to-digital converter (ADC), but a method developed by Peter D. Capofreddi and Bruce A. Wooley of Integrated Circuits Laboratory (Stanford, CA) claims to cut this cost by reducing the time required to achieve the necessary level of measurement accuracy. The method takes the approach of evaluating linearity measurement methods by comparing the efficiency where they generate linearity estimates from information provided by the ADC output. The efficiency of three different methods is compared to the Cramer-Rao Bound—the theoretical limit on the accuracy that can be achieved in a particular measurement configuration. Two open-loop methods—tally and weight along with the code density—were compared by the researchers. It was discovered that the code-density method produced an efficiency close to the theoretical limit while the tally-and-weight method fell short of the optimum for converters with noise levels exceeding one-quarter least-significant bit (LSB). See "The Efficiency of Methods For Measuring A/D Converter Linearity," *IEEE Transactions On Instrumentation And Measurement*, Vol. 48, No. 3, June 1999, p. 763.

Simple Algorithm Extracts SiGe HBT Parameters

A new extraction technique identifies intrinsic and extrinsic parameters of SiGe HBT devices with close agreement to actual measurements.

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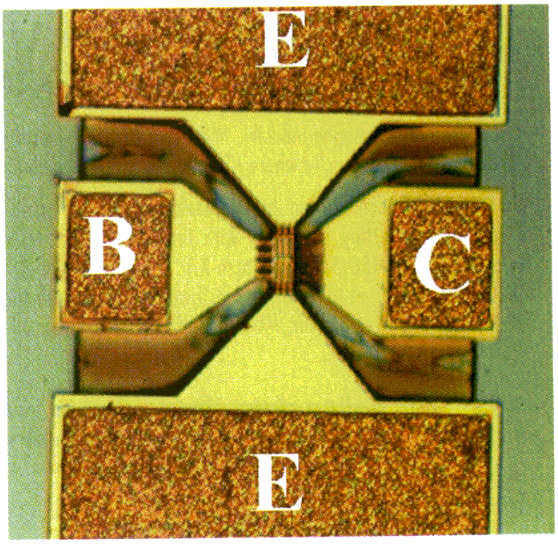
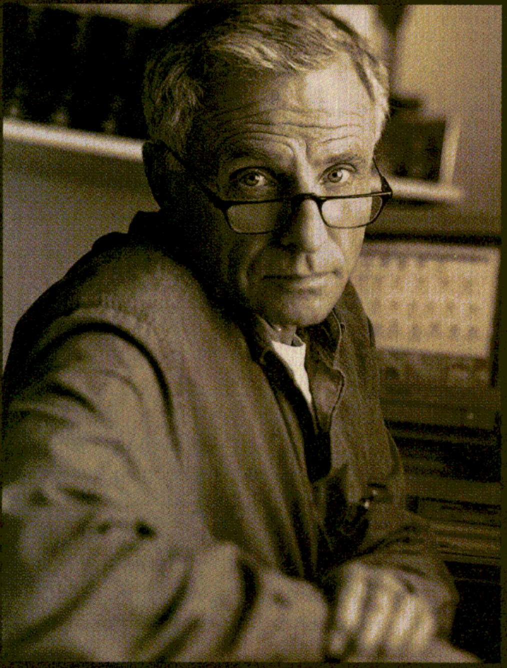
A new and simple extraction method for determining the parameters of small-signal Π topology equivalent circuits of silicon-germanium (SiGe) heterojunction bipolar transistors (HBTs) has been developed. Both intrinsic and extrinsic (parasitic) elements can be extracted with the algorithm. Conventional procedures or methods based on simple bias measurements work very well if the extrinsic elements of the HBT have been previously determined.[1] This approach may be used through different procedures—DC, cut-off measurements, or optimization. However, it is often very difficult to accurately determine the values of parasitic elements of the HBT, since the usual DC and cut-off techniques offer poor performance for SiGe HBT devices.

In order to avoid this drawback, a new technique has been developed which does not require any additional measurements except for the scattering (S)-parameters at different biases. Linear models with a Π topology have been tested to fit the measured S-parameters properly. This method has been applied to several devices but, in particular, to the HBT120Z devices (one finger 1 × 20 μm) and HBT67 (six fingers 1 × 7 μm), which have been employed in two amplifier designs developed by the group (Fig. 1).[2]

This method is an extension to the HBT devices of Shirakawa's[3] and Ooi's[4] work on metal semiconductor field-effect transistors (MESFETs) and high-electron mobility transistors (HEMTs), respectively. The main difference between the MESFET/HEMT approach and this new method is that the MESFET/HEMT method uses nine different biases to evaluate the frequency behavior of the intrinsic elements to properly compute the set of parasitic elements. On the other hand, the method takes the fitting of some important figures of merit of



1. The SiGe HBT67 device manufactured by Diamler Chrysler, shown here in a microphotograph, has six fingers, 1 × 7 μm.



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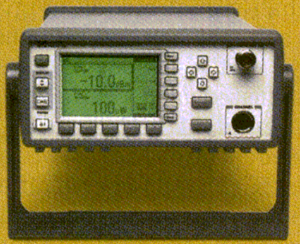

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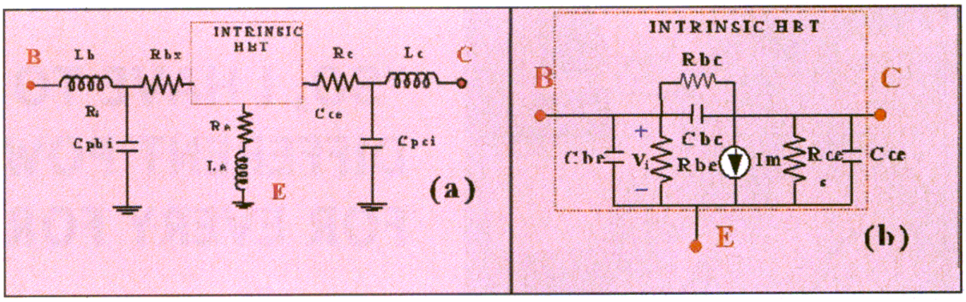
the device into account, such as maximum stable gain (MSG) and Rollet's factor (K) as optimization objectives. A significant reduction of the comparative error function between modeled and measured S-parameters is obtained by using the previously mentioned premises.



2. A small-signal Π equivalent circuit of an HBT device (a) contains intrinsic and extrinsic circuit elements. The intrinsic elements (b) can be determined from the admittance parameters of the device at a number of different bias points.

The proposed circuit topology of the small-signal model, (Fig. 2a), shows a dashed box that contains the intrinsic part of the device (Fig. 2b). The estimated values of the pad parasitic capacitances C_{pbi} and C_{pci} do not exceed hundredths of femto Farads (fF) for a number of different devices tested. Thus, these capacitances may be neglected. The remaining extrinsic parameters (Lb, Lc, Le, Rb, Rc, and Re) in Fig. 2a can be determined easily by an optimization process at the same bias point where intrinsic elements were computed. This will be shown shortly.

INTRINSIC ELEMENTS

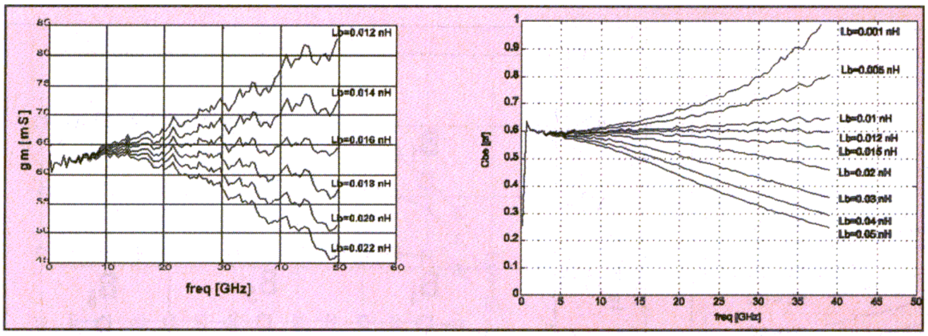
After a conventional de-embedding process to take the effect of parasitic elements [5,6] into account, it is possible to obtain the intrinsic elements from the admittance parameters of the HBT device for each bias point.[7] The relationship between the current source, I_m, and the intrinsic transconductance, g_m, is provided by the following equation:

$$I_m = G_m \cdot V_i$$

where:

$$G_m = g_m \cdot \varepsilon^{-j\omega\tau} \qquad (1)$$

Taking the real and imaginary parts of these admittance parameters (Yij) [I = 1,2 and j = 1,2], it is relatively easy to obtain a set of eight equations with eight unknowns. This process supports the computation of C_{be}, R_m, R_{be}, R_{bc}, C_{bc}, g_m, τ, R_{ce}, and C_{ce}, values by using the following expres-



3. Intrinsic elements of an HBT device such as transconductance (g_m) (a) and base-emitter capacitance (C_{be}) (b) vary with frequency. They also vary with different values of extrinsic elements such as base inductance (Lb) as shown in these curves.

sions:

$$C_{bc} = \frac{-Im[Y_{12}]}{\omega_i} \qquad (2)$$

$$C_{be} = \frac{Im[Y_{11}] + Im[Y_{12}]}{\omega_i} \qquad (3)$$

$$C_{ce} = \frac{Im[Y_{22}] + Im[Y_{12}]}{\omega_i} \qquad (4)$$

$$R_{bc} = \frac{-1}{Re[Y_{12}]} \qquad (5)$$

$$R_{ce} = \frac{1}{Re[Y_{12}] + Re[Y_{22}]} \qquad (6)$$

$$R_{be} = \frac{1}{Re[Y_{11}] + Re[Y_{12}]} \qquad (7)$$

$$g_m = \left\{ \begin{matrix} (R_e[Y_{21}] - R_e[Y_{12}])^2 + \\ (Im[Y_{12}] - Im[Y_{21}])^2 \end{matrix} \right\}^{0.5} \qquad (8)$$

$$\tau = \frac{1}{\omega_i} \cdot tan^{-1}\left(\frac{Im[Y_{12}] - Im[Y_{21}]}{R_e[Y_{21}] - R_e[Y_{12}]} \right) \qquad (9)$$

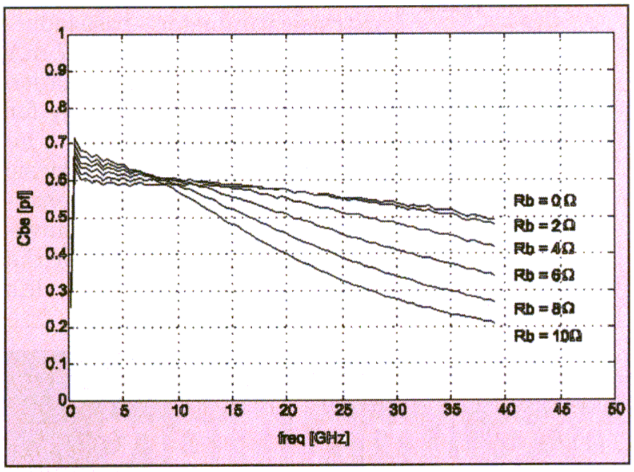
where:

Re and Im = the real and imaginary parts, respectively,

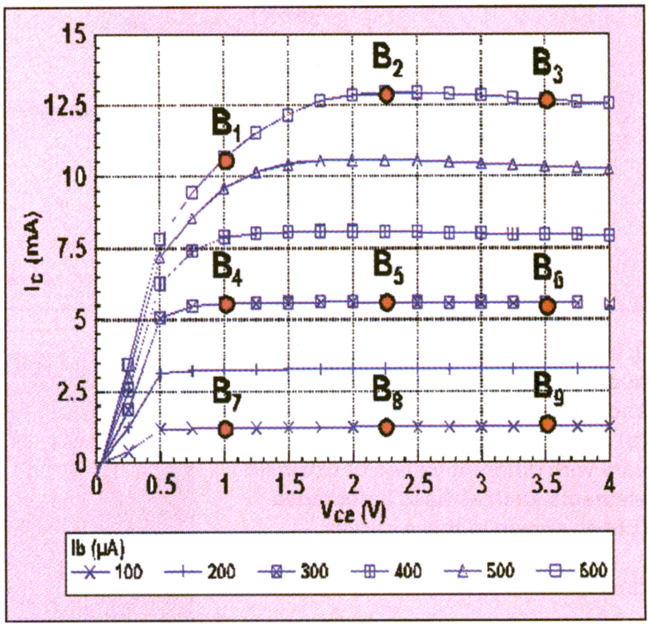
ω = the angular frequency, and

i = 1......N = the number of frequency sampling points.

Figure 3 shows the variation of the intrinsic elements g_m and C_{be} with frequency, for different values of an extrinsic parameter—base inductor L_b—biased at I_c = 1.77 mA and V_{ce} = 2.25 V. The variation with the access resistor R_{bx} for C_{be} is shown in Fig. 4. The remaining intrinsic and extrinsic parameters show similar behavior. Since the small-signal equivalent circuit must be valid for the measured frequencies, this



4. These curves are similar to those of Fig. 3b in that they show how C_{be} varies with frequency and the extrinsic base input resistance R_{bx}.



5. Since parasitic elements vary with bias, they are calculated at nine equally spaced bias points along the I-V characteristic of an HBT67 transistor.

implies that all intrinsic parameters must be frequency independent. Keeping this assumption in mind, the intrinsic parameters could be expressed, for each bias point, as a function of extrinsic elements and frequency. In this way, for the intrinsic parameters, I_π, a set of relationships could be established:[7]

$$I_{pi} = f_i(R_b, R_c, R_e, L_b),\ also$$
$$I_{pi} = f_i(L_e, L_c, C_{pbi}, C_{pci}, \omega_i) \quad (10)$$

However, extrinsic elements that are determined in this form often invalidate the equivalent circuit when the bias point is changed. To take this variation of parasitic elements with bias into account, nine equally spaced bias were chosen along the whole I-V characteristic curves. In this form, the function is redefined as follows:

$$I_{pi} = f_i^*(R_b, R_c, R_e, L_b, L_e),\ also$$
$$I_{pi} = f_i^*(L_c, C_{pbi}, C_{pci}, \omega_i, B_j) \quad (11)$$

where:
i = 1,...,8, and
j = 1,...,9 (with j denoting the number of different bias points used during the optimization process).

Assuming that the intrinsic small-signal model selected is valid over the whole frequency range, and making use of the intrinsic elements for optimization criteria, it is possible to

determine appropriate values for extrinsic elements by iteration without additional measurements for the nine bias points shown in Fig. 5.

Choosing the variance criteria, the first elements of the error function are as follows in equation 12:

$$\varepsilon^{k}_{1,j}\left(Z_{ext}, B_j\right) = \frac{W_{1,k}}{N-1}\sum_{l=0}^{N-1}\left| f_k^*(Z_{ext}, \omega_l, B_j) - \overline{\sum_{l=0}^{N-1} f_k^*(Z_{ext}, \omega_l, B_j)}\right| \qquad (12)$$

where:

$W_{1,k}$ = the weighting factor k = 1,.....8, and

j = 1,......9.

For convenience, Z_{ext} = all parasitic elements, and

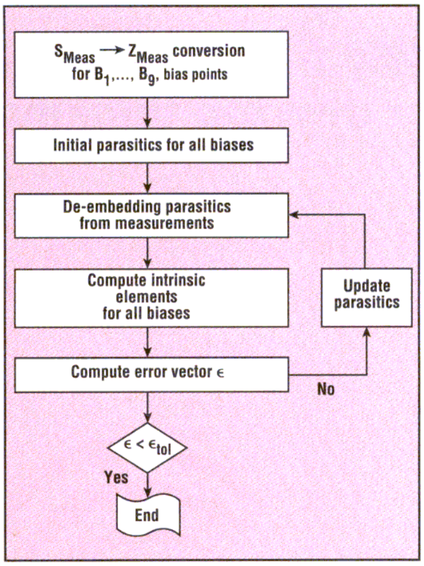
B_j = the variation with bias.

For stable calculations, the difference between the computed and measured S-parameters, the maximum stable gain, and Rollet's factor K must also be checked. The second, third, and fourth components of the error function are selected as:

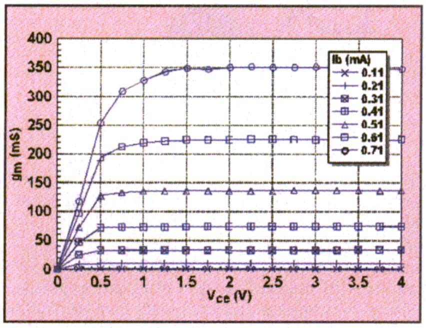
$$\varepsilon_{2,j}\left(Z_{ext}, B_j\right) = \sum_{p=1}^{2}\sum_{q=1}^{2}\sum_{r=1}^{N-1} W_{2pg}\cdot \left| S_{pq}^{Meas}(\omega_r, Z_{ext}, B_j)^2 - S_{pq}^{Mod}(\omega_r, Z_{ext}, B_j)^2\right|^{0.5} \qquad (13)$$

$$\varepsilon_{3,j}\left(Z_{ext}, B_j\right) = \sum_{r=1}^{N-1} W_3 \cdot \left| MSG^{Meas}(\omega_r, Z_{ext}, B_j)^2 - MSG^{Mod}(\omega_r, Z_{ext}, B_j)^2\right|^{0.5} \qquad (14)$$

$$\varepsilon_{4,j}\left(Z_{ext}, B_j\right) = \sum_{r=1}^{N-1} W_4 \cdot \left| K^{Meas}(\omega_r, Z_{ext}, B_j)^2 - K^{Mod}(\omega_r, Z_{ext}, B_j)^2\right|^{0.5} \qquad (15)$$



6. This flowchart is the algorithm for extracting the extrinsic (parasitic) elements of a device at many different bias points (called multibias extraction). It uses an error function computed from measured and modeled S-parameters to obtain more accurate parasitic data.



7. The extracted intrinsic-element transconductance (g_m) plotted against collector-to-emitter voltage (V_{ce}) is shown here as a function of base current I_b.

where:

Meas and Mod = the measured and modeled S-parameters, and

the different Ws = the weighting factors (usually fixed at 0.5). The extended error function for bias j is then composed as follows:

$$\varepsilon_j\left(Z_{ext}, B_j\right) = \begin{bmatrix} \varepsilon_{1,j}^{k}\left(Z_{ext}, B_j\right) \\ \varepsilon_{2,j}\left(Z_{ext}, B_j\right) \\ \varepsilon_{3,j}\left(Z_{ext}, B_j\right) \\ \varepsilon_{4,j}\left(Z_{ext}, B_j\right) \end{bmatrix} \quad (16)$$

where:

k = 1,.....8, and

j = 1,.......9.

The total error function vector for the nine different bias points is defined as:

$$\varepsilon\left(Z_{ext}\right) = \begin{bmatrix} \varepsilon_1\left(Z_{ext}, B_1\right) \\ \cdots \\ \varepsilon_9\left(Z_{ext}, B_9\right) \end{bmatrix} \quad (17)$$

Figure 6 shows the flowchart of the algorithm used for the multibias extraction. The values of parasitic elements are updated in order to minimize the error vector ϵ using the Sequential Quadratic Programming (SQP) method. The program has been written using MATLAB.[8]

When all parasitic elements are finally determined for the nine selected bias points, it is possible to perform the multibias extraction using the values computed by the algorithm for those bias points, under the assumption that they are bias independent for the rest of the bias points. The detailed multibias extraction for the 102 bias points is not presented here. As an example, Fig. 7 shows the behavior of the extracted transconductance, g_m versus V_{ce}, as a function of the base current I_b.

The electrical models obtained by the multibias procedure for the rest of the 102 bias points achieve a fit as good as the particular case shown in Fig. 7 and for S-parameters and Rollet's factor (K). These electrical models are linear models that are dependent on the bias point. This supports the availability of an electrical model valid for a broad range of bias points, which is more useful for the design of medium-power and voltage gain-controlled amplifiers using commercial computer-aided-design (CAD) tools

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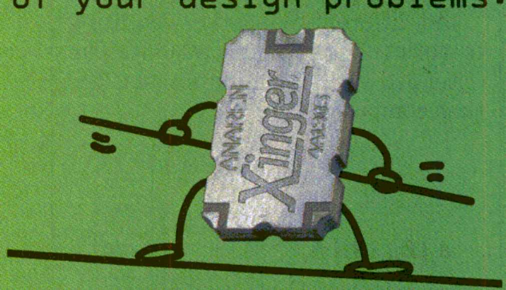

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such as HP ADS or Libra. For different HBTs of the same family, very good agreement between DC measurements and simulations has been found. Excellent agreement between measured and simulated S-parameters over 0.05 to 50 GHz for multibias operation confirms the robustness and validity of the method.

A new extraction technique has been developed to obtain the extrinsic (parasitic), and intrinsic elements of the Daimler SiGe HBT devices. Very good agreement for an HBT67 device between the MESFET/HEMT model and the measurements has been observed throughout the entire 0.05-to-50.05-GHz frequency range for all bias points studied. Linear models that are extracted by this method for other different sized devices have shown satisfactory agreement between measured and simulated data and have permitted validation of the model with encouragement to use these models in new designs.[9] ●●

Acknowledgement

This work has been developed within the framework of a European TMR project (contract No. ERB FMRX-CT96-0050). The authors would like to thank Dr. Luy and Dr. Filimon from Daimler Chrysler Research Center for the samples and measurements of the HBTs that were supplied.

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Design A Wideband VHF Filter With Coupled Resonators

A novel design for a wideband VHF filter uses coupled resonators to clean up noise and harmonic distortion.

Domenico K. Barillari

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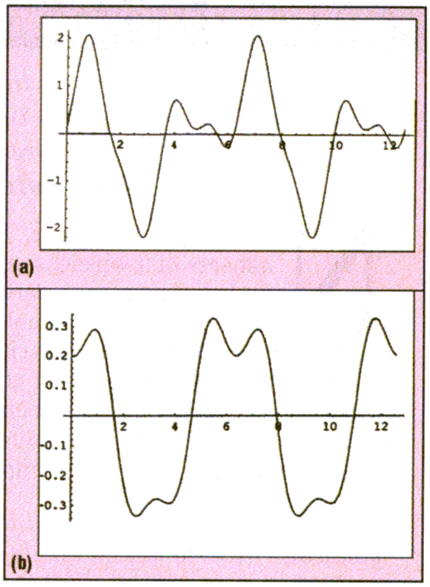
AN engineer cautiously checking a poorly aligned or mismatched amplifier with a wideband oscilloscope and a good probe will likely encounter an objectionable amount of harmonic distortion. If the offending circuit can be repaired or changed, so be it. But the task at hand is often to clean up the existing signals. The author of this article was recently faced with the challenge of filtering the output of a harmonically noisy but otherwise functional very-high-frequency (VHF) signal generator. The filter had to cover the very wide frequency range of 10 to 480 MHz with a minimum of filter sections and stages.

But finding information on VHF and ultra-high-frequency (UHF) filter design techniques, especially with regard to passive systems, can be difficult. The current literature and technical courses seems to focus exclusively on amplifier design. Matching circuits and tuned active circuits are described in many RF texts, but the equally crucial filter design gets short shrift. For the engineer set to the task of working on an intermediate-frequency (IF) stage at a few hundred megahertz, the task becomes one of extrapolating upward from classical engineering texts such as ref. 2, or downward from the extensive microwave literature (Matthaei *et al.*[3]).

To meet the challenge, the author turned to TV/VHF technology and developed a relatively simple electrically tunable coupled-resonator design. The design incorporates a number of ideas that can be of general use to RF engineers working from HF through UHF.

THEORY AND DESIGN

A representative sample of the objectionable waveforms observed from the signal generator, reconstructed with Mathmatica ®[6], is given in Figures 1a and 1b. What is needed here is a lowpass—or better still—a bandpass filter with a wide bandpass and steep upper-rejection skirt. The initial criteria chosen for such a circuit include:



1. These examples of harmonic and harmonic distortion are often seen on degraded equipment.

• A range of 10 to 480 MHz, (very likely implemented with overlapping band sections, each section covering some sub-band).

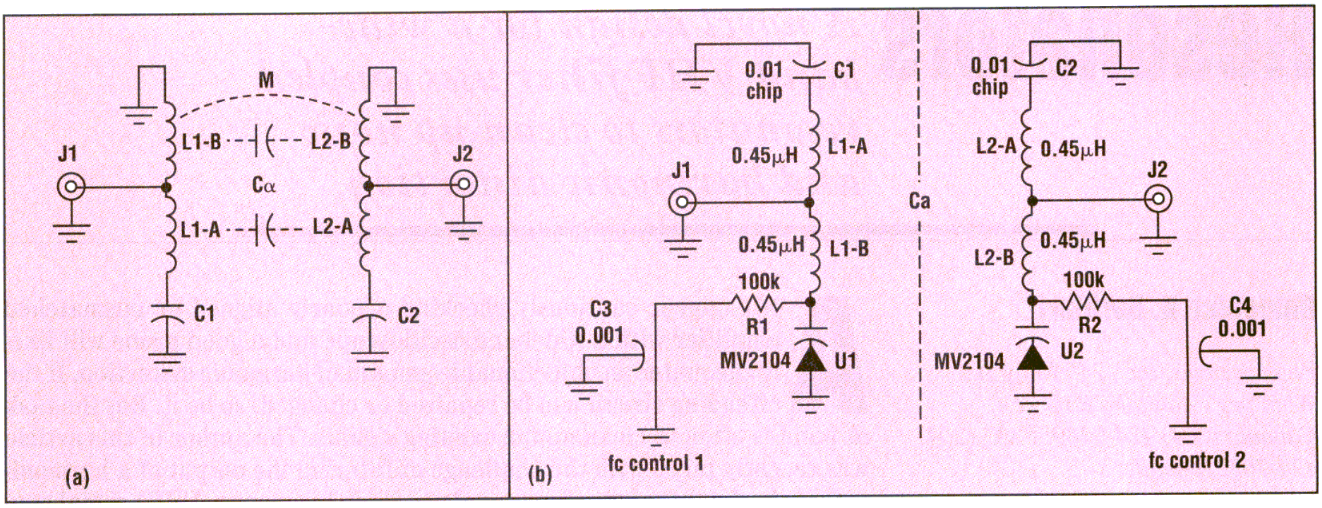
• As few filter sections as possible, possibly with a single tuning control.

• A first harmonic rejection of at least 40 dB at any setting.

• As simple a design as possible for any section, with as few stages as possible.

• Preferably a passive design, but varactor control is an option.

Much research, including construction of prototypes, showed that most lumped-component, reactive-ladder networks having both wideband and sharp roll-off characteristics were very difficult to construct. Further, the variable tuning that would be required in the lower-frequency bands to kill relatively close harmonics would require ganged capacitors or inductors, matched varactors, or



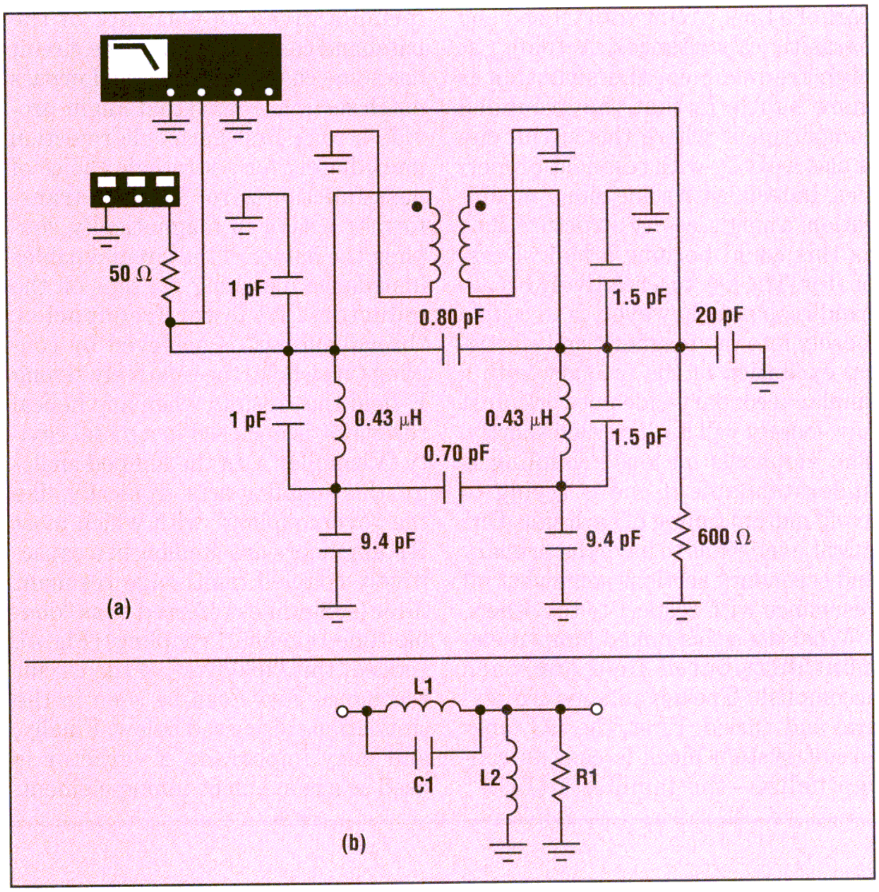
2. These schematics show a basic coupled-resonator filter and a practical circuit.

other such complications. Helical resonators were similarly researched, and although they are admirably suited for fixed narrowband use, they were also difficult to adapt to a continuously tunable design due to the rather severe restrictions on relative component dimensions in such a resonator.

The answer turned out to be the venerable old coupled-resonator design found in most solid-state television-set tuners. These are now quite small and well packed away in the first few sections of those "tin-can" assemblies that comprise the front end of most sets (Fig. 2).

Such coupled resonators operate in the frequency-transition region where a simple schematic of the lumped components does not tell the whole story. This is most clearly brought out here in the presence of a capacitive coupling aperture, which will be discussed shortly. RF energy is introduced via the "primary-side" tapped inductor, which also forms part of a tank circuit with C1 and any parasitic capacitances. Actually, at high frequencies, the situation is more subtle in that the grounded compartment where this circuit sits is also a cavity with resonant properties. Indeed, with some minor modification, a single-cavity structure such as this could become a helical resonator. The key to a relatively broadband response, however, is to couple loosely into the magnetic and electrical excitation of the primary with a similar secondary-side network. Just how loosely will be discussed shortly. The emphasis on loose coupling is understandable if one is trying to avoid mutual tuning of each side. One might even want to have the primary and secondary sections somewhat off resonance with respect to the other.

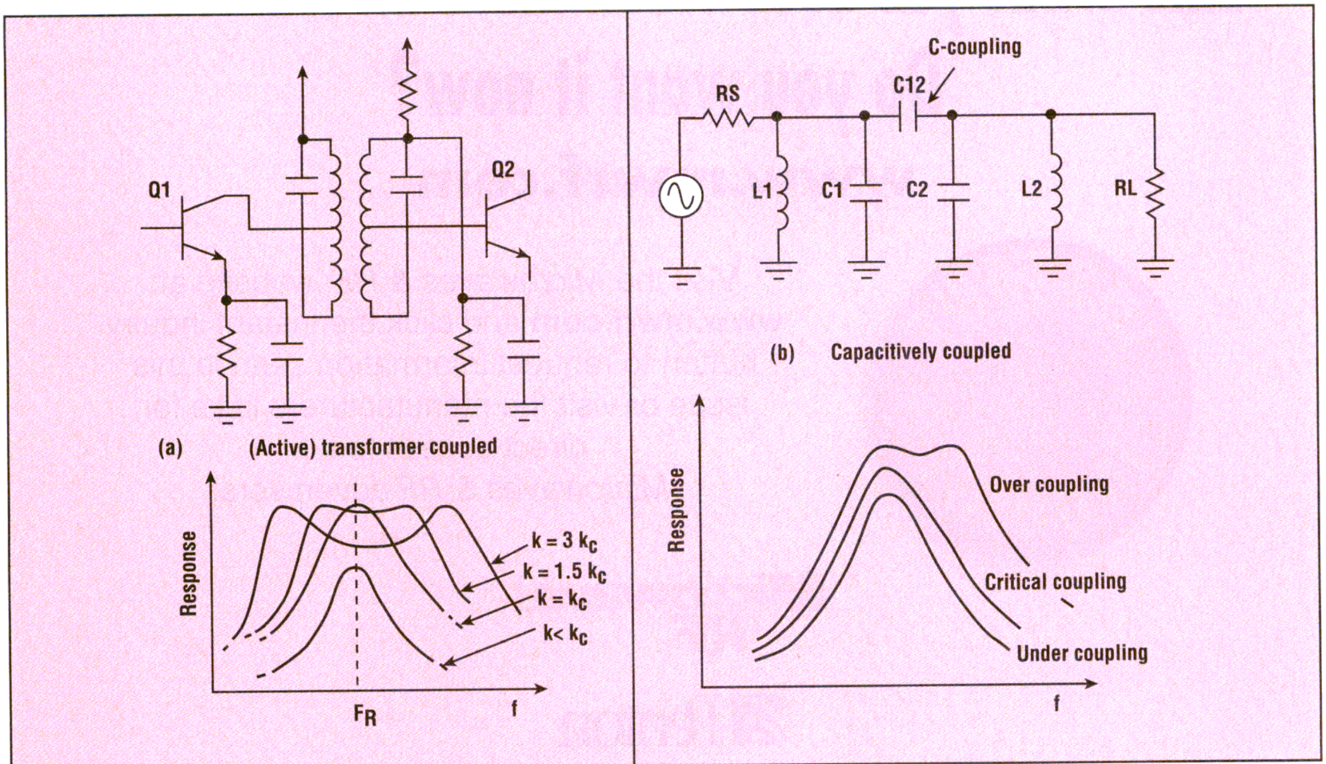
What does this mixed lumped-element/distributed-field approach accomplish? The advantages are several and varied. First, the LC tank-circuit system block is easy to conceptualize—the familiar $1/(LC)^{0.5}$ formula gives an estimate of the passband center. Further, the circuit has some characteristics of a lowpass notch section (Fig. 3) that might provide strong first-harmonic rejection immediately for a strategic choice of pass-function zero. Thirdly, transformer action in magnetically coupling the halves of the circuit enables impedance matching via taps on the inductors. At lower frequencies, "broadbanding" is achieved by coupling (via a hole) the relatively strong E-field that results when any helical resonator is enclosed in a metal cavity (Vizmuller[4]). In the lumped analogy, this coupling acts, ironically, like the "overcoupling" with which most RF engineers are familiar in a capacitively coupled multistage resonant filter (sometimes referred to as "double-tuned element" coupling) [Fig. 4]. Indeed, this "foldover" of the circuit resonance curve can be seen in the simulations discussed below. Finally, and very important, a varactor is used as a convenient tuning element.



3. These schematics show a lumped-component model of the resonantly coupled filter and the lowpass notch section.

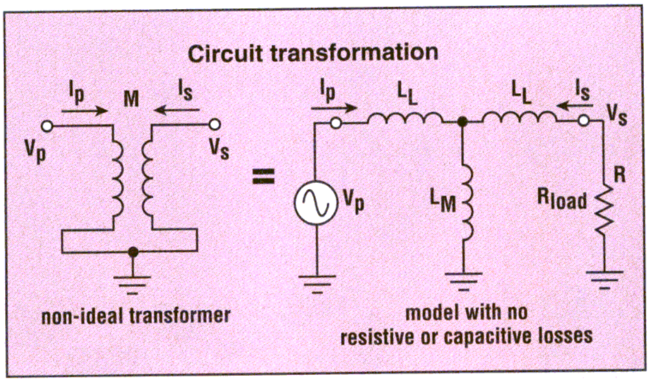
At higher frequencies, magnetic coupling begins to compete with capacitive coupling. This is also an example of "double-tuned-element" coupling (where "double" refers to the mirror symmetry of the circuit[8]) and is fairly easy to understand in terms of a standard transformer model. The primary side "sees" a reflected impedance, which is resistive at the nominal resonance frequency f_R. When the coupling, k, is so that this reflected resistance is equal to the primary's resistance, critical coupling, k_C, is attained. The double hump results when $k < k_C$. Recall that the reflected impedance to the primary is given by $(\Omega M)^2/Z^s$, and that, below resonance, the series impedance of the secondary circuit is capacitive. Thus, we get a reflected inductive value in the primary. This then cancels the primary's capacitive effect below f_R, producing a shifted resonance peak. A similar action takes place above f_R.

The question then becomes: Can an engineer use a readily modeled equivalent circuit to synthesize such a filter? It turns out that a rather straightforward model produces an accurate enough description of the



4. Two lumped-element, tuned-resonator circuits with overcoupling, along with their response, are shown here.

bandpass characteristic to enable a very quick constructive synthesis. Below approximately 200 MHz, one can take the individual values of the lumped components, a prediction of M (the magnetic coupling), and a reasonable estimate of Ca and the (capacitive) parasitic contributions, and solve the system on a relatively inexpensive program with a Bode facility such as Electronics Workbench®[5]. A model that reproduces many of the observed features of the actual filters is shown in Fig. 3. The behavior of the center-tapped inductors/transformer has been approximated by a network where pure self-inductances are series connected to a non-ideal transformer. Further, the electric coupling to the cavity was modeled as relatively large parasitic-like capacitances added at reasonable points in the circuit. When a poor choice of values resulted in unacceptable bandpass shapes, the values were tweaked by examination of the simulation output.



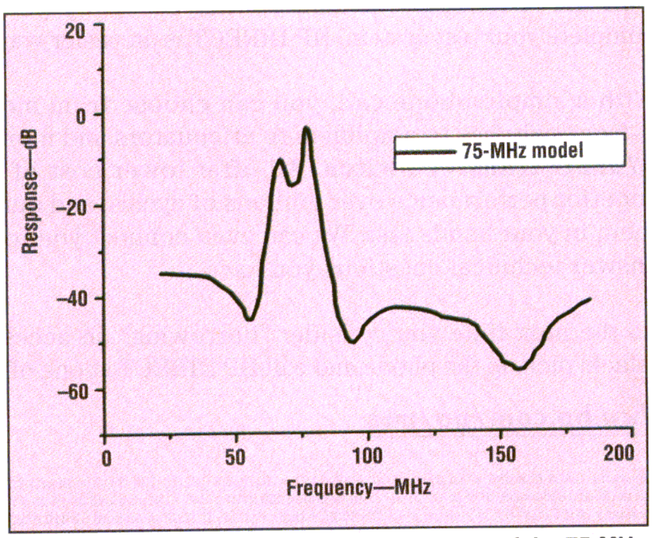
5. These schematics show a transformer with "cold-ends" grounded, and the equivalent T circuit. L_L is the "leakage" inductance and L_M is the "magnetizing" (mutual) inductance.

The mutual coupling of two air-spaced coils has no simple, closed-form solution. One way to predict M is to use the extremely useful tables and formulas of Grover[7]. Since plenty of iteration would be involved in converging on a set of dimensions for some M, the author was motivated to write a small program to calculate L^1, L^2, and M directly from the Nuemann equation. These values then became the parameters for the equivalent-T model of a transformer with the "cold ends" grounded, as shown in Fig. 5.

The reader might be concerned about some of the assumptions involved, but a more intricate model adds little for the purposes of this study. As discussed later, a more fundamental EMF approach might be called for if a more critical analysis if necessary.

As with any resonator design, the loading or "de-Qing" of the system by resistive loads is an extremely important consideration and must be set at the start. Given the argument on reflected impedance given above, this is not surprising. In this project, 50-Ω input and 600-Ω output values were selected. The latter represents a convenient design point for an isolating, untuned, metal-semiconductor field-effect transistor (MESFET)-follower stage that the author prefers as a utility-output driver. This feature is another departure from a strictly passive design, but one that has historically proved to be quite handy.

The result for the simulated "75-MHz" unit are shown in Fig. 6 for various effective capacitive and real mutual-inductor couplings. Dips suggested by theory above are evident whenever either coupling is strong enough, and either may be effective as the practical situation demands. Further, good attenuation at the first harmonic is suggested of the order of the design requirement, and is largely due to the beneficial presence of the band-reject zero near 150 MHz. The latter was purposely emphasized by tweaking components values. Finally, there is a long low-frequency tail, which is largely harmless due to the absence of any low-frequency components in the final application.



6. This graph shows the simulated response of the 75-MHz passband unit.

IMPLEMENTATION

The cavities for the lower-frequency filters were implemented in folded sheet brass of a conveniently thick gauge. After soldering, but before adding other components, a thick gold (Au) plating was electroplated directly to the brass. Thereafter, the feedthrough capacitors and then other components were soldered onto spots pre-wet with solder. An example of the finished 75-MHz unit discussed later is shown in Figure 7a. This design has one varactor per cavity to support experimentation with separate tuning. Figure 7b shows a partially completed 180-MHz unit, constructed in a milled-brass cavity structure more appropriate to higher frequencies. The 400-MHz unit was essentially similar, except for the use of "U-inductors" made from a single fold of thick copper (Cu) stock, approximately 5 mm wide and 50 mm total length.

A number of experimental sections were constructed, most incorporated into a single rack-mount instrument that also contained a manual VHF switch, varactor-tuning control, and gallium-arsenide FET (GaAs FET) buffer for 50-Ω output.

At lower frequencies, the units were tested by feeding

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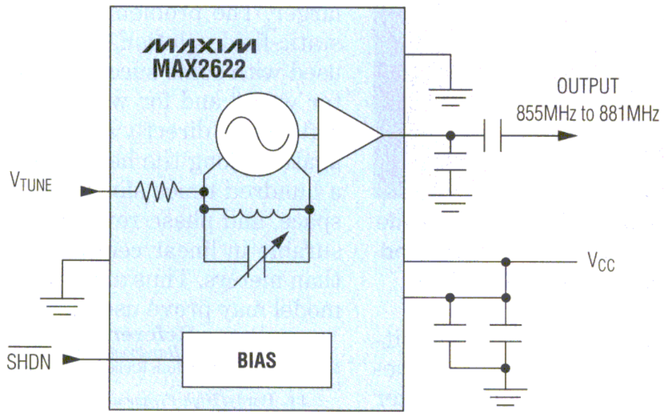

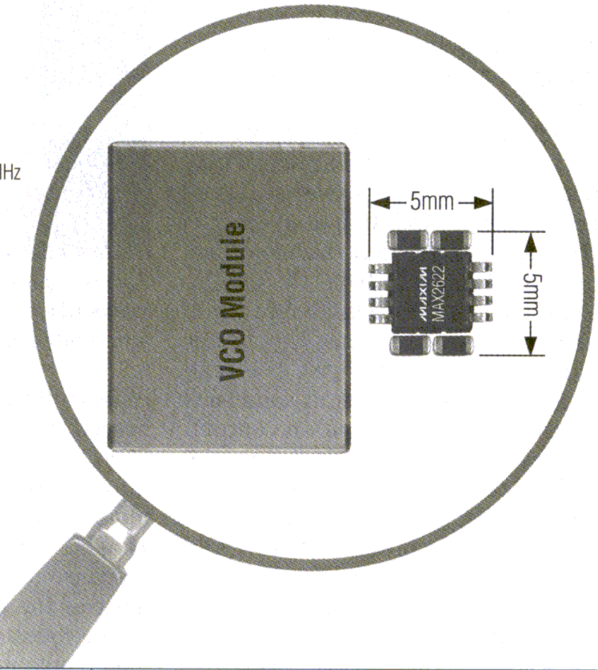

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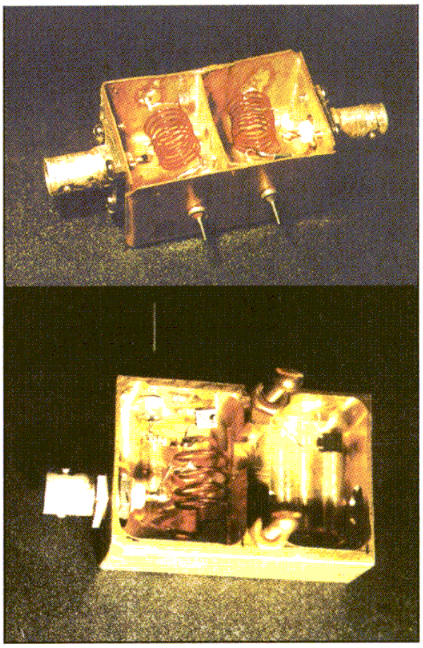
the distorted output of the signal generator directly to the filter with a 50-Ω buffer resistor. Both input and output were monitored with a wideband oscilloscope and high impedance probes to visually inspect the waveforms. Numerical data was taken, however, with calibrated high-impedance RF voltmeters shunted with a suitable load (either 50 or 600 Ω). Visual inspection did indeed seem to show very pure spectral output over most of the passband. This was later confirmed with a spectrum analyzer. On this basis, no further harmonic analysis was performed on the output until well outside the passbands, where severe distortion would, not surprisingly, appear. The input, however, was generally "dirty," and computer analysis was used to extract the harmonic frequency of interest from the oscilloscope shapes. This was simply a convenience measure, since the author did not want to switch the sole real-time spectrum analyzer at hand back and forth between input and output. The response was reported as the ratio of the output to input detected voltages, in dBV, numerically corrected for harmonic content and cross-instrument sensitivity. The procedure provided, in a very direct way, confidence that the filter units were actually behaving as desired.

At higher frequencies, where oscilloscope monitoring is not practical, the response must be measured with a calibrated spectrum analyzer, carefully using a VHF single-pole, double-throw (SPDT) isolation switch between input and output. The harmonic purity of the input was then measured in real time by simply examining the spectral components.

As shown in Fig. 8, over the space of the passband, a number of the essential features of the model were reproduced. Two immediately obvious points of divergence are the weakened rejection-band effect on the low-frequency side and the disappointingly low response in the observed passband. Presumably, the complex self-coupling effects of the cavity cannot be reproduced accurately by the shunt parasitic capacitances exactly as used in this model. Nonetheless, higher harmonics were indeed very well rejected as witnessed in excellent real-time waveform shapes and spectrum-analyzer information. It is difficult to determine whether the zero producing the 155-MHz dip seen in the model response of Fig. 6 is responsible for most of the elimination of that harmonic, because the author's equipment could not accurately follow the very deeply attenuated signals above 90 MHz. Nevertheless, this unit was considered an engineering success. The author is pleased to mention that the other high-band units for this project worked at least as well.



7. These photographs show the inside views of a completed 75-MHz unit and a partially completed 180-MHz filter.

The results obtained in this project has stimulated interest in the more theoretical issue of field strengths and components in the cavity. There exists a vast amount of literature on apertures (such as the Bethe iris) and cavities for the microwave region, but most is inapplicable at wavelengths two orders of magnitude larger. The problem still suggests a static-field solution, which has been used with some success in helical-filter work[3] and for which some EMF software is directly applicable. Propagation along the helix is effectively a hundred times slower than in free space, and phase reversals are measurable in linear centimeters rather than meters. Thus a lumped antenna model may prove useful. ••

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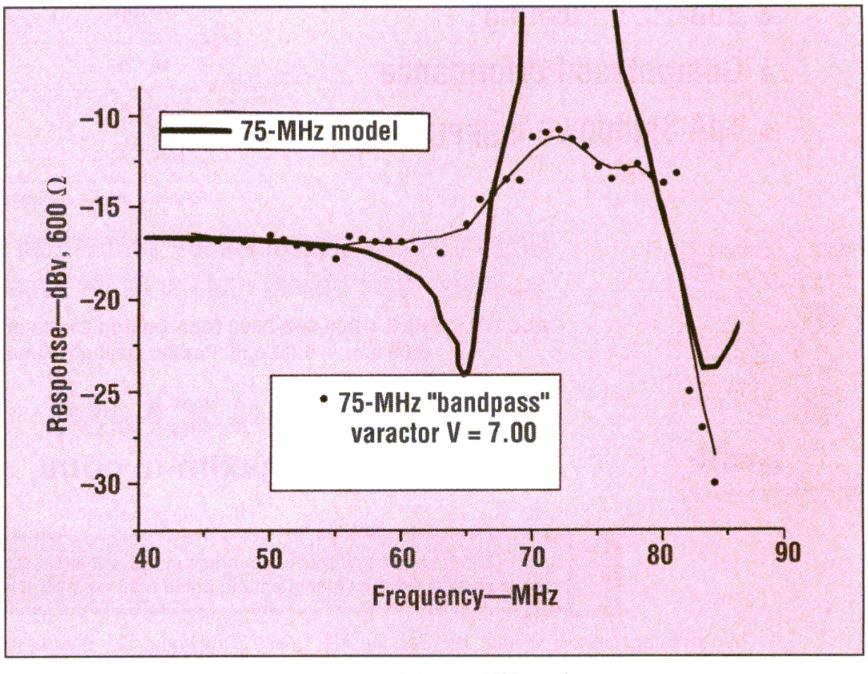

8. This graph shows the response of the 75-MHz unit.

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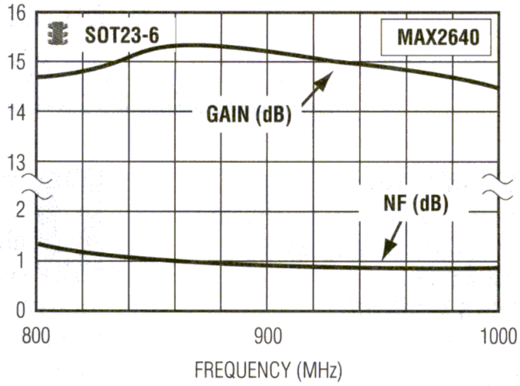

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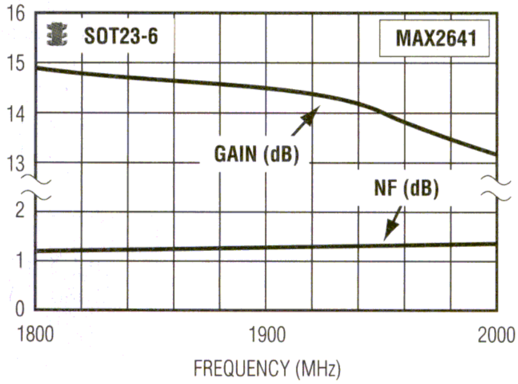
- 3V, 3.5mA

LNA PERFORMANCE TUNED @ 1900MHz



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Measure Digital Satellite Signals Accurately

Digital signals from a downlink require specific measurement techniques to set the proper levels at the head end.

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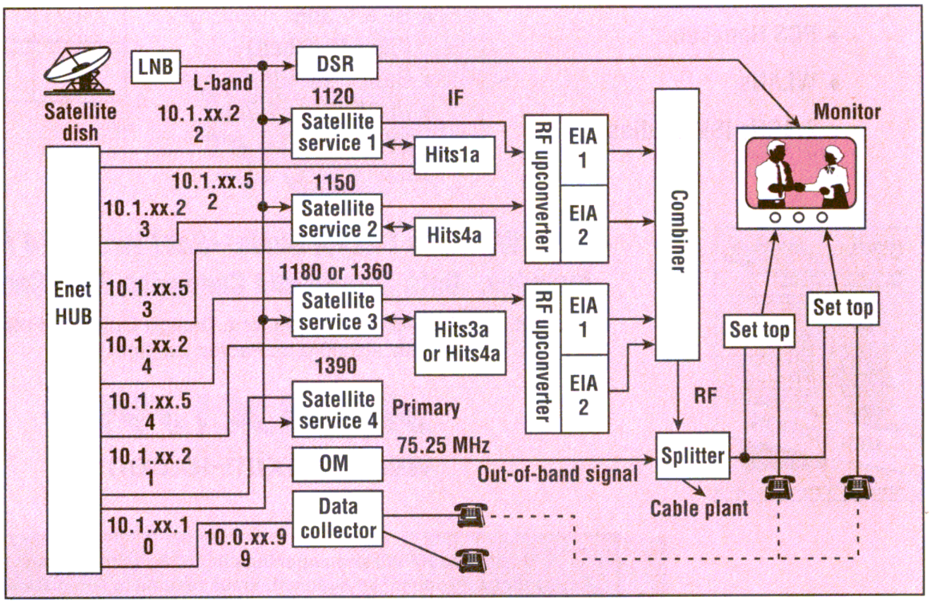
DIGITAL signals that are used increasingly in modern satellite downlinks are far more complex than their analog predecessors and more difficult to measure and set correctly. For this reason, technical and service personnel must understand the proper use of the spectrum analyzer to measure the quadrature-phase-shift-keying (QPSK) and quadrature-amplitude-modulation (QAM) digital signals that appear at a cable head end.

In a satellite downlink system, L-band digital signals come through the antenna, pass through a cable head end and go out to the subscriber sites. The basic components of this head end are the digital satellite receiver (DSR) and/or integrated receiver transcoder (IRT), the RF upconverter, and the out-of-band modulator (OM) [see figure].

The DSR is an integrated QPSK receiver/decoder that accepts L-band inputs, and outputs analog video as well as stereo audio. The input level ranges from –65 to –25 dBm per carrier. An IRT also receives QPSK-modulated L-band inputs from a satellite downconverter and produces 64QAM-modulated intermediate-frequency (IF) signals for cable trans-



In a cable-television (CATV) head-end system, digital downlink L-band signals from a satellite are received and processed before being sent out to subscribers. This type of digital transmission is replacing traditional analog systems.

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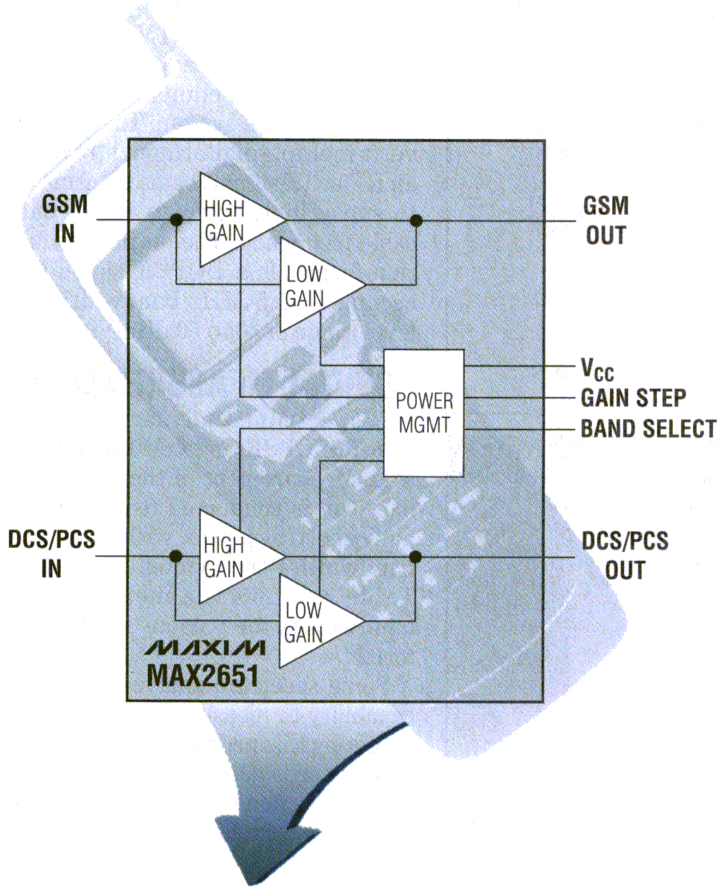

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mission in a 6-MHz-wide RF television channel. QPSK input levels range from –65 to –25 dBm per carrier and the 44-MHz output is nominally 30 dBmV. The RF upconverter converts IF signals from the IRT to RF, with the modulation format dependent on the input. In a digital transmission system, the output is 64 or 256QAM. The IF input range is nominally 30 dBmV and the RF output is adjustable to 61 dBmV. In a digital cable-television (CATV) head end, the OM provides a bridge between the digital head-end equipment and the RF distribution system. The OM converts multiple digital input streams to an RF or IF signal that is transmitted to subscribers over an out-of-band cable frequency. The output modulation format is a DQPSK signal that occupies a 1.5-MHz bandwidth at a level range of 30 to 50 dBmV.

Table 1: Digital signal formats and bandwidth

Modulation format	Channel bit rate	Symbol rate	Measurement bandwidth
QPSK/satellite 24-MHz signal bandwidth	39.02 Mb/s	19.51048 MSamples/s	19.51 MHz
QPSK/satellite 36-MHz signal bandwidth	58.53 Mb/s	29.26572 MSamples/s	29.266 MHz
QPSK/OM 1000	2.048 Mb/s	1.024 MSamples/s	1.024 MHz
64QAM	30.3414 Mb/s	5.0569 MSamples/s	5.057 MHz
256QAM	42.884 Mb/s	5.3605 MSamples/s	5.361 MHz

Table 2: Correction factors for 300-kHz RBW

Modulation format	Symbol rate MSamples/s	Measurement bandwidth (MHz)	Bandwidth correction factor (dB)	Spectrum-analyzer correction factor (if required) (dB)	Total correction factor (dB)
QPSK/satellite 24-MHz signal bandwidth	19.51048	19.51	18.1	2	20.1
QPSK/satellite 36-MHz signal bandwidth	29.26572	29.266	19.9	2	21.9
QPSK/OM 1000	1.024	1.2	6.0	2	8.0
64QAM	5.0569	5.057	12.3	2	14.3
256QAM	5.3605	5.361	12.5	2	14.5

Table 3: Spectrum-analyzer settings for L-Band measurements

Parameter	Spectrum-analyzer settings
Mode	Spectrum analyzer
Center frequency	Center of channel under test
Amplitude units	dBm
Input impedance	75 Ω
Reference level	Positioned to upper one-tenth of display
Span	Wide enough to capture entire signal under test
RBW	300 kHz
VBW	30 kHz or less (preferably 3 kHz)
Video averaging (if applicable)	On
Video filtering (if applicable)	On

DIGITAL MEASUREMENTS

A spectrum analyzer is used to measure QPSK and QAM digital signals. Measurement of the digital signal level is conducted using the signal's spectral power within its 3-dB bandwidth—the symbol bandwidth. Table 1 shows the measurement bandwidths for various modulation formats.

Some spectrum analyzers lack the capability to measure channel power. To get by this problem, the total spectral power is accounted for by applying a correction factor to the measurement. The correction factor is:

$$CF = 10\log(\text{Symbol bandwidth}/\text{Resolution bandwidth}) + \text{Spectrum analyzer correction factor}, \quad (1)$$

where:

the resolution bandwidth = 300 kHz typically.

On analyzers that have a true rms detection mode, the correction factor is not needed.

Two factors account for the spectrum-analyzer correction factor—log detected noise (2.5 dB) and 3-dB/noise-power bandwidths (−0.5 dB). This makes the total correction factor 2 dB.

The log-detected noise-correction factor is caused by the spectrum analyzer averaging its output of the log scale and computing the average of the log instead of the log of the average. This log precessing causes an under-response to noise and noise-like signals by 2.51 dB. The equivalent noise-power bandwidth of the IF filters is actually wider than indicated by approximately 0.5 dB. For a more detailed explanation of these correction factors, see Hewlett-Packard Application Note 1330.[1]

Table 2 shows the calculated bandwidth correction errors for several symbol rates at a resolution bandwidth of 300 kHz.

L-BAND MEASUREMENTS

L-band signals enter either the IRT or the DSR. Before beginning measurements, the spectrum analyzer should be powered up for at least a half-hour and be calibrated. The instrument's manual gives the proper calibration procedure. The information in Table 3 provides the analyzer settings to properly adjust the instrument and measure the digital signals (note: these settings are for a generic spectrum analyzer, and actual instrument settings may vary).

When the signal is displayed properly and the analyzer is adjusted, the marker should be moved to the center of the signal and the power level as displayed on the unit recorded. Table 2 gives the proper bandwidth correction factor for the signal being measured. The following formula can be used to calculate the actual power level of the signal:

Channel Power = Displayed marker level (dBm) + bandwidth or total correction factor
Acceptable L-band input levels are between 265 and 225 dBm. (2)

At the CATV head end, digital signal power from the IRT, RF Upconverter, and OM must be measured within its entire occupied bandwidth. Unlike analog signals which have most of their energy concentrated in the visual carrier, digital signal power is concentrated uniformly throughout the occupied bandwidth. Some spectrum analyzers have a function that enables automatic calculation of the actual power level of the digital signals. If the instrument does not have this function, a correction factor can be added to the observed marker display level. Table 4 provides the analyzer settings for CATV head-end measurements.

Once the signal is displayed and the analyzer is adjusted, move the marker to the center of the signal and record the power as displayed on the unit. Table 2 contains the bandwidth correction factors for the signal being measured. The following formula calculates the actual power level of the signal:

Channel Power = Displayed marker level (dBmV) + bandwidth or total correction factor (3)

Table 4: Spectrum-analyzer settings for CATV head-end measurements

Parameter	Spectrum-analyzer settings
Mode	Spectrum analyzer
Center frequency	Center of channel under test
Amplitude units	dBmV
Input impedance	75 Ω
Reference level	Positioned to upper one-tenth of display
Span	10 MHz (QAM)/3 MHz (QPSK)
RBW	300 kHz
VBW	30 kHz or less (3 kHz)
Video averaging (if applicable)	On
Detector mode (if applicable)	SMPL
Video filtering (if applicable)	On

Accurate analog measurements can be achieved using the sync tip of the signal. Analog signal levels are defined by the sync tip value while digital signals are defined by their average value. In a multiple-signal environment, the combined average value of the signals contributes to the power loading of active devices in the cable system. The average value of the analog signal varies with picture content. A conservative estimate is that the analog average is 3 dB below that of the sync tip.

Before making the measurement, the analyzer must be powered up for a half-hour and be calibrated, as with digital signals. The measurement begins by tracing out the signal with Trace Max Hold function. Then the trace fills in. Next, the maximum value of the signal is recorded to set the digital signal level to the correct value relative to the analog carrier.

SIGNAL POWER

To measure the ratio of signal power between the digital and analog signals, a reference analog channel must first be established to which the digital signals (QAM and QPSK) will be referenced. Then the analog signal is measured and its value recorded. The digital signal is also measured and its value recorded. The digital value is subtracted from the analog to determine how many decibels down the digital signal is with respect to the analog. A cautionary note: When balanced correctly, the OM level will not appear to be the same as the QAM EIA channel on the analyzer because the correction factor is 8 dB for the QPSK signal and approximately 14 dB for the 64 and 256QAM signals.

It should be remembered that digital signals are far more complex than their analog counterparts. Understanding the difference between the two--will enable service personnel to accurately measure digital signals and ensure that proper levels are maintained throughout the digital network. For additional information, refer to General Instrument Tech Tip 464960-001-99 available from the company. ●●

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DESIGN FEATURE

Large-Signal Extraction

Software Simplifies Large-Signal Transistor Modeling

Software for extracting transistor parameters is essential to successful large-signal modeling of MESFETs and HEMTs.

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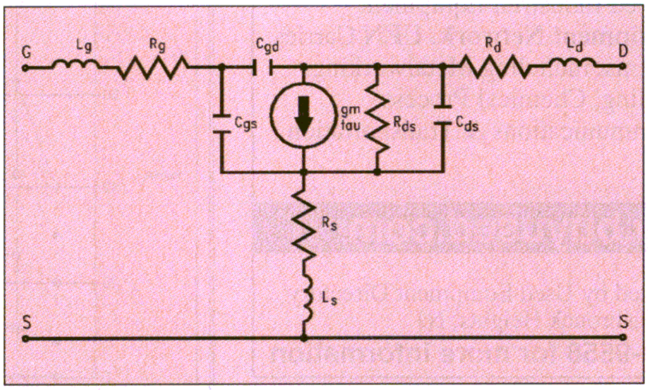
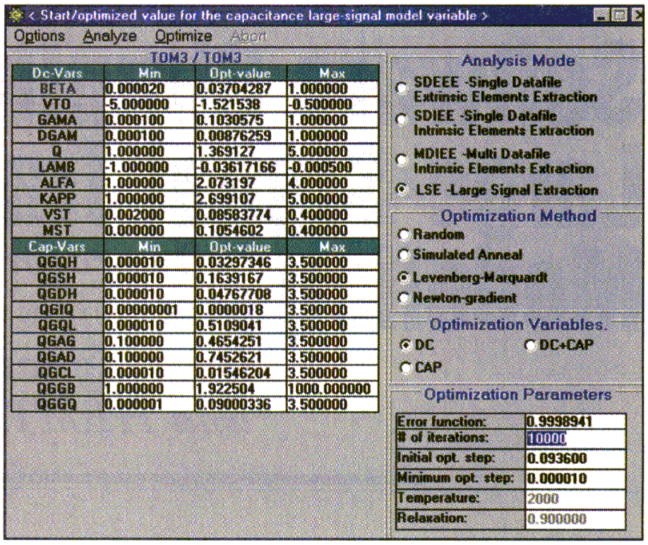
MODEL generation is an essential part of any computer-aided-engineering (CAE) system. Large-signal models are particularly important for studying the inherent nonlinear behavior of semiconductor devices and semiconductor-based active component designs. The LASIMO (large-signal-modeling) software developed by Optotek eases the task of creating and modifying metal-semiconductor-field-effect-transistor (MESFET) and high-electron-mobility-transistor (HEMT) models for nonlinear computer simulation.[1]

A designer using LASIMO is provided with a set of standard transistor models (Curtice, Advanced Curtice, Curtice-Ettenberg, Materka-Kacpryak, Lehovic-Zuleeg, TOM1) as well as the ability to incorporate up to 5 DC and 5 capacitance user-defined models.[2, 3] Model parameters are optimized to match actual transistors by fitting measured and modeled bias dependence of the device characteristics. The user-defined models are implemented as dynamic link libraries (DLLs) which are created and added outside LASIMO, and linked to the program at run time. Each of the DLLs is generated from a set of project files written in the "C" language which are loaded into a C compiler. The default DC model is the Curtice model, and the default capacitance model is the Basic Semi-Junction Model. The user has only to edit a single function that computes the nonlinear parameters. This function receives a set of arguments from LASIMO and the function returns back the required nonlinear parameters. The user can define and access up to 13 large-signal parameter coefficients to be optimized for each DC model and up to 15 large-signal parameters for each capacitance model. These DC and capacitance large-signal parameters can be selected by the user to suit various nonlinear sim-



1. The small-signal model of a discrete GaAs MESFET used in the LASIMO modeling shows the configuration of the DC and capacitance parameters along with the parasitic resistances and inductances.



Dc-Vars	Min	Opt-value	Max
BETA	0.000020	0.03704287	1.000000
VTO	-5.000000	-1.521538	-0.500000
GAMA	0.000100	0.1030575	1.000000
DGAM	0.000100	0.00876259	1.000000
Q	1.000000	1.369127	5.000000
LAMB	-1.000000	-0.03617166	-0.000500
ALFA	1.000000	2.073197	4.000000
KAPP	1.000000	2.699107	5.000000
VST	0.002000	0.08583774	0.400000
MST	0.000000	0.1054602	0.400000
Cap-Vars	**Min**	**Opt-value**	**Max**
QGQH	0.000010	0.03297346	3.500000
QGSH	0.000010	0.1639167	3.500000
QGDH	0.000010	0.04767708	3.500000
QGIQ	0.00000001	0.0000018	3.500000
QGQL	0.000010	0.5109041	3.500000
QGAG	0.100000	0.4654251	3.500000
QGAD	0.100000	0.7452621	3.500000
QGCL	0.000010	0.01546204	3.500000
QGGB	1.000000	1.922504	1000.000000
QGGQ	0.000001	0.09000336	3.500000

Error function:	0.9998941
# of iterations:	10000
Initial opt. step:	0.093600
Minimum opt. step:	0.000010
Temperature:	2000
Relaxation:	0.900000

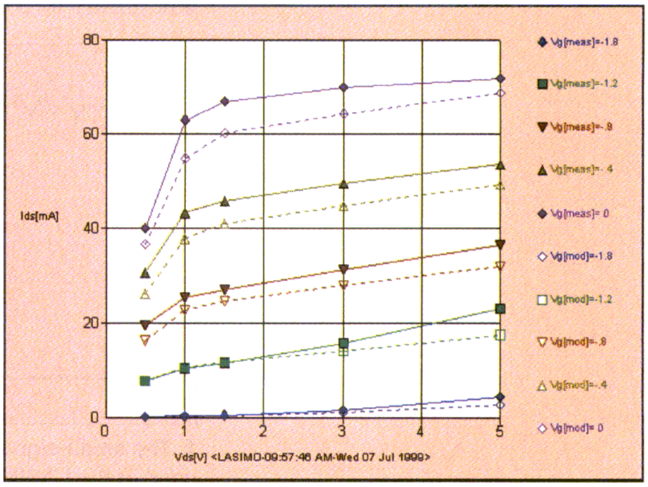
2. This LASIMO file form illustrates the way the optimized TOM3 DC-and capacitance-model parameters given in Tables 7 and 8 are presented on the computer monitor.

ulators.

The following advanced large-signal models have been introduced in the latest upgrade to LASIMO: Triquint Own Model Level 3 (TOM3), TOM3 Modified, and Alpha Own Model (AOM).

THE TOM3 MODEL

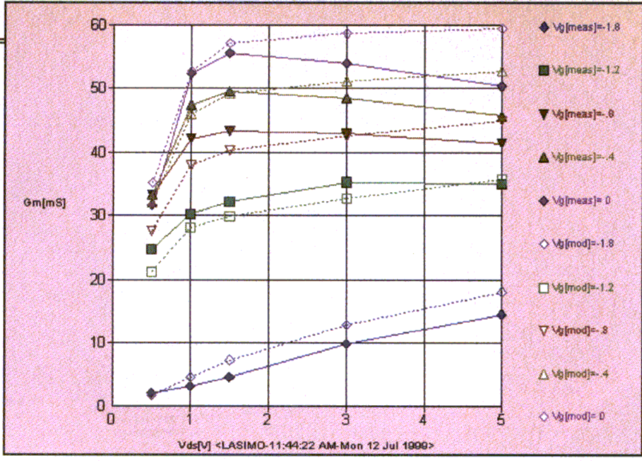
TOM3 is a comprehensive model for GaAs MESFETs which was developed to improve existing MESFET capacitance models for SPICE using conservation of charge in the implanted layer.[4] The capacitance model calculates the gate charge from the drain current and the gate capacitances from the drain conductances. Relating the gate charge to the channel current creates gate capacitances dependent on the channel-current derivatives linking the small-signal model to the large-signal equations. Drain-dispersion and self-heating effects are modeled by a GD model using a set of device equations and a



3. Drain-to-source current (I_{ds}) is shown in modeled and measured form.

specific subcircuit in SPICE.[5,6] Model-parameter extraction is performed using the newly introduced user-defined TOM3 model within LASIMO.

Tables 1 and 2 list the modeled and extracted TOM3 DC and capacitance parameters. Temperature dependence is represented. The TOM3 capacitance model requires the output DC non-linear parameters, I_{ds}, G_m, and G_{ds}. As part of the current upgrade, the capacitance user-defined models have been modified to pass into them the DC model parameters so that the output DC nonlinear parameters are calculated and applied in the capacitance model calculations.



4. These curves show the modeled versus measured results for G_m.

The user-defined TOM3 model returns the nonlinear parameters listed in Tables 3 and 4 (note: for SPICE comparison, the output DC and capacitance parameters must be multiplied by the transistor areas). The nonlinear TOM3 DC and capacitance equations are as follows. For the DC Model, the set of TOM3 DC equations relating the output parameters to Table 3 to the input parameters of Table 1 are:

$$I_{ds} = BETA\ V_g^{\,Q}\ KTanh(1 + LAMB\ V_{dsi}) \quad (1)$$

$$I_{dsRF} = BETA\ V_{gRF}^{\,Q}\ KTanh(1 + LAMB\ V_{dsi}) \quad (2)$$

where:

$$V_g = Q\ V_{st}\ ln\{exp[V_{gsi} - VTO + (GAMA - DGAM)\ V_{dsi}\ /\ (Q\ V_{st})] + 1\} \quad (3)$$

$$V_{gRF} = Q\ V_{st}\ ln\{exp[(V_{gsi} - VTO + GAMA\ V_{dsi})\ /\ (Q\ V_{st})] + 1\} \quad (4)$$

$$V_{st} = \{VST + AVST(TJ - TNOM)\}\ [1 + \{MST + AMST(TJ - TNOM)\}\ V_{dsi}] \quad (5)$$

$$KTanh = (ALFA\ V_{dsi})/[1 + (ALFA\ V_{dsi})^{KAPP}]^{1/KAPP} \quad (6)$$

where:

V_{dsi} = the intrinsic drain-source voltage,

V_{gsi} = the intrinsic gate-source

voltage, and

TNOM= the reference temperature = 298°K.

The G_m and G_{DS} equations are the conventional derivatives of I_{drf} with respect to V_{gs} and V_{ds}, respectively.

For the capacitance model, the set of TOM3 capacitance equations relating the output parameters of Table 4 to the input capacitance parameters of Table 2 are:

$$C_{gs} = C_{gsl}F_t + C_{gsh}(1 - F_t) + (Q_{gl} - Q_{gh})(\partial F_t / \partial V_{gsi}) + QGG0 \quad (7)$$

$$C_{gd} = C_{gdl}F_t + C_{gdh}(1 - F_t) + (Q_{gl} - Q_{gh})(\partial F_t / \partial V_{gdi}) + QGG0 \quad (8)$$

where:

$$C_{gsh} = (g_m + g_{ds})\{QGQH /(I_{dsRF} + QGI0)\} + QGSH \quad (9)$$

$$C_{gdh} = -g_{ds}\{QGQH /(I_{dsRF} + QGI0)\} + QGDH \quad (10)$$

$$Q_{gh} = QGQH\ ln(1 + I_{dsRF} / QGI0) + QGSH\ V_{gsi} + QGDH\ V_{gdi} \quad (11)$$

$$C_{gsl} = F_l\{QGAG\ cosh(QGAD\ V_{dsi}) + QGAD\ sinh(QGAD\ V_{dsi})\} + QGCL \quad (12)$$

$$C_{gdl} = F_l\{QGAG\ cosh(QGAD\ V_{dsi}) - QGAD\ sinh(QGAD\ V_{dsi})\} + QGCL \quad (13)$$

$$Q_{gl} = F_l\ cosh(QGAD\ V_{dsi}) + QGCL\ (V_{gsi} + V_{gdi}) \quad (14)$$

$$F_l = QGQL\ exp\{QGAG(V_{gsi} + V_{gdi})\} \quad (15)$$

$$F_t = exp(-QGGB\ I_{dsRF}\ V_{dsi}) \quad (16)$$

$$\partial F_t / \partial V_{gsi} = -QGGB\{I_{dsRF} + (g_m + g_{ds})V_{dsi}\}F_t \quad (17)$$

$$\partial F_t / \partial V_{gdi} = QGGB\{I_{dsRF} + g_{ds}\ V_{dsi}\}F_t \quad (18)$$

where:

V_{dsi} = the intrinsic drain-source voltage,

V_{gsi} = the intrinsic gate-source voltage, and

V_{gdi} = the intrinsic gate-drain voltage.

A variant of the TOM3 model is also provided where the three parameters assigned to temperature (see Table 1) are not included. For many applications, this room-temperature model can be used quite effectively. Speed of extraction is improved.

Table 1: Input DC parameters to the TOM3 model

Parameter name	Meaning	Unit
BETA	Transconductance coefficient	A/V
VTO	Threshold voltage	V
GAMA	Threshold-shifting parameter	–
DGAM	Extra RF drain-pull coefficient	–
Q	Power-law parameter	–
LAMB	Slope of drain characteristic in the saturated region	/V
ALFA	Slope of drain characteristic in the linear region	1/V
KAPP	Knee-function parameter	–
VST	Subthreshold slope	V
MST	Subthreshold slope-drain parameter	/V
AVST	VST linear temperature coefficient	V/K
AMST	MST linear temperature coefficient	/KV
TJ	Device temperature	K

Table 2: Input capacitance parameters to the TOM3 model

Parameter name	Meaning	Unit
QGQH	Charge parameter	pF. V
QGSH	Charge parameter	pF
QGDH	Charge parameter	pF
QG10	Charge parameter	A
QGQL	Charge parameter	pF. V
QGAG	Charge parameter	V^{-1}
QGAD	Charge parameter	V^{-1}
QGCL	Charge parameter	pF
QGGB	Charge parameter	$A^{-1}V^{-1}$
QGG0	Charge parameter	pF
IDS	Channel current	A
GM	Transconductance	A/V
GDS	Output conductance	A/V

Table 3: Output DC parameters from the TOM3 model

Parameter name	Meaning	Unit
IDS	Channel current	A
GM	Transconductance	A/V
GDS	Output conductance	A/V

THE AOM MODEL

The Alpha Own Model (AOM) is a comprehensive model for GaAs MESFETs which expands on aspects of the Level 1 TOM to account for dispersion, self-heating effects, and charge conservation. A set of capacitance and charge equations are used for consistent small- and large-signal models. Transconductance and output conductance dispersion are modeled by combining a feedback network and a subcircuit which describes the self-heating effects. The new model accurately predicts

Table 4: Output capacitance parameters from the TOM3 model

Parameter name	Meaning	Unit
CGS	Gate-source capacitance	pF
CGD	Gate-drain capacitance	pF

Table 5: Input DC parameters to the AOM model

Parameter name	Meaning	Unit
ALPHA	Current saturation coefficient	1/V
BETA	Transconductance coefficient	A/V
GAMMA	DC drain-pull coefficient	–
VTO	Threshold voltage	V
Q	Power-law exponent	–
CT	Self-heating coefficient	1/W
CFB	Extra RF drain-pull coefficient	–
A	Inverse tailing factor	1/A

the I-V, CV, bias-dependent S-parameter, waveform, power, and linearity characteristics of the MESFET. This model has been implemented in PSPICE.[7,8] Model-parameter extraction is performed using the newly introduced user-defined AOM model within LASIMO. Tables 5 and 6 list the modeled and extracted DC and capacitance parameters. The user-defined AOM model returns the same non-linear parameters listed in Tables 3 and 4.

AN EXTRACTION EXAMPLE

To illustrate the use of the advanced reconfigurable models incorporated in LASIMO, the TOM3 model was fitted to a discrete GaAs MESFET. The transistor was a low-power, low-noise, ion-implanted device, with a gate width of 300 μm and gate length of 0.5 μm.[9] The ion-implantation energy was 150 KeV and the open-channel current after recess was 100 mA. Wafer-probed transistors were measured using a Wiltron 360 vector network analyzer (VNA). The transistors were powered using two Keithley 236 programmable power supplies. The VNA and the power supplies were controlled via the IEEE-488 GPIB bus using LASIMO's data-acquisition (DAQ) module. S-parameters and drain current were measured at the biases of V_{ds} = 0.5, 1.0, 1.5, 3.0, and 5.0V and V_{gs} = –1.8, –1.2, –0.8, –0.4, and 0.0V. A total of 25 data points were thus generated to represent the device's I-V curve. Figure 1 shows the small-signal model chosen to represent the device.

The device-parasitic resistances were measured using in-house Process Control Monitor characterization techniques providing the following values:[10]

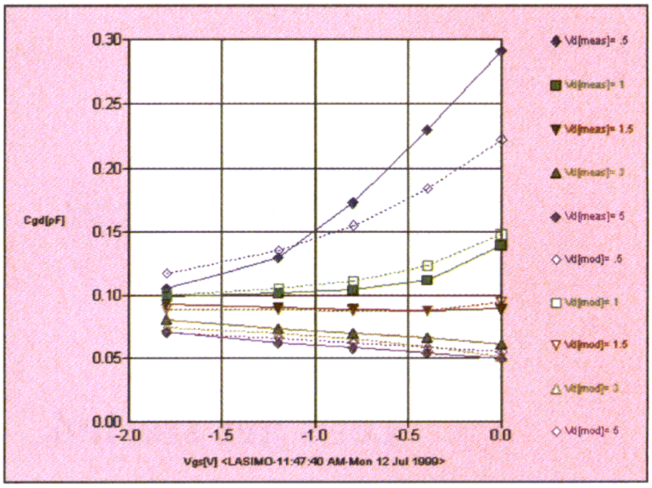
Rs = 1.3 Ω
Rd = 2.7 Ω
Rg = 1.1 Ω

The parasitic inductances were then extracted using LASIMO with the following results:

Ls = 36.35 pH
Ld = 4.21 pH
Lg = 77.46 pH

With the bias-invariant parasitics, LASIMO was used to extract the intrinsic device models at all bias points. In this process, LASIMO gathers in memory the arrays of transconductance G_m, output conductance G_{ds}, gate-source capacitance C_{gs}, and gate-drain capacitance C_{gd}. These arrays are a function of the intrinsic biases V_{gsi} and V_{dsi}. The intrinsic biases are calculated by subtracting the impressed voltages from the parasitics.

LASIMO then optimizes parameter data to the TOM3 device equations representing the non-linear I_{ds}, G_m, G_{ds}, C_{gs}, and C_{gd}. The initial DC model values for optimization must be chosen judiciously so that the optimization process converges rapidly. A two-step optimization process is used where in the first step only the points corre-



5. These are the modeled versus measured results for gate-to-drain capacitance (Cgd).

Table 6: Input capacitance parameters to the AOM model

Parameter name	Meaning	Unit
CO	Intrinsic capacitance at Vds = 0 Vgs = o	pF
SG	Slope parameter of Cgs versus V_{gs}	–
DELTA	Cross product coefficient	$1/V^2$
DC	V_{ds} range that Cgs changes	V
DK	Capacitance Cgd undershoot range parameter	1/V
Cgo	Residual gate-source capacitance	pF
Cgdo	Residual gate-drain capacitance	pF
Cf	Variation range parameter-Cmin/Cgso	

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Table 7: Extracted DC parameters for the TOM3 room-temperature model

Parameter name	Initial values	Intermediate values	Final values
BETA	0.0002	0.03642849	0.03704287
VTO	-0.50	-1.629549	-1.521538
GAMMA	0.062	0.1317646	0.1030575
DGAM	0.01	0.03957766	0.00876259
Q	1.60	1.327517	1.369127
LAMB	-0.044	-0.04223719	-0.03617166
ALFA	2.80	1.889915	2.073197
KAPP	3.30	3.533976	2.699107
VST	0.05	0.05983577	0.08583774
MST	0.0001	0.00009958	0.1054602
Error function	99.55 percent	0.1424853 percent	0.9998941 percent

Table 8: Extracted capacitance parameters for the TOM3 room-temperature model

Parameter name	Initial values	Intermediate values	Final values
QGQH	0.0004	0.03442526	0.03297346
QGSH	0.0007	0.4892903	0.1639167
QGDH	0.00031	0.01748541	0.04767708
QGI0	0.33e-6	0.00805732	0.0000018
QGQL	0.00080	0.3575123	0.5109041
QGAG	0.75	0.9778292	0.4654251
QGAD	0.65	1.301038	0.7452621
QGCL	0.00022	0.02876367	0.01546204
QGGB	10.0	11.133190	1.922504
QGG0	0.0001	0.02723242	0.09000336
Error function	99.32 percent	0.02824027 percent	0.9998941 percent

sponding to the DC curve where VG = 0 V and VDS = +0.5, +1.0, +1.5, +3.0, and +5.0 VDC are optimized, and in the second step the entire I-V characteristic is optimized. The initial model values selected for the first optimization step are chosen arbitrarily. In this case they were selected from default values recommended in the LASIMO manual.[11] The optimized model values from the first step—designated as intermediate values—become the initial values for the second step, resulting in optimized values for the second step. A combination of Random and Levenberg-Marquardt optimizers were selected and care was exercised that values were within limits and that they were realistic. The initial, intermediate, and final TOM3 model parameters are displayed in Tables 7 and 8. The corresponding LASIMO file form is shown in Fig. 2. The fitted parameters demonstrate versatility of TOM3 for giving a best fit for I_{ds}, G_m, and C_{gd} (see figures 3, 4, and 5). ●●

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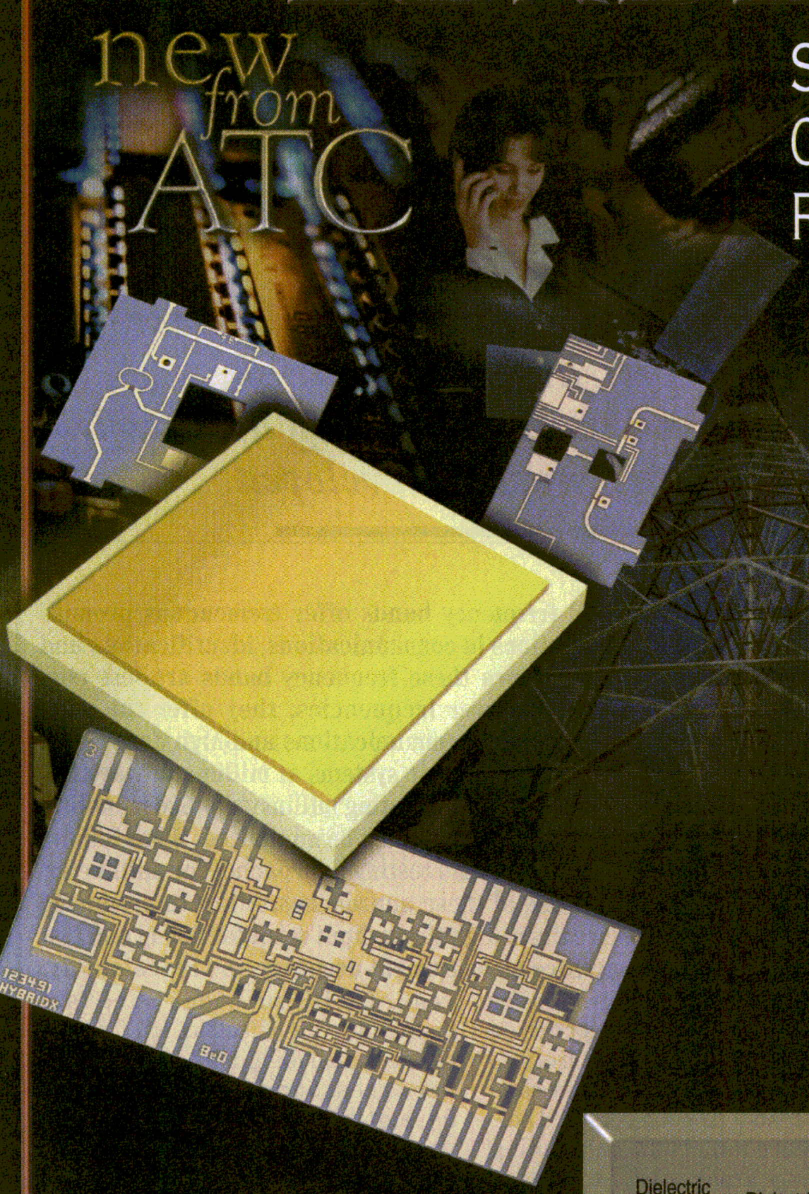

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BA	36	–80 ±30 PPM/°C	0.10 to 4.3
BB	31	0 ±30 PPM/°C	0.09 to 3.7
CA	62	0 ±30 PPM/°C	0.17 to 7.4
CC	130	–750 ±200 PPM/°C	0.37 to 15
DA	165	–1500 ±500 PPM/°C	0.46 to 20
DB	200	±7.5% max. change (non-linear)	0.56 to 24
HC	350	–2000 ±500 PPM/°C	0.98 to 42
EA	650	–4700 ±1500 PPM/°C	1.8 to 77
EC	650	±10% max. change (non-linear)	1.8 to 77
J	1100	+5% to –15% max. change (non-linear)	3.1 to 130
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Meeting The Test Needs Of Millimeter-Wave MMICs

In order for emerging millimeter-wave applications to flourish, improved manufacturing and testing techniques must be developed.

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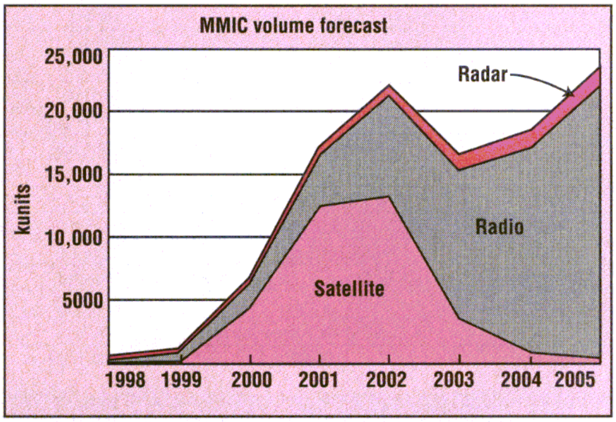
MILLIMETER-WAVE frequency bands offer tremendous promise for emerging applications in communications, identification, and collision avoidance. Since these frequency bands are less congested than bands at lower frequencies, they offer the wide bandwidths needed for multichannel communications and high data rates. But in order to realize practical electronic systems at millimeter-wave frequencies, the costs of manufacturing and testing millimeter-wave monolithic-microwave integrated circuits (MMICs) must be minimized. What follows is a review of current manufacturing and testing approaches for millimeter-wave MMICs and the limitations of these approaches. A key to producing low-cost millimeter-wave circuits is high-throughput manufacturing and test procedures that do not sacrifice measurement accuracy and repeatability.

Most of the projected growth for millimeter-wave MMICs depends on the abilities of IC manufacturers to design and produce products at competitive costs. In some cases, their customers compete in demanding applications areas, such as automotive and wireless communications markets, where low prices are critical to the success of the market.



1. Current forecasts for millimeter-wave applications indicate strong expected growth in satellite, microwave radio, and radar markets.

One of the fastest-growing areas for these MMICs is for local multipoint distribution systems (LMDS), which are fixed wireless services operating at Ka-band frequencies from 28 to 32 GHz.[1,2] Providing more than 1 GHz of usable bandwidth, LMDS licenses were auctioned off by the Federal Communications Commission (FCC) early last year. LMDS provides adequate bandwidth for wireless multimedia services, including high-speed Internet access, and has spurred the development of many device-level products. During 1998 alone, for example, gallium-arsenide (GaAs) manufacturers introduced more than 50 devices for the 28-to-32-GHz band, with many more devices and ICs expected to follow during 1999.

In addition to terrestrial-based LMDS applications, many emerging satellite-communications systems make use of millimeter-wave frequencies, including those from Teledesic, Celestri, and the European Skybridge program.[3] Although the architectures of the low-earth-orbit (LEO) satellite systems differ, they generally employ millimeter-wave frequencies for their uplink communications.

Point-to-point digital radios at 23, 28, and 38 GHz serve as the back-haul links between cellular and personal-communications-services (PCS) antenna sites. With the increase in

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Using Dielectric material with high "Q" results low Insertion Loss.

3. Linearity

Phase is linearly varied by controlling rotated angle of axis which is co-centered with transmission line.

4. High Power Handling Capability

Low IMD with non-contract structure.

Product Code No.	A type : KPH90OSCL000 B type : KPH90OSCL001		
Frequency Range	~ 1GHz	1 ~ 2GHz	2 ~ 3GHz
Insertion Loss (max.)	0.15dB	0.25dB	0.35dB
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		

SMD type is also available.

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smaller cell sites, such as microcells and picocells (often for in-building wireless extensions), microwave/millimeter radio manufacturers face competitive pressures to supply lower-cost back-haul radios to connect these smaller sites.

Traditional millimeter-wave designs have been based on discrete semiconductor devices, such as Gunn and Impatt diodes. But as the growth of millimeter-wave markets drives the need for higher volumes of devices at lower costs, MMIC technology offers a more practical solution than discrete device technology for meeting product-cost targets. Growth in major millimeter-wave markets, including satellite, radio, and radar markets, is expected to increase steadily for the next several years (Fig. 1).[4]

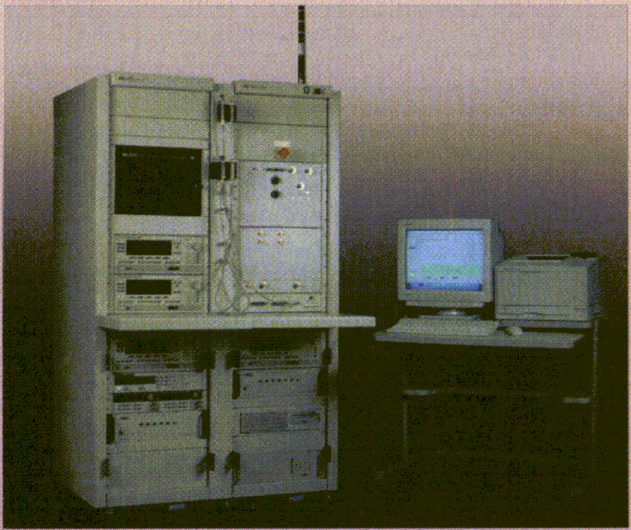
The increases in millimeter-wave device volumes have forced GaAs and indium-phosphide (InP) semiconductor manufacturers to rethink their test strategies. Since cost-of-test pressures tend to grow in direct proportion to the size of a market, device manufacturers must consider test times for MMICs in terms of seconds rather than minutes or hours. Millimeter-wave devices have traditionally been tested with manual, labor-intensive techniques. Since volumes for these devices have been

AUTOMATED TESTING REACHES 45 GHz

The HP 84000 millimeter-wave series is offered in two configurations. The HP E7380A is an on-wafer test system designed to test monolithic microwave integrated circuits (MMICs), and the HP E6520A tests packaged millimeter-wave modules. Both models measure gain, gain compression, efficiency, harmonic distortion, power, spurious signals, S-parameters, supply currents, voltages, noise figure (NF), mixer conversion gain, and Nth order intermodulation.

When incorporated in a high-volume production line, the testers (see figure) can evaluate thousands of devices per day, with little operator invention. Typical total test times per device-under-test (DUT) range from less than 1 s to 1 min. for commercial devices, and 30 to 60 min. per test plan for military and satellite components, depending on the number and type of tests specified.

Measurements of power and noise are made through a high-speed tuned VXI receiver, rather than much slower NF meters and power meters. Noise sources and power-meter sensors are only needed during the calibration process. The system is based on VXI instrumentation, with a high-speed MXI bus to optimize measurement throughput. For high-calibration stability, reflectometers are placed as close as possible to the DUT.



The millimeter-wave series is the latest addition to the 84000 high-frequency production test systems.

The system's object-oriented software is written in C++ and runs on the HP-UX operating system on HP 9000 series workstations. It employs an OSF/Motif-compliant graphical user interface (GUI) that enables the operator to activate and control the system with the keyboard and mouse. The HP-UX operating system provides support for existing local-area-network (LAN) infrastructures, enabling simple access to production data for full off-line statistical data analysis. Standard data output formats are available in ATDF (ASCII test data format), standard test-data format (STDF), as well as comma-delimited ASCII.

The capabilities of the basic system can be extended by adding optional high- and low-frequency test ports, local-oscillator (LO) stimuli, analog and digital DUT control lines, and high-power transmit and receive options. Future versions will include additional measurement capabilities. With the current system, test times of 90 minutes per wafer or less have been achieved.

SPEEDY TEST PLANS

The millimeter-wave production test system minimizes the time required for test-plan development through a graphical forms-based user interface. The GUI enables test plans to be defined or modified efficiently on-line or off-line from the manufacturing test system. Test developers can select from among a wide range of factory-preset stimuli and measurements that can be quickly added to the test plan using the GUI.

The availability of standardized test-plan elements also ensures consistency in test-plan development between production engineers. Test-plan developers have a wide range of available function options, including embedding transforms, test limits, and wafer-mapping functions. Additional development tools include simple definition of test variables and math blocks as well as conditional branching algorithms. Built-in optimization tools help reduce test times through the automatic optimization of test plans. Programming experience is not needed to develop or revise effective test plans with the new production millimeter-wave system.

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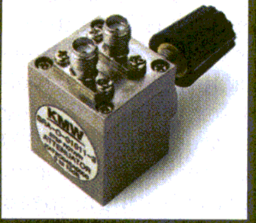
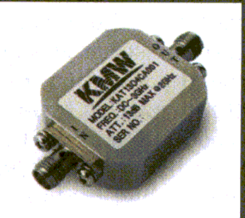
KAT1OO4SA000

KAT2OO4SA000



KAT13O4CA000



KAT13O4CA001

■ Step-Rotary Attenuators

Product Code No.	KAT1OO4SA000	KAT2OO4SA000	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

Contactless Structure for High Power Handling Capability, up to 2W average @2GHz.

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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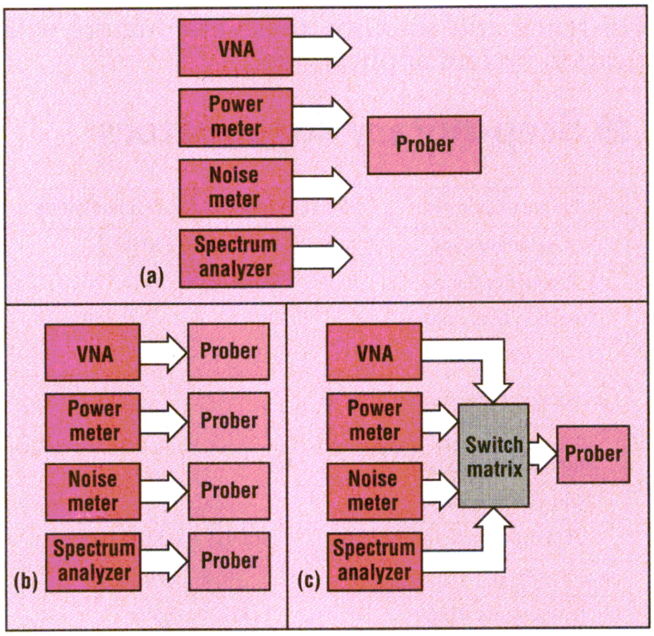
limited, these test methods have been adequate. Typically, measurements have included wafer-probe stations or precision-machined test fixtures as the interface between a device under test (DUT) and the measurement equipment.

The simplest approach has been to manually connect various bench-top instruments to a single wafer-probe station to measure the desired performance parameters (Fig. 2a). As the figure shows, measurement equipment might include a specialized millimeter-wave vector network analyzer (VNA), spectrum analyzer, power meter, and noise-figure (NF) meter, depending on the type of device being tested. In this approach, only one type of measurement, such as an NF measurement, can be performed at one time.

Another approach is the use of multiple wafer-probe stations, each dedicated to a specific measurement (Fig. 2b). A third approach makes use of a low-loss electromechanical switch matrix (Fig. 2c) to route signals to and from the test equipment to the wafer probe and the DUT. The switch matrix eliminates the need to manually connect different instruments for each measurement function, and permits the use of a single wafer-probe station for multiple measurement functions.



2. Current strategies for production testing of wave MMICs include the use of a single wafer-probe station and manual reconfiguration of test equipment (a), multiple wafer-probe stations, each dedicated to a specific test function (b), and multifunction test systems connected to a single wafer-probe station through a switch matrix (c).

EXISTING LIMITATIONS

Each of the millimeter-wave production test approaches has limitations. The first approach, for example, is the most time-consuming due to manual reconfiguration of the test equipment. Each time the equipment is reconfigured, there is risk of operator error, so that a skilled operator is needed to minimize such errors.

In comparing the different measurement strategies, a 4-in. (10.16-cm) GaAs wafer containing approximately 1200 power-amplifier (PA) die serves as a suitable millimeter-wave reference example. The first approach yields reported test times ranging from eight to 16 hours per wafer, depending on the equipment setup time, the time needed for calibration, and the time necessary to correct configuration errors.

In the second approach, the wafer must be handled more often as it is transferred from one wafer-probe station to another. The additional handling increases the risk of contamination and damage to a delicate GaAs wafer. In the first and second approach, the wafer must be re-probed several times, increasing the total test time and the likelihood of damage to the DUT pads and/or the wafer-probe tips. The second approach does drastically reduce or eliminate test-equipment configuration errors compared to the first approach, reducing the risk of having to retest the wafer. Compared to the first approach, the savings in time using the second approach stem largely from the reduction or elimination in equipment setup time and recalibration. Reported test times using the second approach range from four to six hours per wafer.

The third approach offers completely automated "one-touch" testing of a DUT. It is not as commonly used as the first two approaches due to the (sometimes severe) degradation in measurement performance and system stability caused by the switch matrix and the increased complexity of the measurement calibration. The third approach is the most difficult and expensive to automate due to its complexity. Overall system measurement speed is limited by the speed of the individual instruments, especially for noise and power measurements. Using the example above, the third approach yields test times ranging from three to five hours per wafer, depending on the sophistication of the test-system software.

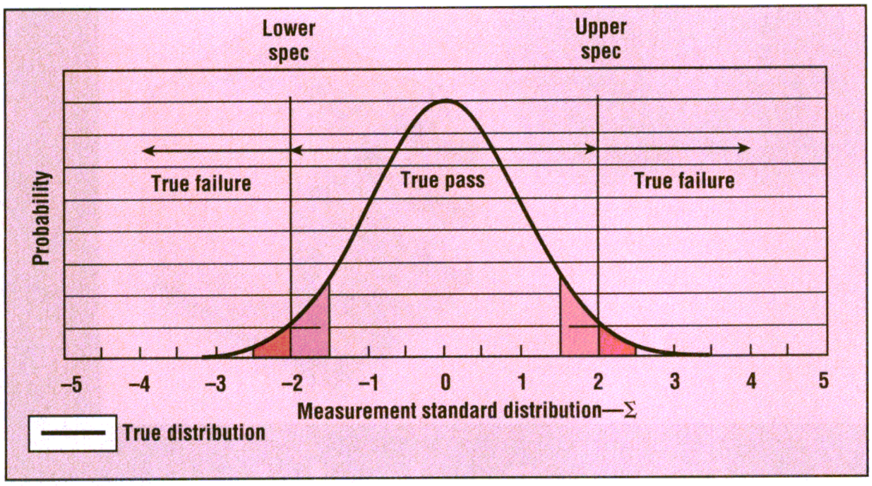
An automated production test system has been developed for the devices currently available and under development for these emerging marketplaces. The HP 84000 production test system, which provides high-speed automated multifunction measurements of MMICs, has been adapted to frequencies up to 45 GHz through a high-frequency millimeter-wave test set (see sidebar).

In addition to shorter device test times, improved test yields can also reduce the cost of testing. Test yields rely heavily upon the guard bands imposed on the measured device data, to compensate for measurement system inaccuracies and instability (Fig. 3). Errors associated with guard bands around production test data can be categorized as type 1 errors, where good parts are erroneously failed (as bad parts), and type 2 errors, where bad parts are erroneously passed (as good parts). Guard bands are normally not symmetric around specification limits due to the higher costs associated with shipping a bad device.

With improved test-system accuracy and stability, these guard bands can be tightened, thus minimizing both types of errors. Enhanced test-

system accuracy and stability can substantially increase test yields and reduce the number of dissatisfied customers and down-line expenses associated with shipping bad devices.

An example may help to demonstrate the impact of improved guard bands on production MMICs. Assume a $10 device with annual production volume of one million devices. An improvement from 2 percent to 1 percent in type I errors (failed good parts) results in an annual revenue increase of $100,000. The financial impact of shipping bad parts (type 2) is difficult to quantify. As an approximation, assume that the down-line financial impact is 10 times the price of the device. Improving type 2 errors from 0.2 percent to 0.1 percent reduces this financial impact from $200,000 to $100,000, a savings of $100,000 annually. In this example, the ability to narrow measurement guard bands using a more-accurate and stable test system results in $200,000 of either increased revenues or lower product warranty expenses.



3. Insufficient guard bands result in excessive type 1 and type 2 errors.

Global millimeter-wave MMIC markets appear ready to explode, primarily because an entirely new application, LMDS, has just become reality. Existing manufacturing and measurement strategies are not geared to produce the large numbers of low-cost ICs to support LMDS as well as other commercial and consumer millimeter-wave markets. ●●

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Calculate The Uncertainty Of NF Measurements

Simple modifications to the basic noise-figure equations can help in predicting uncertainties associated with test equipment.

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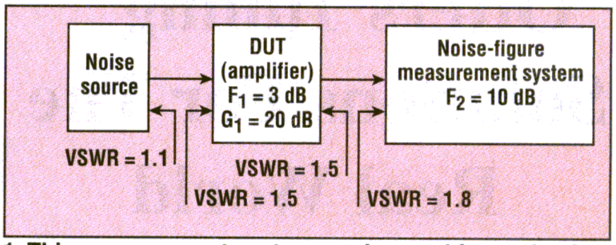
RAPID growth in satellite-communications and mobile-communications markets has increased the demand for monolithic and discrete semiconductor devices with lower noise figures (NFs). With these low-noise devices comes the pressure to reduce the NF measurement uncertainty. What follows is a model for calculating the uncertainty of NF measurements, along with an easy-to-follow example.



1. This measurement system can be used for evaluating the NF of low-noise amplifiers (LNAs) and devices.

It is often assumed that the uncertainty due to mismatch is the largest source of NF measurement uncertainty, although this is rarely the case. In fact, the uncertainty of the measurement system's NF, the uncertainty related to the noise source, the NF and architecture of the measurement system, as well as the gain of the device under test (DUT) have a significant bearing on the overall measurement uncertainty. If any of these parameters are unfavorable, the uncertainty due to mismatch will have little impact on the overall result. Engineers can waste time and money using network analysis to perform S-parameter correction of mismatches. Correction introduces a number of other issues and in reality greater attention to the other parameters in the system can yield more significant improvements.

The procedures and following example that will be shown are designed to increase an engineer's familiarity and understanding of NF measurement uncertainty calculations. A programmed example offers an easy-to-follow model for calculating the uncertainty in any measurement configuration. The model and some further examples clearly show which parameters have the most-significant impact on the uncertainty. A discussion of the main parameters includes the benefits to be gained by improving particular parameters of a measurement system. In addition, a spreadsheet is available upon request from the author (at the e-mail address listed) to automate the uncertainty calculations.

The uncertainty of NF measure-

Table 1: Log to linear transformations for noise figure and gain

Parameter	Log value	Linear = 10^(dB/10)
F_1	3 dB	1.995
F_2	10 dB	10
G_1	20 dB	100
$F_{12} = F_1 + (F_2 - 1)/G_1$	3.19 dB	2.085

Table 2: Calculating noise-figure and gain ratios figure

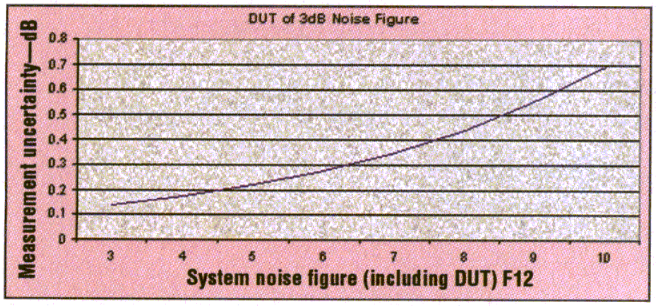
Parameter	Ratio
F_{12}/F_1	1.045
F_2/F_1G_1	0.050
$(F_2-1)/F_1G_1$	0.045
$F_{12}/F_{1)} - (F_2/F_1G_1)$	0.995

ments can be calculated by eq. 7 from the sidebar. One application where NF is important is during the testing of amplifiers. A typical setup would include a noise source, the DUT, as well as an NF measurement system (Fig. 1).

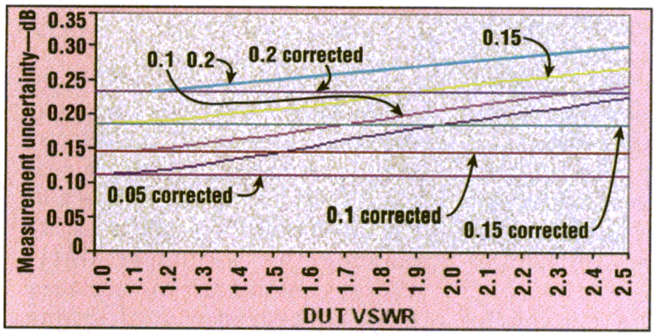
The model provides the measurement uncertainty associated with a particular NF based on the knowledge of individual VSWRs within the system and the specifications of the measurement system. For the model to be accurate, the DUT must have reasonable reverse isolation, so that a mismatch on one port does not drastically affect the impedance seen at the other port.

The first step in using the model requires the operator to refer to the test-equipment owner's manuals to calibrate the NF receiver, apply the DUT, then record the DUT's corrected gain (G_1) and NF (F_1).

Then the NF receiver's autoranging function must be switched off so that the attenuators remain in the same



2. The NF measurement uncertainty increases with increasing test-system measurement uncertainty.



3. Instruments or systems with better NF uncertainty generally outperform systems employing S-parameter correction.

position. Following this, the DUT is removed and the uncorrected NF of the test receiver (F_2) is measured. All of the decibel values must be converted into equivalent linear values and the NF value of F_{12} must be calculated according to Table 1. A series of ratios are then calculated according to Table 2. The various VSWR values can then be converted into reflection coefficients through Table 3. The resulting reflection coefficients can then be used to calculate the various impedance-matching uncertainties of the NF measure-

DERIVING THE NF UNCERTAINTY EQUATION

The general equation for the noise figure (NF) of two cascaded stages is:

$$F_{12} = F_1 + \frac{F_2 - 1}{G_1} \quad (1)$$

where:
F_1 = the linear NF of the DUT,
F_2 = the linear NF of the NF measurement receiver,
F_{12} = the linear NF of the complete system (DUT and measurement receiver), and
G_1 = the linear gain of the DUT.

Since it is the uncertainty of the DUT's NF (F_1) that is of interest, the terms can be rearranged by:

$$F_1 = F_{12} - \left(\frac{F_2 - 1}{G_1}\right) \quad (2)$$

Because F_1 is dependent on the three independent variables F_{12}, F_2, and G_1, differential calculus in the form of Taylor's Theorem can be applied to find the uncertainty of F_1:

$$\delta F_1 = \frac{\partial F_1}{\partial F_{12}}\delta F_{12} + \frac{\partial F_1}{\partial F_2}\delta F_2 + \frac{\partial F_1}{\partial G_1}\delta G_1 \quad (3)$$

where:
δF_1 = the uncertainty of the DUT NF,
δF_2 = the uncertainty of the measurement receiver's NF,
δF_{12} = the uncertainty of the complete system (the DUT and measurement receiver) NF, and
δG_1 = the uncertainty of the DUT gain.

From eq. 2:

$$\frac{\partial F_1}{\partial F_{12}} = 1$$

$$\frac{\partial F_1}{\partial F_2)} = -\left(\frac{1}{G_1}\right)$$

$$\frac{\partial F_1}{\partial G_1} = \frac{F_2 - 1}{G_1^2} \quad (3a)$$

so that:

$$\delta F_1 = \delta F_{12} - \left(\frac{1}{G_1}\right)\delta F_2 + \left(\frac{F_2 - 1}{G_1^2}\right)\delta G_1 \quad (4)$$

RF engineers generally work in logarithmic decibels, so using $NF_1(dB) = 10\log F_1$ and the standard differential coefficient:

$$Log_a \chi = \frac{1}{\chi Lna} \quad (4a)$$

results in:

$$\delta NF_1 = \frac{4.34}{F_1}\delta F_1 \quad (4b)$$

Applying the same procedure for δF_{12}, δF_2, and δG_1, results in:

$$\delta F_{12} = \frac{\delta NF_{12} F_{12}}{4.34}$$

$$\delta F_2 = \frac{\delta NF_2 F_2}{4.34} \quad \delta G_1 = \frac{\delta G_1(dB) G_1}{4.34} \quad (4c)$$

Substituting these into eq. 4 and simplifying yields:

$$\delta NF_1 = \left(\frac{F_{12}}{F_1}\right)\delta NF_1 - \left(\frac{F_2}{F_1 G_1}\right)\delta NF_2 + \left(\frac{F_2 - 1}{F_1 G_1}\right)\delta G_1\,(dB) \quad (5)$$

The three δ terms in the previous equation are due to the NF measurement receiver and the DUT. However, NF instruments rely on a calibrated noise source with a specified excess noise ratio (ENR). Clearly, there will be an uncertainty associated with this ENR and this will contribute to the overall uncertainty equation.

When the DUT is a frequency converting device with F_2 and F_{12} at different frequencies, δENR is included as part of δNF_{12}, δNF_2, and δG_1. However, when the DUT is an amplifier with F_2 and F_{12} being measured at the same frequency, the effect on δG_1 cancels out. The δENR then only influences the first two terms in eq. 5. This produces a fourth term derived from eq. 5 as follows:

$$\left(\frac{F_{12}}{F_1} - \frac{F_2}{F_1 G_1}\right)\delta ENR \quad (6)$$

where:
δENR = the uncertainty of the noise source's ENR.

This term should only be included when measuring amplifiers.

Since the causes of the uncertainties in the four δ factors are different, the terms can be combined in a root-sum-of-squares (RSS) fashion, which provides a realistic overall uncertainty value.

The equation for the overall NF uncertainty is therefore:

$$\delta NF = \left\{\left[\left(\frac{F_{12}}{F_1}\right)\delta NF_{12}\right]^2 + \left[\left(\frac{F_2}{F_1 G_1}\right)\delta NF_2\right]^2 + \left[\left(\frac{F_2 - 1}{F_1 G_1}\right)\delta G_1(dB)\right]^2 + \left[\left(\frac{F_{12}}{F_1} - \frac{F_2}{F_1 G_1}\right)\delta ENR\right]^2\right\}^{0.5} \quad (7)$$

Equation 7 provides the measurement uncertainty associated with a particular NF, using a system's VSWR characteristics and general electrical specifications. By knowing the NF uncertainty, greater confidence can be placed in the measured specifications attributed to production amplifiers and active devices.

Table 3: VSWR-to-reflection-coefficient transformations

	VSWR	Reflection coefficient ρ = (VSWR - 1)/(VSWR + 1)
Noise source	1.10:1	0.048
DUT input	1.50:1	0.20
DUT output	1.50:1	0.20
Instrument	1.80:1	0.286

ment system (Table 4).

The next step involves the calculation of the overall uncertainties using the maximum matching uncertainties and the NF instrument uncertainties. The instrumentation uncertainties should be those specified by the manufacturer. For this example, assume an instrument NF uncertainty (δInstrumentNF) of 0.05 dB, based on instrument-gain uncertainty (δInstrumentGain) of 0.15 dB, and effective-noise-ratio (ENR) uncertainty (δENR) of 0.1 dB. The following calculations show how to calculate the various NF, gain, and ENR uncertainties, although it should be noted that the receiver-only uncertainty (δENR_{RXOnly}) is not used in this example since the DUT is an amplifier:

$$\delta NF_{12}(dB) = [(\delta_{NS-DUT})^2 + (\delta InstrumentNF)^2 + (\delta ENR_{RXOnly})^2]^{0.5} = \sqrt{0.083^2 + 0.05^2} = \underline{0.097} \quad (1)$$

$$\delta NF_2(dB) = [(\delta_{NS-NFI})^2 + (\delta InstrumentNF)^2 + (\delta ENR_{RXOnly})^2]^{0.5} = \sqrt{0.119^2 + 0.05^2} = \underline{0.129} \quad (2)$$

$$\delta G_1(dB) = [(\delta_{NS-DUT})^2 + (\delta_{NS-NFI})^2 + (\delta_{DUT-NFI})^2 + (\delta InstrumentGain)^2 + (\delta ENR_{RXOnly})^2]^{0.5} = [(0.083^2 + 0.119^2 + 0.511^2 + 0.15^2]^{0.5} = \underline{0.552} \quad (3)$$

$$\delta ENR(dB) = \underline{0.1} \quad (4)$$

The next step involves the calculation of uncertainty terms (shown in Table 5) through multiplying the ratios found in Table 2 by the appropriate uncertainties.

There are many ways of calculating the overall uncertainty of a measurement. The traditional root-sum-of-squares (RSS) method will be used as the final step in this example since it is well-understood. RSS should, of course, use linear quantities, but with decibel values of the order that are dealt with here, the error is approximately 0.001 dB. Overall RSS measurement uncertainty is then:

$$RSS\ measurement\ uncertainty = \pm(0.102^2 + 0.007^2 + 0.025^2 + 0.099^2)^{0.5} = \pm 0.144 \quad (5)$$

The NF of the DUT in this example is therefore 3 ± 0.144 dB. The results of the calculations for Table 5 indicate that parameters $\pm\delta NF_{12}$, F_{12}/F_1, and δENR have the most significant influence on measurement uncertainty. These parameters as well as one other factor not appearing in these equations will now be explored in detail.

One of the most significant parameters affecting the uncertainties in Table 5 is δENR, the uncertainty of the noise source. For best uncertainty when measuring low-noise devices, low ENR sources should be used. This results in a lower δInstrumentNF since the low ENR exercises less of the measurement detector's dynamic range. There is a further advantage to using a low-ENR source in that its impedance is more constant between the on and off states. This is because a low ENR source (with ENR of typically 5 dB) is basically a high-ENR source (with ENR of typically 15 dB) with an additional attenuator. Beyond these points, there is not much room for movement with the noise source, since δENR is referred to the National Institute of Standards and Technology (NIST).

The instrument architecture includes any frequency translations that enable measurements at reasonable intermediate frequencies (IFs). The measurement-receiver architecture is either a single-sideband (SSB) or double-sideband (DSB) architecture. Network-analyzer-based instruments use the DSB architecture. This being the case, there is the possibility that power will appear in the undesired sideband causing a measurement error.

The possibility of this error can be reduced by using a narrowband DSB architecture. However, this increases the measurement time dramatically. For example, the theoretical measurement time increase when going from a 4-MHz bandwidth to a 40-kHz bandwidth is 100 times. Instruments employing a SSB structure do not have a problem with uncertainty due to power being in the unwanted sideband since it is filtered out.

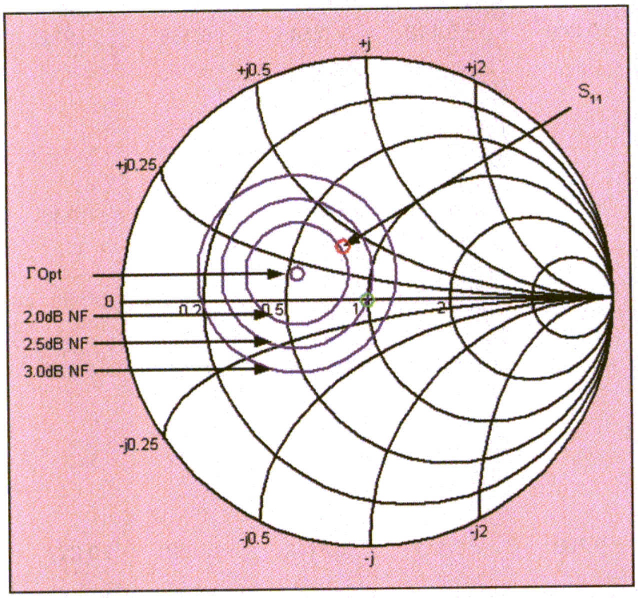
The ratio of system NF [(DUT and measurement instrument)/DUT NF, F_{12}/F_1] was also shown to have a sig-

Table 4: Calculating impedance-matching uncertainties

Ports source – load	Negative uncertainty 1 = -20 log (1 – ρsourceρload)	Positive uncertainty = 20 log (1 + ρsourceρload)	Maximum uncertainty (dB)
Noise source – DUT $_{INPUT}$	0.083	0.082	0.083
Noise source – instrument	0.119	0.117	0.119
DUT $_{output}$ – Instrument	0.511	0.483	0.511

nificant effect on the overall uncertainty. Parameter F_{12} is a function of the instrument NF and the DUT gain and NF. Figure 2 shows how the measurement uncertainty increases with an increase in F_{12}. The data are based on the previous example. In this case, an F_{12} value of 3 dB is the best that can be achieved since this is the NF of the DUT. Low DUT gain and/or high F_2 increase F_{12} and result in a higher uncertainty. Most combinations of instruments and DUTs with gain will produce results in the left-hand portion of the graph. DUTs with negative gain, such as mixers, move into the right-hand portion of the graph, increasing the uncertainty. In this situation, the uncertainty can be improved by inserting a low-noise amplifier (LNA) between the DUT and the measurement receiver.

Parameter δNF_{12} is made up of δInstrumentNF and the mismatch uncertainty. Work can be performed on the mismatch uncertainty using S-parameter correction but δInstrumentNF is a fixed function of the instruments linearity. Table 6 shows the uncertainty of the example DUT and some further typical DUTs against different δInstrumentNF values. The increase in the uncertain-



4. Conjugate matching does affect the NF measured for a DUT. The center of the chart indicated a characteristic system impedance of 50 {CAP OMEGA}.

5. The HP 8970B NF meter has long been a precision tool for simultaneously characterizing the NF and gain of low-noise devices.

Table 5: Calculating uncertainty terms

	Result (dB)
$(F_{12}/F_1) \times \delta NF_{12}$	0.102
$(F_2/F_1G_1) \times \delta NF_2$	0.007
$[(F_2 - 1)/(F_1G_1)] \times \delta G_1$	0.025
$[(F_{12}/F_1) - (F_2/F_1G_1)] \times \delta ENR$	0.099

ty with increase in δInstrumentNF is apparent. It is also clear from this data that an instrument with good δNF and no correction can outperform an instrument with idealized S-parameter correction but slightly worse δNF.

Figure 3 shows the change in measurement uncertainty of a typical DUT against an increase in VSWR. The data are shown with and without idealized S-parameter correction. From Table 7, it can be seen that instruments with good δNF and no correction of mismatches outperform other instruments up to particular VSWRs dependent on the δNF.

Depending on the measurement instrument and the VSWR of the DUT, S-parameter correction of mismatches may then appear to be of some benefit. However, there is one fundamental flaw in using S-parameter correction of mismatches when measuring NF. S-parameters provide almost all of the information needed for a device, with one exception—noise.

It is true that although a device is unlikely to be perfectly matched in practice, S-parameters can be used to obtain the available gain (the gain with the input and output conjugately matched). However, it has been cited that if this available gain is used in the following standard equation to remove the effects of the second-stage NF, that a more-accurate result will be achieved:

$$NF_{DUT} = NF_{DUT+Instrument} - \left(\frac{NF_{Instrument} - 1}{Gain_{DUT}}\right) \qquad (6)$$

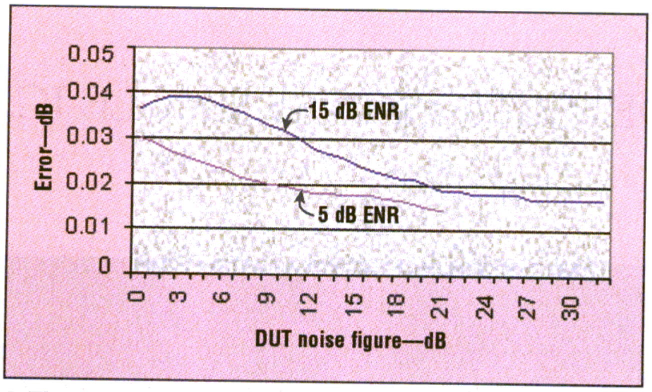
This theory assumes that the NF of the device (NF_{DUT}) and measuring instrument ($NF_{Instrument}$) do not change with impedance. It would be nice if this were the case, but unfortunately NF varies wildly with impedance as a glance at the noise data for any RF device will show. An example is shown below in graphical form in Fig. 4.

An NF measuring instrument will measure the NF and gain with 50 Ω presented to the input of the device (the circle at the center of the chart).

Table 6: Comparing DUT and instrument NF uncertainties

δInstrumentNF	0.05 dB	0.10 dB	0.15 dB	0.20 dB	0.05 dB	0.10 dB	0.15 dB	0.20 dB
	No correction				Idealized S-parameter correction			
DUT (amplifier)	Measurement uncertainty (dB)							
Gain = 20 dB NF = 3 dB Instrument NF = 10 dB In/out VSWR = 1.50:1	±0.144	±0.170	±0.207	±0.249	±0.113	±0.145	±0.186	±0.232
Gain = 13 dB NF = 2.2 dB Instrument NF = 5 dB In/out VSWR = 1.80:1	±0.176	±0.199	±0.232	±0.272	±0.111	±0.145	±0.189	±0.236
Gain = 26 dB NF = 3.5 dB Instrument NF = 10 dB In/out VSWR = 2.0:1	±0.180	±0.200	±0.230	±0.266	±0.112	±0.142	±0.181	±0.225
Gain = 18 dB NF = 0.8 dB Instrument NF = 4 dB In/out VSWR = 2.0:1	±0.181	±0.201	±0.232	±0.268	±0.111	±0.142	±0.182	±0.227

S-parameters could then be used to calculate the available gain, simulating what the gain would be if the input and output of the device were presented with the conjugate of S_{11} (the red circle) and S_{22}, respectively. If the input of the DUT is actually presented with the conjugate of S_{11}, however, its NF would change by approximately 0.5 dB (shown by the blue NF circles in Fig. 4). Any S-parameter correction in NF must therefore be used with great care since it can make the measurements significantly less accurate without knowledge of how the NFs of the DUT and the measurement receiver are altered by conjugate matching.



6. The measurement error of the HP 8970B NF meter is typically less than 0.05 dB.

NOISE PARAMETERS

There are several basic noise parameters that completely describe the noise characteristics of a device. These are the minimum possible NF of the DUT (NF_{min}), the equivalent noise resistance of the device (R_n), and the optimum source-reflection coefficient (Γ_{opt}) for magnitude and phase. These parameters are found by applying different impedances to the device using a tuner for optimization and are completely unrelated to the S-parameters.

There are some other issues that are associated with S-parameter corrections, the first being the need for a network analyzer. This approach to measuring NF is likely to cost several tens of thousands of dollars more than an approach that is without correction. That is not an issue if the area of interest is device characterization, although this will also require bias tees in addition to complex automatic tuner units to present devices with a variety of impedances. NF instruments that are embedded within network analyzers do not then provide a complete solution for device characterization since they can only present the device with a fixed impedance.

NF INSTRUMENTS EMBEDDED WITHIN NETWORK ANALYZERS DO NOT PROVIDE A COMPLETE SOLUTION FOR DEVICE CHARACTERIZATION.

Another problem with using S-parameter correction on NF is the increased time to make the measurement due to the extra calibrations and supplemental measurements. The connecting/disconnecting of cables, bias tees, tuners, and more, also adds reliability issues, measurement uncertainty, and a time penalty.

Considering the number and diversity of operations that are required to make S-parameter-corrected NF measurements, it becomes clear that some form of computer control of all the operations will be required. This adds further cost and complexity. The general complexity of corrected NF measurements/ device characterization suggests that a dedicated test system is required.

MEASURING NF

One current solution that can be used to measure NF is the HP 8970B NF meter from Hewlett-Packard Co. (Fig. 5). Together with an appropriate noise source, the HP 8970B is capable of simultaneously characterizing the NF and gain for receiver systems, their subassemblies, as well as components such as amplifiers, mixers, filters, diplexers, and low-noise block downconverters (LNBs). The low instrumentation uncertainty of the HP 8970B combined with its true SSB receiver architecture and correction to remove the measurement-system noise contribution bring ease and confidence to NF measurements. While the δInstrumentNF for the HP 8970B is specified as 0.1 dB, it is typically much better than this. Typical data from current production HP 8970B units show that the error is typically less than 0.05 dB (Fig. 6) and the advantage of the low ENR source can be clearly seen. ●●

Table 7: Comparing instruments with different NF uncertainties

δInstrumentNF	0.10 dB corrected	0.15 dB corrected	0.20 dB corrected
VSWR where an instrument having δNF = 0.05 dB without mismatch correction provides better measurement uncertainty	≤1.5:1	≤2.0:1	≤2.6:1
VSWR where an instrument having δNF = 0.1 dB without mismatch correction provides better measurement uncertainty	–	≤1.7:1	≤2.4:1
VSWR where an instrument having δNF = 0.15 dB without mismatch correction provides better measurement uncertainty	–	–	≤1.9:1

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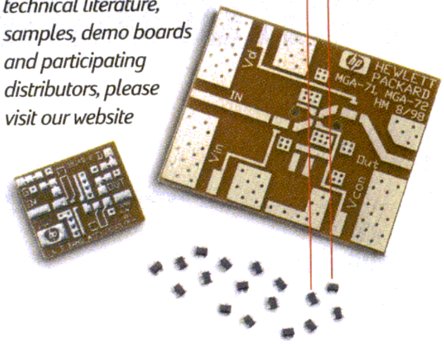
Typical performance @ 2 GHz

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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Predict Effects Of Ice Needles On Satcom Performance

A new method of calculating the scattering effects of atmospheric ice needles can improve higher-frequency satellite communications.

John Howard
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SATELLITE-COMMUNICATION systems operating at frequencies beyond 10 GHz are subject to the scattering effects of ice needles in the lower atmosphere. These ice needles cause depolarization of the transmitted radiation and can severely limit system performance, particularly in the case where two orthogonal polarizations are used as separate communication channels.

This paper presents a new method for calculating the forward scattering coefficients of ice needles. The coefficients are calculated using two methods—a new method based on Maxwell's integral equations, and the Rayleigh-Gans scattering method. The two methods are then compared. In addition, the cross-polarization discrimination values between two orthogonal linear polarizations due to ice needles are compared for the two methods.

First, consider the new Maxwell-based method and its calculations.

The magnetic field vector is given by:

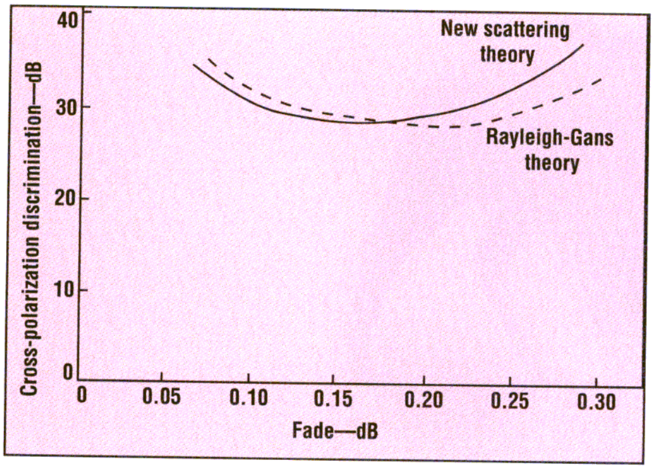
$$H = \frac{1}{m}\nabla \times A \qquad (1)$$

where:
μ = the permeability of the medium of propagation.

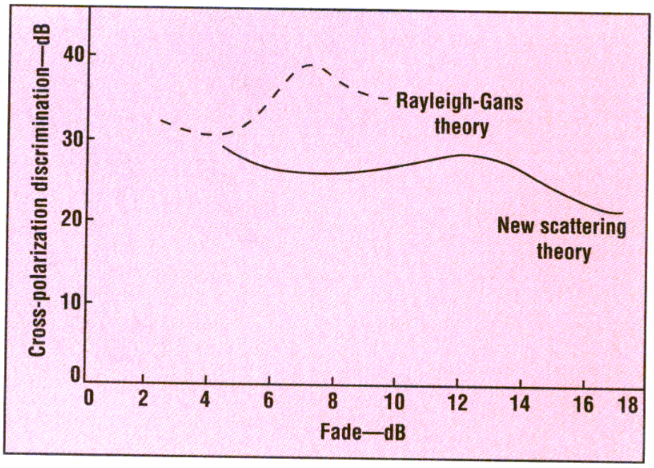
The vector potential A is given by:

$$A = \mu \iiint_{v'} J \frac{e^{-jkr}}{4\pi r} dv' \qquad (2)$$

where:
J = the harmonic source current density,
r = the distance between the



1. This graph shows cross-polarization discrimination of ice needles versus fade at 10 GHz for a 1-km slab and a canting angle of 1 deg.



2. This graph shows cross-polarization discrimination of ice needles versus fade at 30 GHz for a 1-km slab and a canting angle of 1 deg.


Does Your System FLY...
FLOAT...
or ROLL?
CS-2000 VME
VHF/UHF Receiver
CS-5040 VXI
Microwave Tuner
CS-5020C
Microwave Receiver

Comparison between New scattering theory and Rayleigh-Gans theory

Freq.f (GHz)	fs (New scattering theory)	fs (Rayleigh-Gans theory)
10	$0.232123\ 10^{-5} + j.748377\ 10^{-8}$	$0.254882\ 10^{-5} + j.100430\ 10^{-7}$
30	$0.213867\ 10^{-4} - j.165142\ 10^{-6}$	$0.229394\ 10^{-4} + j.903868\ 10^{-7}$

source and observation points,

$k = 2\pi/\lambda$,

λ = the wavelength of propagation, and

v' = the source volume.

Using eq. 2 in eq. 1 and taking the curl of both sides of the resultant equation, the scattered electric field along z, the direction of the ice-needle length, is given by:

$$E_z^s = \frac{\sqrt{\mu_0 / \varepsilon_0}}{4\pi jk} \iiint_{v'} \left[F_1(r) + (z - z')^2 F_2(r)\right] J_z \, dv' \quad (3)$$

where:

prime denotes source and non prime denotes observation points, and

$$F_1(r) = (-r^2 - jkr^3 + k^2 r^4)\frac{e^{-jkr}}{r^5} \quad (4)$$

$$F_2(r) = (3 + 3jkr - k^2 r^2)\frac{e^{-jkr}}{r^5} \quad (5)$$

The harmonic source-current density along z is given by:

$$J_z = j\omega\varepsilon_0 (\varepsilon_r - 1) E_z \quad (6)$$

where:

E_Z = the total electric field (incident plus scattered) along z,

ϵ_r = the permittivity of ice, and

ϵ_0 = the permittivity of free space.

E_Z can be expanded in a mode series given by [1]:

$$E_z = \sum_{n=0}^{N-1} E_n \cos\frac{2n\pi z}{L} J_0\left[K_z p \sqrt{\varepsilon_r - \left(\frac{n\lambda}{L}\right)^2}\right] \quad (7)$$

where:

$k_z = 2n\pi/L$,

E_n = the coefficients of the mode series 0 1 $\rho \leq \alpha$,

α= the radius of the cylindrical ice needle, and

J_0 = the zero-order Bessel function of the first kind.

The forward-scattering complex coefficient may then be computed from:

$$f_s = \frac{E_z^s}{\dfrac{e^{-jkr}}{r}} \quad (8)$$

The scattered field perpendicular to z is calculated in a similar manner, and the forward-scattering complex coefficient for this polarization can then be computed.

RAYLEIGH-GANS

Now consider the calculations required by the Rayleigh-Gans scattering method. Raleigh-Gans scattering may be assumed if the following two conditions are met[2]:

1. The refractive index of the scattering particle (which may be complex and is measured relative to the surrounding medium) is:

$$|m-1| << 1 \qquad (9)$$

where:
m = the refractive index of the scattering particle.

2. The "phase shift" introduced by the scatterer is small.

Therefore:

$$2k\alpha|m-1| << 1 \qquad (10)$$

where:
k = 2π/λ, and
a = the length of the particle, the size of which should be << λ / |m –1|

The forward-scattering complex coefficients are given in this case by:

$$f_s = k^2 p - \frac{2}{3} jk^5 p \qquad (11)$$

where:
ρ = v/4π(m – 1) for the z direction,
ρ = v/4π[(2m^2 – 1)/(m^2 + 2)] for the direction perpendicular to z, and
v = the volume of the ice crystal.

The forward-scattering coefficients for ice needles of various dimensions were computed using the above equations. The table shows a comparison between the two scattering methods for a particular ice-crystal size and a certain incident angle of energy at two different frequencies.

In addition, Figs. 1 and 2 show comparative plots between the two methods for cross-polarization discrimination values versus fade for frequencies of 10 and 30 GHz. The plots show the results of the two methods diverging as the frequency increases. The reason for these differences is that, at the higher frequencies, the Raleigh-Gans scattering assumptions no longer hold.

This investigation shows that, although the Rayleigh-Gans scattering method is much easier to use than the Maxwell-based method, the accuracy of the Rayleigh-Gans method decreases as the frequency increases. For frequencies higher than 10 GHz, the Rayleigh-Gans scattering method does not hold and should not be employed in calculating cross-polarization discrimination values. On the other hand, the new method presented here can be used in its general form to calculate the scattering effects of ice needles or ice columns at all microwave frequencies. ••

References

1. J.H. Richmond, "Digital Solutions of the Rigorous Equations for Scattering Problems," *Proc. IEEE*, vol. 53, pp. 796-804, 1965.
2. H.C. Van De Hulst, *Light Scattering By Small Particles*, Dover, New York, 1981.

Spurious Couplings Degrade Bandpass Filter Performance

Spurious couplings can inhibit achieving the predicted transmission and reflection responses of bandpass filters.

Spurious Couplings, Part 1

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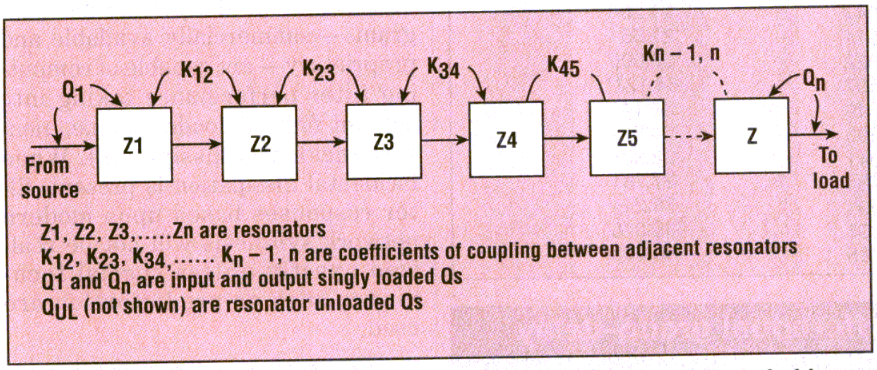
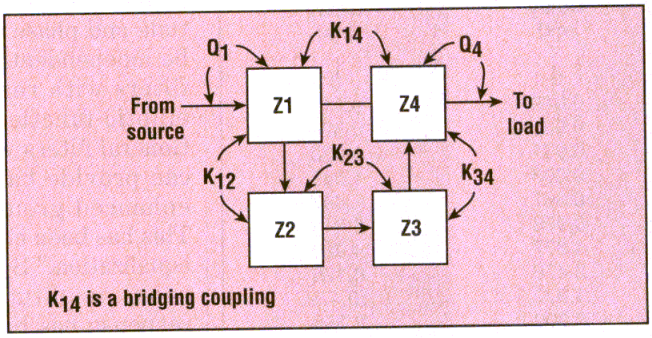
BANDPASS filters are important components in communications equipment. They affect receiver sensitivity by establishing the noise bandwidth, and they provide frequency selectivity that rejects external interference and internally generated spurs. Bandpass filters of small- and moderate-percent bandwidths usually employ direct-coupled adjacent resonators. Spurious couplings—between non-adjacent resonators—can cause degradation in predicted filter performance. These response-shape impairments can render some filter structures unusable.

Multiresonator direct-coupled bandpass filters can be depicted as a ladder network of coupled resonators (Fig. 1). Individual resonators are separated by interior interstage couplings. The first and last filter resonators are interfaced to filter source and load impedances through input and output couplings. These filters can be realized as lumped-circuit L-C circuits as well as transmission-line circuits.

The basic direct-coupled bandpass filter circuits are all pole circuits, which are minimum phase-shift networks. This means that filter amplitude and phase (or group-delay) responses are mutually dependent upon each other. All filter interstage couplings are between adjacent resonators with zero couplings assumed between non-adjacent resonators. The all-pole filters are designed using modern network theory. Popular response shapes are Butterworth, Tchebychev, Equal Element, Gaussian, and others.



1. A direct-coupled multiresonator bandpass filter is represented as a ladder network of coupled resonators.

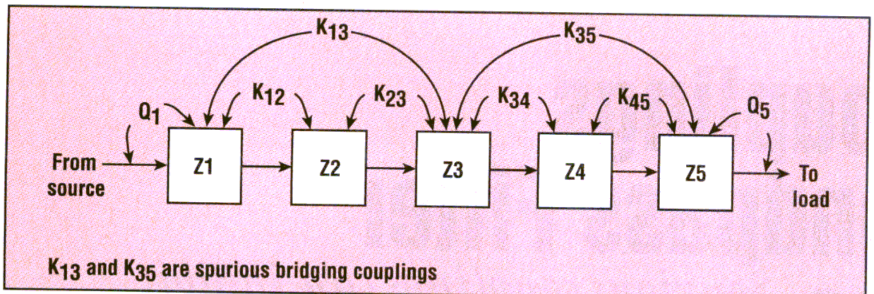


2. The four-pole general filter shown here is a non-minimum phase-shift network.

BANDPASS PARAMETERS

There are two basic design techniques for bandpass-filter design:

1. Start with a lowpass prototype and apply a lowpass-to-bandpass transformation.
2. Use direct-coupled multires-



3. A five-pole direct-coupled bandpass filter has spurious bridging couplings.

Table 1: Comparative amplitude and VSWR responses of four-pole general filters

	$k_{14} = 0$	$k_{23} = 0.541$	$k_{14} = 0.14$	$k_{23} = 0.475$	$k_{14} = -0.14$	$k_{23} = 0.630$
x = normalized frequency variable	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR
0	0.010	1.101	0.010	1.101	0.010	1.098
0.1	0.007	1.083	0.006	1.080	0.007	1.082
0.2	0.001	1.036	0.001	1.026	0.001	1.037
0.3	0.001	1.024	0.002	1.038	0.001	1.025
0.4	0.007	1.080	0.008	1.089	0.009	1.096
0.5	0.010	1.100	0.008	1.087	0.022	1.152
0.6	0.002	1.039	0.000	1.001	0.021	1.150
0.7	0.021	1.150	0.044	1.222	0.002	1.041
0.8	0.258	1.633	0.307	1.707	0.059	1.262
0.9	1.131	2.837	1.046	2.722	0.632	2.165
1.0	3.010	5.828	2.412	4.761	2.610	5.099
1.1	5.716		4.277	8.594	6.290	
1.2	8.740		6.366		10.907	
1.3	11.737		8.462		15.902	
1.4	14.575		10.547		21.286	
1.5	17.220		12.311		27.561	
1.6	19.682		14.022		36.600	
1.7	21.976		15.597		54.121*	
1.8	24.121		17.049		38.141	
1.9	26.136		18.393		34.511	
2.0	28.034		19.641		32.888	
2.5	36.173		24.784		31.781	
3.0	42.714		28.691		33.244	
3.5	48.193		31.835		35.076	
4.0	52.911		34.465		36.903	

Table 2: Comparative differential group-delay responses for four-pole filters

	$k_{14} = 0$	$k_{14} = 0.14$	$k_{14} = -0.14$
x = normalized frequency variable	Delay (ns)	Delay (ns)	Delay (ns)
0	0	0	0
0.2	0.096	0.019	0.110
0.3	0.271	0.031	0.321
0.4	0.515	0.031	0.623
0.5	0.876	0.093	1.045
0.6	1.497	0.367	1.697
0.7	2.607	0.988	2.844
0.8	4.327	1.877	4.929
0.9	6.038	2.476	8.126
1.0	6.074	1.979	10.479
1.1	3.836	0.300	8.668
1.2	0.881	-1.807	4.436

onator bandpass-filter parameters.

Both techniques employ normalized circuit elements. Normalization in frequency and impedance are used. When lowpass prototypes are used, Tchebychev filters are usually normalized to the ripple bandwidth. When direct-coupled bandpass-filter parameters are used, Tchebychev filters are usually normalized to the 3-dB bandwidth. Throughout this article, direct-coupled multiresonator bandpass-filter design parameters, with normalization to 3-dB bandwidths will be used.

Direct-coupled bandpass filters—at baseband, intermediate frequency (IF), and microwave frequencies—can be characterized by three design parameters: input and output singly loaded qs, interstage coefficients of coupling, and resonator unloaded qs. Lower-case parameters are normalized values and upper-case parameters are absolute values. Occasionally, dissipation factors are employed instead of unloaded qs. Dissipation factor is defined as the reciprocal of unloaded q.

In subsequent technical discussions, only lossless filter circuits will be considered. Many computer programs—commercially available and proprietary—are capable of computing filter performance taking into account finite unloaded qs (i.e., nonzero incidental dissipation). When incidental dissipation is present, filter responses based upon modern network synthesis will not be realized exactly unless special techniques such as predistortion are used.

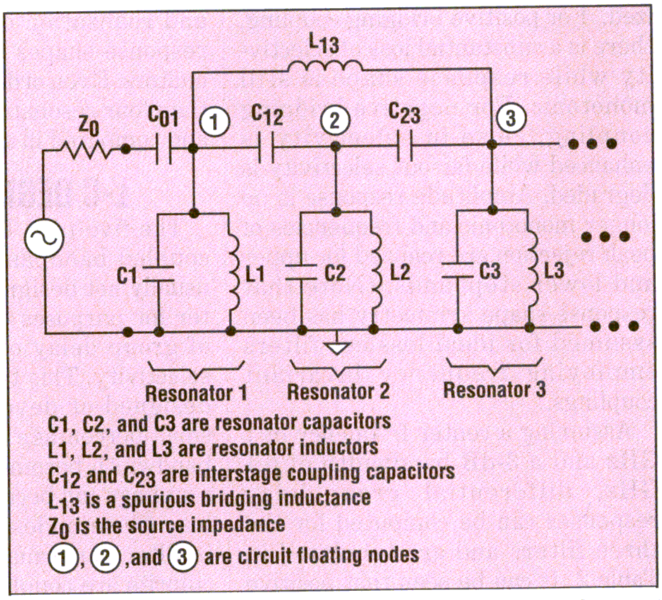
GENERAL FILTER CIRCUITS

General filters are non-minimum phase-shift networks where amplitude and phase (or group delay) can be independently specified. General filters with real zeros can realize elliptic-function response shapes. General filters with imaginary zeros can provide bandpass filters with enhanced group-delay responses. This has been characterized as "self-equalization." Synthesis for simultaneous amplitude and group-delay response has been achieved. Very-small bridging couplings can also be employed to restore response-shape

symmetries.

General filters can be realized using lumped and distributed circuits. Block diagrams of four- and five-pole general filters are shown in Figs. 2 and 3. The four-pole filter is designed as a general filter. The five-pole general filter includes some spurious bridging couplings which degrade filter performance.

General filters are usually designed for enhanced selectivity or self-equalization of group delay. Most general filters use an even number of resonators. For illustrative purposes, the four-pole Tchebychev



4. Three of the L-C resonators of a five-pole Tchebychev direct-couples bandpass filter are shown here in detail. Spurious bridging coupling through the bridging capacitors and inductors can degrade filter responses.

Table 3: Comparative amplitude and VSWR for five-pole filters with positive bridging

	$k_{13} = 0$		$k_{13} = 0.05$		$k_{13} = 0.1$	
x = normalized frequency variable	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR
-2.1	39.474		44.531		51.602	
-1.8	32.174		36.468		42.071	
-1.5	23.204		26.809		31.212	
-1.2	11.584		14.516		17.926	
-1.0	3.010	5.828	4.705	9.716	6.978	17.888
-0.9	0.645	2.182	1.175	2.897	2.077	4.215
-0.8	0.037	1.204	0.065	1.276	0.113	1.382
-0.7	0.004	1.059	0.018	1.139	0.055	1.254
-0.6	0.009	1.097	0.060	1.265	0.184	1.511
-0.5	0.001	1.035	0.053	1.247	0.211	1.557
-0.4	0.002	1.041	0.035	1.197	0.201	1.539
-0.3	0.008	1.092	0.028	1.173	0.201	1.541
-0.2	0.009	1.097	0.034	1.192	0.229	1.585
-0.1	0.004	1.060	0.054	1.249	0.278	1.664
0	0.000	1.000	0.086	1.325	0.332	1.745
0.1	0.004	1.060	0.116	1.388	0.364	1.791
0.2	0.009	1.097	0.126	1.407	0.346	1.766
0.3	0.008	1.092	0.103	1.362	0.274	1.657
0.4	0.002	1.041	0.057	1.258	0.167	1.483
0.5	0.001	1.035	0.016	1.129	0.069	1.288
0.6	0.009	1.097	0.001	1.024	0.015	1.125
0.7	0.004	1.059	0.000	1.000	0.002	1.046
0.8	0.037	1.204	0.022	1.153	0.013	1.114
0.9	0.645	2.182	0.342	1.759	0.172	1.491
1.0	3.010	5.828	1.822	3.823	1.039	2.712
1.2	11.584		9.031		6.812	17.139
1.5	23.204		20.123		17.408	
1.8	32.174		28.665		25.674	
2.1	39.474		35.513		32.238	

bandpass filter with 0.01-dB passband ripple will be considered. For simplicity, the frequency sensitivity of all filter couplings will be neglected. Also, ideal lossless resonators will be assumed. Using nodal analysis, the filter-response shapes for the four-pole general filter (Fig. 2) have been computed for three cases: non-bridging coupling (i.e., basic Tchebychev design), positive bridging coupling between first and fourth resonators ($k_{14} = 0.14$), and negative bridging coupling ($k_{14} = -0.14$). For non-zero bridging coupling, the central interstage coupling, k_{23} is modified to restore the passband impedance matching. For positive bridging coupling, $k_{23} = 0.475$ instead of 0.541. For negative bridging coupling, $k_{23} = 0.630$ instead of 0.541.

Comparative amplitude and VSWR responses for the three different four-pole filters are shown in Table 1. Without bridging coupling, the responses of the lossless Tchebychev 0.01-dB ripple design are realized. For positive bridging coupling, there is a substantial loss of selectivity while response shape is still monotonic. For negative bridging coupling, close-in selectivity is enhanced while far-out selectivity is degraded. Amplitude response is no longer monotonic and frequencies of peak rejection are realized for upper and lower stopbands. (Note that response-shape symmetry has been assumed for ideal lossless filters employing frequency invariant couplings.)

Assuming a center frequency of 1 GHz and a 3-dB bandwidth of 0.1 GHz, differential group-delay responses can be computed for the three filters and are tabulated in Table 2. It can be seen that positive bridging coupling achieves improved group-delay flatness while negative bridging coupling degrades group-delay flatness. Positive bridging coupling is the key to self-equalization while negative bridging coupling is the key to sharper skirt selectivity and realization of elliptic-function response shapes using coupled resonators. Even order filters with more than four resonators can be analyzed and compared in a similar manner.

L-C DEGRADATION

The five-pole Tchebychev direct-coupled bandpass filter (Fig. 3) is usually not designed as a general filter for purposes of self-equalization of group delay or enhancement of selectivity. This type of filter can be designed or developed readily for small-percentage bandwidths. As the bandwidth becomes larger, spurious couplings can become significant and may cause undesirable degradation in filter performance. The principal culprits are couplings k_{13} and k_{35}. It should be noted that spurious couplings can be positive (same sign as the filter) or negative (opposite sign of the filter). Some typical L-C circuit details are shown in the partial schematic of Fig. 4.

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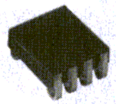

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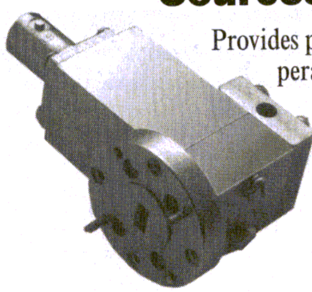

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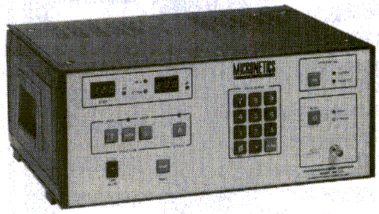

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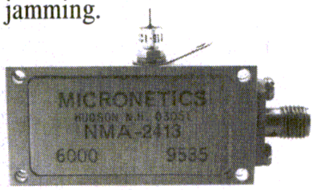

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of the lossless five-pole Tchebychev direct-coupled bandpass filter are shown versus normalized frequency in Tables 3 and 4. Also shown are amplitude and VSWR responses when k_{13} = 0.05, 0.1, –0.05, and –0.1. This also assumes that $k_{13} = k_{35}$. It can be seen that as the absolute value of k_{13} increases, the amplitude and VSWR responses are substantially impaired. The response-shape symmetry that existed for no discernible bridging coupling has been altered. For L-C direct-coupled multiple-resonator bandpass filters, spurious bridging couplings can substantially degrade the filter responses and lossless Tchebychev responses will no longer be attainable. The lack of response-shape symmetry makes external group-delay equalization more difficult and could preclude satisfactory equalization.

Spurious couplings in L-C filters can be improved with careful attention to the physical layout of filter-circuit elements. Increasing distances between components results in increased filter size and probable increase in filter cost. The original filter layout can be made more desirable if small metallic shields are judiciously added to areas where spurious couplings can occur. Parasitic circuit elements can also contribute to spurious couplings.

COMBLINE FILTERS

Combline bandpass filters use direct-coupled quarter-wave transverse-electromagnetic (TEM) resonators with adjacent resonators that have the same open- and short-circuit reference planes. In the past half-century, combline filters have evolved through four distinct phases of implementation:

1. Early combline bandpass filters employed metallic partitions between adjacent resonators with interstage couplings attained through circular apertures in the partitions. Resonator cross-sections entailed cylindrical inner and outer

Table 4: Comparative amplitude and VSWR for five-pole filters with negative bridging

	$k_{13} = 0$		$k_{13} = -0.05$		$k_{13} = -0.1$	
x = normalized frequency variable	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR	Insertion loss (dB)	VSWR
-2.1	39.474		35.513		32.238	
-1.8	32.174		28.664		25.674	
-1.5	23.204		20.123		17.408	
-1.2	11.584		9.031		6.812	
-1.0	3.010	5.828	1.822	3.823	1.039	2.712
-0.9	0.645	2.182	0.342	1.759	0.172	1.491
-0.8	0.037	1.204	0.022	1.153	0.013	1.114
-0.7	0.004	1.059	0.000	1.000	0.002	1.046
-0.6	0.009	1.097	0.001	1.024	0.015	1.125
-0.5	0.001	1.035	0.016	1.129	0.069	1.288
-0.4	0.002	1.041	0.057	1.258	0.167	1.483
-0.3	0.008	1.092	0.103	1.362	0.274	1.657
-0.2	0.009	1.097	0.126	1.407	0.346	1.766
-0.1	0.004	1.060	0.116	1.388	0.364	1.791
0	0.000	1.000	0.086	1.325	0.332	1.745
0.1	0.004	1.060	0.054	1.249	0.278	1.664
0.2	0.009	1.097	0.034	1.192	0.229	1.585
0.3	0.008	1.092	0.028	1.173	0.201	1.541
0.4	0.002	1.041	0.035	1.197	0.201	1.530
0.5	0.001	1.035	0.053	1.247	0.211	1.557
0.6	0.009	1.097	0.060	1.265	0.055	1.511
0.7	0.004	1.059	0.018	1.139	0.002	1.254
0.8	0.037	1.204	0.065	1.276	0.113	1.382
0.9	0.645	2.182	1.175	2.897	2.077	4.216
1.0	3.019	5.828	4.705	9.716	6.978	2.712
1.2	11.584		14.516		17.926	17.889
1.5	23.204		26.809		31.212	
1.8	32.174		26.468		42.071	
2.1	39.474		44.531		51.602	

conductors as in coaxial lines. Early filters were designed or developed for small-percentage bandwidths. Consequently, spurious couplings were usually not a problem in early combline filters. Filter size had to be limited so that the TE_{11} circumferential mode was not propagating.

2. With the advent of strip transmission lines, combline filters were implemented using rectangular bars as resonator center conductors located midway between metallic ground planes. The combline resonators were now direct coupled and had to be foreshortened to preclude all stop behaviors. The original combline technical paper recommended resonator electrical lengths of 45 deg. with resonance achieved using capacitive tuning screws at the open-circuit ends of the resonators. This pushed the first TEM spurious passband up to nominally six times the filter-design center frequency.

3. Subsequently, combline filters used slabline construction which entailed round-rod center conductors between metallic ground planes. This helped to reduce filter fabrication costs. For filter bandwidths approaching 10 percent, nine- and 11-pole filters were designed and manufactured in production machine shops. Typical production runs ranged from 50 to 1000 units.

4. During the past decade, major changes in the communications and electronics marketplace have resulted in large increases in commercial production runs of filters. The availability of temperature-stable high dielectric-constant materials, with low-loss tangents, reduced insertion losses, and reduced temperature sensitivity resulted in major advances in combline filter miniaturization and unit cost reduction. Machined resonators with air dielectric were replaced by electroplated dielectric resonators that were popularly referred to as ceramic filters. The exploding market for wireless communication products and systems has created a real need for combline filters. ••

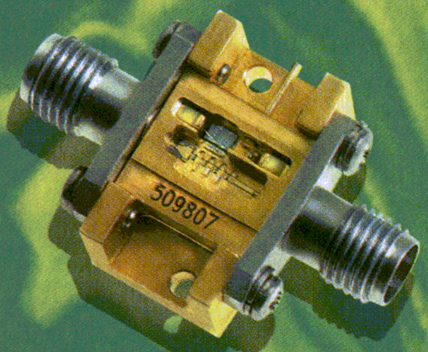
Acknowledgements

In the interest of brevity, all applicable references have been omitted. In the past four decades, many engineers have made important technical contributions in the area of general filters as herein described.

ULTRA-LOW NOISE
AMPLIFIERS
VHF TO V-BAND

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMPR. (dBm, Min.)		DC POWER @ +15 V (mA, Nom.)
OCTAVE BAND AMPLIFIERS									
JS2-00500100-035-5A	0.5 – 1	35	1	0.45*	2:1	2:1	5		250
JS2-00500100-10-5A	0.5 – 1	35	1.2	1	2:1	2:1	5		250
JS2-01000200-035-5A	1 – 2	33	1	0.45*	2:1	2:1	5		250
JS2-01000200-10-5A	**1 – 2**	**33**	**1.2**	**1**	**2:1**	**2:1**	**5**	**$395**	**250**
JS2-02000400-035-5A	2 – 4	28	1	0.45*	2:1	2:1	5		175
JS2-02000400-10-5A	**2 – 4**	**28**	**1.2**	**1**	**2:1**	**2:1**	**5**	**$450**	**175**
JS2-04000800-070-0A	4 – 8	22	1	0.7	2:1	2:1	0		150
JS2-04000800-15-0A	**4 – 8**	**22**	**1.2**	**1.5**	**2:1**	**2:1**	**0**	**$495**	**150**
JS3-04000800-060-5A	4 – 8	30	1	0.6	2:1	2:1	5		175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5		175
JS2-08001200-09-5A	8 – 12	15	1	0.9	2:1	2:1	5		150
JS2-08001200-15-5A	**8 – 12**	**15**	**1.2**	**1.5**	**2:1**	**2:1**	**5**	**$495**	**150**
JS3-08001200-080-5A	8 – 12	25	1	0.8	2:1	2:1	5		175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5		175
JS2-12001800-16-5A	12 – 18	15	1	1.6	2:1	2:1	5		100
JS2-12001800-30-5A	**12 – 18**	**15**	**1.5**	**3**	**2:1**	**2:1**	**5**	**$495**	**100**
• JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5		175
JS3-12001800-30-5A	**12 – 18**	**23**	**1**	**3**	**2:1**	**2:1**	**5**	**$495**	**175**
JS4-12001800-12-5A	12 – 18	30	1	1.2	2:1	2:1	5		200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5		200
JS2-18002600-20-5A	18 – 26	14	1	2	2:1	2:1	5		100
JS2-18002600-30-5A	18 – 26	14	1	3	2:1	2:1	5		100
JS3-18002600-20-5A	18 – 26	22	1	2	2:1	2:1	5		175
JS3-18002600-30-5A	18 – 26	22	1	3	2:1	2:1	5		175
JS4-18002600-16-5A	18 – 26	27	1	1.6	2:1	2:1	5		200
JS4-18002600-26-5A	18 – 26	27	1	2.6	2:1	2:1	5		200
JS2-26004000-35-5A	26 – 40	12	2	3.5	2:1	2:1	5		100
JS2-26004000-45-5A	26 – 40	12	2	4.5	2:1	2:1	5		100
JS3-26004000-35-5A	26 – 40	18	2	3.5	2.5:1	2.5:1	8		175
JS3-26004000-45-5A	26 – 40	18	2	4.5	2.5:1	2.5:1	8		175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	8		200
JS2-26004000-100-20A	26 – 40	17	1.25	10	2.3:1	2.3:1	20		**
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0		175
MULTIOCTAVE BAND AMPLIFIERS									
JS2-00500200-05-5A	0.5 – 2	32	1	0.5	2:1	2:1	5		250
JS2-00500200-20-5A	0.5 – 2	32	1	2	2:1	2:1	5		250
JS2-01000400-07-5A	1 – 4	27	1	0.7	2:1	2:1	5		200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5		200
JS2-02000600-07-5A	2 – 6	24	1	0.7	2:1	2:1	5		125
JS2-02000600-20-5A	2 – 6	20	1	2	2:1	2:1	5		125
JS2-02000800-08-0A	2 – 8	22	1	0.8	2:1	2:1	0		125
JS2-02000800-20-0A	2 – 8	18	1	2	2:1	2:1	0		125
JS3-02001800-25-5A	2 – 18	21	2	2.5	2.5:1	2.5:1	5		150
JS3-02001800-50-5A	2 – 18	21	2	5	2.5:1	2.5:1	5		150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5		200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5		200
JS3-02002600-30-5A	2 – 26	21	2	3	2:1	2.1	5		150
JS3-02002600-40-5A	2 – 26	21	2	4	2:1	2:1	5		150
JS3-06001800-18-5A	6 – 18	23	1.3	1.8	2:1	2:1	5		125
JS3-06001800-30-5A	6 – 18	23	1.3	3	2:1	2:1	5		125
JS4-06001800-135-5A	6 – 18	31	1	1.35	2:1	2:1	5		200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5		200

* Noise figures to 0.35 dB available on a limited basis.
** This unit requires +8V @ 500 mA and -8V @ 90 mA.



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MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMPR. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS (Continued)								
JS3-08001800-17-5A	8 – 18	24	1.2	1.7	2:1	2:1	5	125
JS3-08001800-30-5A	8 – 18	24	1.2	3	2:1	2:1	5	125
JS4-08001800-13-5A	8 – 18	32	1.5	1.3	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	1.5	3	2:1	2:1	5	200
JS3-08002600-30-5A	8 – 26	21	2	3	2:1	2:1	5	150
JS3-08002600-40-5A	8 – 26	21	2	4	2:1	2:1	5	150
JS3-12002600-25-5A	12 – 26	22	2	2.5	2:1	2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2	3.5	2:1	2:1	5	150
JS4-12002600-22-5A	12 – 26	30	1.7	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	30	1.7	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-06-5A	0.1 – 2	32	1	0.6	2:1	2:1	5	250
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	250
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-18-5A	0.1 – 10	26	1.5	1.8	2:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	26	1.5	3.5	2:1	2:1	5	150
JS3-00101200-19-5A	0.1 – 12	25	1.5	1.9	2:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	25	1.5	3.5	2:1	2:1	5	150
JS3-00101800-26-5A	0.1 – 18	23	1.5	2.6	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.5	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-32-5A	0.1 – 26	20	1.8	3.2	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	1.8	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2	5	2.5:1	2.5:1	5	200
JS4-00103000-35-5A	0.1 – 30	20	2.5	3.5	2.5:1	2.5:1	5	200
JS4-00103000-45-5A	0.1 – 30	20	2.5	4.5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	3.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	3.5	8.5	2.75:1	2.75:1	5	200

NOTE: Higher 1 dB compression levels are available on many designs.

For additional information or technical support, please contact either Rosalie DeSousa at (516) 439-9458, e-mail rdesousa@miteq.com or Rizwan Syed at (516) 439-9267, e-mail rsyed@miteq.com.

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CIRCLE NO. 209

SiGe Technology Makes Practical Advances

This novel device technology is making major strides in RF and digital integrated circuits for high-frequency, high-speed communications systems.

JACK BROWNE
Publisher/Editor

SILICON germanium (SiGe) is a semiconductor technology made for wireless applications. It offers the high-speed, high-frequency performance needed for wireless systems, and it provides the potential for integrating analog, RF, and digital functions on a single integrated circuit (IC). Plus, it is based on low-cost Si wafers. Admittedly, it requires more process steps than conventional Si bipolar technologies. But this is a semiconductor process that is gaining acceptance among wireless designers and is bound to make its mark on analog and digital architectures in the years to come.

In a little more than a year, numerous firms have introduced commercial ICs based on SiGe, including TEMIC Semiconductors (Heilbronn, Germany) with last year's announcement of the model U7004B IC for cordless Digital European Cordless Telecommunications (DECT) applications at 2.4 GHz. Integrating a low-noise amplifier (LNA), power amplifier (PA), and transmit/receiver (Tx/Rx) switch driver, the SiGe IC draws 350-mA typical current from a single +3-VDC supply and achieves a typical noise figure (NF) of 2 dB.

Shortly thereafter, SiGe Microsystems (Ottawa, Ontario, Canada) unveiled several amplifier products (see sidebar on p. 124) and Maxim Integrated Products (Sunnyvale, CA) introduced its first SiGe products, a line of downconverter mixers and LNAs for cellular and personal-communications-services (PCS) applications (see *Microwaves & RF*, February 1999, p. 140). Maxim's MAX2680, MAX2681, and MAX2682 downconverter mixers cover a wide range of applications from 400 to 2500 MHz with various NFs and dynamic ranges. The firm's MAX2640 LNA offers more than 15-dB gain and 0.9-dB NF at 900 MHz. The MAX2641 LNA features 14.4-dB gain and 1.3-dB NF at 1900 MHz.

More recently, such firms as Applied Micro Circuits Corp. (San Diego, CA), Intersil/Harris (Melbourne, FL), and Stanford Microdevices (Sunnyvale, CA) have demonstrated the commercial possibilities for digital and RF products based on SiGe. Applied Micro Circuits Corp., for example, last month launched the S3057 IC, a multirate serialization/deserialization interface IC for Synchronous Optical Network (SONET)/synchronous-digital-hierarchy (SDH) and Gigabit Ethernet OC-48 applications at 2.488 Gb/s. The low-jitter IC consumes only 1.5-W power from a single +3.3-VDC sup-

WIRELESS WONDERS?

A new website, http://www.wirelesswonders.com, will be offering an extensive listing of products based on wireless technology. While it may be too soon to find silicon-germanium (SiGe) devices in these products, other advanced technologies, including GaAs heterojunction bipolar transistors (HBTs) and bipolar-complementary-metal-oxide-semiconductor (BiCMOS) integrated circuits (ICs) commonly support a wide range of wireless products, from door chimes to sophisticated medical equipment.

Each year, literally thousands of new products are introduced with some form of wireless technology. Some are simple devices. Others are elaborate systems. In addition to learning about wireless products, visitors to http://www.wirelesswonders.com can first nominate their favorite wireless products of 1999 for top wireless products. Once nominations are accepted, voting for top wireless products begins on the site from December 1, 1999 to January 31, 2000. All entries will be eligible to win one of the seven wireless wonders of the world for 1999, with the drawing to be held at the Wireless Symposium & Exhibition/ Portable By Design Conference in February at the San Jose Convention Center (San Jose, CA).••

ply while working with a reference clock at 155.52 MHz.

Intersil has been converting its successful PRISM line of wireless local-area-network (WLAN) ICs to SiGe bipolar-complementary metal-oxide semiconductor (BiCMOS) as part of the development of the PRISM II chip set. The model HFA3983 SiGe monolithic PA is only one of five ICs aimed at WLAN applications in the 2.4-GHz industrial-scientific-medical (ISM) band. It integrates a 30-dB PA circuit, a logarithmic power-detection function with 15-dB dynamic range, and CMOS-level-compatible power up/down function. The SiGe amplifier features two cascaded low-voltage single-supply stages that combine for +18-dBm typical output power for the direct-sequence, spread-spectrum (DSSS) signals found in WLAN systems. The typical power gain is 30 dB. At the rated output power, the logarithmic detector is accurate within 1 dB. The slope of the detector is 100 mV/dB over its 15-dB dynamic range. The amplifier is designed to work with the PRISM HFA3861 baseband processor and the HFA3783 in-phase/quadrature (I/Q) modulator/demodulator. Similar to many companies entering commercial wireless markets with SiGe ICs, the HFA3983 is fabricated at the SiGe foundry owned and operated by IBM (Hopewell Junction, NY).

General-purpose amplifiers and LNAs in the SGA series from Stanford Microdevices (Sunnyvale, CA) offer low-cost solutions for a variety of wireless applications from DC to 5 GHz. Based on an SiGe heterojunction-bipolar-transistor (HBT) process with 1-μm emitters and f_T of 65 GHz, the amplifier line includes the model SGA-64, which is rated for +20-dBm output power at 1-dB compression from DC to 1.8 GHz. The monolithic microwave IC (MMIC) achieves more than 19-dB gain at 1 GHz and more than 16-dB gain at 2 GHz, as well as drawing 75 mA from a +5.2-VDC supply.

IBM is undoubtedly the single strongest force behind the development and acceptance of SiGe in today's markets. Many firms involved in wireless design and development rely on IBM's Hopewell Junction foundry, which is just north of New York City. RF Micro Devices (Greensboro, NC), for example, has signed an agreement with IBM to use the process to develop SiGe RF building blocks. Partners for IBM's SiGe technology have included Hughes Electronics (Malibu, CA) and Nortel Technology (Ottawa, Ontario, Canada). Also, Philsar Electronics, Inc. (Nepean, Ontario, Canada) has signed a licensing agreement with IBM and plans to use the technology to design low-power-radio ICs. And for some time, IBM has strongly backed the SiGe device developments of CommQuest Technologies (Encinitas, CA) and National Semiconductor (Santa Clara,

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CA), to name a few.

Despite its many partnerships, IBM has also developed several SiGe products of its own, including SiGe discrete devices and ICs. The company's model IBM43RF1111 SiGe LNA, for example, is ideal for cellular and PCS applications. It offers 17-dB gain at 900 MHz and 11-dB gain at 1900 MHz, with a GaAs-like NF of only 1.4 dB at 900 MHz. The unconditionally stable amplifier achieves an input third-order intercept point (IP3) of +8 dBm and draws only 7.5-mA current from a single supply of +2.7 to +3.3 VDC.

Last year, the company reported on the potential of SiGe BiCMOS for creating very-large-scale-integration (VLSI) application-specific ICs (ASICs).[1] Intended to validate design library elements with the SiGe process, the 8.06 × 8.06-mm ASIC may represent the forerunner of a single-chip radio for wireless applications. The ASIC was fabricated with a double-polysilicon self-aligned SiGe BiCMOS process capable of transition frequency (f_T) of 50 GHz at a breakdown voltage of +3.4 VDC and f_T) of 28 GHz at a breakdown voltage of +5.7 VDC.

Earlier this year, the firm also presented work on a transceiver IC for use at 5.2 GHz.[2] Formed with 0.5-μm SiGe BiCMOS technology, the transceiver consists of several mixers, several stages of IF amplification, an RF LNA, and RF driver amplifier. Tests were performed on wafer using a probe from Cascade Microtech (Hillsboro, OR). Designed for operation at 5.2 GHz, the IC achieves a 3-dB bandwidth of 7 GHz, with 11.7-dB downconversion gain, 7.5-dB NF, an input third-order intercept performance of –11.2 dBm, and amplitude imbalance of 0.33 dB across a 300-MHz intermediate-frequency (IF) bandwidth. The transceiver also offers 14.7-dB upconversion gain and output 1-dB compression point of –23 dBm. Using a single +3.3-VDC supply, the IC consumes 122-mW power for the receiver and 114-mW power for the transmitter.

The SiGe story began around 1986 when IBM developed the ultra-high-vacuum/chemical-vapor-deposition (UHV/CVD) processing capability to produce uniform SiGe heterostructures with greater speed than other wafer-processing techniques, such as molecular beam epitaxy (MBE). Although the firm fabricated the first SiGe HBT in 1987 using MBE, the switch to the more efficient UHV/CVD processing approach was made shortly thereafter. In 1993, tremendous publicity was generated by the announcement of a joint effort on the part of IBM and Analog Devices (Wilmington, MA) to produce a 12-b, 1.2-GSamples/s digital-to-analog converter (DAC). Unfortunately, the joint effort by the two firms did not survive the low breakdown voltages and limited yields of the process at that time.

SiGe processes essentially build on established semiconductor process approaches such as Si CMOS, by depositing a thin layer of Ge on top of the Si substrate. Depending on the device characteristics of a particular process, the Ge layer is often graded in thickness, to control parasitic capacitances and inductances and optimize high-frequency performance. The resulting compound semiconductor material has a carrier mobility that is two to three times that of standard Si, while maintaining the relatively low cost of standard Si CMOS processing and materials. The high carrier mobility is of particular benefit to certain types of transistor structures, notably HBTs. Deep trenches are formed in the SiGe materials to achieve the isolation needed for high-quality passive components, such as capacitors. Even compared to GaAs, the proponents of SiGe processing, such as IBM, promise lower-voltage/lower-power operation, less low-frequency noise, improved third-order intercept performance, and tremendous potential for integration of analog and digital functions. Yet, supporters of GaAs claim that the cost differential between GaAs and SiGe is not that great due to the additional processing steps required to create SiGe ICs, and that GaAs is still a better substrate for forming high-performance passive circuit elements.

If the papers at the upcoming 45th Annual IEEE International Elec-

WHAT'S IN A NAME?

No firm could be more closely linked to SiGe technology than one with the name SiGe Microsystems, Inc. (Ottawa, Ontario, Canada). The company, one of the first to announce commercial SiGe products, already supports numerous industries with circuits of varying complexity. The first product, a differential Global Positioning System (GPS) receiver, operates at 1.5 GHz.

More recently, the firm announced the model PA2425 monolithic SiGe power amplifier (PA) for use in the 2.4-GHz industrial-scientific-medical (ISM) band. Ideal for applications in IEEE 802.11 wireless local-area networks (WLANs), Bluetooth systems, and HomeRF products, the high-efficiency PA delivers +25-dBm typical output power (+24-dBm minimum output power) from 2400 to 2500 MHz with 45-percent power-added efficiency (PAE) in Class AB mode. The amplifier exhibits gain variations of typically ∓0.5 dB across the full operating band. The reverse isolation is typically 32 dB, while second, third, and fourth harmonics are –40 dBc.

Supplied in an eight-lead plastic-small-outline-package (PSOP) housings, the amplifier's SiGe structure and heatslug die pad provide high thermal conductivity and resulting low junction temperatures. The effective thermal dissipation enables the amplifier to operate at a 100-percent duty cycle while drawing only 220 mA from a single +3.3-VDC supply. The integrated circuit (IC) includes bias control and power-down functions. SiGe Microsystems, Inc., 1500 Montreal Rd., M50 IPF, Ottawa, Ontario K1A OR6, Canada; (613) 748-1334, FAX: (613) 748-1635, Internet: http://www.sige.com.••

tron Devices Meeting (December 5-8, 1999, Washington Hilton and Towers, Washington, DC) are any indication, SiGe technology promises to become an even greater threat for GaAs and conventional Si processes in the years to come. G. Freeman, D. Ahlgren, and co-workers from IBM Microelectronics will offer insights into the next generation of SiGe devices with a presentation at IEDM on a 0.18-µm SiGe process that is BiCMOS and ASIC compatible. Capable of producing HBTs with f_T performance of 90 GHz at a breakdown voltage of +2.3 VDC, lower-frequency performance is possible at 25 GHz for a breakdown voltage of +5.5 VDC.

Also, a report from Katsuyoshi Washio and associates from the Central Research Laboratories of Hitachi Ltd. (Tokyo, Japan) details a 0.2-µm self-aligned SiGe HBT capable of 6.7-ps emitter-coupled-logic (ECL) gate delays and maximum frequency of oscillation (f_{max}) of 107 GHz. Compatible with standard Si BiCMOS, the new process promises to support high-speed data-communications systems operating in excess of 10 Gb/s. To reduce parasitic capacitance across devices, shallow-trench and deep trench isolations were used.

A paper to be given by K.E. Ehwald and colleagues from the Institute for Semiconductor Physics (Frankfurt, Germany) in conjunction with researchers from Motorola, Inc. (Mesa, AZ) explains how only four additional mask layers are needed to add SiGe:C HBTs to a standard 0.25-µm CMOS process without changes in the CMOS process flow. The resulting BiCMOS process yields peak HBT transition frequency (f_T) of 55 GHz and f_{max} of 90 GHz at a breakdown voltage of +3.3 VDC.

Finally, C.A. King and fellow researchers from Bell Laboratories/Lucent Technologies (Murray Hill, NJ) will offer a technique for lowering the cost of SiGe processing. Their solution is a low-cost graded SiGe-base BiCMOS technology with Lucent's existing 0.25-µm CMOS process at the core. The modified, low-cost process requires the addition of only four lithography levels to the existing process, and results in self-aligned SiGe transistors with peak f_T of 51 GHz and peak f_{max} of 53 GHz at a breakdown voltage of approximately +2.5 VDC.

Several years ago, mention of SiGe technology brought raised eyebrows and much skepticism about its practical application in either high-frequency analog or high-speed digital circuits. Today, the technology boasts a growing list of commercial products, manufacturers, as well as believers in its future for high-frequency, high-speed products (see sidebar on p. 121). ••

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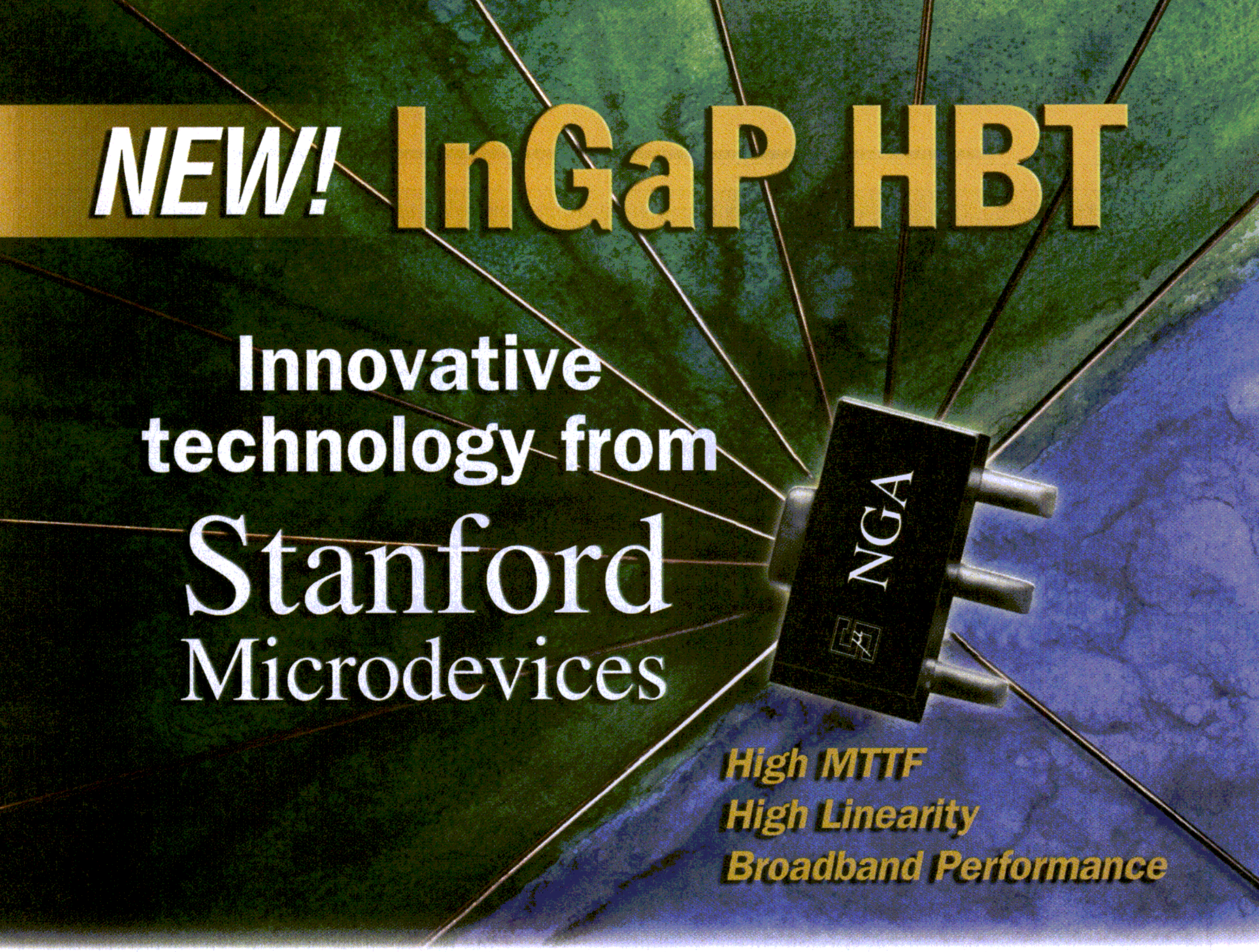
1. Robb Johnson *et al.*, "1.8-million CMOS ASIC Fabricated In A SiGe BiCMOS Technology," *IEDM Technical Digest*, 1998, pp. 217-220.
2. Jean-Olivier Plouchart, Herschel Ainspan, and Mehmet Soyuer, "A 5.2-GHz 3.3-V I/Q SiGe RF Transceiver," Custom IC Conference, Session 10.5, 1999.



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- Low positive voltage supply
- Low thermal resistance package
- High linearity

SPECIFICATION MATRIX

	NGA-489	NGA-589
Frequency (GHz)	DC-8.0	DC -6.0
Gain (dB)	14.5	19.0
TOIP (dBm)	38.5	38.0
N.F. (dB)	4.5	4.5
P1dB (dBm)	17.5	19.0
Supply Voltage	4.2	5.0
Supply Current	80	80

All data measured at 900MHz and is typical.
MTTF @ 150C T_j = 2 million hrs. (R_{TH} = 110 C/W typ.)

NGA-589 DC-6 GHz

High gain and high output make this heterojunction bipolar transistor MMIC amplifier ideal for use in all wireless applications. InGaP HBT technology improves the reliability and performance and minimizes leakage current between junctions. Other features include:

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- Low thermal resistance package
- High linearity
- High gain
- High P1dB

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CIRCLE NO. 370

HSCSD adds power to GSM services

Wireless communications systems worldwide are dominated by the Global System for Mobile Communications (GSM), the single largest service in number of subscribers. That system is about to become even more dominant, with the addition of the new High Speed Circuit Switched Data (HSCSD) service. The service, which is part of the GSM Phase 2+ expansion, offers multislot data transfer at rates up to 14.4 kb/s per channel, resulting in data rates as high as 43.2 kb/s for mobile stations with access to three receiving time slots. An application from Rohde & Schwarz (Munich, Germany), "Measurement software and test cases for new, fast GSM data services," provides details on the new GSM service and software options that are available to upgrade existing test systems for evaluating equipment with the new service.

The application note is part of Issue 163 of *News from Rohde & Schwarz*, which also contains product-based articles on electromagnetic-compatibility (EMC) testing, the ACCESSNET trunked-data system, spectrum analyzers for third-generation (3G) wireless systems, and techniques for simulating channel fading at baseband frequencies. For a free copy of Issue 163 of *News from Rohde & Schwarz*, contact: **Rohde & Schwarz GmbH & Co. KG, Press Office, Muhldorfstrasse 15, D-81671 Munich, Germany; (49) 89-4129-1765, FAX: (44) 89-4129-3208, Internet: http://www.rsd.de.**

CIRCLE NO. 194 or visit www.mwrf.com

Evaluating telecom network performance

Interoperability of telecommunications networks is critical to reliable, error-free performance in diversified networks supporting such services as telephone, data, Internet, and video communications. An application note from Wandel & Goltermann GmbH & Co. (Eningen, Germany), "How to x-ray a TMN with the QMonitor," details the maintenance and evaluation of telecommunication-management networks (TMNs).

Since the International Telecommunications Union (ITU) standardized its TMNs for use by all manufacturers of telecommunications equipment, a universal standard has been available for network testing and management. One of the benefits of the TMN is its versatile applicability—it can manage entire networks or network elements as well as logical applications, such as services and business applications. Application note 67 examines the use of Wandel & Goltermann's QMonitor software package for analysis of TMN performance problems.

The application note offers suggestions on troubleshooting management-protocol definitions in advanced networks, and outlines how to assess TMN performance. Copies of the 12-page note are free, from: **Wandel & Goltermann GmbH & Co., Marketing International, Postfach 1262, D-72795 Eningen, Germany; (49) 712186-1616, FAX: (49) 712186-1333, e-mail: info@wwgsolutions.com, Internet: http://www.wwgsolutions.com.**

CIRCLE NO. 195 or visit www.mwrf.com

Using a low-cost UHF receiver IC

Receiver design has become somewhat simplified during the last few years with the availability of low-cost integrated circuits (ICs) for many of the front-end functions. A case in point is the MICRF001 QwikRadio receiver IC from Micrel, Inc. (San Jose, CA). It is a complete ultra-high-frequency (UHF) receiver intended for use from 300 to 440 MHz. Application note 22, "MICRF001 Theory of Operation," provides 20 pages of detailed information on the operation and application of the MICRF001.

Written by Tom Yestrebsky, application note 22 explains how the MICRF001 can be used with only one ceramic resonator and two capacitors to form a complete UHF receiver. The application note provides background theory of operation on the IC, along with information on superheterodyne receiver architectures and super-regenerative receivers. It points out the advantages of the MICRF001's architecture compared to standard superheterodyne receivers, and details the control signals required for proper operation. Copies of application note 22 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. 196 or visit www.mwrf.com

CIRCLE NO. 380

Wideband Choke Biases Amplifier Circuits To 8 GHz

This broadband RF choke can save valuable gain and output power when supplying power to monolithic microwave amplifiers.

Engineering Department

Mini-Circuits, P.O. Box 350166, Brooklyn, NY 11235; (718) 934-4500, FAX: (718) 332-4661, Internet: http://www.minicircuits.com.

WIDEBAND monolithic amplifiers operating to 8 GHz have become commonplace. Unfortunately, bias circuitry for these amplifiers has not been readily available until the recent introduction of the model ADCH-80A wideband microwave choke from Mini-Circuits (Brooklyn, NY). The choke supports RF/DC applications from 50 to 8000 MHz and can help to significantly reduce resistive RF signal losses in the bias circuitry of broadband microwave amplifiers.

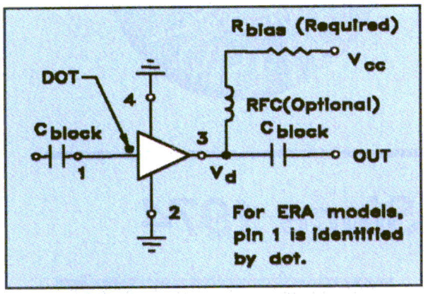
Widespread use of wideband monolithic-microwave amplifiers, such as the ERA series from Mini-Circuits, has required circuit designers to develop broadband power-supply circuitry that supports the bias requirements of these integrated circuits (ICs). Monolithic microwave amplifiers are typically biased through current injection at the RF output port. Since DC power and RF output signals share this port, it must exhibit strong DC and RF characteristics. An inadequately designed DC biasing circuit will degrade the RF performance. One recommended bias approach involves the use of a resistor and RF choke in series with the DC supply. The RF choke serves to minimize the RF losses of the resistor.

The biasing scheme for a typical monolithic microwave amplifier is shown in Fig. 1, with the biasing resistor denoted by R_{bias}. The value of this resistor is determined by the device's required voltage, the supply voltage, and the desired operating current of the amplifier design. For example, for a model ERA-1 monolithic amplifier, the device voltage is +3.6 VDC. Assuming that the supply voltage is +12 VDC, the value of the biasing resistor will be:

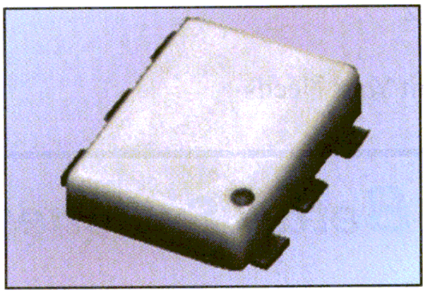
$$R_{bias} = (V_{cc} - V_d)/I_d = 210\ \Omega$$

where:
V_{cc} = the supply voltage, and
I_d = the biasing current.

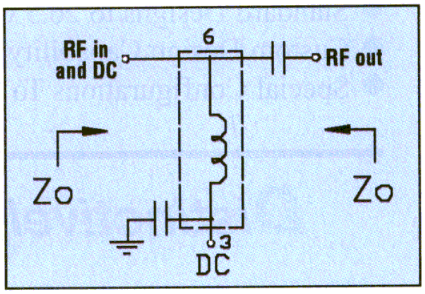
The use of a 210-Ω bias resistor without an RF choke in series results in a 1-dB loss of amplifier gain and output power. If the supply voltage is dropped to +5 VDC, then the required bias resistor, R_{bias}, will be 35 Ω. The use of this smaller resistor results in a loss of 4.6 dB in gain, a loss in output



1. This simple schematic diagram represents a basic biasing scheme for the ERA line of monolithic microwave amplifiers.



2. The model ADCH-80A RF choke is housed in a miniature surface-mount package compatible with automated assembly equipment.

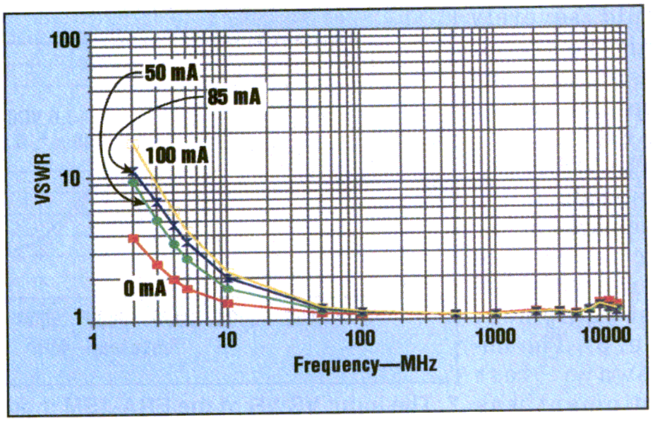


3. This simple evaluation circuit is used to interconnect the model ADCH-80A RF choke to 50-Ω test equipment.

power, and degraded return-loss performance for the amplifier. However, the addition of the RF choke in series with the bias resistor adds an inductive reactance to the resistor's shunt impedance, minimizing the effect of the resistor on the performance of the monolithic amplifier. This simple example illustrates how a wideband choke such as the model ADCH-80A can dramatically improve the performance of wideband amplifiers.

Commercially available inductors can be used as RF chokes. The low-frequency limit for this approach is determined by the component's inductance. Higher values of inductance deliver lower-frequency performance. The high-frequency limit for an inductor is determined by its series resonant frequency. The series resonant frequency tends to decrease with increasing values of inductance. Thus, when using an inductor as an RF choke, the RF bandwidth will be sharply limited by the value of inductance, complicating the job of the circuit designer. Any design changes that are performed by the inductor manufacturer will have an unknown effect on the amplifier circuitry.

The ADCH-80A (Fig. 2) was designed to optimize the performance of RF amplifiers over wide frequency ranges (Table 1). Its frequency range of 50 to 8000 MHz is wide enough to support amplifiers such as the ERA-1, which is specified to 8 GHz. The equivalent inductance of the ADCH-80A is 1 μH. In comparison, a typical commercial 1-μH inductor has a series resonant frequency as low as 90 MHz which

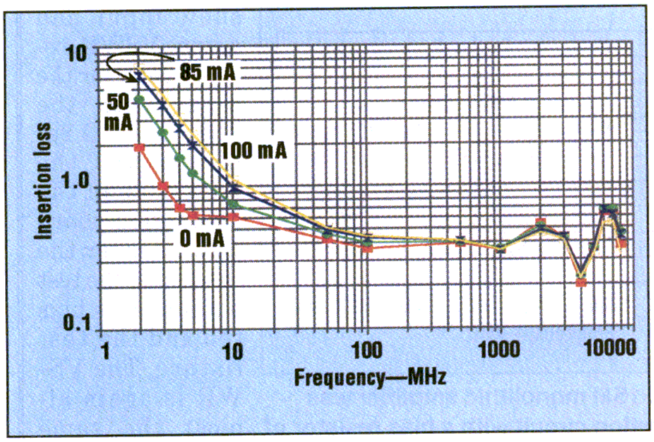


5. The return-loss performance of the ADCH-80A was evaluated through 8 GHz with applied currents of 0, 50, 85, and 100 mA.

Table 1: The ADCH-80A RF choke at a glance

Frequency (MHz)	50 to 8000
Insertion loss (dB)	0.4 typical 1.0 maximum
VSWR	1.15:1 typical, 1.35:1 maximum
DC current (μA)	100 maximum
Inductance (μH)	7 at 0 μA typical
	1.8 at 50 μA
	1at 100 μA

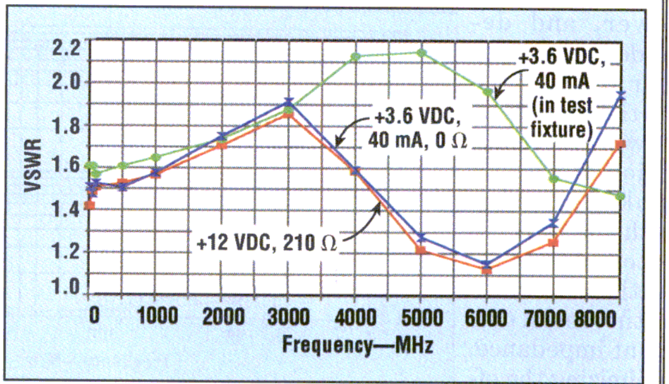
Note: Two different models ADCH-80A and ADCH-80 are available with the above specs. Choice of the model depends on the PCB layout in the application.



4. The insertion loss of the ADCH-80A RF choke was measured across a broad bandwidth with applied currents of 0, 50, 85, and 100 mA.

would severely limit the usefulness of a wideband amplifier such as the ERA-1.

The ADCH-80A was evaluated in an application test circuit designed for use with 50-Ω test equipment (Fig. 3). The microwave test equipment was used to measure insertion loss (Fig. 4) and return loss (Fig. 5) with currents up to 100 mA. The insertion loss and voltage standing-wave ratio (VSWR) were found to change very little with changes in current.

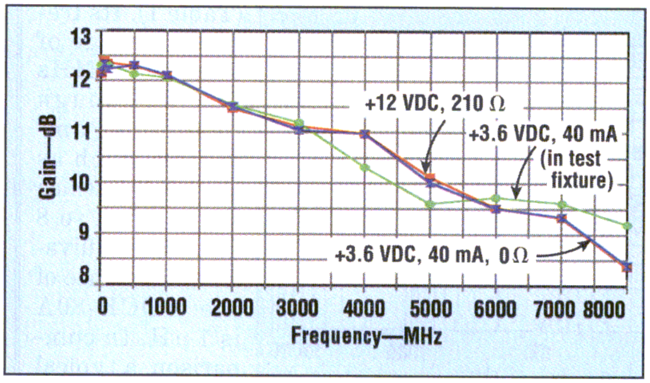


7. The input VSWR of the ERA-1SM monolithic amplifier was measured in the evaluation circuit with a bias resistor of 210 Ω and +12-VDC supply, in the evaluation circuit with a bias resistor of 0 Ω and a supply of +3.6 VDC, and in the test fixture with a supply of 40 mA and +3.6 VDC.

EVALUATING PERFORMANCE

To illustrate the use of the RF choke, evaluation boards were built using the ERA-1SM amplifier, based on the schematic diagram that appears in Fig. 1. Figure 6 shows the gain of the amplifier with two values of V_{CC}: +12 and +3.6 VDC. In the first case, the biasing resistor is 210 Ω. In the second case, the biasing resistor is 0 Ω. The gain levels in both cases are very close, demonstrating the effectiveness of the RF choke in wideband circuits. In practice, a finite biasing resistor is recommended for introducing a potential drop of a few volts to ensure bias-current stability, unless a constant-current source is used for biasing (see the Mini-Circuits publication, "Biasing ERA Amplifiers," which is available on the company's website at http://www.minicircuits.com/appnote/an60010.htm).

Figure 6 also shows the gain of the ERA-1SM amplifier in a 50-Ω test fixture. In this case, the biasing current is supplied through the bias tee that is a part of the microwave test system's S-parameter test set. Note that the gain is very close to that measured in the RF choke evaluation board except in the range of 7 to 8 GHz. Part of the difference is due to the longer transmission lines which increase the loss of the evaluation board.



6. The gain of the ERA-1SM monolithic amplifier was measured in the evaluation circuit with a bias resistor of 210 Ω and +12-VDC supply, in the evaluation circuit with a bias resistor of 0 Ω and a supply of +3.6 VDC, and in the test fixture with a supply of 40 mA and +3.6 VDC.

Figures 7 and 8 show input and output VSWR, respectively, for the cases with the 210- and 0-Ω resistors and the RF choke in the evaluation board as well as for the case using the test equipment's bias tee and the test fixture. The VSWR is again almost the same with the 210- and 0-Ω biasing resistors. There is

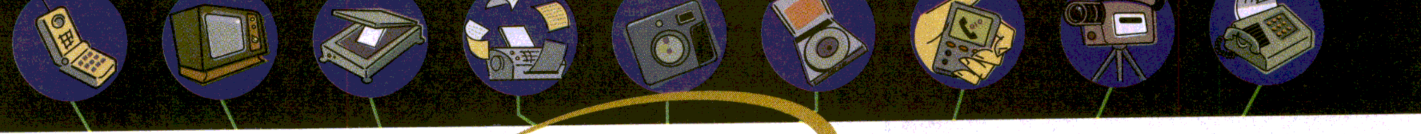


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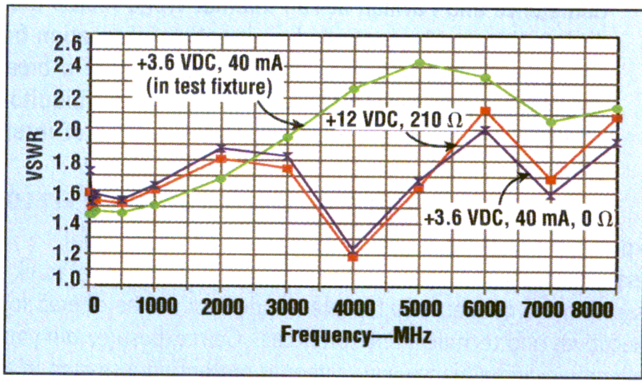
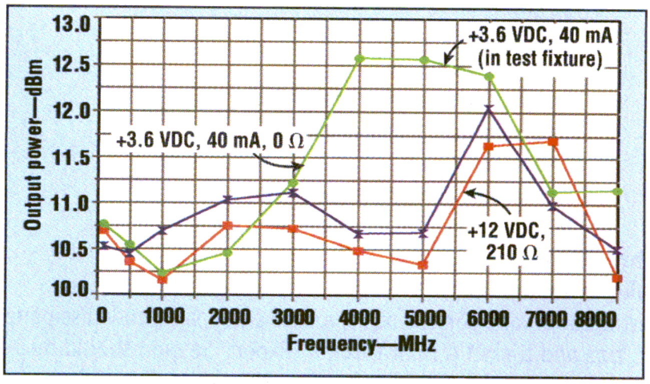
Table 2: Comparing linear performance

Device	Frequency (GHz)	Third-order intercept in evaluation board with RF choke (dBm)	Third-order intercept in test fixture (dB)
ERA-1	2	+25.0	+25.5
ERA-2	2	+25.5	+25.6
ERA-3	2	+23.0	+23.1
ERA-4	1	+33.2	+33.2
ERA-5	1	+33.5	+33.8
ERA-6	1	+37.5	+37.0

slight improvement in the evaluation board. Part of this improvement is due to fixed-stub matching. This is required to compensate for the difference in the ground pattern of the evaluation board compared to the test fixture. Figure 9 shows the 1-dB compression for the three cases. Another characteristic that can be affected by the RF choke, if its magnetic core is nonlinear, is the third-order intermodulation performance. Table 2 shows the third-order intercept-point (IP3) performance of the ERA-1 amplifier measured with the test fixture and with the evaluation board and the RF choke. No measurable difference was found. Table 2 also offers data on the ERA-2 through ERA-6 amplifiers, again showing the closeness of the third-order-intercept (TOI) performance in the various cases. These measurements indicate that the RF choke is not degrading the TOI performance of these amplifiers.



8. The output VSWR of the ERA-1SM monolithic amplifier was measured in the evaluation circuit with bias resistors of 210 and 0 Ω as well as in the test fixture with an instrumentation-grade bias tee.



9. The output power at 1-dB compression for the ERA-1SM amplifier was measured in the evaluation circuit with bias resistors of 210 and 0 Ω as well as in the test fixture with an instrumentation-grade bias tee.

The ADCH-80A RF choke is designed to simplify the bias-circuit design requirements for monolithic and discrete amplifiers operating through 8 GHz. Although it is not DC coupled for use with fiber-optic amplifiers, it should serve the majority of wireless applications through 6 GHz. It is well-suited for biasing the entire Mini-Circuits ERA series and similar amplifiers operating up to 8 GHz.

CIRCLE NO. 51 or visit www.mwrf.com

PRODUCT TECHNOLOGY

SiC MESFET

SiC MESFET Delivers 10-W Power At 2 GHz

This innovative semiconductor technology offers the outstanding thermal conductivity and electron velocity needed for creating high-power, high-frequency devices.

JACK BROWNE
Publisher/Editor

SILICON carbide (SiC) has often been lumped together with Si germanium (SiGe) as a semiconductor technology of the future. But similar to SiGe, SiC devices have become commercially available this year from a company called Cree, Inc. (Durham, NC). And unlike SiGe, SiC transistors are built for power and speed. The first commercial product is the CRF-20010 SiC metal-semiconductor field-effect transistor (MESFET). It makes use of SiC's high breakdown voltage and high thermal conductivity to yield 10 W of linear output power for driver amplifiers from 400 to 2500 MHz.

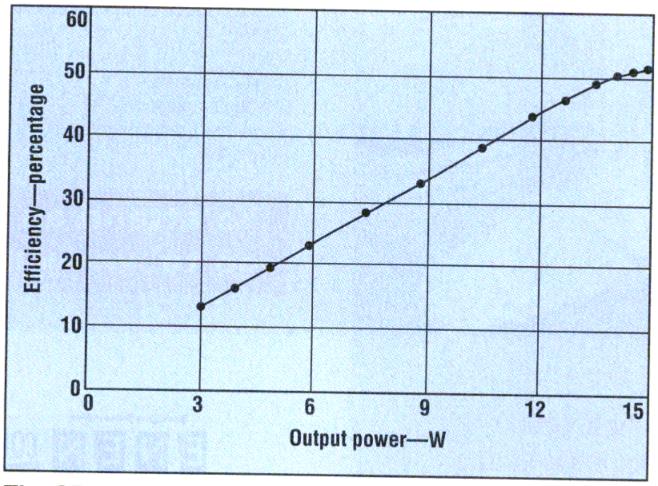
The CRF-20010 is based on the Cree's proprietary +48-VDC SiC MESFET technology. The CRF-20010 is designed to work directly with a base station's +48-VDC supply, without additional DC conversion. The company, founded in 1987, is perhaps best known for the application of SiC to the development of SiC blue light-emitting diodes (LEDs) and photodetectors. Cree has been a supplier of SiC LED components, as well as wafer materials to device manufacturers for several years, but has also been aware of the material's tremendous potential for RF and microwave devices.

The company, which last month received research contracts valued at \$6.8 million from the Office of Naval Research and the Air Force Research Laboratories for the development of wide-bandgap SiC microwave and power-switching devices, offers two grades of substrates—4H-SiC and 6H-SiC. The 4H-SiC materials are used to develop devices to increase the output power of a solid-state phased-array radar, to reduce the weight and volume of communications systems, and to create devices with power densities that are approximately four times higher than possible with Si or GaAs devices.

Cree has already fabricated Schottky diodes with blocking voltages to 2000 V and current levels of 1 A as well as PIN diodes with blocking voltages of better than 6.2 kV. MESFETs based on 4H-SiC substrates have yielded transition frequencies (f_T) as high as 22 GHz and maximum frequency of oscillation (f_{max}) of 50 GHz. Power densities of 4.6 W/mm at 1.8 GHz have been achieved in Class A designs, while Class B power densities of 3 W/mm have reached 0.8 and 1.8 GHz with a power-added efficiency (PAE) of up to 60 percent. MESFETs that are based on SiC can operate at temperatures as high as +500°C.

SiC has several characteristics that make it an ideal substrate material for high-power, high-frequency transistors. The company's two substrate materials exhibit wide energy bandgaps—3.03 eV for 6H-SiC (predominately used for optoelectronic devices) as well as 3.26 eV for 4H-SiC (predominately used for microwave and power devices) compared to 1.12 eV for Si and 1.43 eV for GaAs. SiC devices can operate at high temperatures without suffering from intrinsic conduction effects due to the wide energy bandgap. It is the wide energy bandgap that makes SiC ideal for blue LEDs in addition to photodetectors.

SiC can withstand an electric field more than eight



The CRF-20010 SiC MESFET provides better than 50-percent PAE at its highest output-power levels.

Comparing semiconductor characteristics				
Parameter	Si	GaAs	4H-SiC	6H-SiC
Energy bandgap	1.12 eV	1.43 eV	3.26 eV	3.03 eV
Breakdown E field	2.5 x 105 V/cm	3 x 105 V/cm	2.2 x 106 V/cm	2.4 x 106 V/cm
Thermal conductivity	1.5 W/cm-K	0.5 W/cm-K	3.0-3.8 W/cm-K	3.0-3.8 W/cm-K
Saturated electron drift	1.0 x 107 cm/s	1.0 x 107 cm/s	2.0 x 107 cm/s	2.0 x 107 cm/s

times greater than that of Si or GaAs without undergoing avalanche breakdown. Also, the material is an excellent thermal conductor. At room temperature, SiC has a higher thermal conductivity than any metal, enabling SiC devices to dissipate extremely high power levels per unit size. This leads to higher RF power being available in similar-size packages to those used to house Si or GaAs devices. Finally, the high saturated electron drift velocity of SiC, which is twice that of Si and GaAs (see table), enables devices that are fabricated with the material to operate at RF as well as microwave frequencies.

OUTPUT POWER

The CRF-20010 is Cree's first semiconductor device based on its own substrate materials. The transistor is designed for 10 W of Class A output power from 400 to 2500 MHz. It achieves 12-dB gain at 2 GHz, with impressive PAE of 50 percent (see figure). It is ideal for code-division-multiple-access (CDMA) cellular systems as well as time-division-multiple-access (TDMA) driver-stage applications. The transistor attains extremely low levels of third-order intermodulation distortion (IMD) when "backed off" from its nominal 10-W output-power level. One example is when it operates at 5-W linear output power. The third-order IMD is –40 dBc. At the rated output power, the third-order IMD is better than –30 dBc.

The device achieves a maximum drain-source voltage of +100 VDC, with maximum gate-source voltages of +3 and –15 VDC. The maximum total device dissipation at +25°C is 75 W, and a maximum operating junction temperature of +250°C.

The CRF-20010 is the first device in a full line of RF transistors for wireless applications. Higher-power Class AB devices will take aim at output-power levels of 30, 60, and 120 W at frequencies up to 4 GHz. **Cree, Inc., 4600 Silicon Dr., Durham, NC 27703; (919) 313-5300, FAX: (919) 313-5451, Internet: http://www.cree.com.**

CIRCLE NO. 52 or visit www.mwrf.com

Switch Solution Integrates GaAs And Si ASIC

This +3-VDC switch assembly is available for high-linearity and high-power applications in a low-cost plastic package.

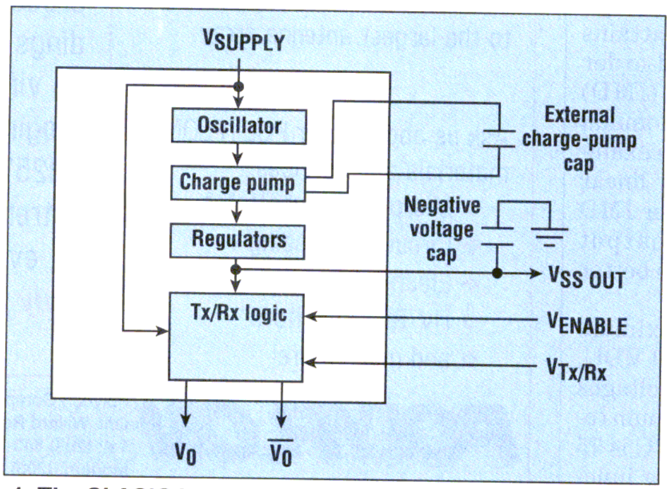
Steve Belletete
Senior Design/Applications Engineer
Paul Martyniuk
RF Engineer
Alpha Industries, 20 Sylvan Rd., Woburn, MA 01801; (781) 935-5150, FAX: (781) 935-2359.

SWITCH designers are often challenged to produce transmit (Tx) and receive (Rx) switches that can operate at low voltages and still maintain linear performance. This task will only become tougher as wireless products shrink in size. Alpha Industries (Woburn, MA), meanwhile, has integrated a silicon (Si) application-specific integrated circuit (ASIC) with gallium-arsenide (GaAs) technology to produce an Si DC-to-DC converter with a decoder into the same package as a GaAs switch. The packaged unit is suitable for applications from 0.5 to 2.5 GHz.

Currently, the best solution for switching applications is GaAs field-effect-transistor (FET) technology. This technology supports some integration along with low current consumption. Even with advancements in GaAs technology—metal semiconductor FETs (MESFETs) or pseudomorphic high-electron mobility transistors (PHEMTs)—there are, however, remaining limitations to power-handling capability and logical function integration.

GaAs MESFET technology requires greater than +6-VDC differential bias between the drain/source and gate of the FET to meet linearity requirements at Global System for Mobile Communications (GSM) power levels. GaAs PHEMT technology requires slightly less voltage, but it still does not meet handset designers' desire for a true single-supply +3-VDC solution. One method of obtaining the necessary bias is to use the negative voltage required by the power amplifier (PA) to increase the bias on the switch. Recent advancements in PA design have excluded a negative voltage that is available for the switch.

The integration of an Si DC-to-DC converter with a decoder into the same package as the GaAs switch has produced an Si ASIC that is available for GSM Tx and Rx functions. The Si ASIC consists of an oscillator, a charge-pump-based DC converter, voltage regulators, temperature-robust bandgap biasing, and driver-logic circuitry (Fig. 1). The internal voltage regulator enables the designer to have the flexibility of powering the ASIC directly from the battery while still using +3-VDC logic levels to switch states. As an additional benefit, isolation is achieved for the radio power supply from the inherently noisy internal oscillator circuitry.

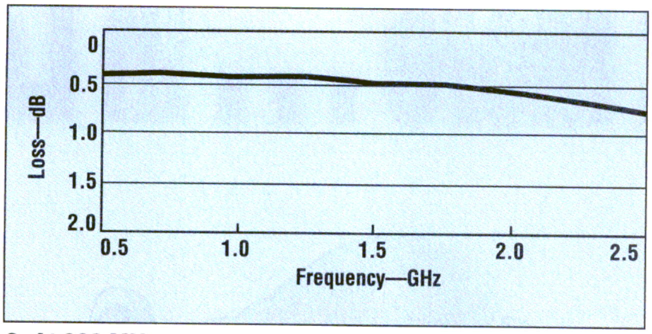


1. The Si ASIC is made up of an oscillator, a charge pump, voltage regulators, and Tx/Rx logic.

The negative voltage that is generated by the charge-pump circuitry is used to bias the switch during transmit mode. With a large differential voltage, the switch maintains its linearity at high RF power levels—the GSM power requirement of +34.5 dBm for example. The frequency of the charge pump was chosen to be 600 kHz to keep the required external switching capacitor low in value and, at the same time, keep the oscillator frequency low enough so that it does not interfere with other components in the radio. Lower-value ceramic

capacitors can be used to replace expensive and large tantalum (Ta) capacitors that are typically required by other DC-to-DC converter ASICs.

The logic portion of the ASIC driver circuitry reduces overall system complexity by switching from the Rx-to-Tx function with one control line. A single positive control line is the result. It helps conserve the number of control lines that have to be allocated on the central processing unit (CPU) and routed out to the switch. With the use of internal logic, the current consumption during the receive state can be minimized by powering down the charge-pump and oscillator circuitry. A large differential voltage is not required during the receive mode since the power levels are much lower. The additional value of logic circuitry comes free since it does not contribute significantly to the Si die size.



2. At 900 MHz, the insertion loss of the switch is less than 0.5 dB.

By using the best technology available for each function, Alpha is able to keep die size small enough to be integrated into a single TSSOP-16 package. By using the appropriate technology for each function, the cost of the overall solution remains in line with handset manufacturers' cost requirements.

Figure 2 shows insertion loss versus frequency. At 900 MHz, the insertion loss is less than 0.5 dB. It can also be seen that insertion loss of the switch spans approximately from –0.30 to –0.75 dB across the 0.5-to-2.5-GHz frequency range. The isolation of the switch ranges from –20 to approximately –41 dB over the 0.5-to-2.5-GHz range. At 900 MHz, the isolation is greater than 20 dB. At the distributed-communications-systems (DCS) frequency, the isolation is approximately 30 dB.

Alpha's Si ASIC requires 350 μA during transmit mode and 25 μA during receive mode. Switching between Rx and Tx states is accomplished with a single logic input. The switch and its associated ASIC will be used as a building block for other higher integration products. **Alpha Industries, 20 Sylvan Rd., Woburn, MA 01801; (781) 935-5150, FAX: (781) 935-2359.**

Acknowledgments

The authors would like to thank Dave Baldwin, Erick Olsen, and Sheila Rosado for their contributions to this article.

CIRCLE NO. 55 or visit www.mwrf.com

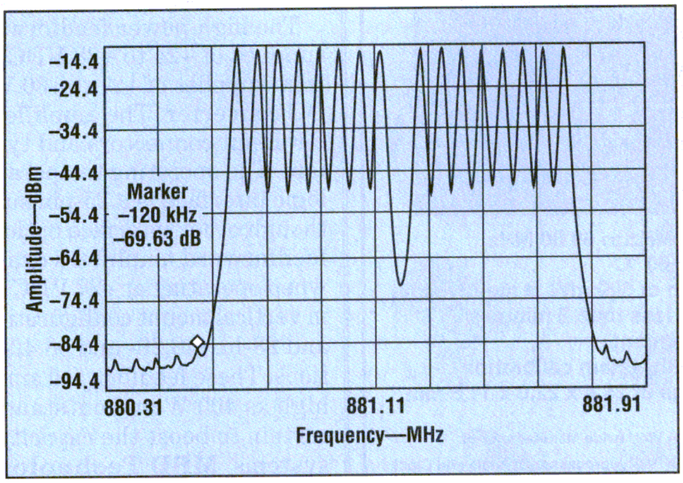
Feedforward Amplifiers Power Base Stations To 400 W

The low-distortion benefits of feedforward amplification are now available at power levels to 400 W at cellular frequencies from 869 to 894 MHz.

JACK BROWNE
Publisher/Editor

FEEDFORWARD techniques are now well-accepted as a way of boosting multicarrier signals in cellular and personal-communications-services (PCS) base-transceiver systems (BTS). Feedforward amplifiers are well-suited for amplifying multiple protocols simultaneously, such as Advanced Mobile Phone Service (AMPS) channels concurrent with time-division-multiple-access (TDMA) and code-division-multiple-access (CDMA) channels. The technology offers linear amplification with minimal addition of intermodulation distortion (IMD). With a new line of feedforward amplifiers from MPD Technologies, Inc. (Hauppauge, NY), cellular and PCS system designers can now add up to 400 W of feedforward output power to accommodate the most aggressive multicarrier system configurations.

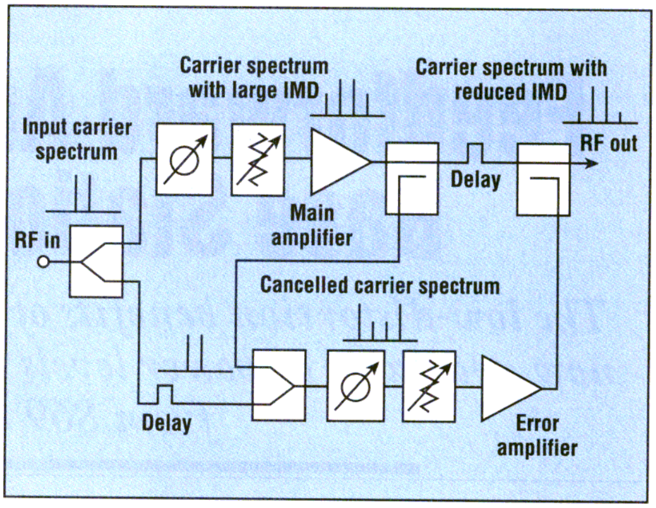
The new feedforward amplifier system has a modular configuration that supports minimum output-power levels of 100, 200, 300, or 400 W, depending on the number of installed 120-W power-amplifier (PA) modules. The highest output-power unit employs four 120-W feedforward power modules which are combined through passive power combiners. Regardless of which output-power level is selected, the forward and reverse IMD performance for these amplifiers is rated at −65 dBc. Spurious performance is also rated at −65 dBc. Spectrum-analyzer measurements centered at 881.11 MHz reveal the low spurious and IMD levels when tested with multitone signals (Fig. 1).



1. The output spectrum of the 400-W feedforward power amplifier was measured with a spectrum analyzer centered at 881.11 MHz and the reference level set at −14.4 dBm as well as 50 video averages (to lower the displayed noise levels).

The output-power requirement for a BTS is a function of the number of carriers to be transmitted and the amount of power needed to transmit each carrier. For example, a feedforward power amplifier with a 35-W output-power rating can handle 70 transmit carriers of 0.5 W each, or 35 carriers of 1 W each. In a BTS, the signals from several radios are combined at low power levels and then simultaneously amplified by a multicarrier amplifier connected to an antenna and its matching circuitry. Compared to the use of multiple single-channel amplifiers, this approach reduces the losses in the high-power section of the BTS by eliminating the output filter/combiner assemblies and thereby reducing the necessary system amplifier power. When more channels are needed, radios and low-power combiners can be added without necessarily changing the multichannel amplifier (assuming that enough output power is available) or wideband antenna circuitry.

Feedforward techniques make it possible to essentially cancel IMD by using the distortion as part of a correction circuit. In a basic feedforward amplifier design (Fig. 2), the distortion signals must be first isolated from the desired spectrum. These signals must then be linearly amplified by an error amplifier to produce a correction term. The correction signals are then injected 180 deg. out of phase at the out-



2. This block diagram illustrates how error signals are isolated, amplified, then delayed 180 deg. out of phase for mixing with output signals in order to minimize IMD.

put of the feedforward amplifier in order to achieve IMD cancellation.

Class AB laterally-diffused-metal-oxide-semiconductor (LDMOS) transistors are used in the main amplifier section of MPD's feedforward amplifier designs, with similar devices employed in the error amplifiers. Control circuitry continuously monitors the operating points of the main and error amplifiers, with dynamic adjustments made to minimize IMD under all environmental and load conditions. Adjustments are made according to a proprietary algorithm, under high-speed microprocessor control. Control and status data are available through relay contacts, parallel bus, or serial bus. The flexible data access supports local control and status monitoring at a cell site, or remote control and monitoring through modem connections.

VOLTAGE-SUPPLY VERSATILITY

The high-power feedforward amplifiers operate from supplies of +24 to +28 VDC, although they can also run from supplies of +21 to +30 VDC with an optional DC-to-DC converter. The amplifiers incorporate SMA female RF input connectors and type-N female output connectors. The operating temperature range is 0 to +50°C. The logic interface is a 25-pin subminiature D connector, although other interface options are available. The 400-W feedforward amplifiers draw nominal current of 200 A when operating at +27 VDC. The amplifiers are available in vertical mount configurations as well as in standard 19- and 23-in. (48.26- and 58.42-cm) rack-mount configurations. These feedforward amplifiers, with power levels as high as 400 W and outstanding IMD performance, are certain to boost the capacity of a wide range of cellular systems. **MPD Technologies, Inc., subsidiary of Microwave Power Devices, Inc., 49 Wireless Blvd., Hauppauge, NY 11788-3935; (516) 231-1400, FAX: (516) 231-0712, e-mail: sales@mpd.com, Internet: http://www.mpd.com.**

CIRCLE NO. 56 or visit www.mwrf.com

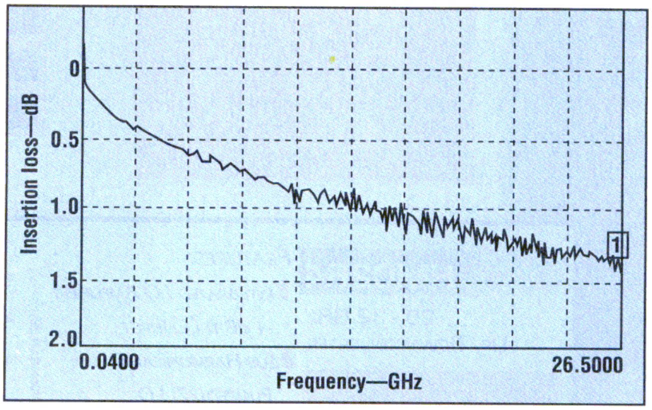
Improvements Yield Cables With Lower Loss And Enhanced Stability

By reworking the materials in their cables, a major manufacturer has created products with less weight, less loss, and greater stability than their previous versions.

JACK BROWNE
Publisher/Editor

COAXIAL cables represent one of the "staple" components of RF/microwave system design. While they are essential, coaxial cables rarely benefit from significant technology improvements due to the simple nature of their construction—conductors and insulators. Yet, by carefully considering and implementing advancements in materials, Insulated Wire, Inc., (Ronkonkoma, NY) was able to make significant improvements in amplitude and phase stability as well as insertion-loss performance throughout their coaxial cable lines.

The company's 1801 series coaxial cables are a good example of the improvements resulting from the product transformation. These cables, which formerly measured 0.230 in. (0.5842 cm) in diameter, yielded insertion loss of 0.37 dB/ft. The new version of the 1801 series cables measures only 0.180 in. (0.4572 cm) in diameter, with associated insertion loss of 0.33 dB at 18 GHz. In terms of long cable runs, for example 100 ft, this can translate into savings of 4-dB insertion loss at the system level. For typical 4-ft. 1801 series cable assemblies with SMA connectors, the VSWR is 1.20:1 to 18 GHz.



The insertion-loss performance of the 2301 series coaxial cable was measured from 40 MHz to 26.5 GHz for a 3-ft. assembly with SMA connectors using a model 360 vector network analyzer (VNA) from Anritsu Co.

In terms of applications that demand small size and lightweight, such as avionics systems, the re-designed 1801 cables can provide tremendous savings in mass and volume. In addition, the lower insertion loss may enable the use of lower-power amplification, with corresponding savings in active components.

As a result of re-defining and re-designing their cable products, the company's 2301 series cables now exhibit the lowest insertion-loss performance on the market, at 0.3 dB/ft. to 26.5 GHz. An evaluation of a 3-ft.-long 2301 series cable assembly with SMA connectors reveals the impressive insertion-loss performance, especially at frequencies below 10 GHz (see figure). Measurements were made with a model 360 vector network analyzer (VNA) from Anritsu Co. (Morgan Hill, CA).

Despite the fact that the new cables are smaller and lighter than their predecessors, they are nonetheless double-shielded designs with low RF leakage and impressive phase stability. The cables typically vary only by 400 PPM with temperatures from –20 to +120°C, compared with earlier cables that had temperature variations of 700 PPM for the same temperature range. The process control also results in repeatable low-VSWR performance through the company's cables.

The improved coaxial cables are available as bulk cables or as cable assemblies with a customer's choice of coaxial connectors. The cables meet all of the material requirements outlined in MIL-C-17 specifications. Inner conductors are silver (Ag)-plated copper. The dielectric is a low-loss polytetrafluoroethylene (PTFE) material. The outer conductor is pure Ag, applied in a novel manner that yields extremely low contact resistance. A variety of materials is available for the outside protective jackets. **Insulated Wire, Inc., 2065 Smithtown Ave., Ronkonkoma, NY 11779; (516) 981-7424, FAX: (516) 981-7990, Internet: http://www.insulatedwire.com.**

CIRCLE NO. 57 or visit www.mwrf.com

PLL Synthesizers Suit Low-Power Wireless Systems

This new generation of integer-N frequency synthesizers provides the low-current performance needed to shrink wireless handsets.

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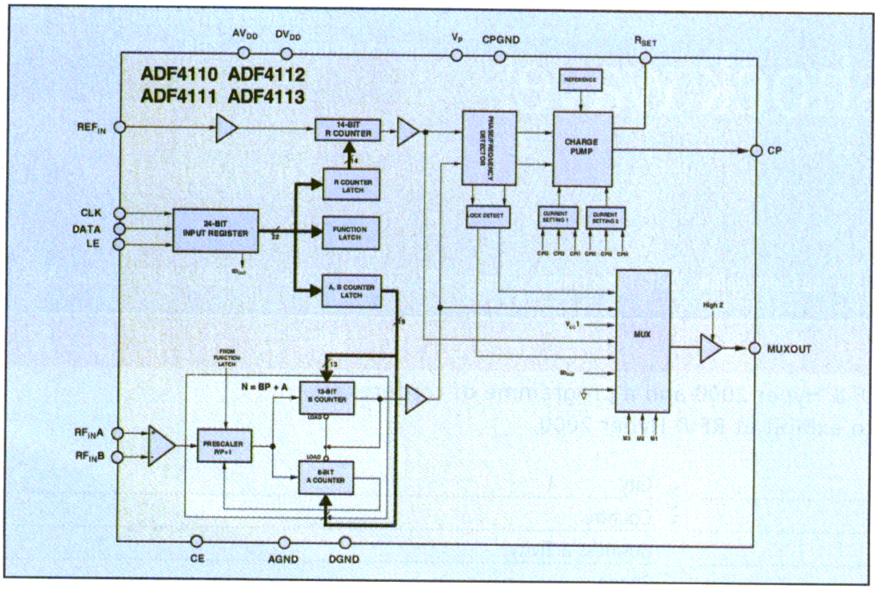
PHASE-LOCKED-LOOP (PLL) frequency synthesizers maintain frequency stability in wireless handsets and, as a matter of course, consume battery power in the process. But wireless handsets are shrinking and so are their batteries. For handset designers, this shrinkage means a reduced power budget, which forces them to seek components that draw as little current as possible. To answer this need, Analog Devices, Inc. (Wilmington, MA) has developed a new generation of integer-N PLL synthesizer that conserves power without compromising performance.

The synthesizers are segregated into two families—the ADF4110 family and the ADF4210 family. The main difference between the two is that the ADF4110 family has a complementary RF input, while the ADF4210 family has an RF input and an intermediate-frequency (IF) input.



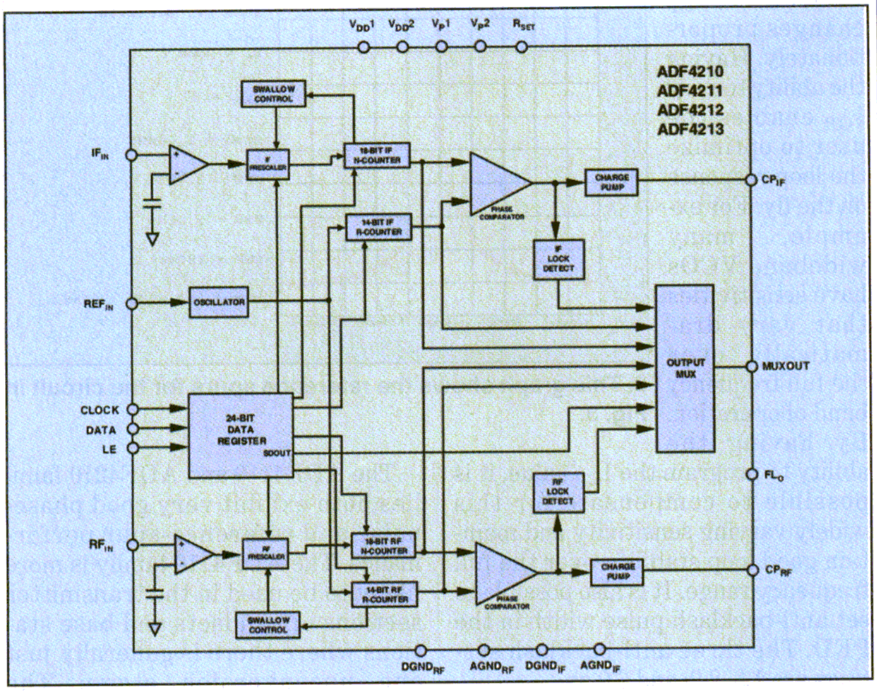
1. This block diagram shows the functional components of the ADF4110 family of PLL synthesizers.

This article examines various wireless architectures and shows how this new type of synthesizer fits those architectures. The discussion presents phase-noise and reference-spur performance, and shows how features such as a programmable charge-pump current can help keep the loop bandwidth constant under varying loop conditions.

The ADF4110 family of PLL synthesizers (Fig. 1) has four members—the ADF4110, ADF4111, ADF4112, and ADF4113. Functionally, they are identical. The major difference among them is maximum operating frequency. The ADF4110 operates to 550 MHz; the ADF4111 to 1.1 GHz; the ADF4112 to 2.8 GHz; and the ADF4113 reaches 3.7 GHz.

The synthesizer's frequency-reference input (REF_{IN}) feeds a 14-b counter, which can derive all of the commonly used channel-spacing frequencies from off-the-shelf temperature-controlled crystal oscillators (TCXOs). In the case of Global System for Mobile Communications (GSM) network, for example, a 13-MHz TCXO is often used as the system reference. Setting the reference divider to 65 provides the required 200-kHz channel spacing.

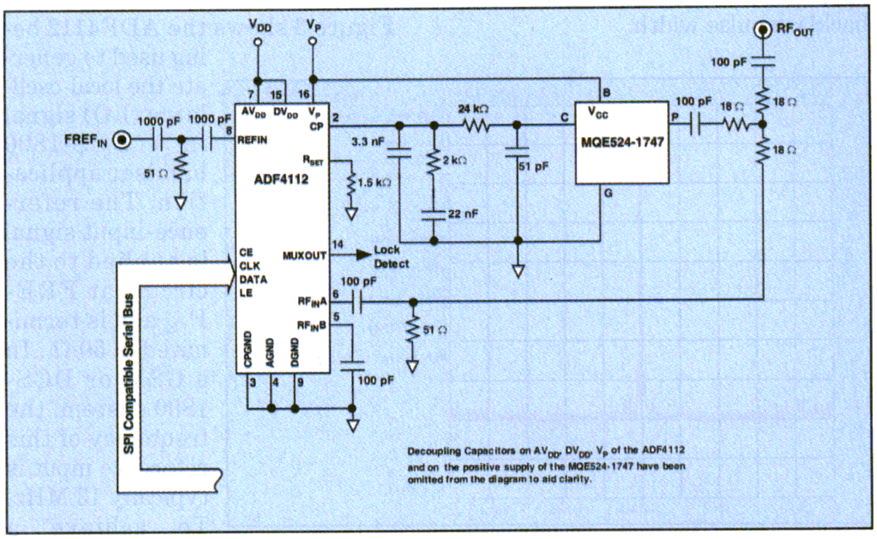
The complementary RF input ($RF_{IN}A$ and $RF_{IN}B$) supports a balanced signal. More often, an unbalanced RF signal from a voltage-controlled oscillator (VCO) is fed to $RF_{IN}A$, and $RF_{IN}B$ is grounded



2. This block diagram shows the functional components of the ADF4210 family of PLL synthesizers.

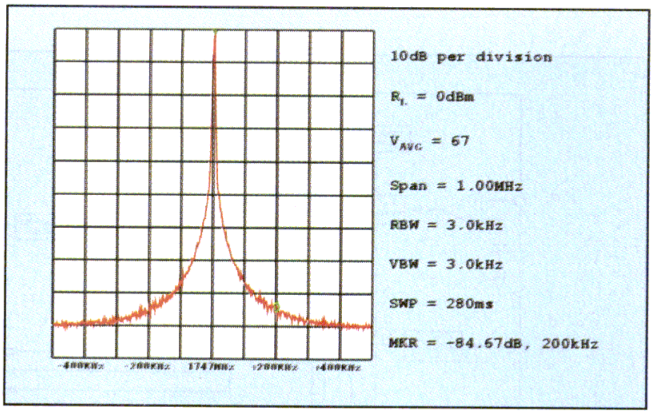
through a small bypass capacitor. The RF input leads to the N divider, which sets the output frequency. The divider consists of a dual-modulus prescaler (programmable for divide ratios of 8/9 or 32/33) followed by a 6-b A counter and 13-b B counter. The combination of programmable prescaler, 6-b A counter, and 13-b B counter yields three overlapping divide-ratio ranges. These ranges are 56 to 65591 (prescaler of 8/9); 240 to 131119 (prescaler of 16/17); and 992 to 262175 (prescaler of 32/33).

The phase/frequency detector (PFD)/charge-pump output current (I_{CP}) is programmable in binary steps from 5 mA to 625 μA. Additionally, the absolute value of I_{CP} can be varied by changing the value of the R_{SET} resistor. For full-scale current of 5 mA, R_{SET} must be 4.7 kΩ. As this



3. This schematic shows how the ADF4112 can be used as part of the local oscillator (LO) for a DCS-1800 handset transmitter.

resistor value is varied, I_{CP} changes proportionately. Having the ability to vary I_{CP} enables the user to optimize the loop response on the fly. For example, many wideband VCOs have sensitivities that vary dramatically over the full frequency band of operation. By having the ability to program the I_{CP} value, it is possible to compensate for this widely varying sensitivity and maintain good loop stability over the full frequency range. It is also possible to set anti-backlash pulse width in the PFD. The three anti-backlash settings are 1.3, 2.9, and 6.0 ns.

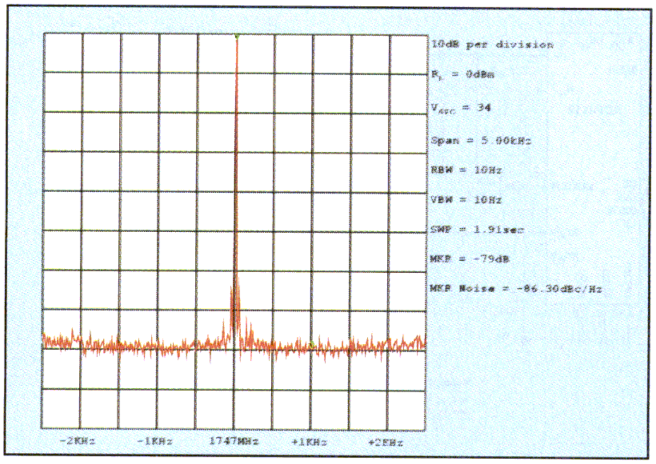


5. This graph shows the reference spurs for the circuit in Fig. 3.

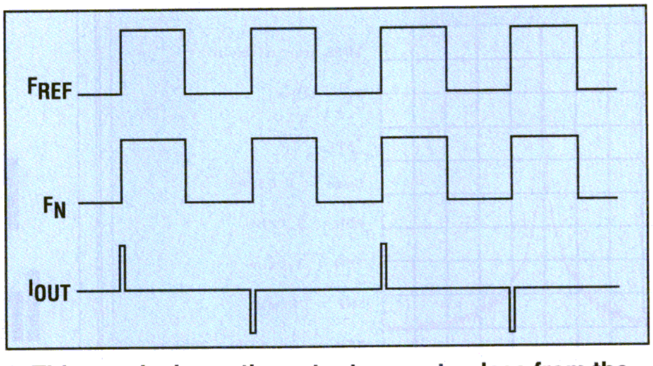
The ADF4210 family of PLL synthesizers (Fig. 2) has three members—the ADF4210, ADF4211, and ADF4212. Each of these has an IF synthesizer capable of operating to 510 MHz. As with their cousins, the members of this family differ mainly in their maximum RF operating frequency. The ADF4210 operates to 1.2 GHz; the ADF4211, 2.0 GHz; and the ADF4212, 3.0 GHz. And similar to the ADF4110 family, the ADF4210 family offers programmable prescaler, programmable I_{CP}, as well as programmable anti-backlash pulse width.

The ADF4110 and ADF4210 families both exhibit very good phase-noise and reference-spur performance. The ADF4110 family is more likely to be used in the transmitter sections of handsets and base stations where there is generally just one upconversion stage. The ADF4210 family is best suited to receiver designs where two downconversion stages are needed. In all of these applications, low power consumption is critical. The ADF4212's typical I_{DD} of 7 mA and maximum frequency of 3 GHz makes it ideal for wideband code-division-multiple-access (WCDMA) receivers. The ADF4211's typical I_{DD} of 5 mA and maximum frequency of 2 GHz is ideal for DCS-1800 receivers. The ADF4112 has a typical power I_{DD} of 3 mA and is ideal for DCS-1800 transmitters.

Figure 3 shows the ADF4112 being used to generate the local-oscillator (LO) signal in a DCS-1800 handset application. The reference-input signal is applied to the circuit at $FREF_{IN}$ and is terminated in 50 Ω. In a GSM or DCS-1800 system, the frequency of this reference input is typically 13 MHz. To achieve a channel spacing of 200 kHz (the



4. This graph shows the phase-noise performance of the circuit in Fig. 3.



6. This graph shows the output current pulses from the PFD charge pump.

DCS-1800 standard), the reference input must be divided by 65 using the ADF4112's on-chip reference divider.

The ADF4112 is capable of reaching 2.8 GHz. In this integer-N synthesizer, N can be programmed from 56 to 262,000 in discrete integer steps. In the case of the handset transmitter, which requires an output range of 1710 to 1785 MHz, and where the internal reference frequency is 200 kHz, the desired N values range from 8550 to 8925.

The charge-pump output of the ADF4112 (pin 2) drives the loop filter. When calculating the loop-filter component values, a designer must consider a number of items. In this example, the loop filter is designed so that the overall phase margin for the system is 45 deg. Other PLL system specifications include the following items: Kd (the charge-pump current of the ADF4112) = 5 mA; Kv (the sensitivity of the MQE524-1747) = 50 MHz/V; a loop bandwidth of 12 kHz; a reference frequency (FREF) of 200 kHz; N = 8700 (midrange value), and an extra reference-spur attenuation of 10 dB. All of these specifications were used to calculate the loop-filter component values shown in Fig. 3.

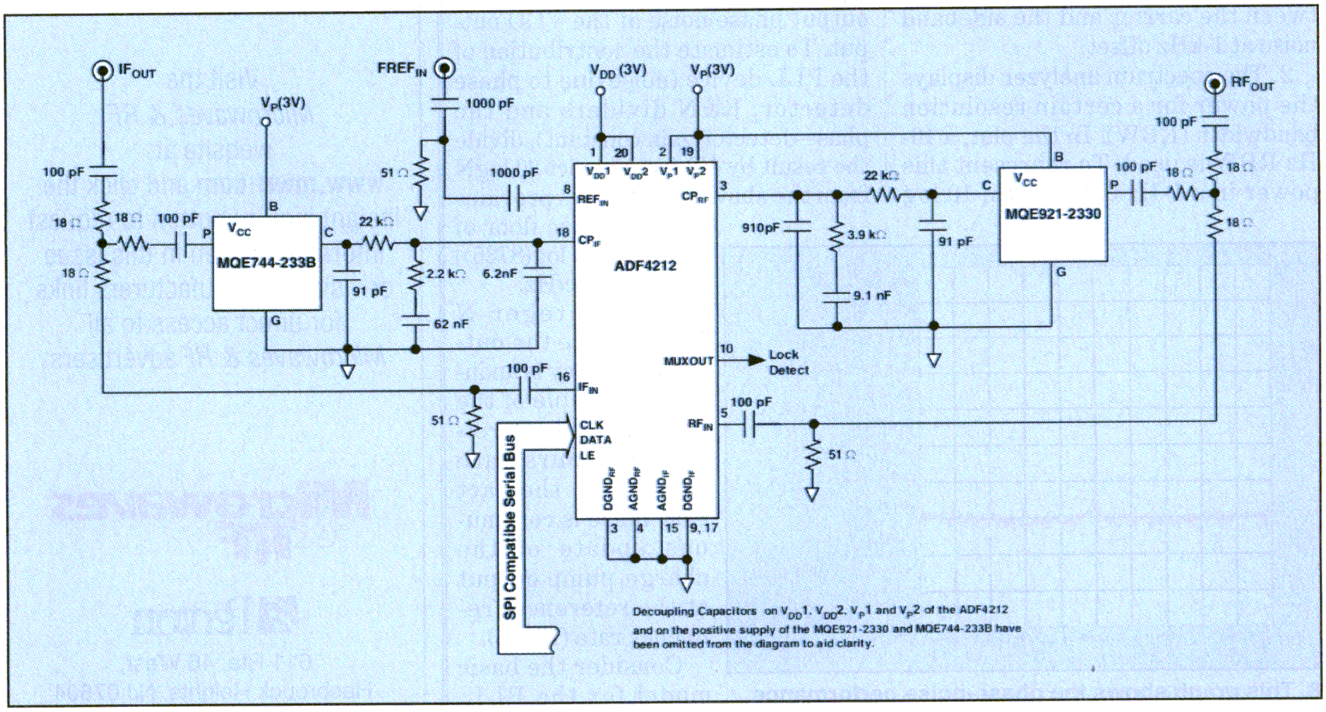
The loop-filter output drives the VCO. The VCO output is fed back to the RF input of the PLL synthesizer and also drives the RF output terminal. A T-circuit configuration using 18-Ω resistors provides 50-Ω matching impedance at the VCO output, RF output, and RF_{IN} terminal of the ADF4112.

In a PLL system, it is important to know when the system is locked. In Fig. 3, this is accomplished by using the ADF4112's MUXOUT signal. The MUXOUT pin can be programmed to monitor various internal signals in the synthesizer. For example, the MUXOUT can be programmed to monitor the lock-detect (LD) signal, and can be used to initiate the power ramp sequence.

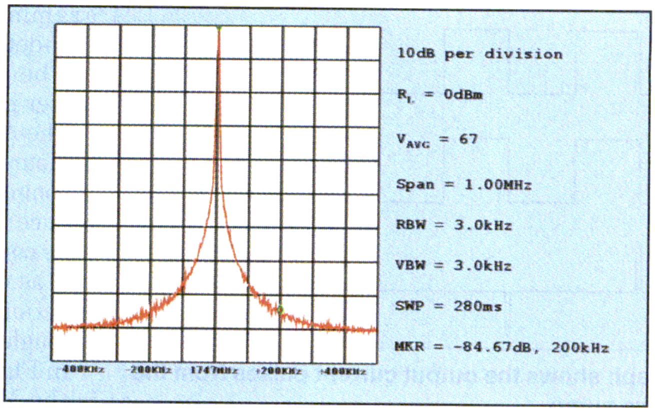
The ADF4112 uses a simple four-wire serial interface to communicate with the system controller. The reference counter, N counter, as well as various other on-chip functions are programmed through this interface.

In Fig. 3, the output VCO is the MQE524-1747 from Murata Electronics (Smyrna, GA). This device was specifically chosen for the transmitter section due to its high output-power level. It runs from a supply of +3.8 VDC and delivers a typical output level of +10 dBm. For newer designs that might require a lower operating voltage, the MQE9PE-1747 is recommended. It runs from a supply voltage of +2.75 VDC and delivers a typical output-power level of +7.5 dBm. Note that the sensitivity of the MQE9PE-1747 is 70 MHz/V, compared to 50 MHz/V for the MQE524-1747. The designer must take this into account when calculating the loop-filter components.

Figure 4 illustrates the phase-noise plot for Fig. 3. The frequency



7. This schematic demonstrates how the ADF4212 can be used as part of the LO for a WCDMA receiver.



9. This graph shows the reference-spur performance for the circuit in Fig. 7.

and phase noise were measured over a 5-kHz span. As already stated, the reference frequency used was FREF = 200 kHz and the output frequency was 1747 MHz (N=8735). If this were an ideal PLL synthesizer, a single discrete tone would be displayed along with the spectrum analyzer's noise floor. But the real-world circuit exhibits the tone and the phase noise generated by the PLL components. The loop-filter values were chosen to provide a loop bandwidth of approximately 12 kHz. The flat part of the phase noise for frequency offsets less than the loop bandwidth is actually the phase noise generated by the synthesizer. It is specified at a 1-kHz offset. The value measured was –86.30 dBc/Hz. This is the phase-noise power in a 1-Hz bandwidth. The authors arrived at this figure by considering the following factors:

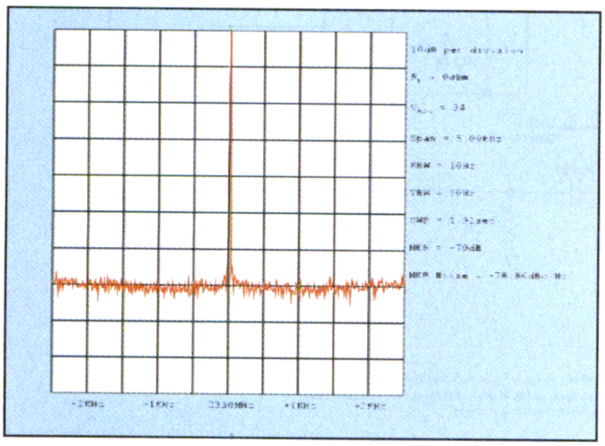
1. The spectrum-analyzer marker shows the relative power in dBc between the carrier and the sideband noise at 1-kHz offset.

2. The spectrum analyzer displays the power for a certain resolution bandwidth (RBW). In the plot, a 10-Hz RBW is used. To represent this power in a 1-Hz bandwidth, 10 log (RBW) must be subtracted from the value in item 1.

3. It is necessary to add a correction factor that takes into account the implementation of the RBW, the log-display mode, and detector characteristic of the spectrum analyzer.

Phase-noise measurement with spectrum analyzers such as the HP 8561E from Hewlett-Packard Co. (Palo Alto, CA) can be made quickly by using the marker-noise function, MKR NOISE. This function takes into account the previously mentioned three factors and displays the phase noise in dBc/Hz.

The phase-noise measurement mentioned previously is the total output phase noise at the VCO output. To estimate the contribution of the PLL device (noise due to phase detector, R&N dividers and the phase-detector gain constant), divide the result by N2 (or subtract 20 logN from the above result). This provides a phase-noise floor of [–86.30 – 20 log(8735)] = –165.1 dBc/Hz.



8. This graph shows the phase-noise performance for the circuit in Fig. 7.

In an integer-N PLL (where the output frequency is an integer multiple of the reference input), reference spurs are caused by the fact that there is continuous update of the charge-pump output at the reference-frequency rate (Fig. 5).

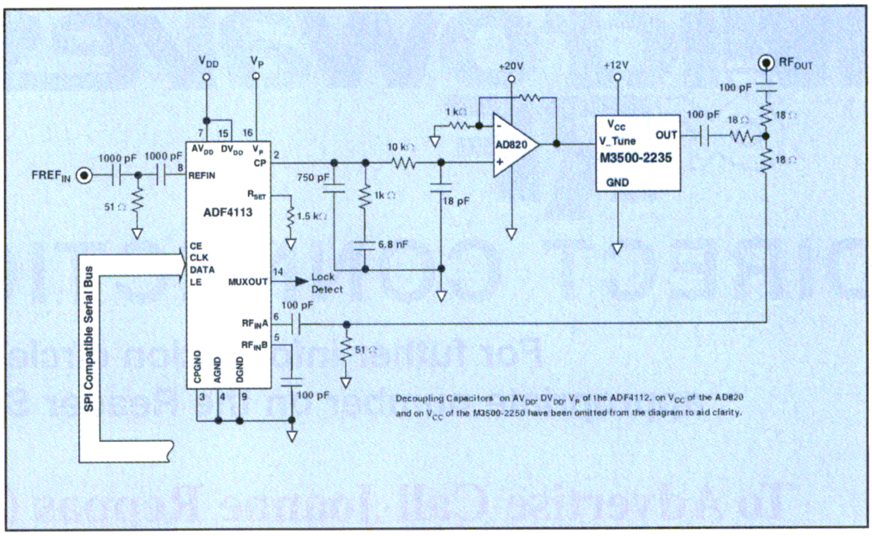
Consider the basic model for the PLL. When the PLL is in lock, the phase and frequency inputs to the PFD (F_{REF} and F_N) are essentially equal. In theory, one would expect that there would be no output from the PFD in this case. In effect, the loop would be broken and the output could drift. To prevent this, the PFD is designed so that in the locked condition, the current pulses from the charge pump are equal in magnitude and duration (Fig. 6). Any difference in up and down duration will provide charge to the loop filter at the reference rate, thus causing ripple on the VCO control input, which, in turn, causes ouput spurs.

It is possible to detect reference spurs using a spectrum analyzer simply by increasing the span to greater than twice the reference frequency (Fig. 5). In this case, the reference frequency is 200 kHz, and the diagram shows that reference spurs at ±200 kHz from the center frequency output are –83 dBc.

Figure 7 shows the circuit for the LO section of a WCDMA receiver. In this design, the system power supply is +3 VDC. The total power typically consumed by the VCOs and the synthesizer is estimated to be 70 mW (45 mW by the VCOs and 25 mW by the synthesizer). There are two downconversion stages. The first stage takes the signal in the 2070-to-2130-MHz band and converts it down to 230 MHz. This is then further translated to the baseband by the MQE744-233B VCO.

Figure 8 shows that the phase noise is –78.86 dBc/Hz at an offset of 1 kHz. Working back to calculate the synthesizer noise based on this value, the result is –161 dBc/Hz. It is interesting to note that this is several decibels away from the –165-dBc/Hz value resulting from the DCS-1800 design (Fig. 3). This is because the circuit in Fig. 7 is a full +3-VDC design. The output levels from the VCOs are typically 0 dBm, compared to +10 dBm from the VCO in Fig. 3. The first reference spurs (200 kHz) are below –82 dBc, and the second reference spurs (400 kHz) are below –90 dBc (Fig. 9).

Many of the wireless applications for synthesizers and VCOs in PLLs are narrowband in nature. These include the various wireless standards



10. This schematic demonstrates how the ADF4113 can be used in a wideband LO.

such as GSM, DCS-1800, CDMA, and WCDMA. In each of these cases, the total tuning range for the LO is approximately 100 MHz. However, there are also wideband applications where the LO could have up to an octave of tuning range. For example, cable-TV (CATV) tuners have a total range of approximately 400 MHz. Figure 10 shows a wideband-LO application that uses the ADF4113 (3.8-GHz bandwidth) to control and program the Micronetics M3500-2235.

THE SYNTHESIZERS ARE SEGREGATED INTO TWO FAMILIES—THE ADF4110 AND ADF4210. THE MAIN DIFFERENCE IS THAT THE ADF4110 FAMILY HAS A COMPLEMENTARY RF INPUT, WHILE THE ADF4210 FAMILY HAS AN RF INPUT AND AN IF INPUT.

The loop filter was designed to have an RF output frequency of 2900 MHz, a loop bandwidth of 100 kHz, a PFD frequency of 1 MHz, an ICP of 5 mA, and a VCO KD of 90 MHz/V (which is the sensitivity of the M3500-2235 at 2900 MHz).

In narrowband applications, there is generally a small variation in output frequency (generally less than 10 percent) and also a small variation in VCO sensitivity over the range (typically 10 to 15 percent). However, in wideband applications, both of these parameters have a much-greater variation. Figure 10, for example, shows –25- and +17-percent variation in the RF output from the nominal 2.9 GHz. The sensitivity of the VCO can vary from 120 MHz/V at 2750 MHz to 75 MHz/V at 3400 MHz (+33 percent, -17 percent). Variations in these parameters will vary the loop bandwidth. This, in turn, can affect stability and lock time. By changing the programmable I_{CP}, it is possible to compensate for these varying loop conditions and ensure that the loop is always operating close to optimal conditions. The circuit shown in Fig. 10 was verified over the frequency range of 2200 to 3400 MHz, and maintained lock under these conditions.

This new generation of integer-N PLL synthesizer provides low-noise performance for wireless applications without the large power consumption associated with frequency synthesizers. **Analog Devices, Inc., One Technology Way, Norwood, MA 02062-9106; (800) 262-5643, Internet: http://www. analog.com.**

Acknowledgments

The authors would like to acknowledge the following people: Bill Hunt (Analog Devices, Inc.) and Brendan Daly (Analog Devices, Inc.) and Lorraine Kearn (Murata Manufacturing Co.) for technical support on the Murata VCOs.

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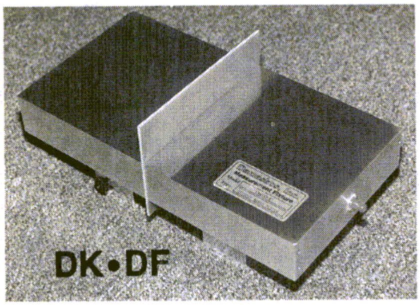

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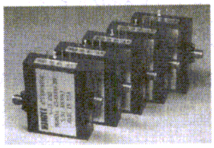
6-12 GHZ Voltage Controlled Attenuator



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- Transfer Function (Monotonically Increasing): 10 dB /Volt , 0-6 Volts
- Transfer Function Accuracy:
  - 0 - 0.9 dB @ ±0.1 dB Max
  - 1 - 30 dB @ ±0.5 dB Max
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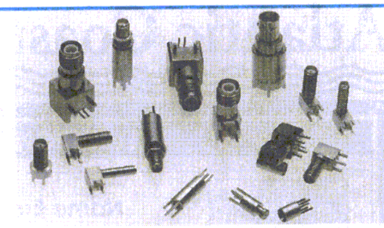

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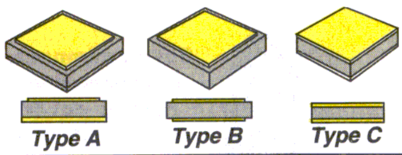

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High-power couplers cover CDMA band

Two high-power, surface-mount, 90-deg. hybrid couplers operate across the 2.0-to-2.2-GHz code-division-multiple-access (CDMA) cellular band. The couplers present an insertion loss of less than 0.2 dB and feature a phase balance of ±2 deg. The model S03A2150N1 coupler handles 100 W with a VSWR of less than 1.15:1. It offers an amplitude balance of ±0.1 dB and an isolation of greater than 22 dB. The model S03B2150N2 coupler handles 200 W with a VSWR of less than 1.25:1. It provides an amplitude balance of ±0.2 dB and an isolation of greater than 20 dB. Applications for these wideband, hybrid couplers include CDMA amplification and distribution. **RF Power Components, Inc., 125 Wilbur Pl., Bohemia, NY 11716-2482; (516) 563-5050, FAX: (516) 563-4747, Internet: http://www.rf-power.com.**

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Diplexer screens GPS band

The model 800020 diplexer covers the Global Positioning System (GPS) satellite band with 0.7-dB loss at the L1 and L2 passbands. Its channel-to-channel isolation is 30 dB. The diplexer is housed in a 1.0 × 0.5 × 0.4-in. (2.54 × 1.27 × 1.02-cm) surface-mount package (SMP) that will survive an Sn62 reflow environment. Location of the miniature input and output pads can be specified at the time of purchase to accommodate many board configurations. **Bree Engineering, 1296 Linda Vista Dr., San Marcos, CA 92069; (760) 510-4950, FAX: (760) 510-4959, Internet: http://www.breeeng.com.**

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Panel antennas cover 800/900 MHz

A line of directional panel antennas is designed for 800-to-900-MHz applications. The line consists of four models. Model MP8066PT covers frequencies from 806 to 866 MHz. Model MP8246PT spans frequencies from 824 to 896 MHz. Model MP8906PT covers frequencies from 890 to 945 MHz. And model MP8068 covers frequencies from 806 to 960 MHz. The first three models mentioned provide 6-dB gain, 70-deg. horizontal beamwidth, and 60-deg. vertical beamwidth. Model MP8068 provides 8-dB gain, 35-deg. horizontal beamwidth, and 65-deg. vertical beamwidth. All models include a 12-in. (30.48-cm) RG-58/U pigtail and a choice of connectors. They are designed to withstand severe weather conditions, and their 2.25-in.-(5.715-cm-) deep housing makes them appropriate for indoor or outdoor use. **Maxrad, Inc., 4350 Chandler Dr., Hanover Park, IL 60103-6763; (630) 372-6800, FAX: (630) 472-8077, Internet: http://www.maxrad.com.**

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RF ICs serve mobile communications

A new family of radio-frequency (RF) integrated circuits (ICs) includes single and dual phase-locked-loop (PLL) frequency synthesizers, a universal RF mixer, a direct modulator, and a series of dual-band low-noise amplifiers (LNAs). The model PMB2341 single PLL frequency synthesizer operates at frequencies to 2.5 GHz and delivers phase-noise performance of –87 dBc/Hz at 500-Hz offset from a 900-MHz carrier. The model PMB2347 dual PLL synthesizer generates intermediate frequencies (IFs) to 500 MHz and RF signals to 2.8 GHz with similar performance. The model PMB2335 universal RF mixer operates at frequencies to 3 GHz. The model PMB2212 direct modulator contains a mixer and automatic gain control (AGC) with a dynamic range of 40 dB. The model PMB2362 dual-band LNA operates at 900/1800 MHz, and the model PMB2363 dual-band LNA operates at 800/1900 MHz. The ICs are fully compliant with code-division-multiple-access (CDMA), time-division-multiple-access (TDMA), personal-communications-services (PCS), and Global System for Mobile Communications (GSM) standards. **Infineon Technologies Corp., 1730 North First St., San Jose, CA 95112; (408) 501-6000, Internet: http://www.infineon.com.**

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Satellite tuner IC replaces canned tuners

The model MAX2108 integrated-circuit (IC) tuner for digital satellite communications directly downconverts L-band signals to baseband in-phase/quadrature (I/Q) channels. It eliminates the intermediate-frequency (IF) local oscillator (LO), IF mixer, and surface-acoustic-wave (SAW) filter found in traditional superheterodyne tuners. The chip also includes a low-noise amplifier (LNA) with automatic gain control (AGC), I/Q downconversion mixers, A 90-deg. phase shifter, and baseband buffers. This IC tuner's input third-order intercept point (IP3) of +8 dBm at minimum gain enables its RF input to be directly connected through a matching network to the F connector of a 75-Ω cable without the need for a PIN-diode attenuator and amplifier. The chip operates at temperatures from 0 to +70°C and is available in a 24-pin quarter-sized outline package (QSOP). **Maxim Integrated Products, 120 San Gabriel Dr., Sunnyvale, CA 94086; (408) 737-7600, Internet: http://www.maxim-ic.com.**

CIRCLE NO. 76 or visit www.mwrf.com

Tiny VCOs target WLL systems

Two compact voltage-controlled oscillators (VCOs) are available for wireless-local-loop (WLL) systems. Model MW520 covers frequencies from 1865 to 2055 MHz, and model MW521 covers frequencies from 2248 to 2448 MHz. They provide +9 dBm of output power and a phase-noise performance of –90 dBc/Hz at 10-kHz offset. These VCOs operate from a +5-VDC power supply over a temperature range of –40 to +85°C. They are housed in a 10 × 7.4-mm package compatible with pick-and-place and solder-reflow manufacturing methods. **Micronetics Wireless, 26 Hampshire Dr., Hudson, NH 03051; (603) 883-2900, FAX: (603) 882-8987, Internet: http://www.micronetics.com.**

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Front-end ICs simplify cable and satellite systems

Two integrated circuits (ICs) simplify cable and satellite systems by integrating their front-end functions. The model STV0297 is a single-chip quadrature-amplitude-modulation (QAM) demodulator for cable systems that receive compressed television, audio, or data services. It includes an analog-to-digital converter (ADC) that can handle up to 256 QAM signals in a direct intermediate-frequency (IF)-sampling architecture, eliminating the need for an external downconverter. Model STV0299 is a multistandard quadrature-phase-shift-keying (QPSK) and binary-phase-shift-keying (BPSK) demodulator for use in digital satellite receivers and set-top boxes. **ST Microelectronics, Inc., Lexington Corporate Center, 10 Maguire Rd., Bldg. 1, Third Floor, Lexington, MA 02421; (781) 861-2650, FAX: (781) 861-8678, Internet: http://www.st.com.**

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NEW LITERATURE

Crystal oscillators

A 20-page catalog features a line of precision quartz-crystal oscillators. A series of temperature-controlled crystal oscillators (TCXOs), voltage-controlled crystal oscillators (VCXOs), and oven-controlled crystal oscillators (OCXOs) is featured. The catalog includes performance specifications and application information. **MTI-Milliren Technologies, Inc.;** *(978) 465-6064, FAX: (978) 465-6637, Internet: http:// www.mti-milliren.com.*

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Cable assemblies

A 32-page catalog presents high-frequency low-loss cable assemblies up to 40 GHz, cable assemblies for interconnects in land mobile, cellular and paging systems, as well as custom laboratory test kits to meet specific customer applications. Specifications along with cable and connector selection guides for RF and microwave coaxial-cable assemblies are included. **Florida RF Labs;** *(800) 544-5594, (561) 286-9300, FAX: (561) 283-5286.*

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DC-to-DC converters

Surface-mount/plug-in transformers and inductors, DC-to-DC converters, and AC-to-DC power supplies are covered in a 160-page catalog. Specifications include input voltage range, output voltage, output voltage ripple, and maximum output power. Pricing and ordering information are included. **PICO Electronics, Inc.;** *(800) 228-9747, (914) 738-1400, FAX: (914) 738-8225, e-mail: HLSC73A@prodigy.com, Internet: http://www.picoelectronics.com.*

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Switch modules

Matrices, multiplexers, power relays, coaxial matrices, and custom modules are the focus of a 12-page brochure. General specifications include power bandpass, insertion loss, and isolation/crosstalk. Features, descriptions, and warranty information are provided. **Cytec Corp.;** *(800) 346-3117, e-mail: sales@cytec-ate.com, Internet: http: //www.cytec-ate.com.*

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Frequency counters

A 96-page catalog describes applications, features, and specifications for oscilloscopes, frequency counters, sweep/function generators, audio-test products, power meters, digital LCR meters, and high-definition television (HDTV). Digital-video-disk (DVD), mini-disk (MD), compact-disc (CD), and CD-read-only-memory (CD-ROM) instruments are also covered, along with digital and analog power supplies. Quality-assured probes, accessories, and carrying cases developed to meet a spectrum of needs are included. **Leader Instruments Corp.;** *(516) 231-6900, FAX: (516) 231-5295.*

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Cable simulators

Filters, coil/inductors, and video signal-processing equipment are listed in a catalog. Delay-line pulse and video, cable equalizers, and cable simulators are also offered. **Faraday Technology Ltd.;** *+44 (0)1782 661501, FAX: +44 (0)1782 630101, e-mail: sales@faradaytech.co.uk, Internet: http://www.faradaytech.co.uk.*

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Trimmer potentiometers

Trimmer potentiometers are covered in a 128-page catalog and design manual. Single-turn and multi-turn sealed cermet as well as single-turn open frame cermet types are covered. The through-hole trimmer category includes single-turn sealed cermet, single-turn carbon element, single-turn sealed wirewound, and multi-turn sealed cermet types. A list of part numbers is included. Selection guides, cross-reference tables, guidelines, and precautions are listed. Specifications and a glossary of terms are provided. **Tocos;** *(847) 884-6664, FAX: (847) 884-6665, e-mail: sales@tocos.com, Internet: http:// www.tocos.com.*

CIRCLE NO. 85 or visit www.mwrf.com

Test and measurement

A 432-page catalog provides solutions for mobile radio, electromagnetic compatibility (EMC), and general-purpose measurements as well as turnkey test and measurement systems. Test equipment for sound and TV broadcasting is also offered. **Rohde & Schwarz GmbH & Co. KG;** *+4989/4129-1765, FAX: +4989/4129-3208, Internet: http://www.rsd.de.*

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Power amplifiers

Low-noise broadband amplifiers, low-noise amplifiers (LNAs), broadband power amplifiers (PAs), high-power amplifiers, broadband general-purpose amplifiers, general-purpose amplifiers, microwave and millimeter-wave active multipliers, as well as millimeter-wave amplifiers are featured in a brochure. Electrical characteristics include frequency range, noise-figure (NF) range, bandwidth, and power output at 1-dB gain-compression point. **Quinstar Technology, Inc.;** *(310) 320-1111, FAX: (310) 320-9968, e-mail: sales@quinstar.com, Internet: http://www.quinstar. com.*

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Ground telemetry

A 48-page product-selection guide offers personal-computer (PC) products and systems, VME products and systems, a real-time telemetry processing system, and rack-mount products. Descriptions, features, and application information are included, along with an index **Aydin Telemetry;** *(215) 497-8000, FAX: (215) 968-3214, (215) 968-4175, e-mail: telemetry@aydin.com, Internet: http://www.aydin. com/ telemetry.*

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Oscilloscopes/analyzers

New and used test equipment are offered in a 52-page catalog. Power supplies, spectrum analyzers, network analyzers, RF signal sources, and pulse generators are offered. Logic analyzers, oscilloscopes, meters, and data-acquisition (DAQ) equipment are provided. **TestEquity;** *(800) 298-3457, (805) 498-9933, FAX: (805) 498-3733, e-mail: sales@testequity.com, Internet: http://www.testequity.com.*

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Ceramic filters

A series of data sheets overviews ceramic filters for C-band applications. Specifications include operating bandwidth, insertion loss, stopband rejection, and VSWR. Product descriptions and performance data are included. **K&L Microwave, Inc.;** *(410) 749-2424, FAX: (410) 749-5725, e-mail: klsales@klmicrowave.com, Internet: http://www.klmicrowave.com.*

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Antenna products

Double-ridged waveguide horns, a standard-gain and octave horn, tuned dipole, and conical-log spirals are covered in a 100-page antenna catalog. Electrical specifications include frequency range, maximum continuous power, peak power, and nominal impedance. Product descriptions, performance data, and a reference section are presented. An antenna uncertainty value table is also provided. **EMC Test Systems, L.P.;** *(800) 253-3761, (512) 835-4684, FAX: (512) 835-4729, e-mail: info@emctest.com, Internet: http://www.emctest.com.*

CIRCLE NO. 91 or visit www.mwrf.com

AFC subsystems

A 32-page catalog overviews automatic-frequency-control (AFC) subsystems, automatic-gain-control (AGC) amplifiers, modulators, and demodulators. Intermediate-frequency (IF), DC-coupled, and distribution amplifiers; group-delay equalized IF filters; and multifunction components are also featured. Specifications are included. **MITEQ, Inc.;** *(516) 436-7400, FAX: (516) 436-9219, (516) 436-7430, Internet: http://www.miteq.com.*

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Surface-mount devices

A catalog specifies surface-mount devices, including rectifiers, bridge rectifiers, ultra-high-frequency (UHF) diodes, zener diodes, Schottky diodes, switching diodes, transistors, and junction field-effect transistors (FETs). Performance data and outline drawings are provided. **Electronic Devices, Inc.;** *(914) 965-4400, FAX: (914) 965-5531, e-mail: edi-sales@cwix.com, Internet: http://www.edidiodes.com.*

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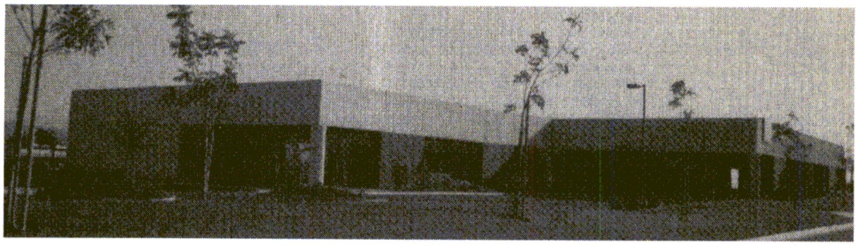

Slightly more than 15 years ago, MicroWave Technology, Inc. opened its doors in Fremont, CA to manufacture gallium-arsenide (GaAs)-based monolithic microwave integrated circuits (MMICs) and microwave integrated circuits (MICs). Funded by venture capital, the 31,000-sq.-ft. facility was founded by former Avantek (Santa Clara, CA) employees Masa Omori and Art Herbig, who brought considerable expertise in device design and semiconductor processing to the startup company.

Microwaves & RF November Editorial Preview

Issue Theme: Frequency-Control Technology

News

Frequency stability and control is assumed in most modern communications systems. But maintaining frequency for closely spaced, digitally modulated communications channels is not a trivial matter, and requires quartz crystals and other resonant materials with tight machined tolerances. What are the demands being placed on manufacturers of crystal oscillators by emerging wireless technologies? Don't miss this Special Report on crystal oscillators.

Design Features

Design Features in November address frequency-control topics that are of interest to designers and users of high-frequency oscillators. Articles include a study on techniques for extending the bandwidth of direct-digital-synthesizer (DDS) sources above the Nyquist frequency, as well as a comparison of the performance of strip crystals with conventional quartz crystals. Additional articles include methods for modeling phase noise in phase-locked loops (PLLs) and ways to measure residual noise in quartz crystals.

Product Technology

November's Product Technology section will unveil new beginnings—for a major high-frequency company and for their main computer-aided-engineering (CAE) suite of software tools. Additional articles will examine the use of sapphire resonators for low-phase-noise oscillators, introduce a line of millimeter-wave filters, scrutinize a new set of features for one of the industry's most powerful peak power meters, and unveil an innovative switch-filter channelizer.

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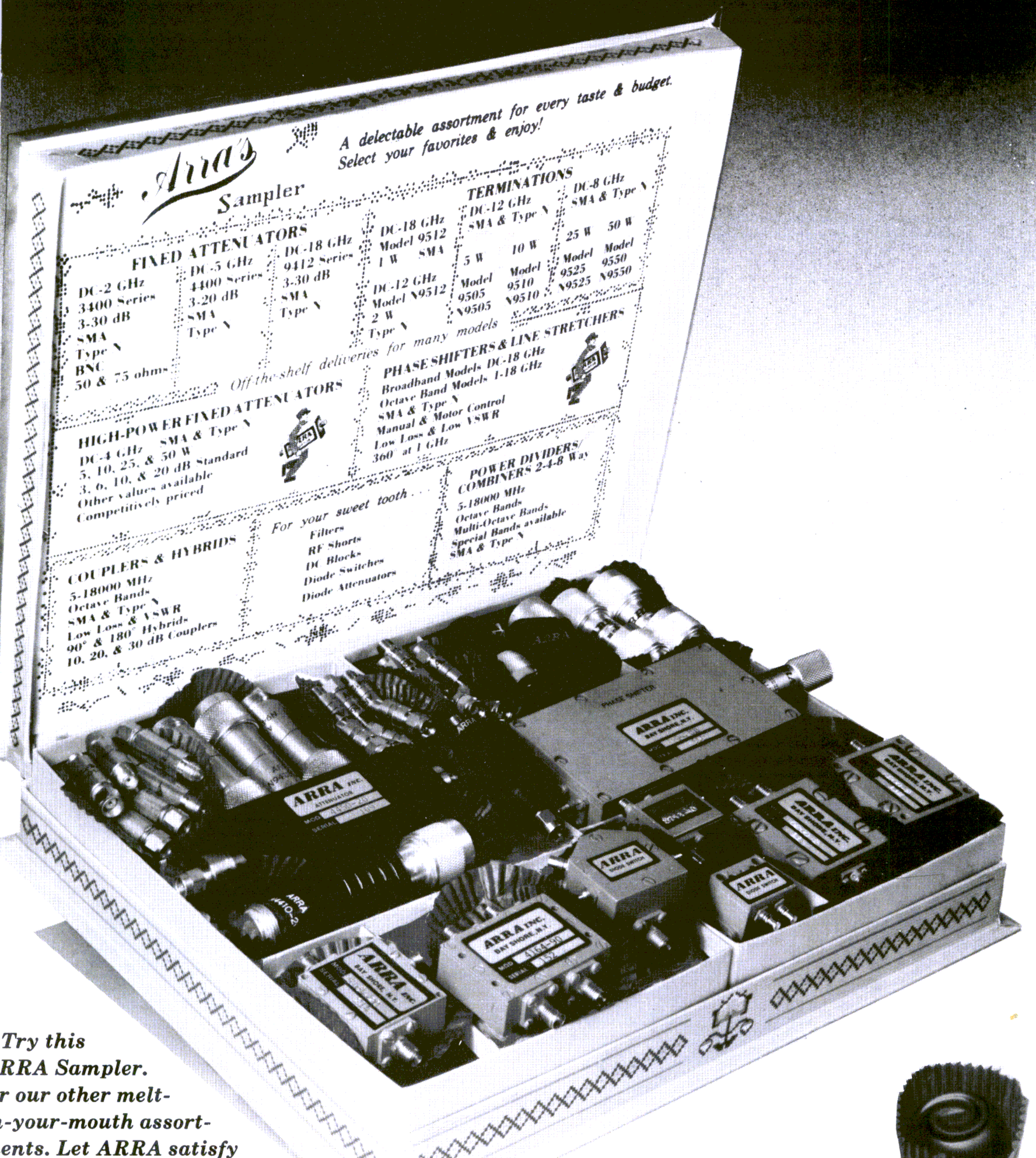

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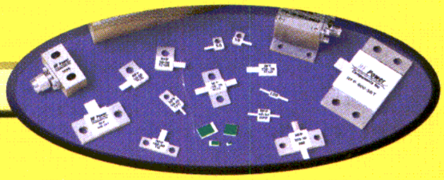
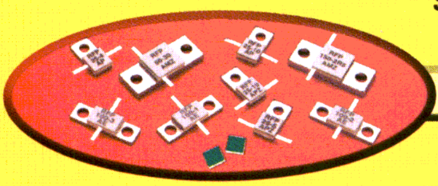
- Complete value added and custom design capabilities



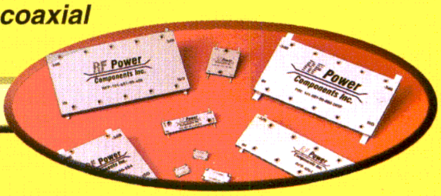
Resistors/Terminations

2-800 Watts, DC - 6 Ghz,
SMD, flanged, coaxial



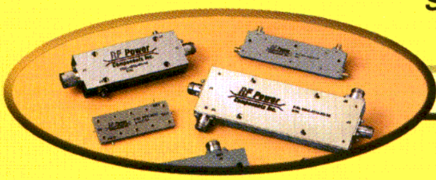
Attenuators

20-100 Watts, DC - 4 Ghz,
SMD, flanged, coaxial



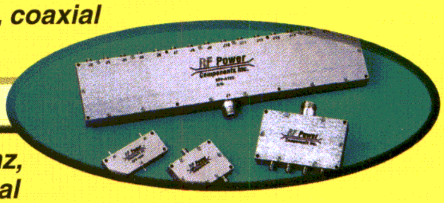
90° Hybrid Couplers

100-800 Watts, 50 - 4200 Mhz,
SMD, caseless, coaxial



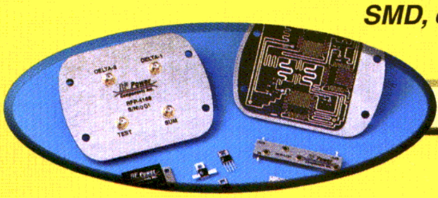
Directional Couplers

100-2000 Watts, 20 - 6000 Mhz,
SMD, caseless, coaxial



Combiners/Dividers

50-1500 Watts CW, 25 - 2000 Mhz,
SMD, caseless, resistive, coaxial



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