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FOR DESIGNERS AT HIGHER FREQUENCIES

Microwaves & RF

MTT-S Preview Radar & Antennas Issue

NEWS
Scanning new antenna developments

DESIGN FEATURE
Evaluating the range of Bluetooth antennas

PRODUCT TECHNOLOGY
Novel notch filter is programmable

Receivers upgrade noise-figure testing

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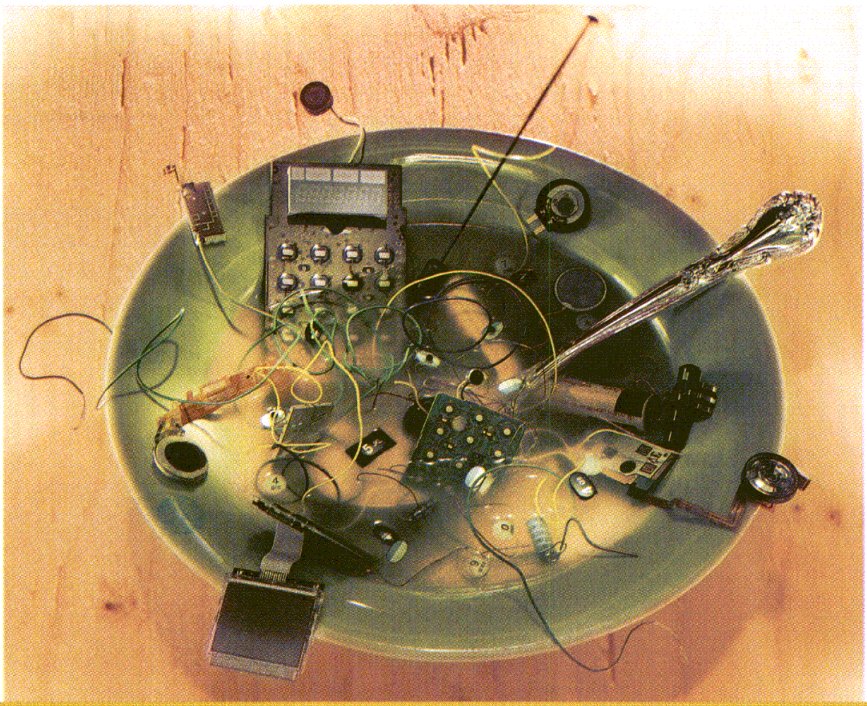

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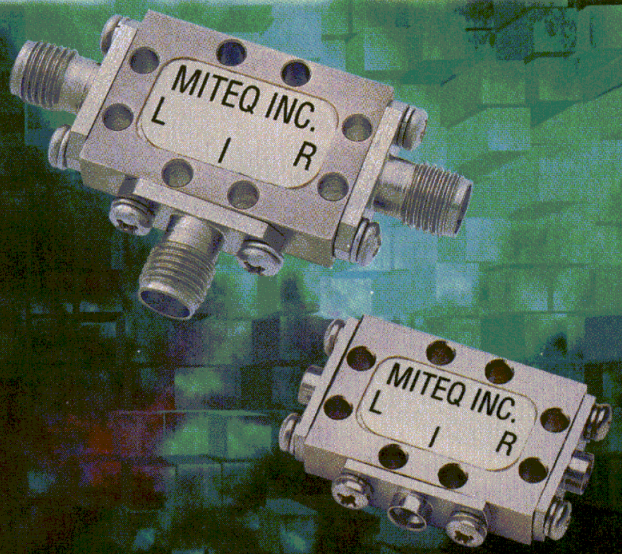
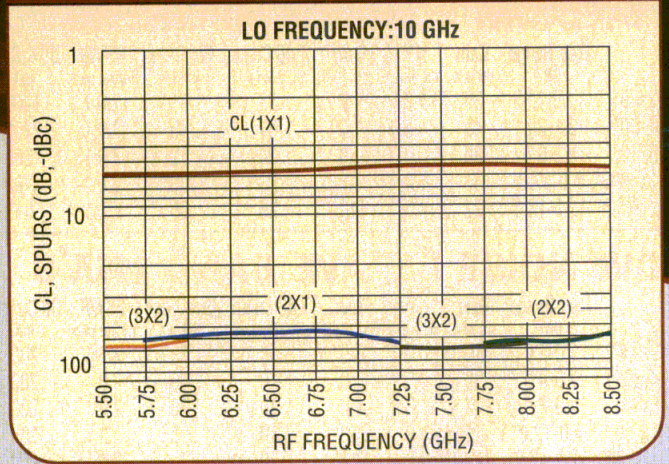
LOW SPURIOUS SPACEBORNE MIXERS

FEATURES:

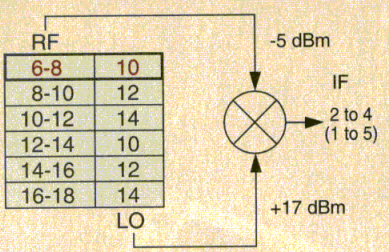
- Broadband operation
- Minimal variation in conversion loss
- High IP3 and 1 dB compression versus LO power



CONVERSION LOSS/SPURIOUS



TYPICAL OPERATING BANDS



RF	LO
6-8	10
8-10	12
10-12	14
12-14	10
14-16	12
16-18	14

SPECIFICATIONS - Model TBR0618HA1/TBR0618HA1-S

RF/LO Input Frequency Range	6 to 18 GHz
IF Output Frequency Range	0.05 to 5 GHz
Conversion Loss	6 dB Typical
Spurious	-55 dBc
Third Order Intercept Point	+23 dBm Typical
1 dB Compression Point	+13 dBm Typical

For further information, please contact Mary Becker
at (631) 439-9423 or e-mail mbecker@miteq.com

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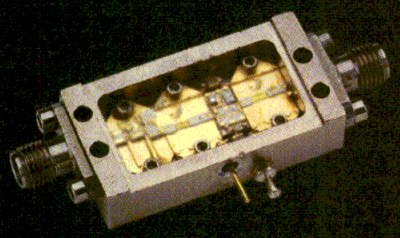
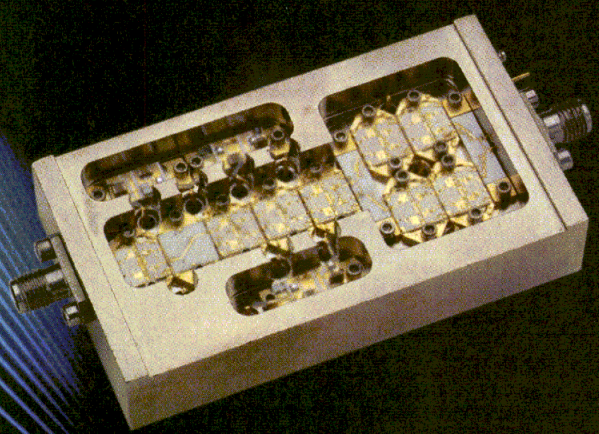
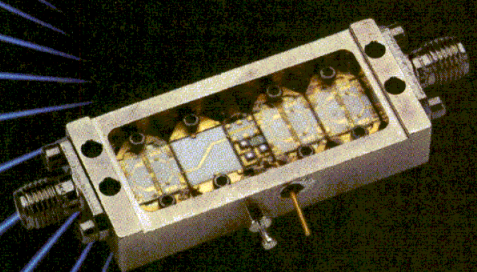
AMPLIFIERS
for every application







ULTRA BROAD BAND

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

MULTI OCTAVE AMPLIFIERS

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

MEDIUM POWER AMPLIFIERS

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

LOW NOISE OCTAVE BAND LNA'S

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

NARROW BAND LNA'S

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

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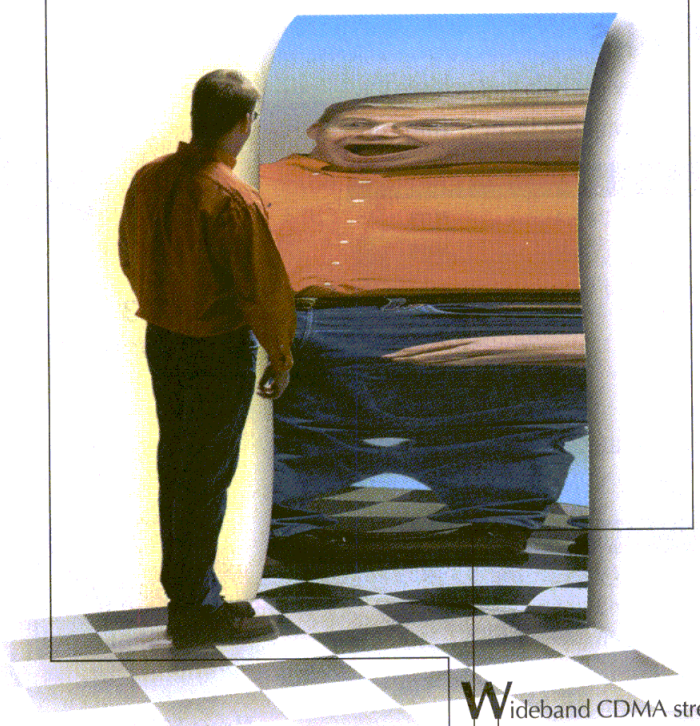

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3GPDP

tas® Telecom Analysis Systems

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New NEC Bipolar Transistors

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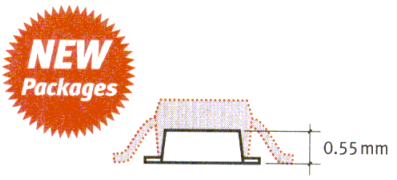
New, Smaller Packages



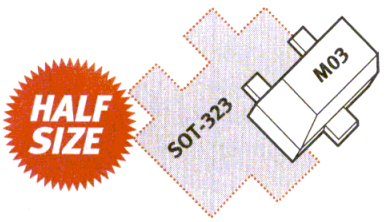
- **Flat Lead** design reduces parasitics and improves electrical performance
- **Low Profile** package is ideal for PCMCIA and other space-constrained designs

Oscillators & Buffer Amps

With the best *1/f* performance available, these devices help you achieve the phase noise your design demands. They're also available in Twin Transistors.

Part Number	Corner Freq*	V_{CE}	I_C	Package
NE856M03	3 KHz	3 V	30 mA	M03
NE685M03	5 KHz	3 V	5 mA	M03

**Review Application Note AN1026 on our website for more information on 1/f noise characteristics and corner frequency calculation.*

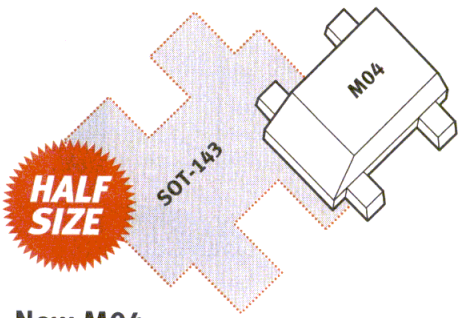


New M03
Half the footprint area of a SOT-323

LNAs

Need low noise and high gain in an ultraminiature package for your hand-held wireless products? These new high frequency NPN transistors deliver!

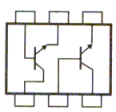
Part Number	Description	NF	Gain	Freq	Package
NE687M03	11 GHz f_T LNA	1.2 dB	13 dB	1 GHz	M03
NE661M04	25 GHz f_T LNA	1.2 dB	22 dB	2 GHz	M04
NE662M04	23 GHz f_T LNA	1.1 dB	20 dB	2 GHz	M04



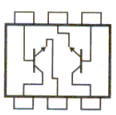
New M04
Half the footprint area of a SOT-143

Twin Transistor Devices

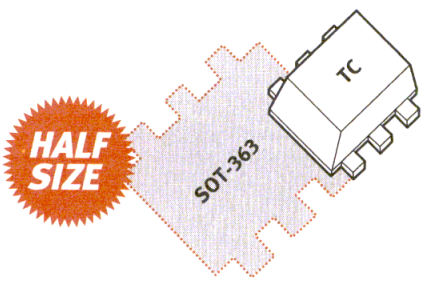
Cascode LNAs, cascade LNAs and oscillator/buffer combinations are just three possible uses of these versatile devices. *Matched Die* versions pair two adjacent die from the wafer to help simplify your design, while *Mixed Die* versions — an NEC exclusive — let you optimize oscillator performance while achieving the buffer amp output power you need. 40 different combinations available.



Part Number	Description	Q1 Spec	Q2 Spec
UPA810TC	Matched Die/Cascade LNA	NE856	NE856
UPA814TC	Matched Die/Cascade LNA	NE688	NE688



Part Number	Description	Q1 Spec	Q2 Spec
UPA826TC	Matched Die/Osc-Buffer Amp	NE685	NE685
UPA840TC	Mixed Die/Osc-Buffer Amp	NE685	NE681



New TC Twin Transistors
Half the footprint area of a SOT-363

Data Sheets and Application Notes are available at **www.cel.com**

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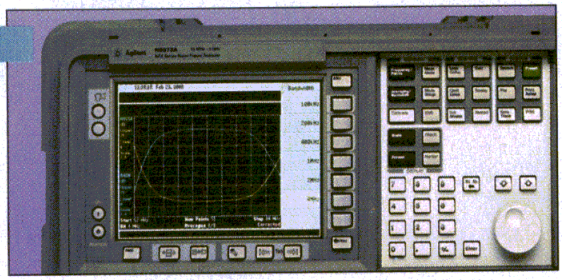
COVER FEATURE

199

Analyzer Adds Speed And Accuracy To NF Testing

The successor to the venerable 8970 series of noise-figure meters brings enhanced accuracy and operating speed to noise testing.



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218 Scopes Test Mixed Analog And Digital Designs

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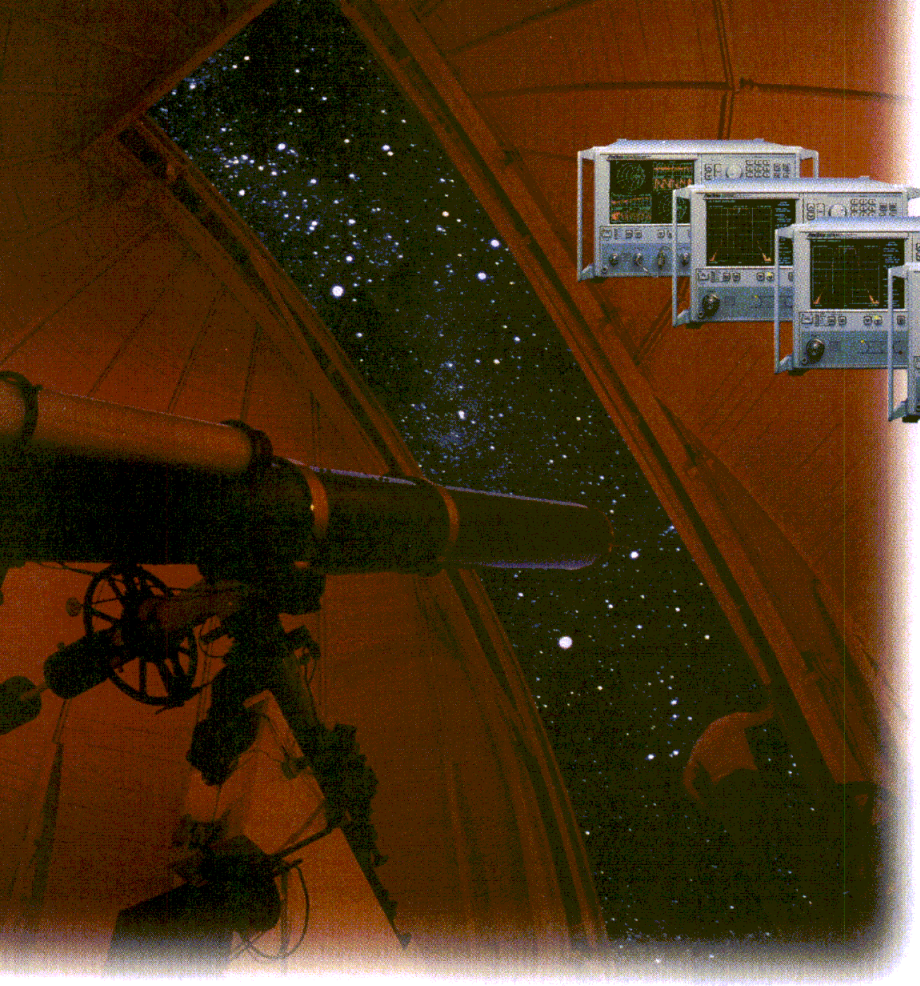

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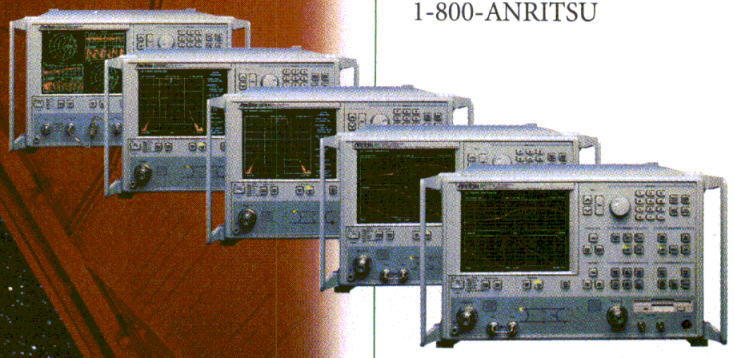

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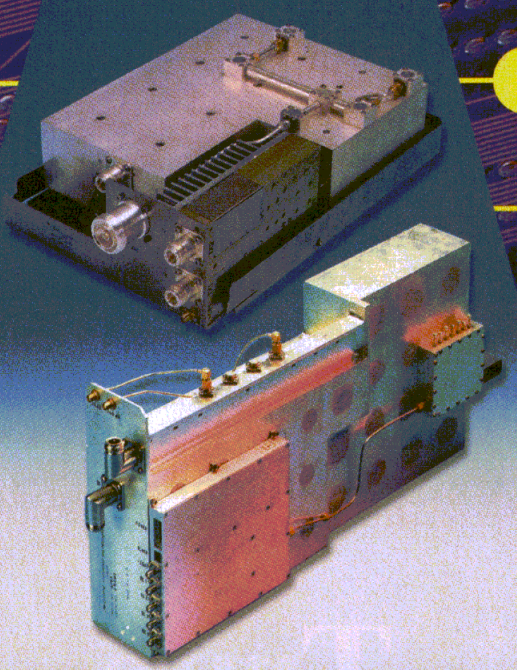

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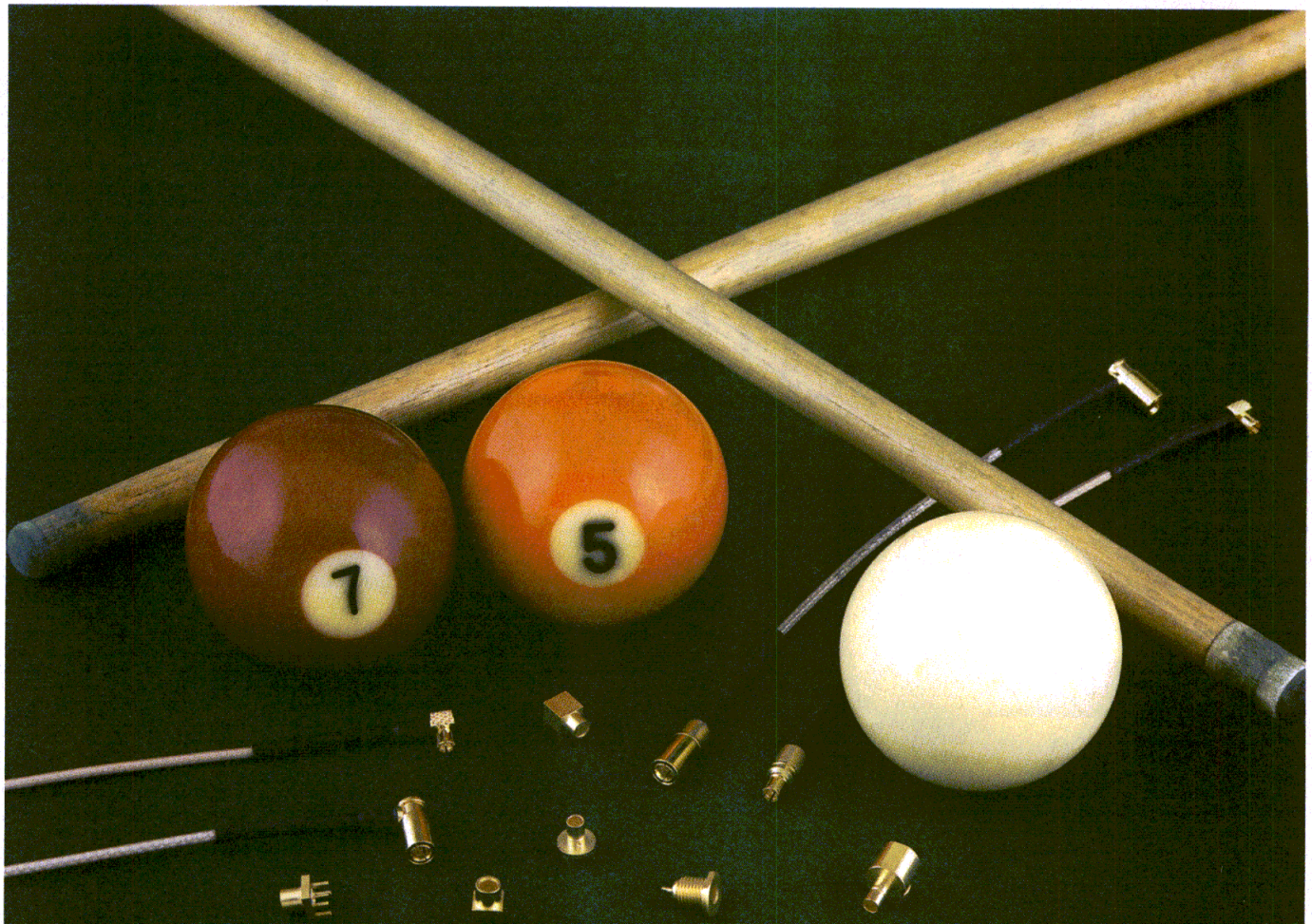

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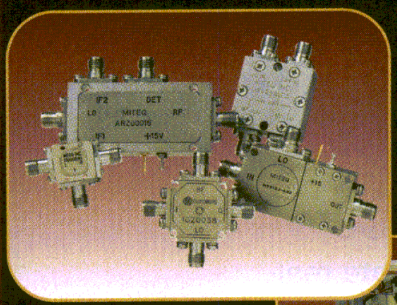

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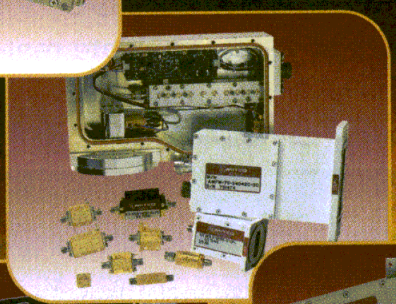

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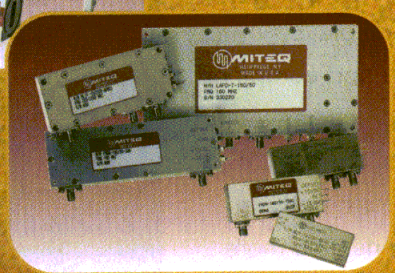

INTEGRATED SUBASSEMBLIES TO 60 GHz

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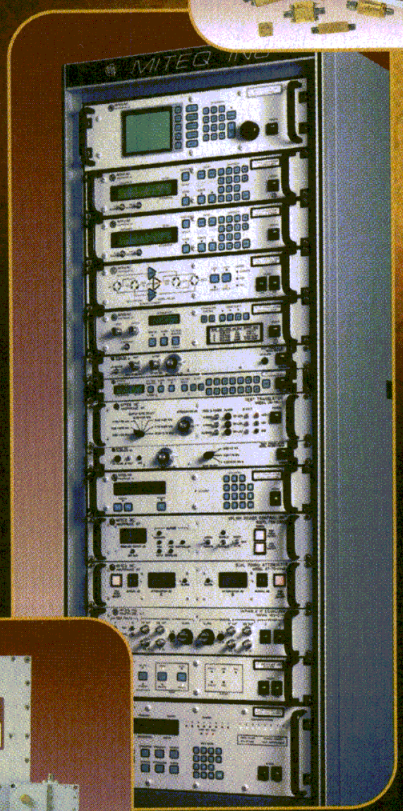

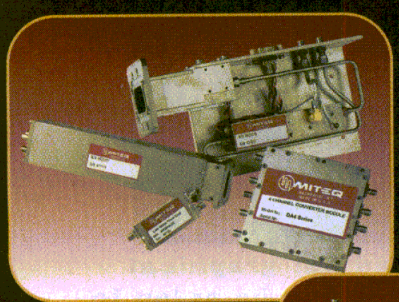

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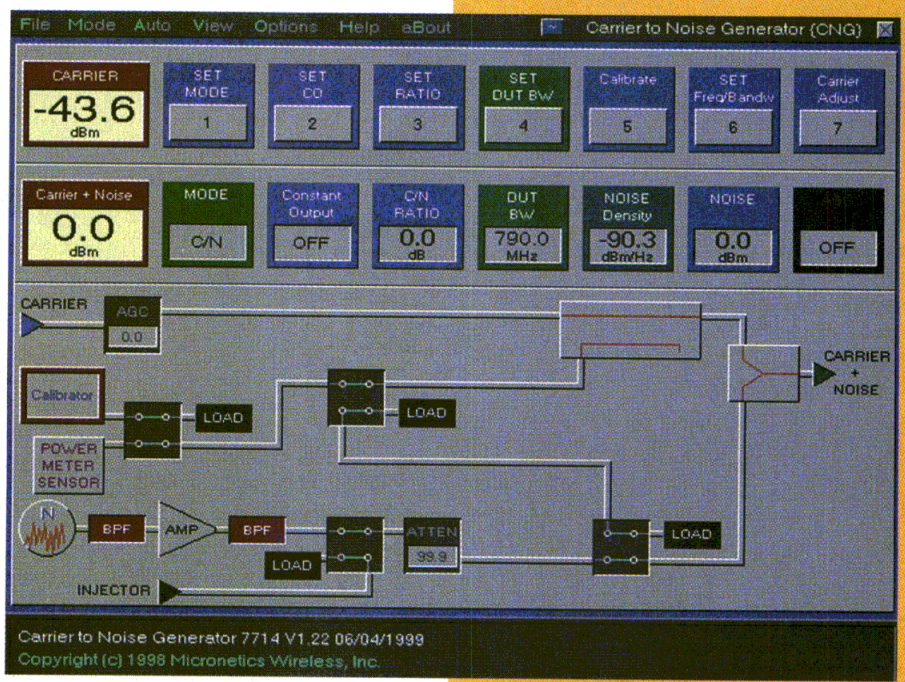

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QUIET LEADER

To the editor:

I enjoyed your excellent editorial about Dick Newman that appeared in the February issue (p. 17). While I did not know him well, the thoughts that you expressed were so well put and they thoroughly reflect the sort of values we need to inculcate in all of our young engineers.

Colin M. Jones
P.E.
Honolulu, HI

REFERENCE ERROR

To the editor:

In the April 2000 article "Design Waveguide Bandpass Filters," I downloaded the article from the web and noticed an inadvertent error which was not in the copy I had proofread for Don Keller. On page 103, the paragraph on "corrugated waveguide bandpass filters and waffle-iron filters" ended with "described in reference 2." This should have been "described in reference 3." The book by Matthaei, Young, and Jones was the correct reference. S.B. Cohn's article per reference 2 does not cover corrugated waveguide bandpass filters or waffle-iron filters. I referenced the book to avoid an excessively long list of references.

Richard Kurzrok
RMK Consultants

RF WEBSITES

To the editor:

April's editorial (p.17) mentions Pete Conrad's reviews of RF websites. Although we are small, we operate a website that many RF laboratories, government agencies, universities, and experimenters look at. We would like to add our website to the list: http://www.shfmicro.com.

Your editorial mentions something dear to me—a home page that loads fast. We receive many compliments from folks who do not enjoy seeing a 100K GIF load with nothing more than a button at the bottom to "enter" a website. The editorial also mentioned content. We offer a data sheet on-line or, at least, the instruction sheet which comes with the product or a plagiarized data sheet from the manufacturer.

We find that Acrobat PDF files really help our small customers decide what to purchase from us. Many engineers are grateful to find that even basic information is right there with the product, along with a small photo. Photographing small devices is a challenge that we met by purchasing an expensive color capture card, and finding a modern piece of equipment at our local K-Mart. Finally, we used an Intel-Play QX-3 children's computer-microscope to present our recent photos on the web. The total cost was $90 and it plugs-and-plays from the USB port of our computer.

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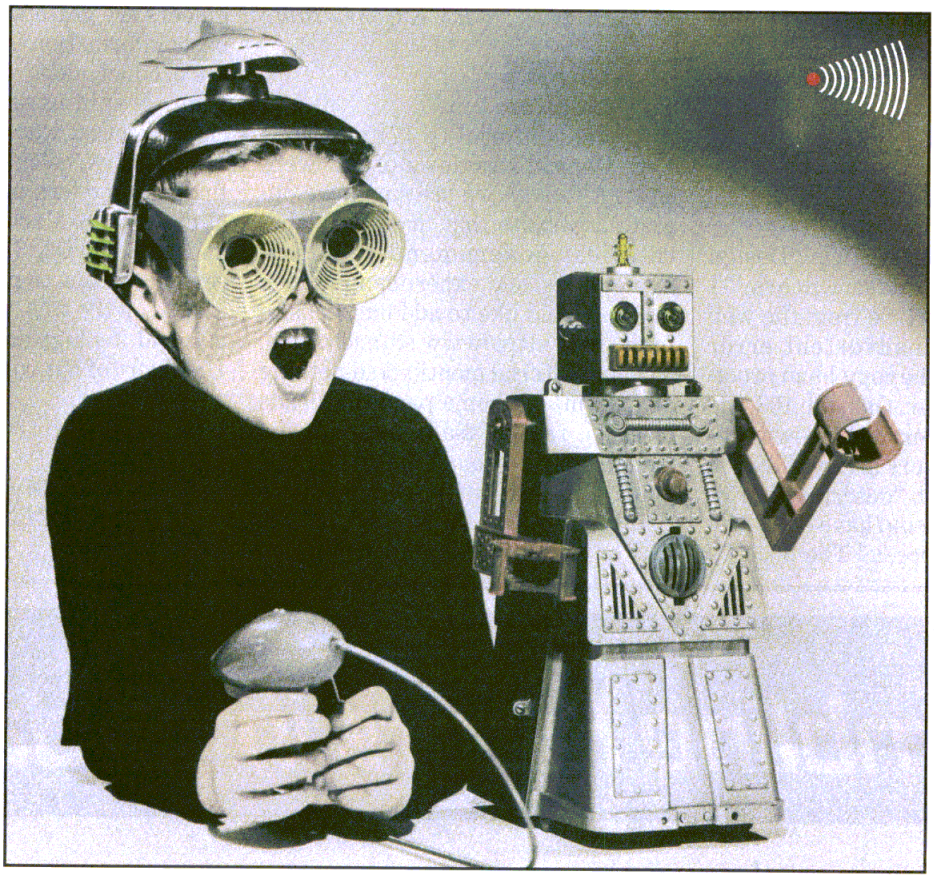

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DON'T OVERLOOK THOSE "SMALLER" SYMPOSIA AT MTT-S

Boston, MA is the site of this year's RF/microwave extravaganza, the 2000 International Microwave Symposium. Scheduled for June 11-16, 2000 at the John B. Hynes Veterans Memorial Convention Center, the IEEE/International Microwave Theory & Techniques Symposium (MTT-S) offers a total of 456 technical presentations, culled from a total of 845 submissions—easily the largest annual meeting for microwave technology in the world. But often lost in the magnitude of six parallel technical sessions over three days, are the smaller conferences that are integrated within the week's events.

This year's MTT-S program is impressive (see p. 150), covering everything from the latest amplifier technologies to trends in wireless communications. It has content for military-component designers as well as communications systems designers. It is massive, with multiple simultaneous sessions, and certainly has too much information for one sitting. Often lost amidst the sheer volume of papers at MTT-S are several conferences which, if held as stand-alone symposia, would probably garner a great deal more attention than they do when integrated into the MTT-S.

For example, the 2000 Radio Frequency Integrated Circuits (RF ICs) Symposium, chaired by David Lovelace of Motorola (Tempe, AZ), offers an amazing snapshot of technical advances in monolithic integrated-circuit (IC) technologies. Taking place from June 11-13, the symposium kicks off with two solid workshops—one covering silicon-germanium (SiGe) bipolar-complementary-metal-oxide-semiconductor (BiCMOS) processes for RF ICs and one on Bluetooth technology, tracing the history and development of this promising 2.4-GHz personal-wireless-connectivity standard. Approximately 50 technical papers will be presented that Monday (June 12).

Another small, but important, symposium at the 2000 MTT-S is one of extreme interest to engineers involved with high-frequency measurements—the 55th Automatic RF Techniques Group (ARFTG) conference. Chaired by Mike Fennelly of Roos Instruments (North Andover, MA), this one-day conference, scheduled for Friday, June 16, at the Hynes Convention Center, also boasts high-caliber presentations from some leading high-frequency organizations, such as Agilent Technologies (Santa Rosa, CA), Bell Laboratories (Murray Hill, NJ), and National Institute of Standards and Technology (Boulder, CO).

A sampling of the ARFTG presentations covers quasioptical measurements at Ku-band using focused Gaussian beams, novel noise-power-ratio measurements for multicarrier power amplifiers (PAs), a cost-effective method of modeling large-signal device behavior, simulation of large-signal on-chip amplifier modules excited by wideband-code-division-multiple-access (WCDMA) signals, and multiport noise characterization of differential amplifiers.

The MTT-S offers a great deal. But both of these "smaller" conferences offer impressive lineups that should not be overlooked within the big picture.

Jack Browne

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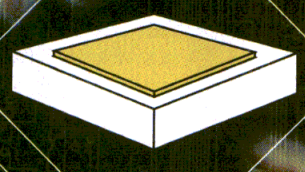

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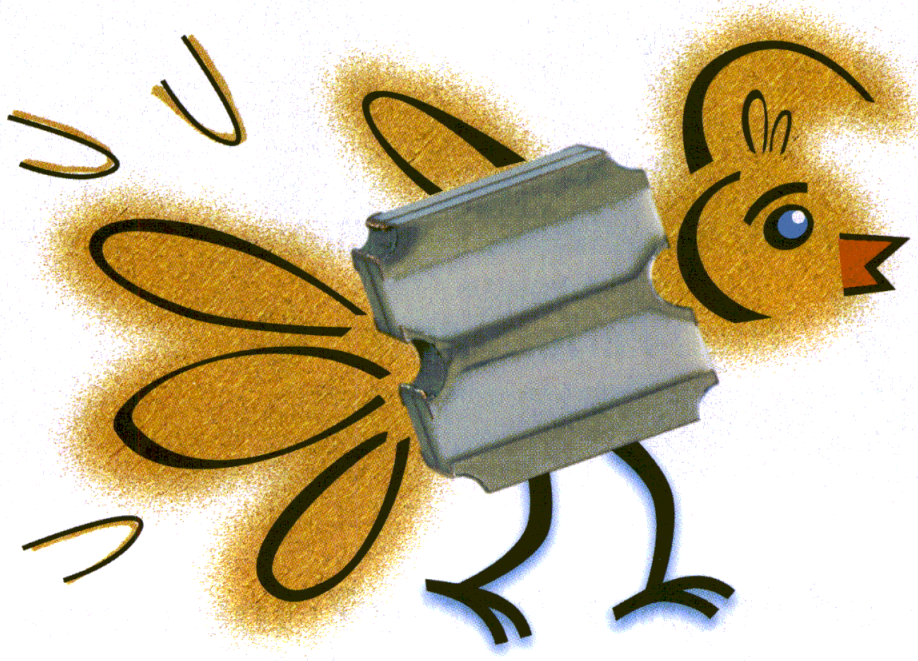

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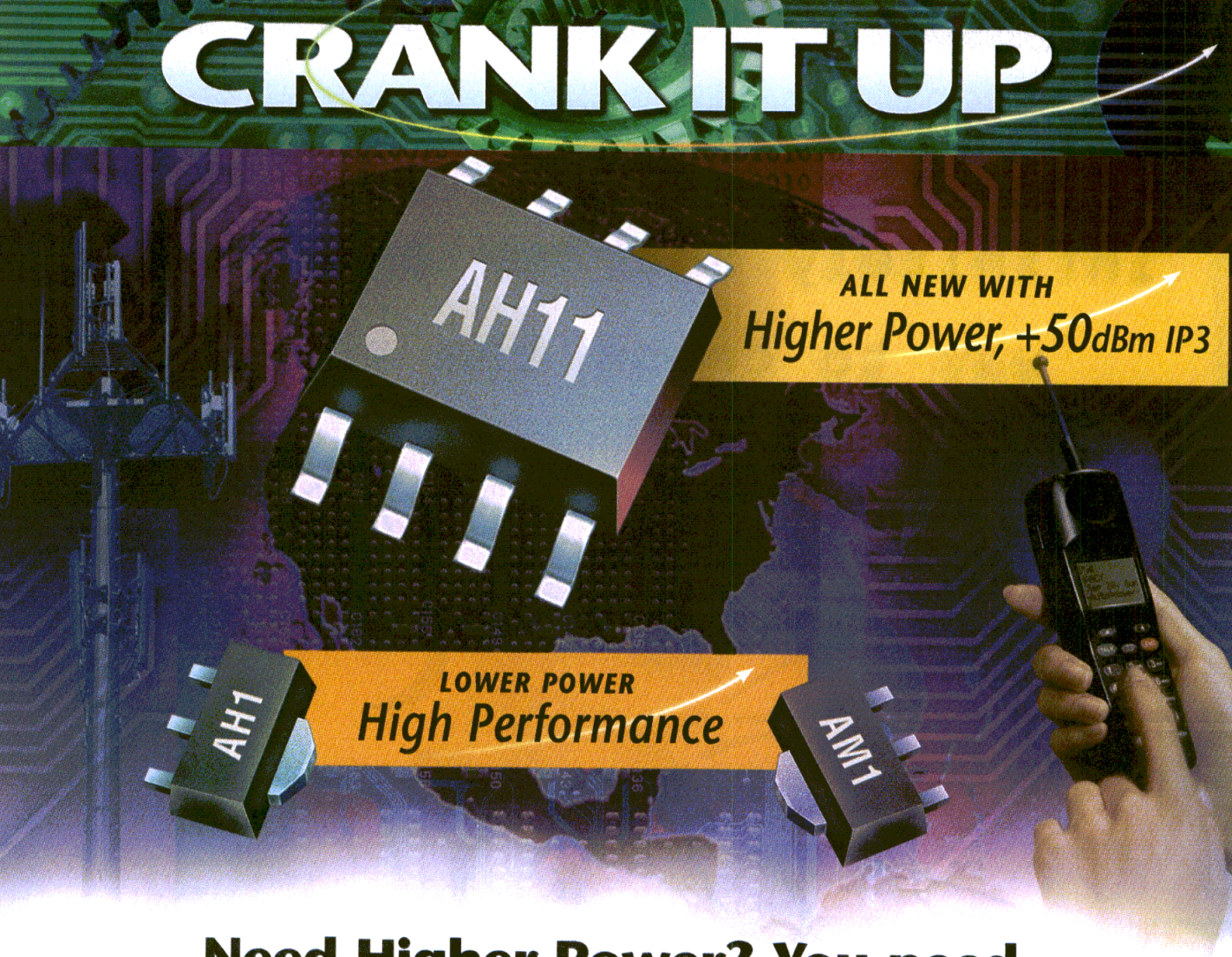

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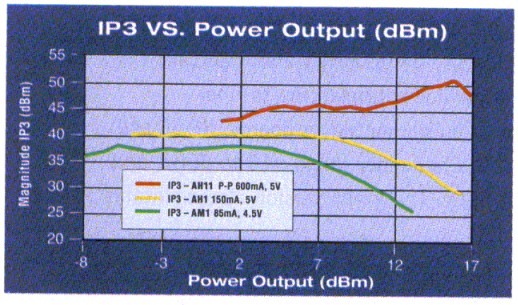

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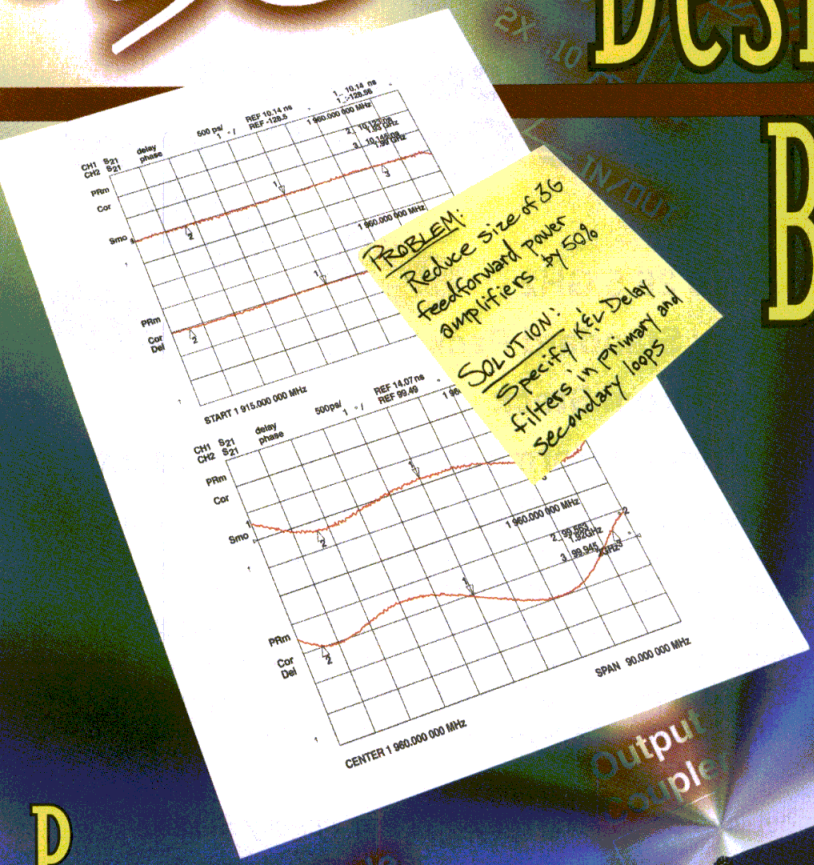

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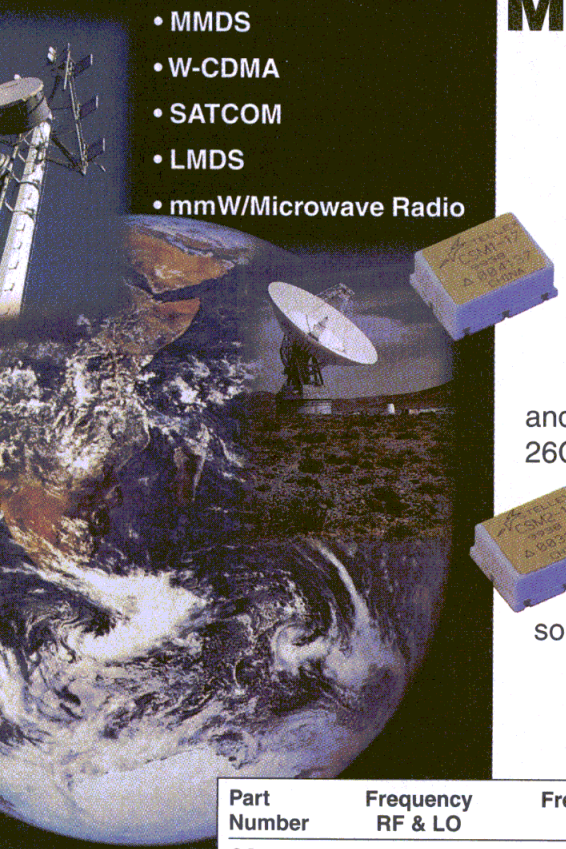

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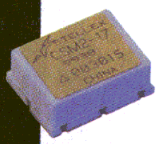

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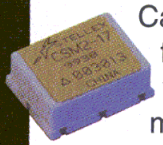

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CSM1-13	10 to 1,500 MHz	1 to 500 MHz	+13 dBm	40 dB	22 dBm	7.5 dB	Surface Mount
CSM1-17	10 to 1,500 MHz	1 to 500 MHz	+17 dBm	40 dB	27 dBm	7.5 dB	Surface Mount
CSM2-10	10 to 2,800 MHz	10 to 2,000 MHz	+10 dBm	30 dB	20 dBm	7.5 dB	Surface Mount
CSM2-13	10 to 2,800 MHz	10 to 2,000 MHz	+13 dBm	30 dB	22 dBm	7.5 dB	Surface Mount
CSM2-17	10 to 2,800 MHz	10 to 2,000 MHz	+17 dBm	30 dB	27 dBm	7.5 dB	Surface Mount
MC4107	2 to 10 GHz	DC to 2 GHz	+7 dBm	40 dB	11 dBm	6.0 dB	Open Carrier
MC4110	2 to 10 GHz	DC to 2 GHz	+10 dBm	40 dB	14 dBm	6.0 dB	Open Carrier
MC4113	2 to 10 GHz	DC to 2 GHz	+13 dBm	40 dB	17 dBm	6.0 dB	Open Carrier
MC4120	2 to 10 GHz	DC to 2 GHz	+20 dBm	40 dB	23 dBm	6.5 dB	Open Carrier
MC4507	4 to 22 GHz	DC to 4 GHz	+7 dBm	32 dB	11 dBm	6.0 dB	Open Carrier
MC4510	4 to 22 GHz	DC to 4 GHz	+10 dBm	32 dB	14 dBm	6.0 dB	Open Carrier
MC4513	4 to 22 GHz	DC to 4 GHz	+13 dBm	32 dB	17 dBm	6.0 dB	Open Carrier
MC4520	4 to 22 GHz	DC to 4 GHz	+20 dBm	32 dB	23 dBm	6.5 dB	Open Carrier

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CIRCLE NO. 261

Program Delivers Location-Enhanced Traffic And Traveler Information

SAN RAMON, CA—U.S. Wireless Corp., a provider of wireless-location information, recently introduced the RoadShare™ partnership program to deliver location-enhanced traffic and traveler information services. In concert with its partners, U.S. Wireless aims to provide Intelligent Transportation Systems (ITS) consumers with real-time traffic data, services, and systems while providing government agencies with traffic information and data to build and monitor more-efficient roadways.

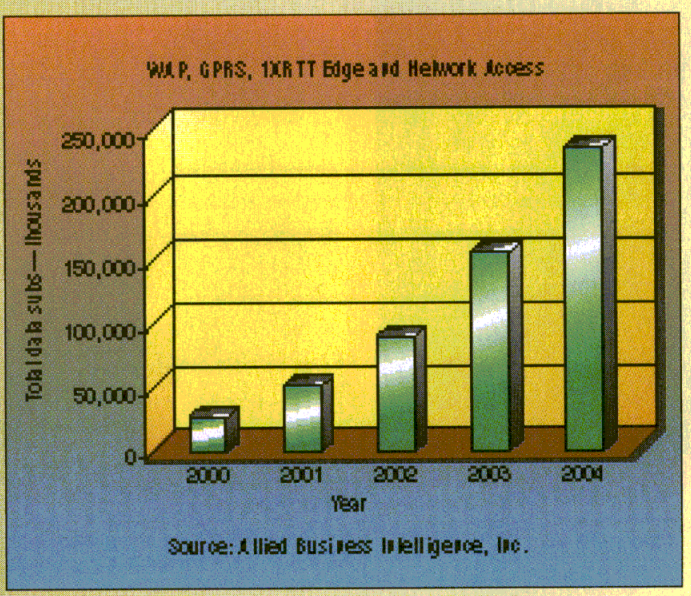
The RoadShare partnership program will allow developers of ITS and other transportation-related services to design localized traffic and traveler services, using the U.S. Wireless location network. The U.S. Wireless location network provides real-time estimates of traffic velocity and congestion without the need for inroad sensors, aircraft, or video cameras. By monitoring the flow of cellular traffic on roads and highways, the U.S. Wireless network supports a variety of intelligent transportation services, including live traffic monitoring, location-enhanced traveler updates and alerts, as well as statistical and real-time "best route" information.

"The RoadShare partnerships combine the capabilities of U.S. Wireless' mobile information platform with the expertise of leaders within the Intelligent Transportation Systems industry," says Uday Nagendran, director of Intelligent Transportation Systems at U.S. Wireless. "The compilation of real-time traffic speed and congestion information will enable numerous traffic and traveler information services, as well as future in-vehicle interactive services."

Large Wireless Data Market Is Emerging

OYSTER BAY, NY—There is finally some evidence that a wireless data market will bloom as third-generation (3G) wireless services begin to emerge, according to a study from Allied Business Intelligence, Inc. (ABI).

Actual direct Internet access to a handset will become available to more than 100 million wireless users during 2000. Truncated-access methods will be put in front of hundreds of millions of users during the next year. By 2005, there will be more than 240 million data users, increasing from 26 million in 2000 (see figure), according to the findings in the study "3G Wireless: Global Data & Voice Networking Strategies—WAP, GPRS, IXRTT, EDGE and Internet Access."



"With GPRS going up in much of Western Europe and truncated-access services, such as those based on WAP, gaining considerable momentum worldwide, there is a data market finally at hand which will prove valuable to users and profitable to suppliers," says Larry Swasey, ABI's vice president of communications research and author of the report. "As 3G radio-channel element upgrades take place and data strategies are placed into mobile wireless networks, the handset will become a much more valuable tool to both the operator and the user."

Thirty-seven operators made wireless-application-protocol (WAP) announcements as of January this year. The competitive nature of most wireless markets will force other operators to deploy truncated-access services. Many handset manufacturers have already vowed to have WAP-based handsets on the market in 2000. The introduction of WAP handsets began in 1999. Overall, the Americas will account for approximately 40 percent of the data market by 2005.

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Wireless Networking System Available For Home And Small-Business Users

NEW ORLEANS, LA—Lucent Technologies recently announced the ORINOCO™ home networking system, a high-speed (11-Mb/s), wireless networking system for home and small-business users. The ORINOCO home networking system allows users to easily and inexpensively set up a high-speed-data network that enables them to share resources such as files, scanners, and printers. The ORINOCO system also allows multiple users to share the same Internet connection and to connect anywhere in a home or small office.

The ORINOCO system became available in March to service providers, original equipment manufacturers (OEMs), and selected online retailers. The system works with Windows™ and Apple computers.

Operating in the unlicensed 2.4-GHz band, the ORINOCO system operates indoors and outdoors, transmits through walls and floors, and provides a range up to 1500 ft. ORINOCO is fully IEEE802.11 high rate compliant, which means that it can work with all other 802.11-compliant wireless networking systems including the Apple AirPort. The 11-Mb/s speed provides Ethernet-quality networking experience—all without cords or cables.

The built-in installation wizard and Dynamic Host Configuration Protocol (DHCP) automatically sets up the user's station parameters for network and Internet connection. ORINOCO also includes built-in Network Address Translation routing that enables multiple users to share a single dial-up connection or broadband Internet connection.

ORINOCO consists of PC-card radios that slip into the Personal Computer Memory Card International Association (PCMCIA) slots of laptops and portable computers. Industry-standard-architecture (ISA) and Personal Communications Industries (PCI) adapters enable the PC cards to work inside of desktop computers, and the RG-1000 Residential Gateway access point acts as a central base station in a house or small business. The Residential Gateway includes a built-in 56K modem that can be directly plugged into a phone line, and a 10baseT Ethernet jack for connecting to an existing wired Ethernet network or an xDSL or cable-modem Internet connection.

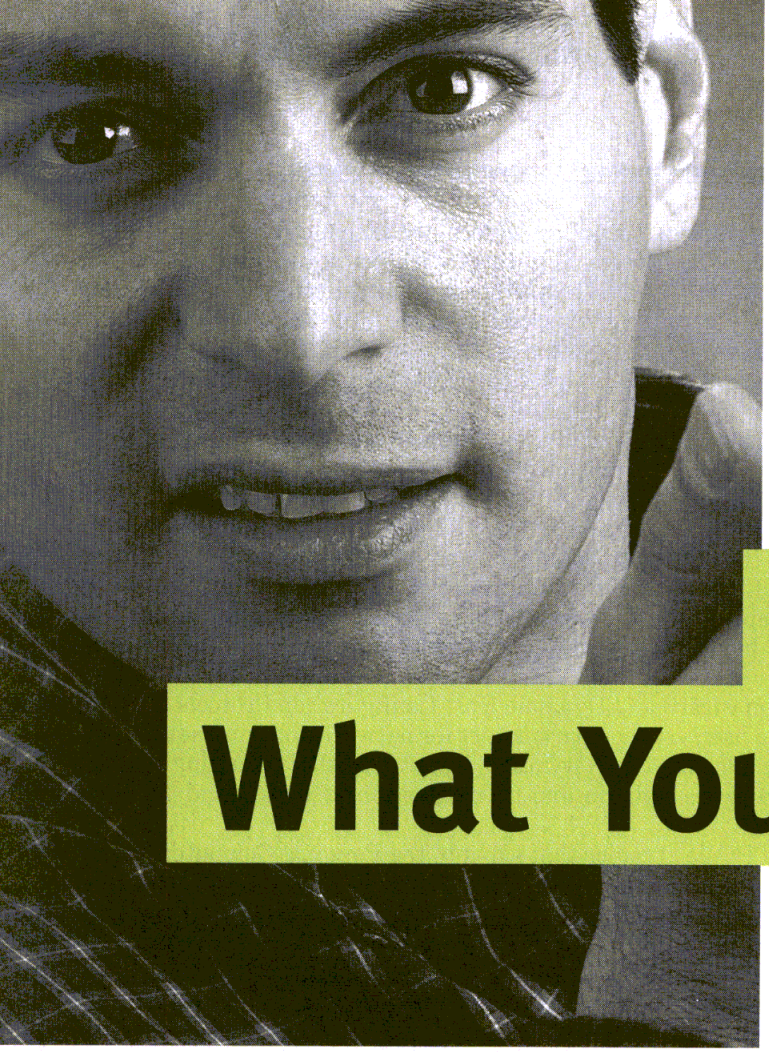
Bluetooth CDMA Solutions Are Jointly Developed

SAN DIEGO, CA—Qualcomm and Ericsson Microelectronics have announced an agreement to jointly develop and market the world's first wireless technology solution supporting the Bluetooth standard and the code-division-multiple-access (CDMA) digital wireless standard.

"The true promise of wireless technology lies in its ability to expand the potential of personal appliances. This has been the driving force behind Ericsson's work in pioneering the Bluetooth technology," says Bo Andersson, vice president and general manager of Ericsson Microelectronics. "Qualcomm's expertise in providing tightly integrated CDMA technology makes them a perfect partner."

Under the agreement, Ericsson Microelectronics will develop a Bluetooth-compatible radio unit and Qualcomm CDMA Technologies will develop the Bluetooth digital baseband processing to be integrated into its future mobile-station-modem (MSM™) chip-set and software solutions. The companies will work together to optimize the RF and digital designs for operation within a CDMA handset environment to create the most complete and high-performance solution available for CDMA-based wireless data applications. Ericsson Microelectronics and Qualcomm CDMA Technologies will jointly promote the integrated solution to handset and mobile device-manufacturing companies.

This design solution will connect end users to a new world of benefits. The range of possible applications is unlimited, including the ability for end users to walk into their offices and have the handset easily update their schedule and phone book to synchronize to the personal computer (PC). On a business trip, the end users could take a picture with a digital camera and send it to the office through the handset. Other examples are that the handset could be used as an intercom in the office, a cellular phone on the move, and as a cordless handset in one's home. Moreover, the end user will be able to exchange electronic business cards, unlock one's car, or set up radio preference. High-capacity, high-data-rate-capable CDMA networks will deliver wide-area-voice and Internet services to the Bluetooth-enabled terminal devices.



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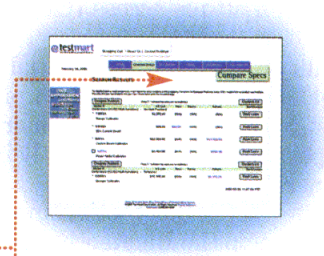

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CIRCLE NO. 234

Partnership Approved For Radiation-Hardening Commercial Microelectronics

SAN DIEGO, CA—Space Electronics, Inc. (SEi), a subsidiary of Maxwell Technologies, Inc., announced that it has signed a strategic partnership with U.S. Semiconductor to commercialize U.S. Semiconductor's RHI-NO™, an innovative, wafer-thinning technology and production-thinning process for radiation-hardening commercial microelectronics.

Radiation-hardened microelectronics are required for a broad and rapidly-growing variety of space and avionics applications. The parties are jointly seeking government and industrial research-and-development (R&D) contracts to further develop the technology for space and military systems.

"U.S. Semiconductor's wafer-thinning process will offer another technology for us to leverage commercial silicon for the space market," says David J. Strobel, chief technical officer at Space Electronics, Inc.

SEi's RAD-PAK® technology enables the application of advanced microelectronic devices to meet the strictest requirements. The company's rad-hard integrated circuits (ICs) are appropriate for a wide range of commercial and military-space platforms and experiments. SEi's contributions to space programs include low-cost application-specific IC (ASIC) solutions; board products; communication devices; data-conversion devices; first-in, first-out (FIFOs)/field-programmable gate arrays (FPGAs); linear devices; logic; memory; microprocessors/microcontrollers; peripheral devices; and power semiconductors.

R&D Spending Will Significantly Increase In 2000

COLUMBUS, OH—Total research-and-development (R&D) expenditures in the US are expected to increase to more than $266 billion in 2000, according to the annual Battelle-*R&D Magazine* R&D forecast. That is approximately 8 percent more than the $247 billion that the National Science Foundation estimates was spent in 1999.

"A combination of situations has created what could very well be the continuation of a major resurgence in R&D support and activity in the United States," says Jules J. Duga, a Batelle senior researcher and co-author of the report.

The continual growth of R&D funding is largely the result of the industry's growing realization that structural and operational changes are not the only road to profitability. Continual investment in research—especially in development—will be required for long-term survival.

Federal R&D spending will change very little. The forecast estimates that the federal government will spend $66.4 billion on R&D in 2000, only slightly more than the $65.9 billion spent in 1999. Industry will increase its R&D commitment, spending $187 billion in 2000, an increase of more than 10 percent from the $169 billion spent in 1999. The remainder of R&D support—more than $12.6 billion—will be provided by universities and other nonprofit organizations, representing an increase of 6 percent or more over 1999 funding. Once the dominant supporter of R&D, providing more than 50 percent of total funding, the federal government now contributes approximately 25 percent. There is little indication that this trend will be reversed in the near future.

The federal budget continues to give very mixed signals about its support of R&D, with major differences between budget authorization and the actual expenditures of funds.

Kudos

Astrolab, Inc. announced that it has been awarded "Certified To Stock Supplier" status by EFW, Inc. of Fort Worth, TX. This award was based on materials delivered to EFW for the past year and is presented to those suppliers who have continuously provided high-quality, timely delivery as well as rapid responsiveness...Microchip Technology recently recognized four suppliers for outstanding performance in quality, reliability, and service for 1999. Microchip's outstanding suppliers are: Dai-ichi Seiko Co., auto-assembly molding equipment (Kyoto, Japan); Kleinfelder, environmental consultants (Phoenix, AZ); Komatsu, silicon (Si)-wafer supplier (Nagaski, Japan and Miyazaki, Japan); and NS Electronics Bangkok, assembly contractor (Bangkok, Thailand)...Dr. Frank L. Whetten, associate professor of electrical engineering at Embry-Riddle Aeronautical University in Prescott, AZ has been chosen to participate in the prestigious Boeing-A.D. Welliver Faculty Summer Fellowship Program.

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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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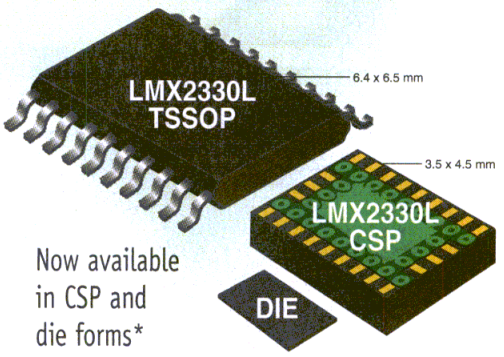

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CIRCLE NO. 274

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Antenna Technology Tunes To Wireless Markets

JACK BROWNE
Publisher/Editor

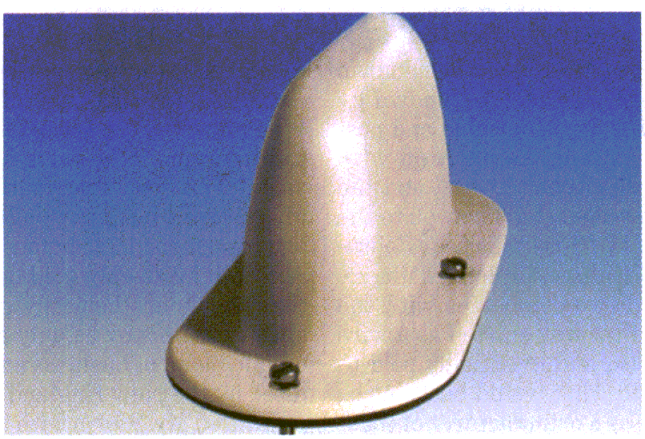
ANTENNAS are the gateways to wireless communications. They are the interface between free space and the low-noise electronics of a receiver as well as the high-power circuitry of a transmitter. While military markets drove the design of high-frequency antennas more than a decade earlier, the growth of wireless applications has prompted the development of many unique and compact passive and active antennas over the last few years. In many cases, antenna specifiers must find components that can increase the range and/or coverage of their wireless system, and do so without adding high visibility to the system.

Antennas are characterized by a number of key parameters, including bandwidth, beamwidth, directivity, efficiency, gain, polarization, radiation pattern, and voltage-standing-wave ratio (VSWR). To ensure that an antenna either receives the minimum detectable signal possible or transmits the maximum amount of power efficiently, it should be closely matched in impedance (usually 50 Ω) to its associated receiver and/or transmitter electronics. A less-than-ideal impedance match will result in reflections between the associated electronics and the antenna.

In most circuits, the radiation of energy into free space causes interference with adjoining components, such as a local oscillator (LO) radiating energy into an amplifier. But antennas are constructed for this free-space radiation of energy, according to carefully designed radiation patterns. An ideal point-source isotropic omnidirectional antenna radiates energy over 360 deg., creating a pattern in the shape of a perfect sphere. Since the energy is equally radiated in all directions, it is said to have 0-dB directivity. Oftentimes, this ideal isotropic antenna is used as a reference for comparison to real-world antennas in terms of gain and directivity. For example, an antenna with 2 dB of more gain in a particular direction than the ideal isotropic antenna is said to have 2-dBi gain. If a dipole antenna is used as the reference (with the same differential in gain between the actual antenna and the ideal reference), then the antenna is characterized as having 2-dBd gain.

In reality, even the best omnidirectional antennas can only approximate the radiation pattern of an ideal isotropic antenna, resulting in some level of antenna directivity relative to the ideal isotropic radiator. The relative level of an antenna's maximum radiated energy to that of a reference isotropic antenna is called the antenna's directivity, which is essentially the effectiveness of an antenna to focus its radiated energy in a particular direction. High values of directivity indicate antennas with highly focused energy beams.

An antenna's beamwidth is a measure of the angular width (in degrees) of the antenna's radiation pattern, at some drop in the amplitude of the pattern, such as 3 dB, relative to the maximum level of the radiation pattern. In general, the beamwidth is specified separately for an antenna's hori-



1. Model ASP-372 is a low-profile vehicular antenna designed for UHF TETRA applications from 380 to 400 MHz. (Photograph courtesy of Antenna Specialists, an Allen Telecom Company, Cleveland, OH.)

zontal or azimuth plane and for its vertical or elevation plane.

An antenna's bandwidth refers to the useful frequency range of an antenna, defined by some loss in the level of the radiation pattern (such as 3 dB) at the edges of the bandwidth. Gain is essentially a measure of an antenna's overall efficiency, given as the product of the antenna's directivity and efficiency. For an ideal antenna with 100-percent efficiency, the gain would be equal to the directivity (such as the ideal reference isotropic antenna).

Polarization describes the orientation and sense of an antenna's E-field vector. An antenna with linear polarization has an E-field vector that follows a linear fixed trajectory as a function of time, with the vector amplitude varying sinusoidally with an angular velocity. When the vector forms a circle, the polarization is said to be circular. When the vector's circular path is clockwise (observed in the direction of propagation), it is said to be right-hand circular polarization. When it is counterclockwise, it is left-hand circular polarization.

A variety of different antenna types is used in modern wireless systems, from the giant parabolic dish antennas of satellite communications systems to tiny dipole antennas used in cellular communications handsets. Basic radiating elements, such as monopoles or dipoles, are often combined to form larger array antennas. Reflector antennas are formed by surrounding one or more radiating elements with a reflective surface to focus the received or transmitted energy. As the reflector increases in size with respect to wavelength, the beamwidth of the reflector becomes more narrow. For miniaturization, microstrip antennas are formed by means of metallized circuit traces on thin dielectric substrates. As the thickness of the dielectric material increases, the bandwidth also increases. While the bandwidth of a microstrip antenna is somewhat limited, it can be made extremely compact and low in cost.

A recent trend in antenna development has been the design of "smart" antennas by combining antenna hardware and digital signal processing (DSP) with spatial-processing software to electronically adjust antenna patterns according to changing operating conditions. These types of antennas, which borrow from traditional military phased-array technology, can be formed by means of switched-beam, adaptive arrays, and adaptive beamforming approaches (see "Smart-Antenna Technology Is The Wave Of The Wireless Future, *Microwaves & RF*, May 1999, p. 29). Switched-beam antennas employ a fixed-phase feed network known as a beamforming network with different outputs corresponding to beams with fixed directions. The beams are selected or optimum performance according to operating conditions. In an adaptive array, the patterns are controlled dynamically, by means of feedback to control the phase and amplitude of the beams. Adaptive beamformers use DSP techniques and iterative algorithms to adjust the contributes of the radiating elements. A variety of companies is currently involved with smart antenna technology, including ArrayComm (San Jose, CA), Metawave Communications Corp. (Redmond, WA), and RangeStar (San Jose, CA).

2. The MFB line of 2.4-GHz omnidirectional antennas is shown here. (Photograph courtesy of MAXRAD, Hanover Park, IL.)

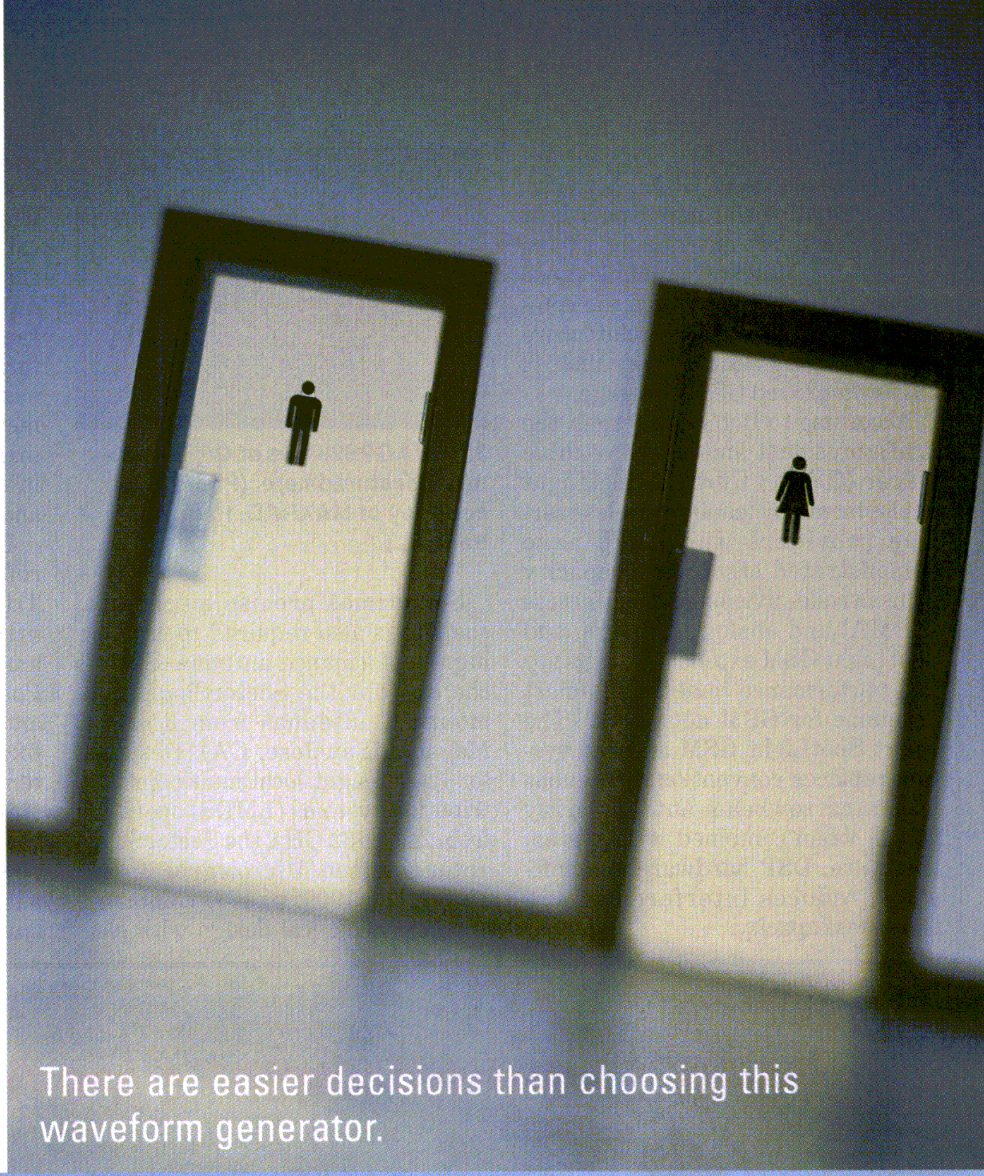
Metawave Communications Corp. recently applied its beam-switching

INDUSTRY LOSES A PIONEER

Storm Products (Westmont, IL) is not a firm known for antennas, although it may well be known by antenna specifiers for its precision cable and passive components that are used in conjunction with antennas. Sadly, Storm Products employees, and the high-frequency industry as a whole, bid farewell to the company's founder John A. (Al) Storm this past February. He died on February 25th after a prolonged illness.

Al Storm actually started two companies with the name Storm Products. The first was started as a wire distributor in the early 1950s, and eventually sold to Raychem. In 1958, he started what would become the current Storm Products Co., as a manufacturer of coaxial and fiber-optic cables and assemblies.

Al Storm was one of the early pioneers of the wire- and cable-distribution business. He was one of the first to offer value-added services—dye/stripe, twisting and cut-and-strip—as part of the distribution activity. When his interest turned to manufacturing, he continued to promote the value-added concept, as well as the value of creative imagination and individual initiative. Mr. Storm's business philosophy embraced employee empowerment long before it became fashionable, and reflected his deeply held beliefs that life should be fun, that people are basically good, and that business is a game with money as only a way of keeping score. His unique blend of business and "people" sense was recognized by the New England Wire and Cable Club, which presented Mr. Storm with the Scott Award for honesty, integrity, and outstanding service to the industry. Al Storm is survived by his wife, June, and three daughters, Janet, Sandy, and Nancy.



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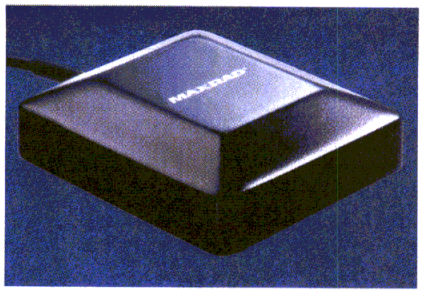
CIRCLE NO. 292

technology to improve the capacity and performance of GSM networks, in the form of the new SpotLight GSM. The smart antenna for Global System for Mobile Communications (GSM) networks leverages the company's SpotLight 2000 architecture (see *Microwaves & RF*, May 1999, p. 30) for improved GSM coverage.

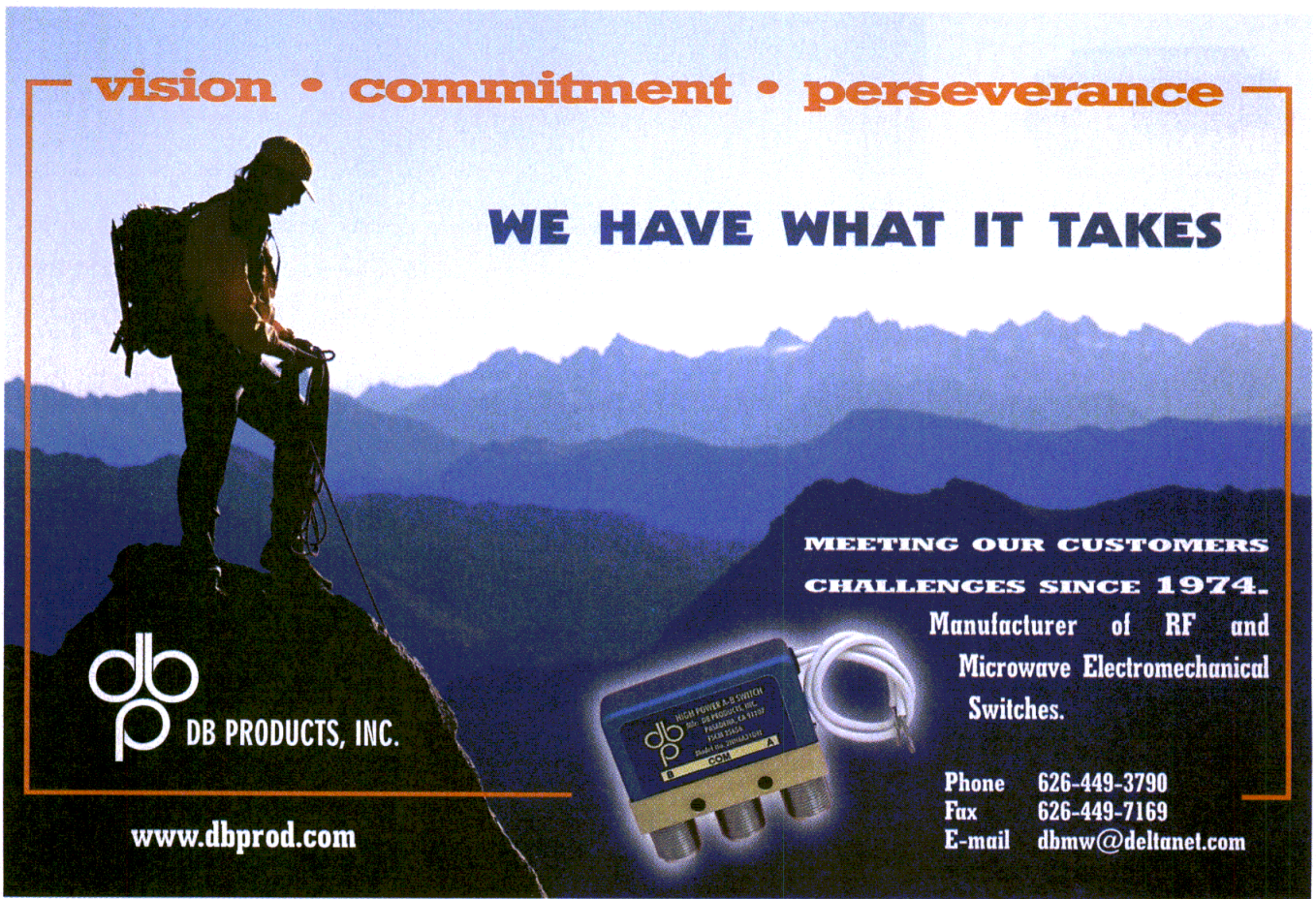
According to Bob Hunsberger, the firm's president and chief executive officer (CEO), "Current SpotLight deployments by leading wireless service providers worldwide have demonstrated significant capacity gains in code-division multiple access (CDMA) and analog networks, and SpotLight GSM extends the capacity and performance benefits of smart antennas for GSM networks." The smart SpotLight GSM antenna system replaces conventional antennas with a narrowbeam antenna array which, when combined with system hardware, DSP hardware and software, reduces interference, and increases capacity.



3. The AGPS26 line of GPS antennas is featured here. (Photograph courtesy of MAXRAD, Hanover Park, IL.)

Sometimes precise mechanical shaping is also required to create a high-performance antenna. Such is the case for the SectorShape hub-mounted antennas from REMEC Magnum (San Jose, CA). Designed for line-of-sight local multipoint distribution systems (LMDS) operating from 24.5 to 32 GHz, the SectorShape antennas (see *Microwaves & RF*, November 1999, p. 177) combine a unique mechanical design with electronically controlled downtilt characteristics to simplify mounting and alignment in LMDS systems. Vertical beam patterns are electronically shaped to remove nulls and provide uniform coverage as a function of range. The odd-shaped ridged reflectors used in the SectorShape antennas were developed with the help of a method-of-moments (MOM) electromagnetic (EM) field solver working in conjunction with mathematical-analysis software.

In some cases, the word "smart" refers to an antenna's installation. Trimble Navigation (Sunnyvale, CA, http://www.trimble.com) recently announced the Acutime 2000 Global Positioning System (GPS) "smart" antenna for integrating GPS into any fixed or mobile communications system. The antenna system provides a time and position source that includes a receiver, antenna, and power supply. When powered up, it automatically tracks satellites and surveys its position within a few



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meters. It then switches to overdetermined time mode and can generate a pulse-per-second timing signal synchronized to universal time code (UTC) within ±100 ns. It can also make use of the company's proprietary TimeTrigger technology which allows the smart-antenna system to receive a pulse from the host system and then respond back to the host with a time tag.

In terms of "more conventional" passive-antenna designs, the trend remains for smaller, less-expensive antennas. For example, the ASP-372 series of low-profile transit antennas from Antenna Specialists (Cleveland, OH, http://www.antenna.com) is designed for Trans-European Trunked Radio (TETRA) systems operating from 380 to 400 MHz. The compact antenna (Fig. 1) replaces traditional whip-style antennas for ultra-high-frequency (UHF) wireless communications in vehicles and construction equipment. The low-profile antenna, which requires no field tuning during installation, handles up to 100-W power with 2-dBi gain and VSWR of less than 2.0:1.The antenna measures 3.5 × 8.0 × 3.5 in. (8.9 × 20.3 × 8.9 cm) with a brass radiator and plastic radome.

In addition, the company now offers a series of in-building repeater kits to improve mobile wireless reception in poor coverage areas. The EAC-50 repeater kit includes a bi-directional amplifier, antennas, and related hardware. The minimum system gain is 50 dB at 800 MHz for Advanced Mobile Phone Service (AMPS) systems, with typical noise figure of 8 dB for uplinks and 10 dB for downlinks.

The transportation industry is also a driving force behind the MFB line (Fig. 2) of omnidirectional antennas from MAXRAD (Hanover Park, IL, http://www.maxrad.com). Designed for traffic-monitoring applications in the transportation industry and for wireless Internet and inventory-tracking systems operating at 2.4 GHz, the antennas achieve 10-dB gain with less than 1.50:1 VSWR.

The company manufactures a variety of different antenna types, including the MP series of directional-panel antennas for wireless local-

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area networks (WLANs) and personal-communications-services (PCS) applications at 800 and 900 MHz. The antennas are designed for 6-dBd gain with as much as 70-deg. horizontal beamwidth and 60-deg. vertical beamwidth. In addition, the firm's AGPS series of Global Postioning System (GPS) antennas (Fig. 3) are based on a low-profile microstrip patch design. Available in 26- and 31-dB gain versions, the antennas are designed for 360-deg. aximuth and 0-to-90-deg. elevation coverage.

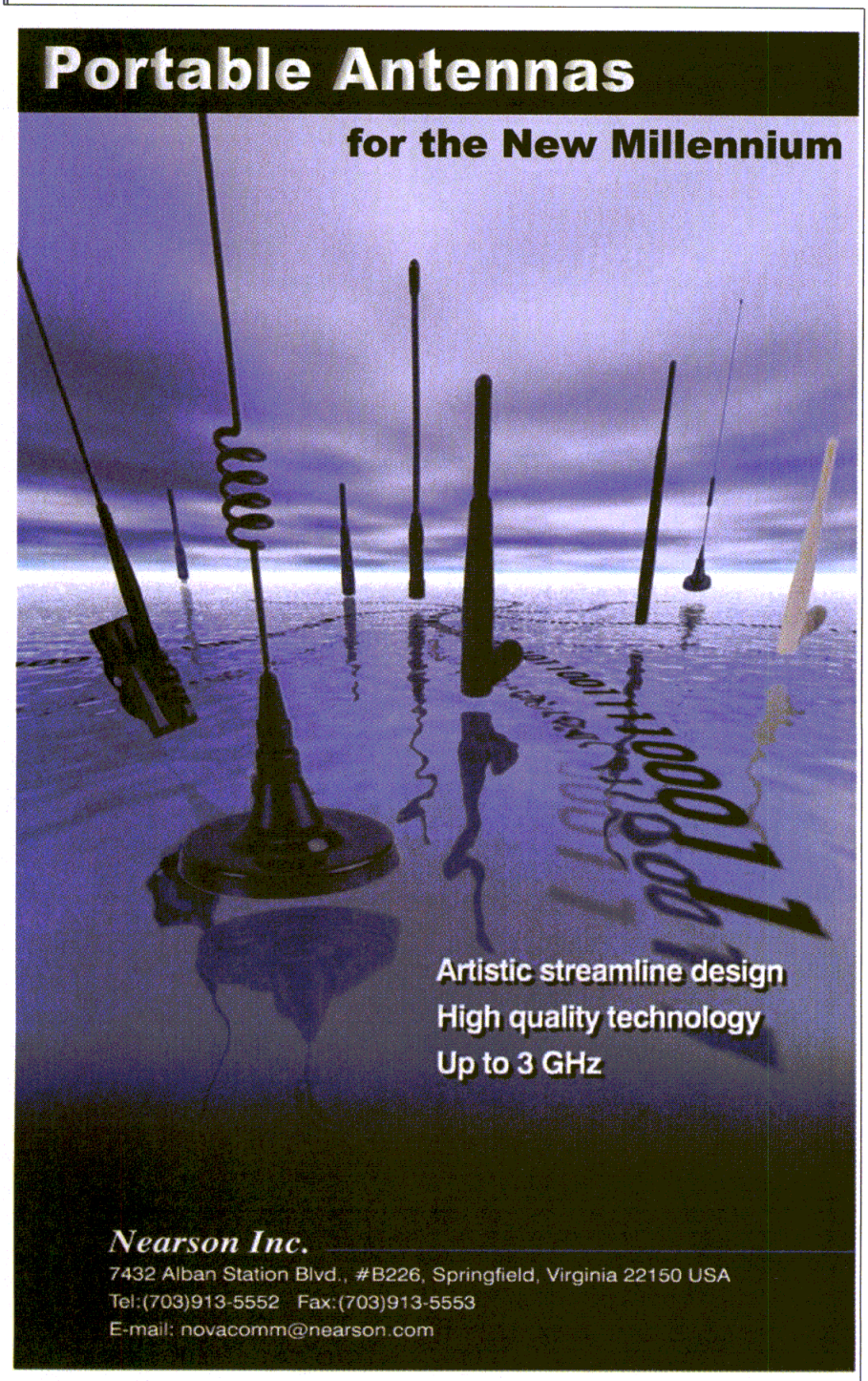
Pacific Wireless (Aptos, CA) announced a directional antenna for 2.4-GHz digital wireless communications systems. The design incorporates a parabolic grid reflector (model PMANT-19) with a passive-feed dipole (model PMPF-1). The antenna achieves 19-dBi gain from 2.3 to 2.5 GHz with a return loss of 12 dB. It measures 16.7 × 9.4 × 23.3 in. (42.4 × 23.88 × 59.95 cm) and weighs less than 4 lbs. (1.8 kg).

Seavey Engineering Associates (Pembroke, MA, http://www.seavey-antenna.com) recently introduced the model 9911-800, a dual-polarized broadband diagonal-horn antenna designed for low sidelobe performance with constant beamwidths. The antenna, which operates from 4.5 to 7.0 GHz and is suited for radiometry and data-gathering applications, features 21-dBil nominal gain using dual-orthogonal linear polarization. It features 25-dB minimum cross-polarization with 20-dB minimum isolation. The nominal half-power beamwidth is 14 deg while the VSWR is nominally 1.70:1. The antenna has an overall length of 32 in. (81.28 cm) and weighs approximately 17.5 lbs. (7.875 kg).

The company also announced the model 9887-800, a trihedral corner reflector for use from 150 to 300 MHz (Fig. 4). It can be rotated for horizontal or vertical polarization, and can handle power levels up to 325 kW at 0.01-percent duty factor. The minimum gain is 8 dBi while the maximum VSWR is 2.0:1.

Hirschmann (Pine Brook, NJ, http://www.hirschmann-usa.com) has developed the model E-6003 disk-shaped antenna (Fig. 5) for Cell-Net Data Systems as part of an automated meter-reading system for the utility industry. The antenna, which operates in the unlicensed band from 908 to 928 MHz, has a 5-in. (12.7-cm) diameter with a low 1.5-in. (3.81-cm) profile. It employs vertical polarization with monopole-equivalent horizontal and vertical beamwidths. The antenna's unique radiator system is enclosed in a weather-resistance radome.

The firm has also developed a fixed-mount GPS antenna that enables the capture and deployment of synchronized time signals for timing applications in the cellular, paging, and PCS applications. The antenna permits GPS signals to be used at fixed locations for establishing precise time/space reference measure-



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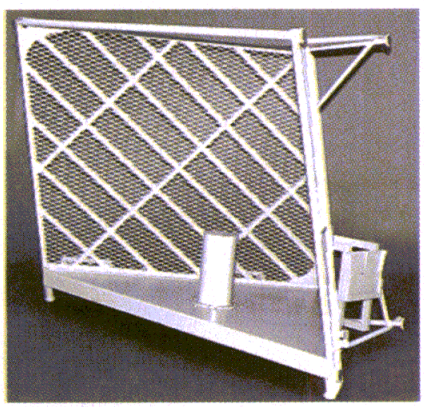
ments. The antenna can also be used to geographically locate signal sources, such as when a wireless caller dials the 911 emergency number and forgets to give his or her location. The caller can then be located by analysis and interpretation of the time measurement data received at two cell sites. The antenna, which operates at 1575 ± 4 MHz, features 3.5-dB gain. It also integrates an LNA with 26-dB gain.

Integral Technologies, Inc. (Vancouver, British Columbia, Canada, http://www.itechfin.com) has its contrawound-toroidal-helical antenna (CTHA) under evaluation by a number of military, government, and law-enforcement customers. The CTHA is designed for use in vehicle, ship, and aircraft communications applications from 800 to 2400 MHz. It is realized as a printed-circuit pattern that occupies an area of less than 2 × 2 in. (5.08 × 5.08 cm) at 800 MHz and less than 1 × 1 in. (2.54 × 2.54 cm) at 1500 MHz (for GPS applications). The printed-circuit configuration eliminates the need for a standard dipole antenna, reducing cost and enhancing reliability. Range improvements of as much as 300 percent compared to monopole antennas.

When is an antenna more like a transmission line? When it is a radiating cable. Several firms offer this type of transmission-line antenna for communications within buildings and in tunnels. The RADIAX RCT series



4. Model 9887-800 is a trihedral-corner-reflector antenna designed for high-power applications from 150 to 300 MHz. (Photograph courtesy of Seavey Engineering Associates, Inc., Pembroke, MA.)

5. The E-6003 antenna suits remote data collection in the utility industry. It operates in the ISM band. (Photograph courtesy of Hirschmann, Pine Brook, NJ.)

of radiating cables from Andrew Corp. (Orland Park, IL, http://www.andrew.com), for example, compete against traditional radiating cables which are broadband but lossy. In contrast, the RCT cables are optimized for improved efficiency over specific frequency bands. Version LT cables are optimized for use from 70 to 134 MHz and 174 to 500 MHz while version TC cables are optimized for use in cellular and GSM bands. The optimized cables provide 10-to-15-dB performance improvements over broadband radiating cables, requiring less additional amplification for in-building and tunnel communications systems.

In order to cut antenna-measurement time, Orbit/FR, Inc. (Horsham, PA, http://www.orbitfr.com) recently announced its ASCENT antenna measurement system, which it claims can drastically cut the time needed to perform near-field measurements on a wide range of cellular and wireless antennas. The system offers measurement accuracy of ±0.3 dB at 800 MHz and measurement time of less than 90 s at 2 GHz.

With time to market a strong motivation, EMS Wireless (Norcross, GA, http://www.emswireless.com), a division of EMS Technologies, Inc., recently announced the opening of an enhanced production facility located in Norcross, GA. The new ISO-9001 certified facility now has three production lines, and fully integrates the assembly, quality assurance, and packing processes, increasing the firm's capacity to manufacture base-station panel antennas. ••

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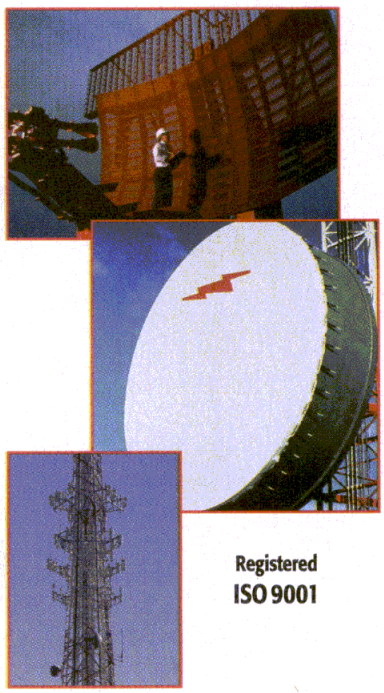

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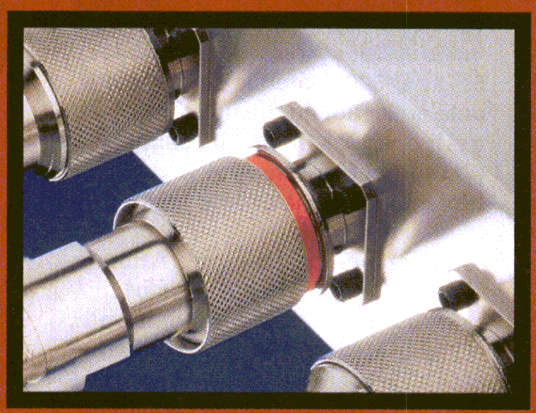

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The potential for Micropower Impulse Radar (MIR) technology is great, but troubling questions of Department of Energy policy cast a shadow.

Macro Applications Seen For Micro Radar

GENE HEFTMAN
Senior Editor

IT has been hailed as one of the most important technological developments of the past decade, capable of supplying unique products to industries as diverse as automotive, security, medical, infrastructure, and the military. Invented in 1993 at the US Department of Energy (DOE) Lawrence Livermore National Laboratories (LLNL) by Tom McEwan (there are serious legal disputes to that claim), Micropower Impulse Radar (MIR) represents a major departure from conventional radar technology. An MIR is different in that it emits very-short, sharp pulses of approximately 0.2 ns (rise times are measured in picoseconds) as opposed to the longer bursts of a pulsed continuous-wave (CW) radar. A CW radar sends out narrowband electromagnetic (EM) energy in the microwave frequency range whereas an MIR is an ultra-wideband (UWB) system. Since the MIR pulse is of very short duration, little power and relatively simple circuitry can be used to generate the signal. This means that an MIR can be much smaller than conventional radar and can be built with standard electronic components, making MIRs very inexpensive compared with other types of radar-based products.

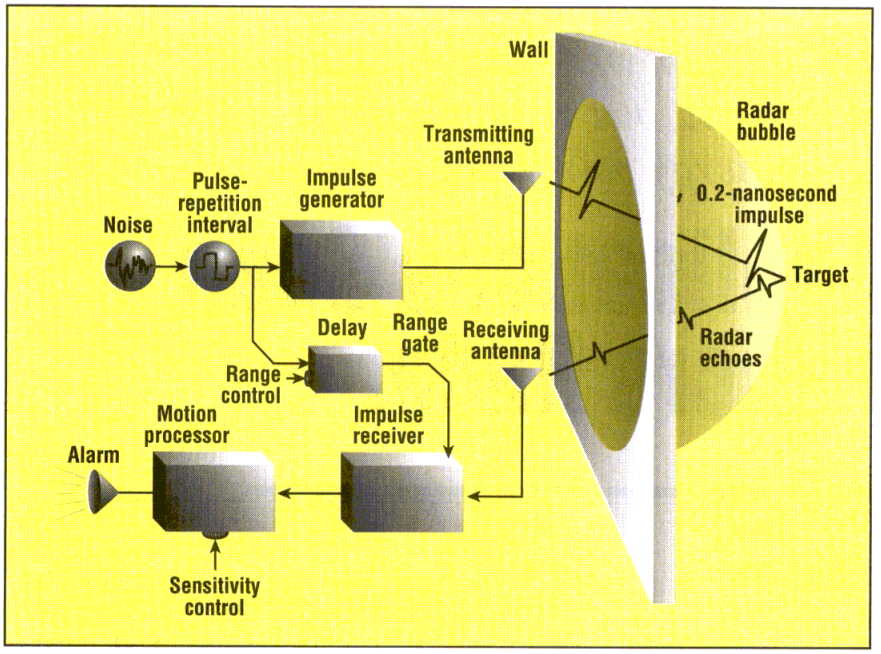
While MIR is an outstanding technical achievement, its potential benefits to society may be delayed or lost due to what the Committee on Science of the House of Representatives last year called a "technology transfer at a DOE laboratory that may not have lived up to the high professional and ethical standards expected of a federally funded facility." In a report that resulted from congressional hearings in 1998-99, the Committee was highly critical of a number of LLNL practices surrounding the MIR program. These included improper patenting and technology transfer practices, misleading Congress and the public about the genesis of early UWB work, and misleading licensees of MIR that resulted in financial damage to many of the companies. The Committee recommended aggressive investigation of all the allegations brought before it, but many of the issues, such as patenting procedures in federally funded laboratories, may take years to be ironed out.

Meanwhile, the benefits of MIR are not in dispute. The positive aspect of a UWB pulse is that its broad bandwidth can generate a large amount of information about reflected objects. On the negative side, the lower energy content of a short pulse is that makes it is more difficult to measure the return signal. To overcome this problem, a large number of pulses is transmitted and the return signals are averaged. Averaging thousands of pulses reduces the signal-to-noise ratio (SNR) and improves the receiver sensitivity.

Figure 1 shows the architecture of an MIR motion sensor as invented and described by McEwan. A key element is the range gate in Fig. 1, which opens the receiver briefly after a fixed delay time to sample and average the received echoes in a narrow-time window. The time delay can be adjusted to obtain target information at different distances. The sensor shown in the figure can be implemented on a printed-circuit board (PCB) that measures approximately 1.0×1.5 in. (2.54×3.8 cm). Small size together with inexpensive electronic components opens up a wealth of applications for MIR.

Recent comments in the media about the prospects for MIR have fairly gushed with enthusiasm. One report states that the average consumer will soon own a personal radar system. With the wide range of products that can be implemented with the technology, it is possible that many consumers will have some kind of MIR-enabled device in the near future.

One of the frequently mentioned products of the future is the "electronic dipstick," a fluid-level measuring device that has no moving parts and is impervious to all types of moisture and corrosion. It operates on the principle of sending a signal down a

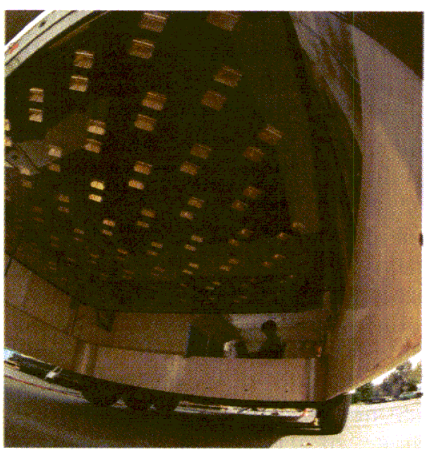


1. Short, fast pulses transmitted by an MIR cover a very broad bandwidth and are returned from a target to the system's receiving antenna. A range gate is used to sample the return echoes and prepare them for averaging to provide information about the target.

metal wire and measuring the transit time of the reflected EM pulses from the top of the dipstick to the fluid surface. The dipstick is said to be very accurate—to within 0.1 percent of its length.

Another consumer product, for home-security purposes, is the MIR motion sensor or concealable burglar alarm. It is based upon the emission of low-voltage impulses and can operate on small batteries. The average emission power is below 1 μW and is spread over a range of several gigahertz, which virtually eliminates interference and interception. And the emissions can penetrate foliage and non-conductive barriers such as gypsum, wood, and concrete walls. Applications besides intrusion detection and security include detecting objects in darkness, curb and blind-spot sensing for vehicles, and traffic-control embedded sensors. The motion sensor comes in a small package—approximately 4 × 3 in. (10 × 7.6 cm)—and is low in cost.

At the other end of the MIR spectrum is the High Performance Electromagnetic Roadway Mapping and Evaluation System (HERMES), an MIR-based sensing system that is mounted on a trailer (Fig. 2). Developed as a research project at LLNL under the sponsorship of the Federal Highway Administration, HERMES is a bridge inspector that is intended to diagnose the problems of deteriorating bridge decks faster. It is also cheaper and more effective than any other method. HERMES is pulled by a vehicle at traffic speeds over a bridge deck to collect information about the roadway's subsurface. Its sensors gather information at a depth of 30 cm (approximately 12 in.) into the concrete, and an on-board computer system processes the data into three-dimensional (3D) images that pinpoint problems in the concrete and provide engineers with quantative information about deterioration in the bridge deck.



2. HERMES consists of 64 MIRs that are mounted in an 8 × 8 array beneath a trailer that is pulled by a vehicle. Its function is to check the integrity of bridge decks by mapping defects in the concrete using radar data supplied by an MIR module.

HERMES uses 64 high-speed impulse radar modules that are mounted beneath the trailer to send out very fast, short pulses in the 1-to-5-GHz range. As the transmitted energy propagates through the bridge deck, concrete defects alter the return signal because of the change in dielectric constant. The changes are recorded by the on-board computer and the 3D map of the deck (very similar to a medical CAT scan) is created through sophisticated imaging software. One of the engineering challenges of the project was to fit all 64 radar modules underneath a standard-width trailer, and have them separated sufficiently to avoid causing interference in transmitting and receiving pulses among the modules. The problem was solved by mounting them in 8 rows of 8 modules each and separated by insulating foam to minimize crosstalk.

Projections for the HERMES bridge inspector indicate that it can save approximately $100 million in inspection costs—an important economy in the $20 billion annual cost of highway infrastructure maintenance and repair.

While HERMES and other government-sponsored MIR research projects bring important technological advances, it is the private sector that must ultimately convert this information into innovative products that solve real-world problems. The jury is still out, but the Committee on Science report indicates that many companies that engaged in the MIR technology transfer program with LLNL have come away severely disillusioned with their attempts to partner with a US Government laboratory. Unless the situation is redressed, future government funding for national research and private initiative to develop new technology could be damaged. ••

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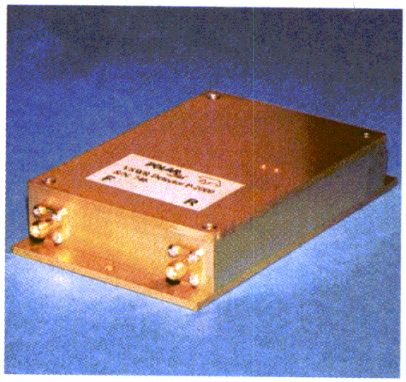

The model P-2000 VSWR detector operates from 25 MHz to 2 GHz and boasts 50-dB forward-power dynamic range and 20-dB return-loss range. The detector is ideal for use with a high-directivity directional coupler to monitor the antenna in mobile-phone base stations or other transmitters that exhibit large variations in output power. The P-2000's output voltage is proportional to the measured return loss in decibels. Its wide bandwidth also makes it well-suited for spread-spectrum applications. It operates from a single +12-VDC power supply, and is available with an optional analog-to-digital converter (ADC) for digitization of the monitor signal. **Polar Lab, Lucernevej 100, DK-2610 Rodovre, Denmark; +45 3672 7245, FAX: +45 3672 7245, Internet: http://www.polarlab.com.**

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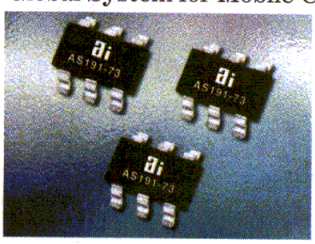

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The model 150W1000 broadband test amplifier provides at least 150 W of RF power from 80 to 1000 MHz for testing to International Engineering Consortium (IEC) requirements. The amplifier's manufacturer claims that the 150W1000 can provide the extra margin of power needed when testing to IEC 10 V/m electromagnetic- compatibility (EMC) requirements at three meters from 80 to 90 MHz. It is also suitable for automotive bulk-current-injection (BCI) applications. Its digital control panel provides local and remote control of the amplifier. The control panel displays forward and reflected power, control status, and internal amplifier status. Features include gain control, internal/external automatic level control (ALC), pulse input, and RF output-level protection. **Amplifier Research, 160 School House Rd., Souderton, PA 18964; (215) 723-8181, Internet: http://www.aramps.com.**

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Waveguide switch serves satcom

The model WS8189 is the first in a family of waveguide switches that operate in the Ku- and direct-broadcast-satellite (DBS) bands and are designed for use in ground-based satellite-communications (satcom) applications. The switch includes manual override and Form-3C inhibit contacts as standard features. Switch time at room temperature is typically 60 ms and 100 ms under worst-case conditions. Actuator current is typically 0.7 A at room temperature, and 0.9 A under worst-case conditions. The design of

the model WS8189 is an economized version of the company's miltary design. The entire switch, including the actuator mechanism, is weatherized to withstand harsh environmental conditions. **Sivers Lab AB, Box 1134, SE-164 22 Kista, Torshamnsgatan 9, Sweden; +46 8 477 68 00, FAX: +46 8 751 00 19, Internet: http://www.silversla b.se.**

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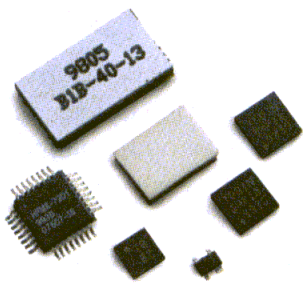

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What Is Ahead For Microsoft?

Strike two on Microsoft Corp. in its anti-trust battle with the Justice Department which occurred on April 3. The government once again found that the software giant maintained monopoly power in the software market through anticompetitive means. The first strike came five months earlier on November 5, 1999 at the conclusion of the original trial, which also declared the company a monopoly (see "Feds Curb Microsoft's Power," *Microwaves & RF*, December 1999, p. 45). This time, however, the consequences for Microsoft could be severe, as a separate proceeding gets underway to decide what penalties will be imposed.

The latest blow to the company's fortunes was a result of a breakdown in settlement talks that had been taking place for four months and, despite 20 drafts of a settlement offer, had produced no compromises. At the center of the case is the Justice Department's assertion that Microsoft had hurt consumers by stifling competition in the software marketplace, particularly at the expense of the Netscape Internet browser. In his conclusions regarding the case, Thomas Penfield Jackson, the original trial judge, found that Microsoft had unlawfully tied its Web browser to its operating system in violation of the Sherman Antitrust Act. This allowed the company to gain market share for its browser and put Netscape at a disadvantage.

At the conclusion of the trial, there was a conciliatory atmosphere regarding penalties that the government might seek. It was believed that a settlement could be reached that would neither break up Microsoft nor inflict severe financial damage, but would encourage more competition in the software industry. But the failure to settle puts the company in much greater peril.

Although Justice Department officials will not reveal what steps they would recommend to Judge Jackson, the judge himself is apparently in a retaliatory mood as evidenced by a section of his decision, which reads, "Microsoft's anticompetitive actions trammeled the competitive process through which the computer software industry generally stimulates innovation to the optimum benefit of consumers."

Still, the penalties to be imposed are far from certain. The number of options before Jackson range from breaking up Microsoft at one extreme to ordering the company to change its business tactics at the other. One of the milder possibilities is to bar the company from using price to punish clients who deal with competitors. But if the government asks for a remedy that breaks up Microsoft, it could change the shape of the industry and be called strike three for the most successful company of the computer age. ●●

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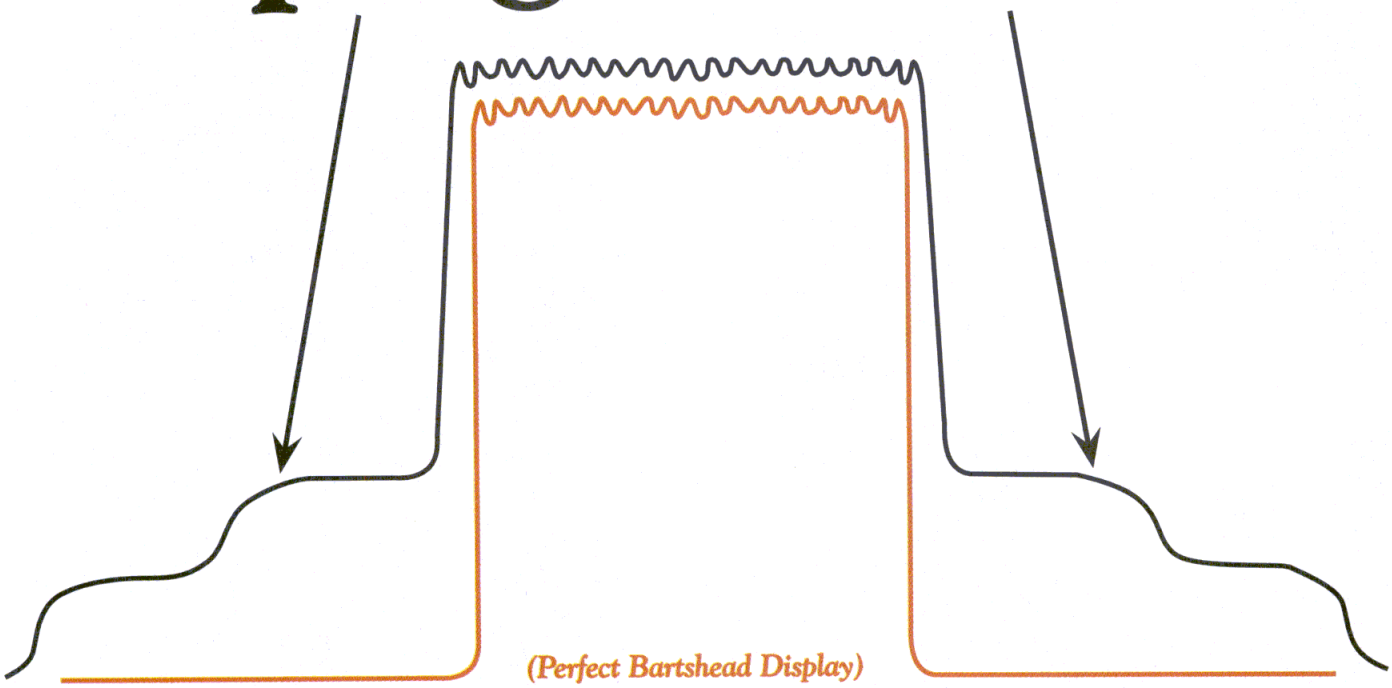

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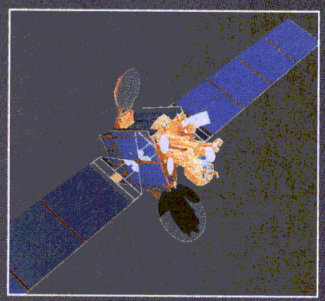

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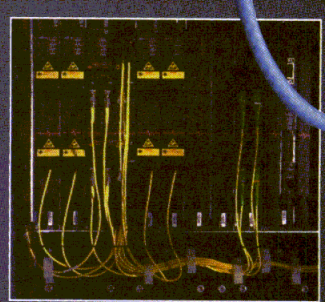

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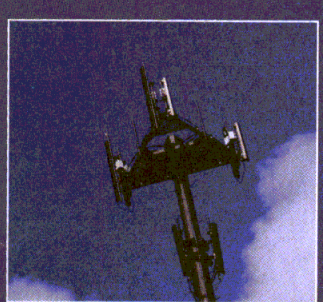

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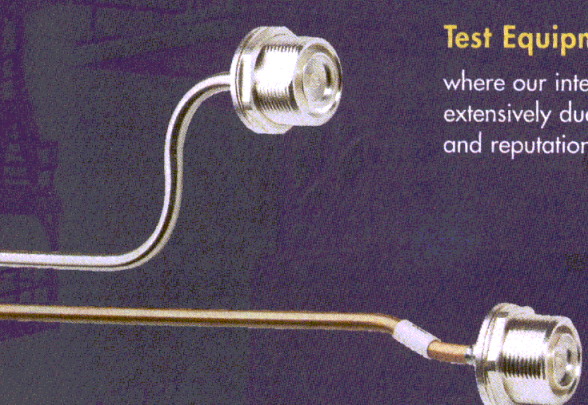

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Contracts

Motorola, Inc. and Powerwave Technologies, Inc.—Have entered into a five-year exclusive agreement that designates Powerwave as the sole aftermarket supplier of high-power, multicarrier power amplifiers (PAs) for Motorola's cellular base stations.

Alpha Industries—Has entered into a contract with Ericsson to develop multichip modules for leading-edge digital wireless telephone standards, including EDGE and other future, highly linear applications.

Microwave Power Devices, Inc.—Announced that it has been awarded a $13.3 million contract from Unique Broadband Systems, Inc. for amplifiers to be used in a terrestrial repeater system.

ITT Industries' Avionics Division—Was awarded a $7.5 million contract from the US Army Special Operations Command in Fort Eustis, VA for the production of 31 high-power remote transmitters (HPRTs). The HPRT was designed and developed by ITT Industries to provide Army Special Operations helicopters with electronic protection required against advanced RF threats that are encountered in today's conflict zones.

ANADIGICS—Announced that it has received an order for more than 10 million RF switches and has begun production shipments to one of the largest US-based manufacturers of wireless handsets. The AWS5504S14 and AWS5506S14 single-supply +3-VDC RF switches will be used in an 800-MHz digital code-division-multiple-access (CDMA)/Advanced Mobile Phone Service (AMPS) dual-mode phone that is experiencing increased cellular market penetration, particularly in the US.

Harris Corp.—Announced an agreement with SCIENTECH, Inc., a worldwide technical and consulting company that serves the needs of energy utility and related markets, for the purchase of equipment and services for the Illinois State Toll Highway Authority Project. The agreement, which is valued at approximately $3 million, requires the use of Harris Megastar microwave radios to support communication throughout the system.

Surrey Satellite Technology Ltd. (SSTL)—Has been awarded a $120,000, 100-day study contract following the successful selection of the company's 350-kg mini-satellite platform for NASA's RAPID II catalog. The Magnetospheric Multiscale (MMS) Mission Study has been initiated to investigate the range of suitable concepts for a five-spacecraft mission to investigate the Earth's magnetosphere. The mission will open new frontiers toward understanding the Earth's magnetic-field interaction with solar radiation.

Rohde & Schwarz—Has been awarded a contract from Finnish operator Digita totaling 2.5 million DM for the supply of TV transmitters. In all, five 5-kW digital-video-broadcast (DVB) transmitters of the new series NV7000 will be supplied.

RF Micro Devices, Inc.—Announced that it has received production orders for Global System for Mobile Communications (GSM) and digital-communications-system (DCS)/personal-communications-services (PCS) power amplifiers (PAs) from Groupe Sagem, a French manufacturer of GSM mobile phones and GSM mobile handsets.

Fresh Starts

Ultima Communication and Silicon Wave—Announced a cross-license agreement of protocol software for the implementation of Bluetooth™ wireless technology. The agreement allows both companies to deliver complete software solutions to original equipment manufacturers (OEMs) that are looking to add Bluetooth functionality to their products while minimizing the development cycle.

RangeStar Wireless—Announced that it is partnering with the Plastic Moldings Corp. (PMC) of Cincinnati, OH, a supplier of solutions using advanced plastic technologies, to manufacture RangeStar's cutting-edge embedded (internal) antennas for original equipment manufacturers (OEMs). Through this alliance, PMC will manufacture RangeStar's Global Positioning System (GPS) embedded antennas for integration into wireless devices.

Aethercomm, Inc.—Has moved to a new facility in San Marcos, CA. Aethercomm designs, develops, and manufactures sophisticated RF, microwave, and millimeter-wave amplifiers, transmitters, receivers, upconverters, and downconverters.

QUAKE Wireless, Inc.—Has entered into a strategic alliance with Samsung Electro-Mechanics Co. Ltd., a producer of electonics parts and components. Under the terms of the agreement, Samsung will manufacture QUAKE's satellite-communications products that enable companies to track, monitor, and provide two-way data-communications capabilities between fixed and mobile assets worldwide.

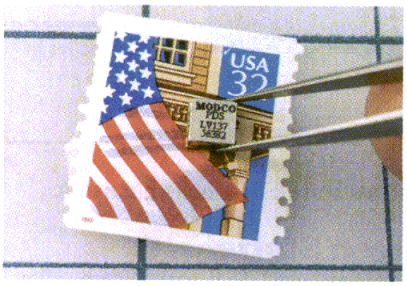
G.T. Microwave—Announced the appointment of nine foreign representatives in Belgium, Denmark, Finland, Germany, Greece, India, Korea, The Netherlands, Spain, and Turkey. This doubles G.T.'s existing foreign coverage with local, sales, and technical liaison personnel in each country.

AbsoluteValue Software, Inc.—Has collaborated with Intersil to develop wireless local-area-network (WLAN) driver software that enables Linux operating systems to communicate with Intersil's PRISM-based WLAN cards.

Cardinal Components, Inc.—Announced an alliance with Anderson ECD (Electronic Component Distribution) in San Jose, CA. Anderson is one of the few multi-franchised distributors in the US exclusively specializing in crystals and oscillators.

Peregrine Semiconductor and Symbol Technologies—Announced that they have signed an agreement to jointly develop complete solutions for the Bluetooth wireless high-speed personal-area-networking (PAN) market. Peregrine's UTSi™ (Ultra-Thin Silicon) complementary-metal-oxide-semiconductor (CMOS) technology and expertise in RF integrated-circuit (RF IC) development will be combined with Symbol's baseband processing, software, and systems expertise to develop complete Bluetooth subsystems.

Cidera, Inc.—Announced that BellAtlantic's Data Solutions Group has signed on as the largest customer for Cidera's Usenet News Service. The deal upgrades BellAtlantic's Usenet News capacity and improves delivery.



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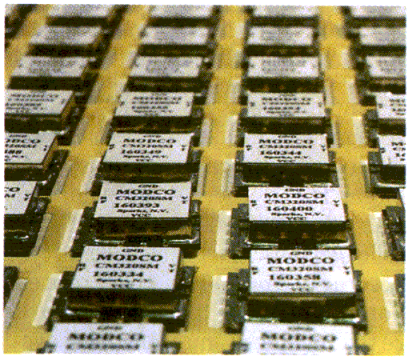

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Superconductor Technologies, Inc.—Martin S. McDermut to chief financial officer and vice president of finance and administration; formerly vice president of finance and administration and chief financial officer at International Remote Imaging Systems, Inc.

AMT—Randy Jenkins to vice president of corporate sales; formerly director of new business development. Also, Scott Rumery to director of engineering; formerly employed at Hewlett-Packard Co.

Del-Tron Precision, Inc.—Tom Tomsky to sales engineer for California and the Far East; formerly responsible for microscopy sales and applications of two- and three-dimensional (2D and 3D) optically-based automated geometric metrology systems in the Western US and Far East.

TOMSKY

STAFFORD

K&L Microwave, Inc.—Jon Stafford to director/general manager of the Bridlington, UK facility; formerly business development director and a business unit manager at Nera Telecommunications.

Mitel Semiconductor—Pierre Nadeau to vice president of the communications business; formerly general manager at LSI Logic Europe.

Quad Systems Corp.—Ross Fraser to general manager and director of Quad Europe Ltd. (QEL); formerly head of marketing and sales for automated optical and X-ray inspection operations in the United Kingdom for Macrotron.

American Microsystems, Inc. (AMI)—H. Gene Patterson to CEO and president; formerly CEO and president of Symbios Logic.

King Safety Products—Robert S. Clubb to director of national accounts, irrigation; formerly manager of national accounts with Columbia Energy. Also, Glenn M. Hoffman to plant manager; formerly plant and manufacturing manager at Souvenir Group, Inc. And, Carl G. Oberle to director of sales and marketing for the Consumer Division; formerly senior account supervisor for the Medical Marketing Division of the Atkinson Group, Inc.

TEGAM, Inc.—Roger Zimmerman to director of operations; formerly employed as operations manager at Bird Electronics.

Tellabs—Joan E. Ryan to chief financial officer; formerly senior vice president and chief financial officer of Alliant Foodservice, Inc.

IFR Systems, Inc.—Mitch Stone to vice president of sales for the Americas region; formerly vice president of sales and marketing at Datum, Inc.

Champion Technologies, Inc.—Raj Alluri to applications engineer; formerly prime reliability engineer. Also, Dick Thompson to sales manager; formerly employed at M-tron Industries.

Interconnect Devices, Inc. (IDI)—Thad Sketers to western regional sales manager; formerly product specialist.

SAL, Inc.—Dan Fleming to president and CEO; formerly vice president of customer support and technology at SVGL.

FLEMING

SHILLITO

TriPoint Global Communications, Inc.—William L. (Bill) Shillito to executive vice president of Prodelin, a TriPoint Global Co.; formerly Prodelin's vice president of operations and engineering.

CTS Corp.—Catherine C. Edwards to director of human resources for the resistor/electrocomponents business; formerly human resources and organization development manager at Avery Dennison Corp.

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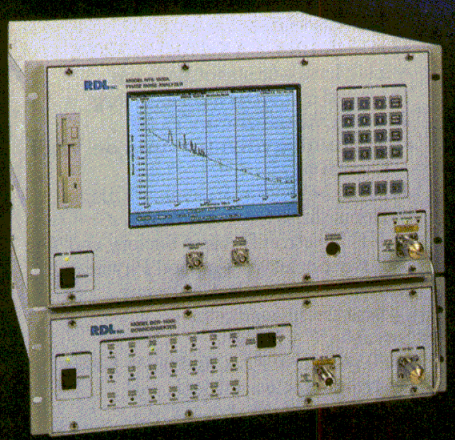

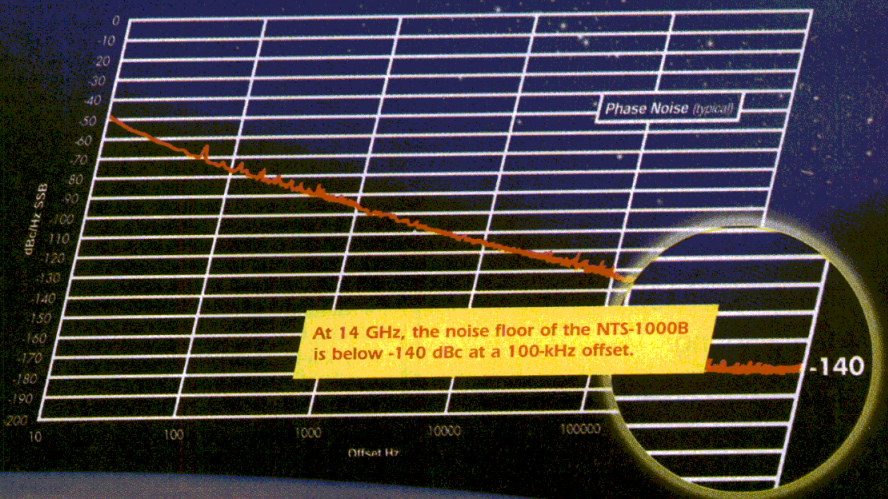

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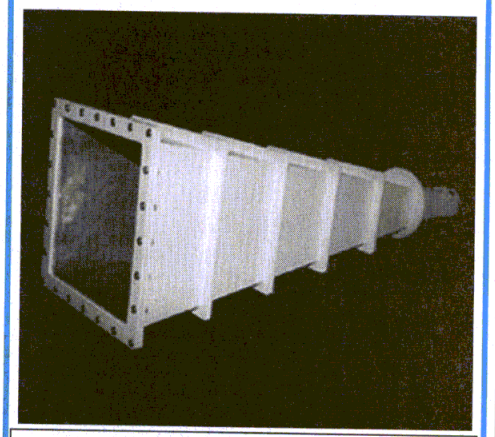

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CIRCLE NO. 387

New parameters improve ground-wave measurement model

Ground-wave losses of radio signals in urban areas can affect the performance of radio systems, but available methods of measuring ground-wave propagation can be inaccurate. A new ground-wave model for medium frequencies (MF) and high frequencies (HF) has been developed by Luo Lichun of the North Electronic Equipment Institute (Beijing, People's Republic of China) that improves on a well-known method (the Norton model) by adding three so-called building-complex parameters and a height-gain factor. The building-complex parameters represent the impact of the buildings on the ground wave in urban areas since buildings are an important cause of propagation losses. Two types of predictions can be made with the new model—coarse and fine. A coarse prediction only gives information about a general propagation area. A fine prediction takes into account the building parameters and other factors along the receiving path, a method not available with other models. Tests conducted in the Beijing urban area using the fine model confirm the improvement in ground-wave measurement. See "A New MF and HF Ground-Wave Model for Urban Areas," *IEEE Antennas & Propagation Magazine*, February 2000, Vol. 42, No. 1, p. 21.

Design microwave oscillators using Z-parameters

The complexities of single-frequency, bipolar, microwave oscillator design are best handled with an analytic approach in which optimum values of feedback elements, load impedance, and maximum power output are expressed in terms of the transistor's Z-parameters says Andrey V. Grebennikov of the VLSI Circuit & Test Department, Institute of Microelectronics (Singapore). He shows the development of a simplified large-signal model for a bipolar transistor based on its small-signal Z-parameters and DC characteristics. In addition to single-frequency oscillators, the paper describes the design of common-base, bipolar voltage-controlled oscillators (VCOs) based on an optimum bipolar oscillator. The methods developed can be integrated into RF circuit simulators. See "An Analytic Method of Microwave Bipolar Single-Frequency and Voltage-Controlled Oscillator Design," *International Journal of RF and Microwave Computer-Aided Engineering*, September 1999, Vol. 9, No. 5, p. 403.

Improving the noise performance of HEMTs

High-electron-mobility transistors (HEMTs) could play a greater role in front-end amplifiers used in millimeter-wave receivers if their low-noise performance could match that of non-transistor, state-of-the-art devices. HEMTs operate very effectively at cryogenic temperatures, but improvements can be made. Two processing methods to boost performance (gain and cutoff frequency) are proposed by Richard Lai of TRW, Inc., who developed the techniques for indium (In)-based HEMTs. One approach involves increasing the In content of the In gallium-arsenide (GaAs) channel layer, but in a non-uniform (graded) manner so as to grow the thickest possible channel that will yield the highest electron-transport properties. The second technique involves reducing gate lengths which leads to lower noise, higher cutoff frequencies, and higher gain. See "Two Approaches to Improvement of InGaAs/InAlAs/InP HEMTs," *NASA Tech Briefs*, February 2000, Vol. 24, No. 2, p. 42.

CMOS DAC delivers spectrally pure performance

Communications systems that use complementary-metal-oxide-semiconductor (CMOS) digital-to-analog converters (DACs) require devices having a high degree of spectral purity. That is, there should be a very low level of spurious frequency components in the DAC's output signal. A high-resolution (14-b), high-speed (100-MS/s) DAC designed specially for its spectral performance is reported by Alex R. Bugeja and Bang-Sup Song of the Department of Electrical and Computer Engineering at the University of Illinois (Urbana-Champaign, Urbana, IL) and Patrick L. Rakers and Steven F. Gillig of Motorola, Inc. (Schaumberg, IL). The architecture is called a return-to-zero (RZ) type using current-source switching implemented differentially, a technique, used in all high-speed, high-resolution DACs. This stage tracks the DAC output at midcycle to avoid switching transients, and then resets it to ground. Isolation of the differential circuits from each other preserve the settled position of the DAC when the track phase begins. In tests of spurious-free dynamic range (SFDR) performance against other high-performance DACs, the RZ architecture is superior at high frequencies. See "A 14-B, 100 MS/s CMOS DAC Designed for Spectral Performance," *IEEE Journal of Solid-State Circuits*, Vol. 34, No. 12, December 1999, p. 1719.

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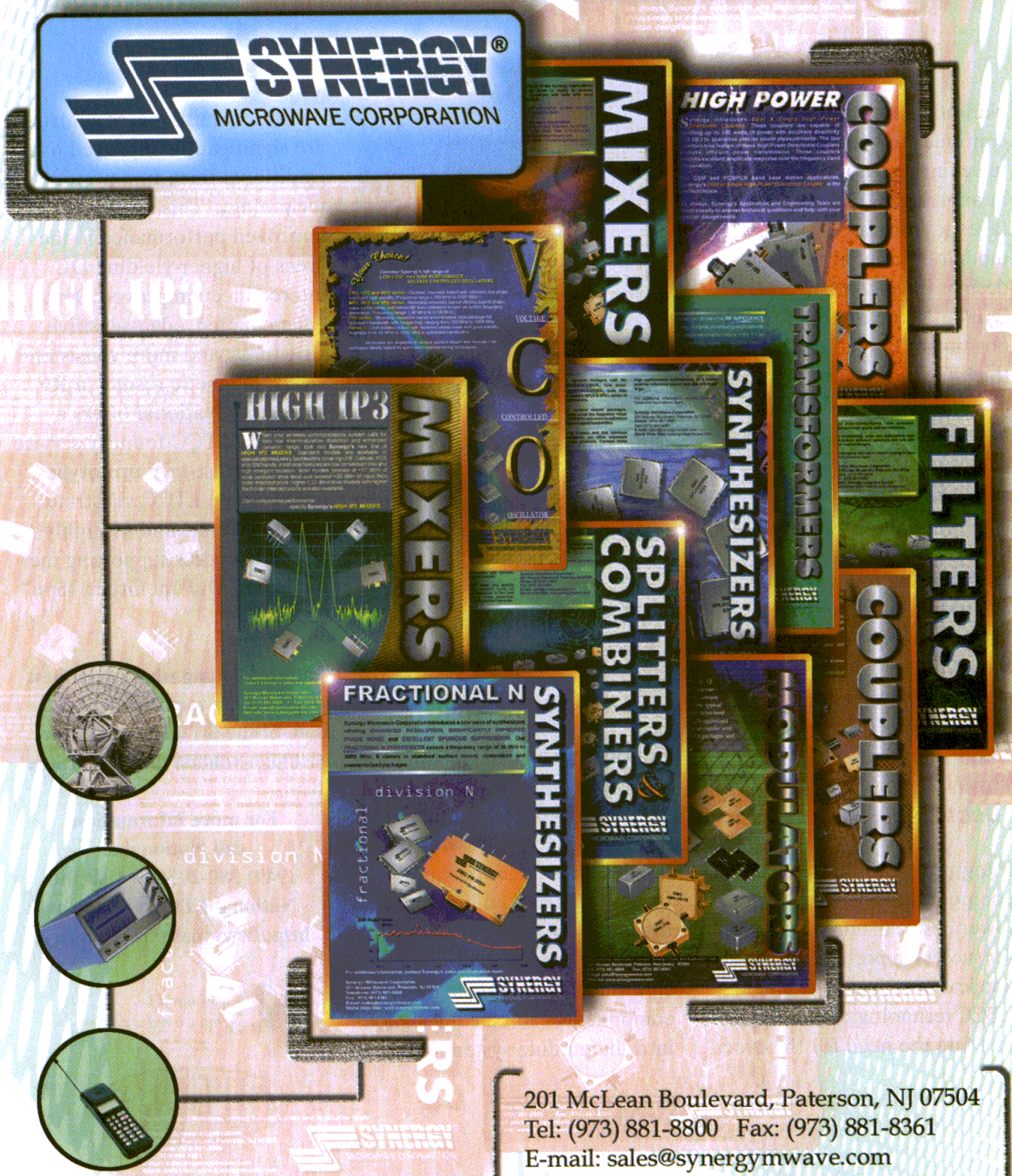

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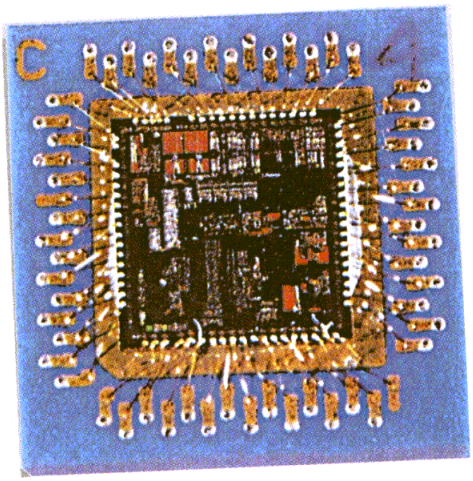

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CIRCLE NO. 219

Bluetooth Integration Poses Challenges For Developers

In-house knowledge and design skills may not be enough to overcome the obstacles for implementing a total Bluetooth solution.

Doug Rosener
Senior RF Engineer
RangeStar Wireless, Inc., 9565 Soquel Dr., Aptos, CA 95003; (831) 661-4200, FAX: (831) 661-4202, e-mail: drosener@rangestar.com.

WHAT is the hottest new development in the wireless industry? Many industry insiders would say Bluetooth, a wireless connection that, when embedded into consumer electronic devices, will make cables obsolete and allow devices to communicate and transfer information as they never have before. For example, Bluetooth will allow individuals to surf the Internet wirelessly on a laptop using their cellular phones—while those phones sit in the individuals' pockets.

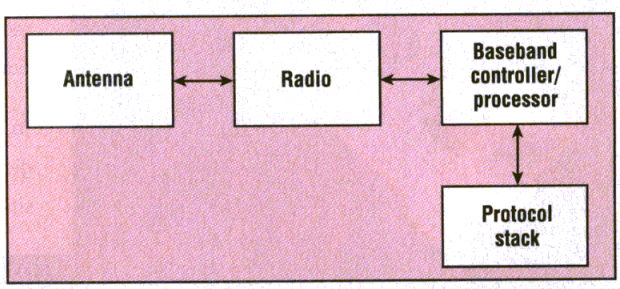
The potential applications of Bluetooth are virtually unlimited and its adoption is expected to explode. In fact, according to industry experts, Bluetooth will be the fastest-growing industry standard ever. One industry analyst firm, the Gartner Group, predicts that more than 75 percent of new wireless handsets shipped in 2004 will be Bluetooth enabled—and that is only one type of Bluetooth-enabled device.

An unprecedented 1600 members have joined the Bluetooth Special Interest Group (SIG), including key industry players such as Motorola, Microsoft, Intel, IBM, Lucent Technologies, Nokia, Toshiba, 3Com, and Ericsson. These leading industry manufacturers have attempted to speed the development of Bluetooth devices and applications.

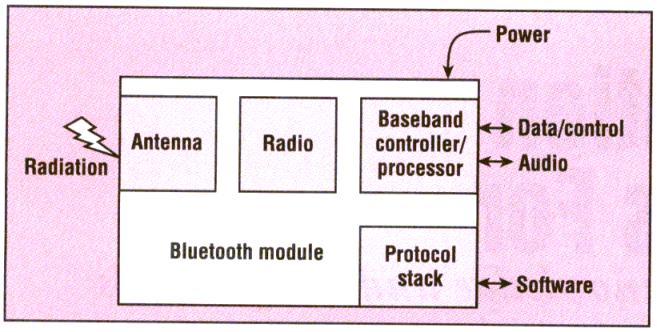
Ironically, the majority of the Bluetooth SIG members (more than 75 percent) do not have previous RF experience, which is likely to result in delays in bringing Bluetooth-enabled devices to market. These companies will need assistance—and information—to help them reduce development time, lower costs, and integrate cutting-edge antenna technology into Bluetooth applications. There are many hurdles to enable Bluetooth in new or existing products. To offer a total solution, an inordinate amount of information and skills must be mastered. Bluetooth adopters will be confronted with many engineering subtleties such as antenna placement and electromagnetic-compatibility (EMC) considerations. While in-house knowledge is desirable, high cost and time-to-market considerations may make it an impractical alternative.

BLUETOOTH BASICS

Technically, Bluetooth is a specification for low-cost, short-range wireless



1. A basic Bluetooth system consists of four functional blocks—the antenna, a radio, a baseband controller/ processor, and a software-protocol stack.



2. The key interfaces for integrating a Bluetooth module to the equipment where it resides are shown here. These include the data/control, power, audio, software stack and antenna, and ensuring that antenna radiation falls within acceptable safety limits.

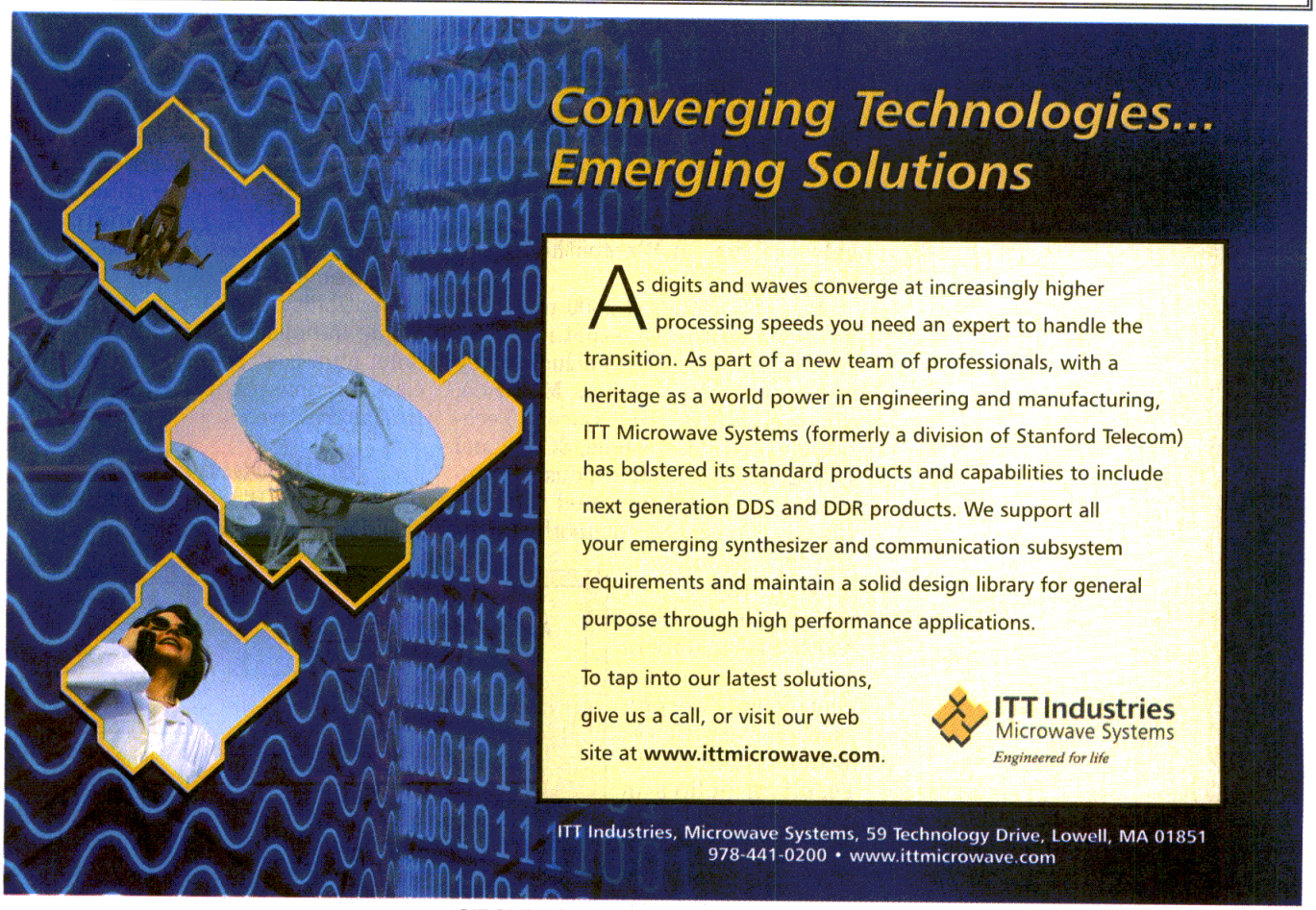
Table 1: Integration interfaces

Hardware	
Data/control	Called the Host Interface in Bluetooth documentation. It can be either UART or USB.
Audio	Analog or digital signals in possibly both directions
Power	Supplies power to radio and processors(s)
Radiation	Both receive and transmit
Hardware	Host Controller Interface (HCI)

devices that encompasses physical and software layers. It uses well-understood frequency-hopped multiple access (FHMA), time-division-duplexed (TDD) radios in the industrial-scientific-medical (ISM) band (2.4 to 2.5 GHz). It not only defines new protocol interfaces for wireless serial ports and audio devices, but also integrates with existing protocol layers such as transmission-control protocol/Internet protocol (TCP/IP) and Infrared Data Association (IrDA).

Basic Bluetooth functional components consist of an antenna, a radio, a baseband controller/ processor, and a software-protocol stack. The baseband controller/processor (referred to in Bluetooth specs as the host controller) may be integrated with the radio in some cases. Its purpose is to process incoming data and audio streams for the radio, as well as to instruct the radio in setting up links. The antenna is shown separated from the radio because it has independent properties that affect the product as a whole (Fig. 1).

Collectively, these components form a Bluetooth module. Several



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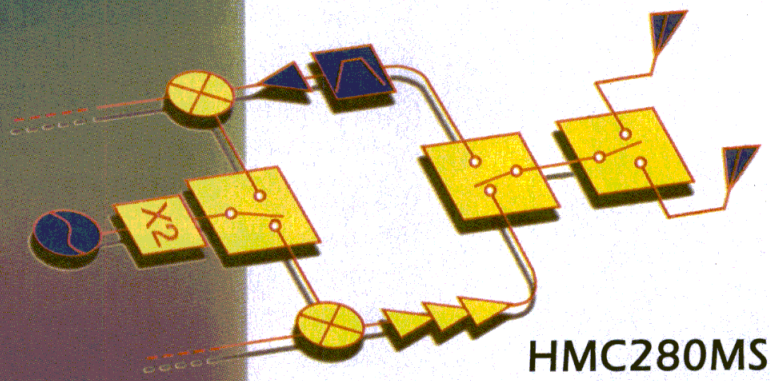

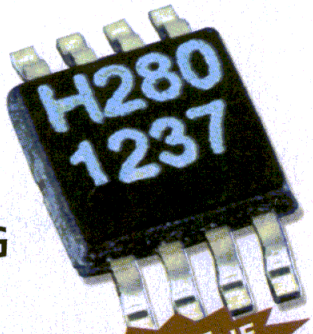

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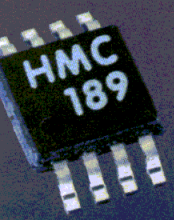

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QPSK MODULATOR	HMC267QS16G	5 to 6	30 dBc Carrier Suppression, 0/+5V
T/R SWITCH	HMC224MS8	5 to 6	1.2 dB IL, 30 dB Isolation, +33 dBm P1dB
T/R SWITCH	HMC223MS8	4.5 to 6	1.2 dB IL, 26 dB Iso, +33 dBm P1dB
SP4T SWITCH	HMC222C12	DC to 6	+5V, 2 Bit TTL, 2.3 dB IL, 27 dB Isolation
FREQUENCY MULTIPLIER	HMC189MS8	4 to 8 Fout	X2 Fo, +10 dBm Drive
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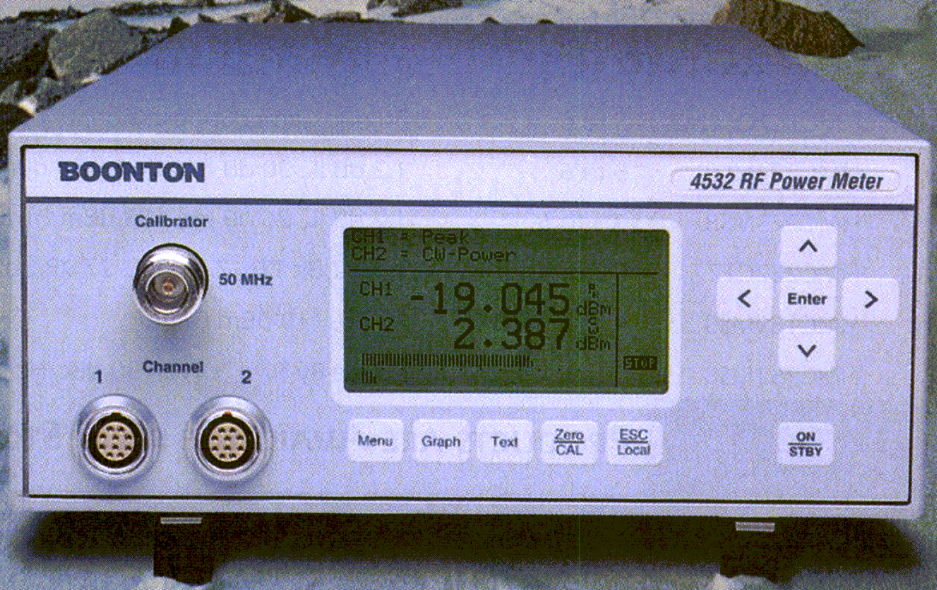

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CIRCLE NO. 349

manufacturers offer single-, double-, or triple-chip hardware modules. Some companies offer complete solutions (hardware and software), while others offer only software. Of course, there is also the option to design and develop one or more components.

Assuming that the Bluetooth components are designed or available, the components must be interfaced/integrated into the product. Table 1 shows the main interfaces to integration. In some cases, not all interfaces will be used. The host-computer interface (HCI) is at the bottom of the software-protocol stack and is meant to be hardware independent. It drives the baseband controller and the link manager.

INTEGRATION ISSUES

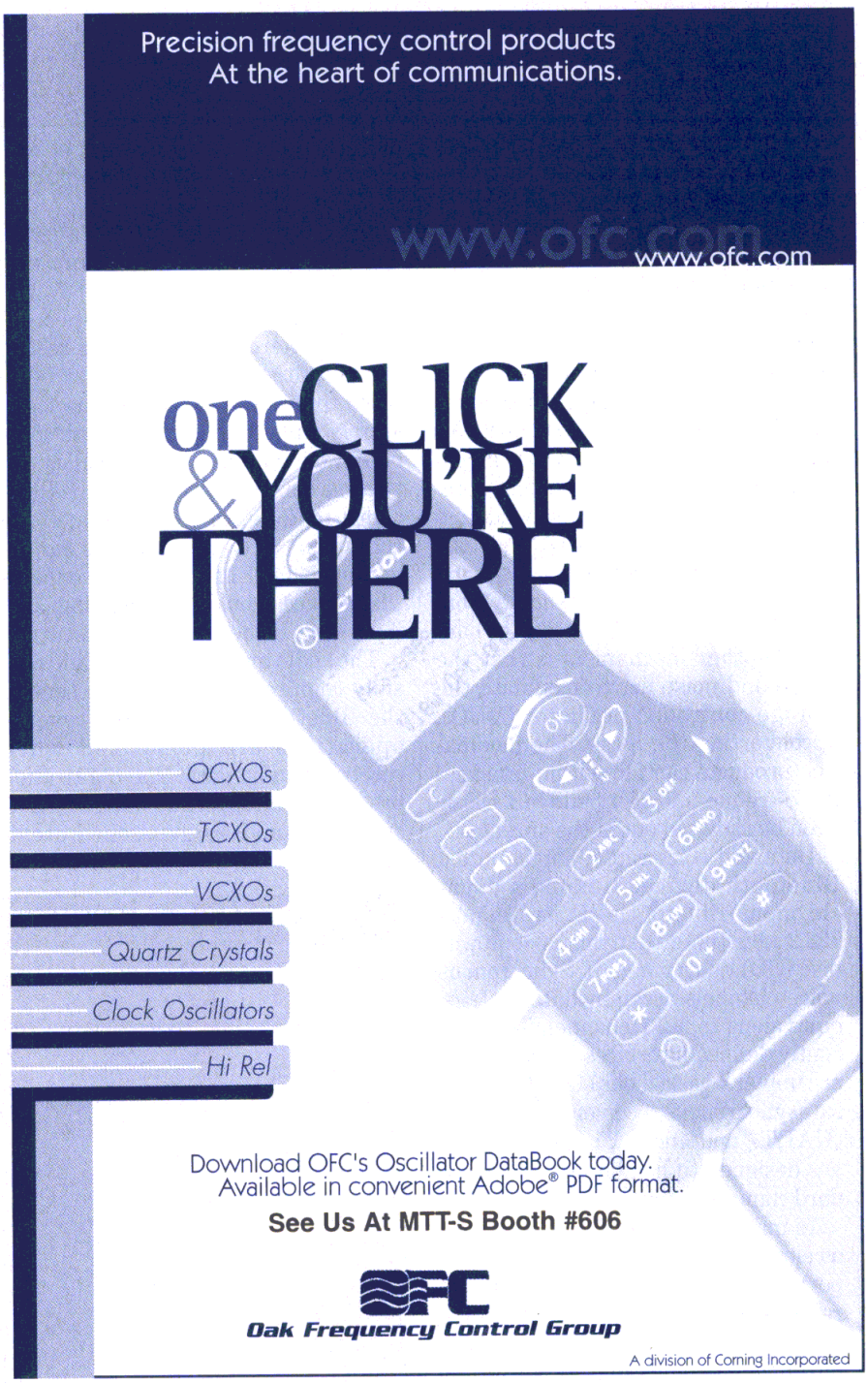
There are many challenges to integrate Bluetooth components into a design. Figure 2 illustrates, in block diagram form, the various interfaces listed in Table 1. These are the most pressing and difficult issues of bringing a Bluetooth product to market.

The principal Bluetooth specifications number more than 1500 pages.[1] The integrator must have experience dealing with this format of information to ensure compatibility to other devices. Bluetooth involves significant RF design, which is not a core competency of most companies. Even if a company has RF capabilities, it may find itself developing components that are already available from other sources.

Due to the highly specialized Bluetooth requirements, most product manufacturers will opt to buy a module. Even then, a litany of integration problems remains. In particular, if the antenna is not part of a purchased module, it can create numerous problems.

The advantage gained by using an outside source for a Bluetooth component—while trying to develop the other components in-house—is often offset by the lack of in-house knowledge. This is especially crucial for integration, as interaction often occurs between components, and changes must be made in a timely manner. Due to intellectual-property protection, many manufacturers are hesitant to offer complete support for components. In addition, as the component becomes dated, knowledgeable engineers are moved to new projects, and it becomes more difficult to get solutions to difficult problems.

The TDD transmitter in a Bluetooth transceiver sends power in 625-μs-to-3.125-ms bursts with repetitive spacing as small as 625 μs (occurring when master and slave are exchanging packets). These pulses result in signals with fundamental frequencies up to 800 Hz (1/2(625 μs). Audio is present in many Bluetooth applications and, unfortunately, 800 Hz is in the audio range. The spectrum of the generated noise hinges on the pulse rise time and can be quite broadband.[2] This broadband noise can cause logic upsets and reduction in receiver sensitivity. Radio modules generally have common data, audio, and power lines. If the data, audio, and power grounds are



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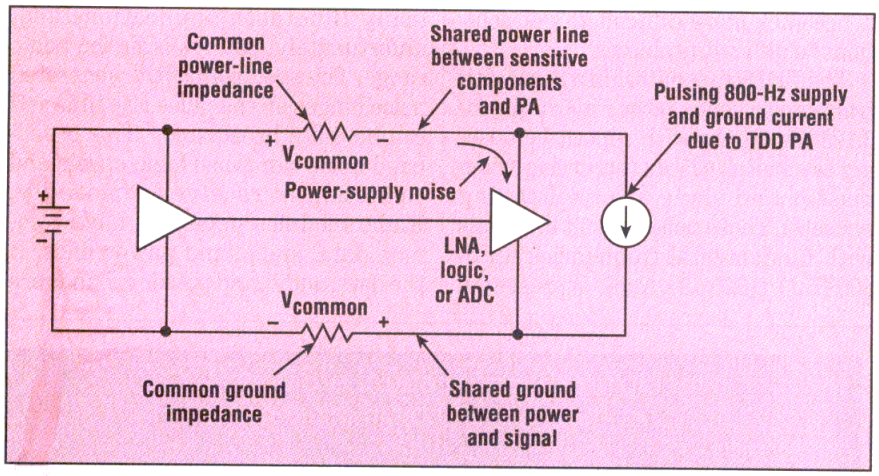

3. Proper power-supply routing and grounding are necessary to prevent noise from entering the signal path of a Bluetooth module. Noise can cause logic upsets and interfere with transmission and reception of signals.

shared, data and/or audio quality can be compromised (Fig. 3).

Power-supply ripple and noise [conducted electromagnetic interference (EMI)] can raise the noise floor of the receiver due to leakage and upconversion. For example, a 50-kHz switcher can produce harmonics into the megahertz range. Due to imperfect common-mode rejection of baseband analog-to-digital converters (ADCs), the switcher noise leaks into the baseband signal and is indistinguishable from the 1-MHz bandwidth Bluetooth channel. Another mechanism is that the ripple and noise act to modulate the voltage-controlled oscillator (VCO), upconverting the noise. The modulation products can fall within intermediate-frequency (IF) bandwidths, or even be radiated out of the antenna.

Data and voice can be inputs into a Bluetooth device. Maximum asynchronous Bluetooth data rates are 723.2 kb/s [note that this is compact-disc (CD)-quality audio]. Synchronous voice channels accommodate 64 kb/s, equivalent to a T1 channel. The currently defined interfaces for Bluetooth are universal serial bus (USB), universal asynchronous receiver /transmitter (UART), and audio. These interfaces are designed to handle industry-standard connections and formats.

In many cases, the data interface is as simple as hooking up the standard ports. However, many situations may require some design work. The logic between the Bluetooth module and the parent electronics may be different technology, requiring buffers and transceivers.

If the Bluetooth module uses a different power supply than the rest of the circuitry, protection must be in place to prevent exceeding the logic-supply rails. Otherwise a high-logic signal on the parent electronics can reverse bias the substrate on the Bluetooth module, or vice-versa. Long cable runs between modules can be pathways for EMI. The interconnects themselves should have filtering, and, if possible, be hot swappable.

The audio must be digitized and possibly compressed before it is sent to the Bluetooth module. The reverse operations are performed when receiving audio data. The digitization process requires an audio codec (ADC and DAC) and, as mentioned previously, these are susceptible to ground and supply noise.

Audio is passed typically as linear 16-b pulse-code modulation (PCM), 8-b compressed PCM (μ-law or A-law) or as continuously variable slope-delta-modulation (CVSD) data. PCM is the standard waveform coding of landline-telephone communications (ITU Recommendation G.711). CVSD is used in many voice-over IP (VOIP) applications. It takes a 16-b linear PCM (digitized) signal as input.

All waveform-coding techniques require setting an audio reference level. This is a common challenge in the cellular world, where switching between cell stations must not cause audible volume changes to the user. In Bluetooth integrations, this will be important as well.

ANTENNA PATTERNS

High-quality radio links require the desired link gain and the desired pattern of the radiation. Loss of gain reduces transmitter-coverage area and receiver sensitivity. Alternatively, if gain loss is compensated by increasing transmit power, then it reduces a Bluetooth device's operating time or power efficiency. In general, unanticipated gain changes of 0.5 dB or more are usually unacceptable.

Gain is affected by losses in the circuit, including mismatch loss (for example, the antenna does not look similar to 50 Ω). If the antenna is not placed close to the power amplifier (PA), losses can occur in the printed-circuit board (PCB) or transmission line and connector mismatch losses leading to the antenna. Standard FR-4

Table 2: Specific absorption-rate limits

Area	Reference	Maximum allowed (mW/g)	Frequency	Conditions
US, Canada, and Australia	FCC adopted the standard set by ANSI/IEEE C95.1-1992	1.6	100 kHz to 6 GHz	Any 1 g of tissue
Elsewhere (including EU and most other European nations)	International Commission on Non-Ionizing Radiation Protection (ICNIRP)—1996	2.0	10 MHz to 10 GHz	Averaged over the highest contiguous 10 g of tissue

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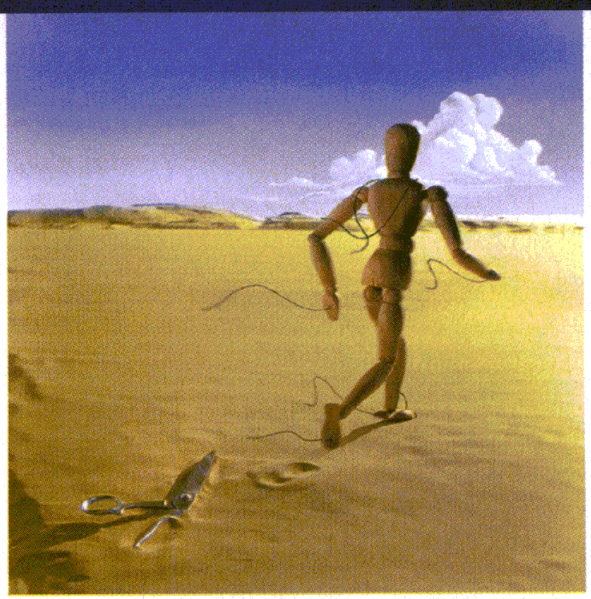

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PCB material can lose several decibels per foot in the ISM band. Quality coax can lose several tenths of decibels/feet, and connector mismatch can contribute several tenths of decibels as well.

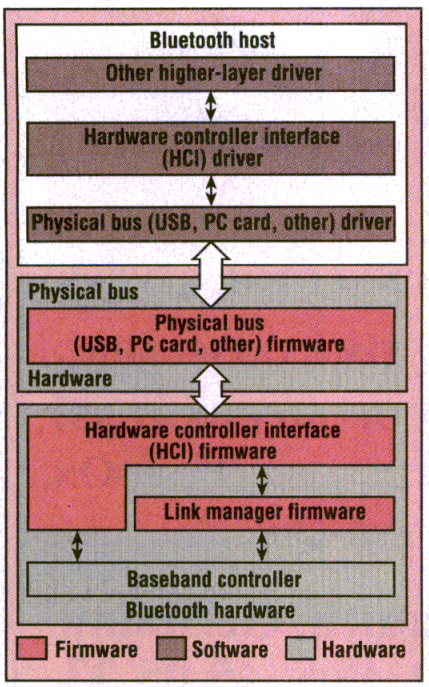
Gain also varies with the radiation pattern. Radiation patterns often have nulls, or areas where gain is low. For example, a desktop Bluetooth device should typically have a hemispherical circularly polarized pattern. This ensures that compatible devices can work at any angle and polarization above the desktop.

The antenna-radiation pattern (and, therefore, gain) is affected by the proximity of other objects. If the antenna is not placed appropriately, nulls can form so that communication is poor or impossible at certain orientations. Proximity can also disturb the antenna tuning. This can result in larger-than-anticipated mismatch losses with accompanying reduction in gain.

Users can modify radiation patterns by absorbing the RF energy. An important design requirement for devices in close proximity to people is specific absorption rate (SAR). This is the ratio between the minute amounts of RF power absorbed in an infinitesimal mass of tissue at a particular point.[3] Proper design and placement of the antenna ensures meeting regulatory requirements. Table 2 shows the current requirements.

The location of the Bluetooth module can effect the module operation due to EMI from neighboring data and switching power circuits. The power supply has already been discussed, but clock circuits and other application components can interfere.

The Bluetooth specifications (Vol. 1) define a protocol stack (Fig. 4). If the stack is written in-house, the main stack interface for integration is the HCI. This is well-defined, platform independent, and aids in porting the stack to other hardware.[4] Many solu-



4. One of the more complex elements of a Bluetooth design is designing the software stack. A successful design requires interfacing the software, firmware, and hardware elements with the main interface being the HCI.

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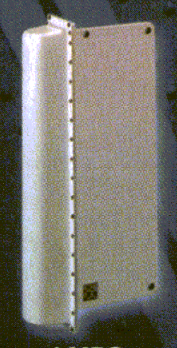

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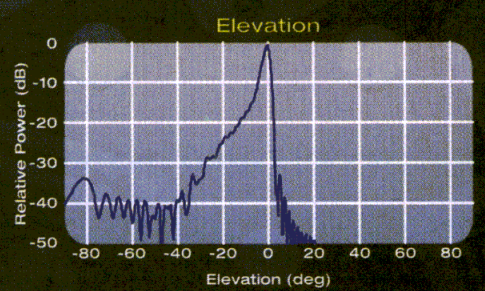

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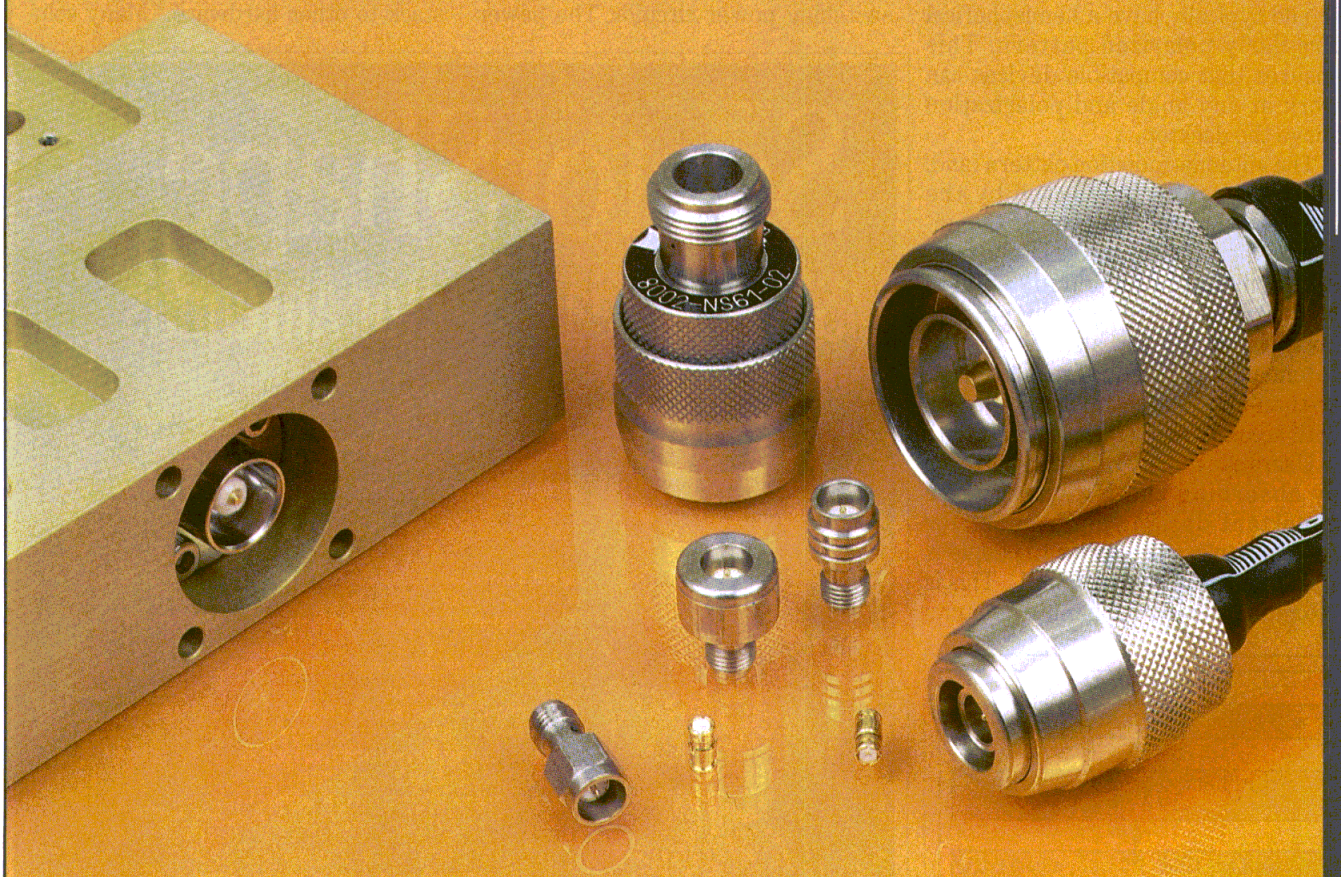

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tion suppliers are offering stacks with Java/application-programming interfaces (APIs). Ideally, these make the software integration task easy. However, problems can still parallel the integration of an electrical module. For example, if the stack is purchased, the integrator is at the mercy of the manufacturer when bugs are found. Do-it-yourselfers will find that a complete Bluetooth-compatible stack will require 10 person years or more to develop.

The Bluetooth specifications (Vol. 2) define profiles. Profiles describe the messaging and procedures followed by the various stack layers for specific applications. Applications and devices that adhere to these profiles ensure themselves of compatibility with similar devices. Stack testing during integration requires protocol analyzers. Commercial Bluetooth analyzers are only now arriving on the market. Typical integration testing would involve sending test patterns, monitoring with an over-the-air protocol sniffer, and monitoring at master-slave ends to ensure that messages are transmitted and responded to correctly.

RangeStar and their partners in the wireless industry are merging all aspects of Bluetooth adoption—antenna design, integration, software development, radio-module design, testing, and production—to offer a complete solution, available in their *Total Bluetooth Solution.*

Any product containing Bluetooth capabilities must meet certain requirements to ensure compatibility with other Bluetooth products. Of special interest to the integrator are EMC and Bluetooth. Compliance is verified through certification with various regulatory groups. Other regulations deal with government approval and safety regulations.

Meeting EMC or Bluetooth compliance requirements can affect product design. Therefore, those integrating components must be aware of design issues when compliance fails. Often, measures that improve EMC often disturb design goals. For instance, an embedded antenna requires gaps in enclosure shielding, which also may allow unintentional radiation to escape. Shielding (gasketing, stamped metal enclosures, or milled boxes) can impact mechanical layout in cost and footprint.

All products with the Bluetooth brand must undergo Bluetooth qualification. Compliance requirements are stated in the Bluetooth specification.[5] Product qualification testing is performed at a Bluetooth Qualification Test Facility (BQTF) and checked by a Bluetooth Qualification Body (BQB). The actual test requirements are still in draft form, but are due in mid-2000. Testing includes the profile behavior, the protocols themselves, and the radio link. ••

References

1. Bluetooth Core (Vol. 1) and Profile (Vol. 2) Specifications, v1.0B, December 1999, available at http://www.bluetooth.com/.
2. H.W. Johnson and M. Graham, *High Speed Digital Design*, Prentice Hall, 1993, pg. 3.
3. K. Fujimoto and J.R. James, *Mobile Antenna Systems Handbook*, Artech House, 1994, pg. 254.
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5. Bluetooth Core Specification, v1.0B, December 1999, Part I-2.

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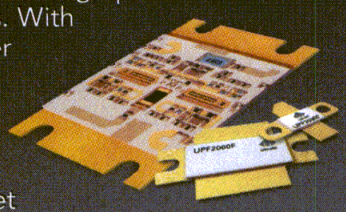

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Analyze IC Interactions At The System Level

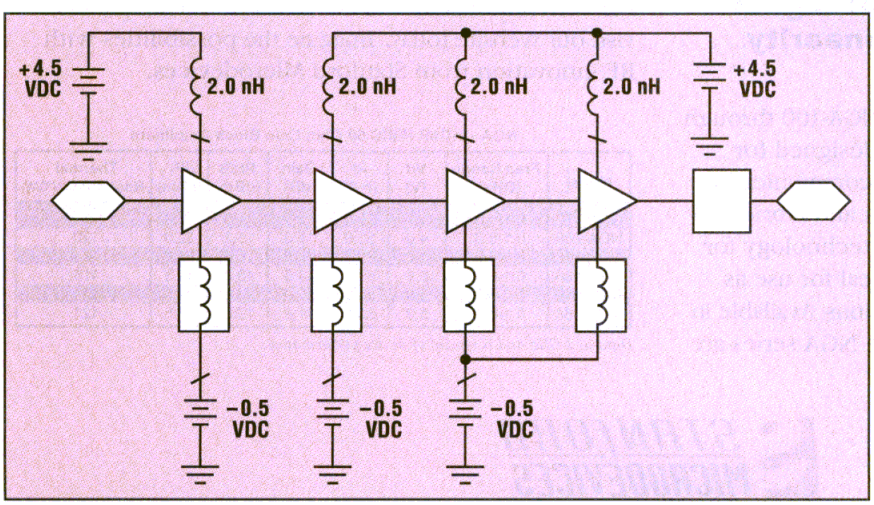
By using an LMDS system as an example, the interactions between a MMIC and the system as a whole will be studied.

Julio Perdomo
Senior Design Consultant
Agilent Technologies, Inc., 1400 Fountaingrove Pkwy., Santa Rosa, CA 95403-1799; (707) 577-1400, Internet: http://www.agilent.com.

SUCCESSFUL chip-set creation for commercial transmitters requires intensive design and simulation efforts. For the integrated-circuit (IC) designer, the process must account for the effect of parasitic circuit elements in active and passive components, as well as frequency dependence and layout effects. A successful design effort requires an ability to accurately simulate the interaction between the various components within the completed transceiver. This article will illustrate this process by examining the design and system evaluation of a millimeter-wave monolithic microwave IC (MMIC) for use in a local-multipoint-distribution-system (LMDS) broadband wireless application. The analysis includes the overall design process, with particular focus on the interaction between the MMIC and the system. Accordingly, system simulation becomes an indispensable tool to help determine specifications and verify performance of the MMIC in the system.

LMDS is a broadband wireless point-to-multipoint communications system used to provide digital two-way voice, data, Internet, and video services via a transmitter/receiver located at the customer premises. The system is defined to operate in Federal Communications Commission (FCC)-licensed spectrum from 28 to 32 GHz in the US. Currently, most LMDS systems and standards are based on either time-division-multiple-access (TDMA) or frequency-division-multiple-access (FDMA) approaches, while modulation methods employed are generally quadrature phase-shift keying (QPSK) or quadrature amplitude modulation (QAM). Modulation options are the same for TDMA and FDMA.

A typical customer-premises LMDS transmitter designed to use North American spectrum allocations might exhibit spectral regrowth (which represents maximum output power) of –32 dBc and output power at 1-dB compression ($P_{1\,dB}$) of +22 dBm. If linearity is one of the performance limitations, the best performance representation is spectral regrowth, and the output-power specification is only significant



1. This schematic diagram shows a four-stage MMIC amplifier designed for operation from 20 to 43 GHz in LMDS applications.

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NGA-186	0.1-6.0	4.1	50.0	12.5	14.6	32.9	120
NGA-286	0.1-6.0	4.0	50.0	15.5	15.2	32.0	120
NGA-386	0.1-5.0	4.0	35.0	20.8	14.5	25.8	144
NGA-486	0.1-6.0	5.0	80.0	14.8	18.3	39.5	118
NGA-586	0.1-6.0	5.0	80.0	19.9	18.9	39.6	121
NGA-686	0.1-6.0	5.9	80.0	11.8	19.5	37.5	121

Data at 1 GHz and is typical of device performance.

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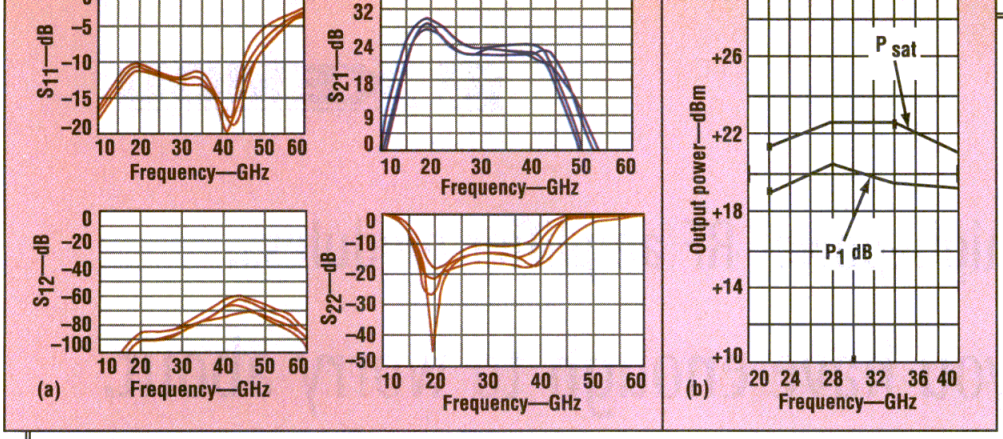

2. The plotted (a) small-signal and (b) power performance for the MMIC amplifier show that the small-signal results are less than 10 dB for S_{11} and S_{22}, more than 20 dB for S_{21}, with isolation of less than 50 dB.

if the required linearity specification is met. Finally, gain is typically stipulated to be greater than 30 dB.

Using well-established design methodologies, the MMIC amplifier for the example system is developed based on current modeling and simulation techniques. At this point, model accuracy is perhaps the most difficult variable to quantify, yet verifying models over the target operating range is a difficult and costly proposition, especially for active devices in a nonlinear setting. At the very least, limitations and operating ranges for models must be clearly established. For the LMDS example, the MMIC amplifier can be defined by small-signal gain of better than 21 dB, bandwidth of 25 to 32 GHz, third-order intercept point (IP3) of better than +28 dBm, saturated output power of better than +21 dBm, and output power at 1-dB compression of better than +19 dBm.

Regardless of the type of linear or nonlinear model used, the first verification is usually performed at the device level, where DC, S-parameters, noise figure, and load-pull characteristics can easily be measured. Critical nonlinear behaviors such as third-order intermodulation (TOI) product, harmonic content and intermodulation distortion (IMD) are usually affected by matching and biasing conditions, and therefore are not typically measured or specified at the MMIC level.

At the schematic level, the MMIC includes four stages of amplification and an output-matching structure (Fig. 1). For simulation purposes, the power supply, its parasitics, and connections (conforming to the layout of the chip) are also included at this level.

The output-matching circuit is designed to present an optimum load to a 450-μm field-effect transistor (FET) in the fourth stage for maximum output power. FETs in

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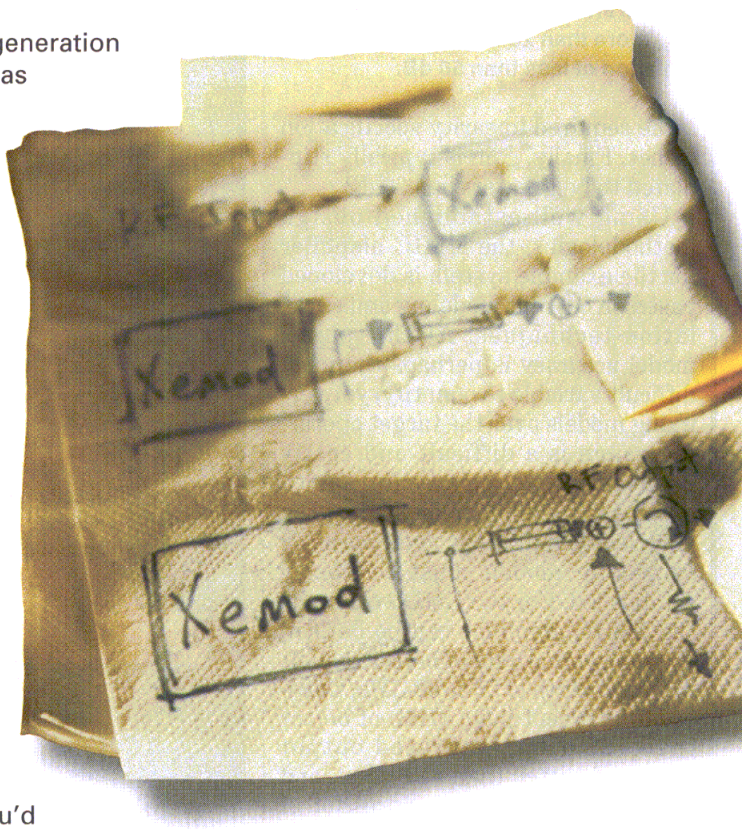

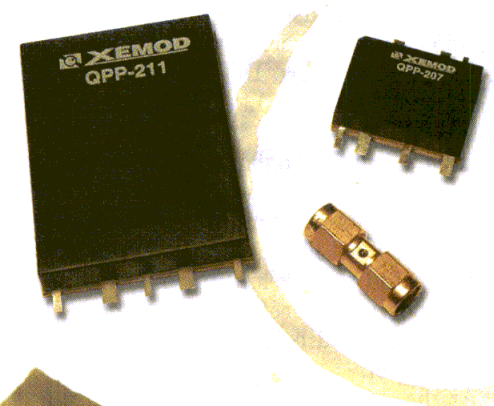

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the third and previous stages are sized to ensure that compression occurs only at the last stage, under Class A operation. This provides the best conditions for linearity.

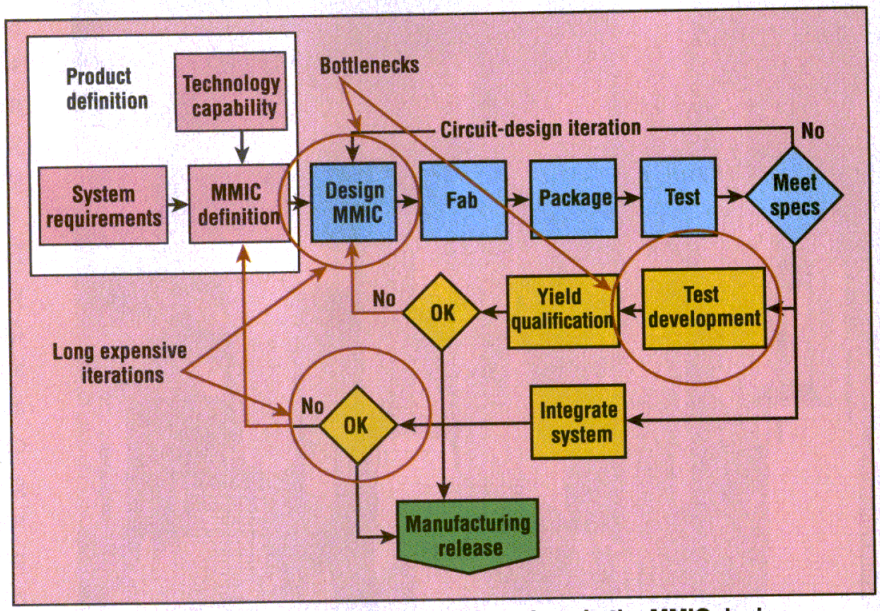
At the transistor level, the MMIC amplifier design was achieved using a custom design kit developed for the amplifier's intended pseudomorphic-high-electron-mobility-transistor (PHEMT) MMIC fabrication process. The custom elements in the library include models for transmission lines, backside via holes, FETs, and other passive components. Custom design kits, such as the one employed in this example, can be readily used with standard libraries supplied with computer-aided-engineering (CAE) tools such as the Advanced Design System (ADS) software suite from Agilent Technologies (Santa Rosa, CA).

Although performance specifications are provided in the data sheet for the MMIC in this example, no information is available regarding spectral re-growth or adjacent-channel-power-ratio (ACPR) performance. Consequently, the data sheet provides insufficient information to make it possible to determine system performance, and system simulation must be performed.

Figures 2a and 2b show the results of small-signal performance and power-performance simulations for our MMIC amplifier. Note that the 1-dB compression point at 31 GHz is +19 dBm. At this point, a determination must be made regarding how close the amplifier can be operated to this value before our system linearity requirements are exceeded.

Even if data about nonlinear behavior is available, correlating it with the other key nonlinear specifications for the system is difficult without simulation. For example, behavior such as TOI performance at specific operating frequencies versus output power is easily predicted and analyzed.

Once the product-definition stage is completed, the focus turns to improving the design process. In particular, optimizing the MMIC design flow inevitably centers on three primary loops—circuit design, yield qualification, and system integration

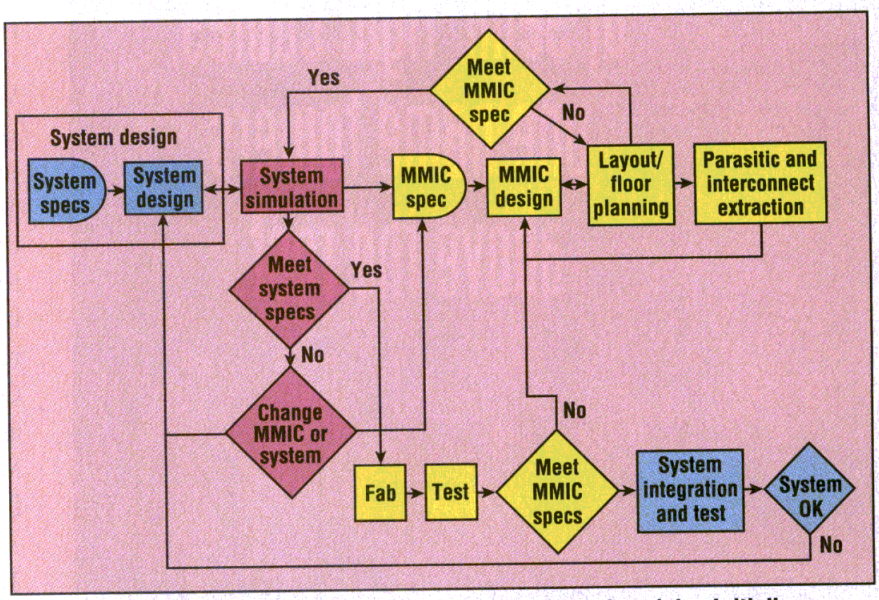


3. The longest and potentially most costly iterations in the MMIC design process are system integration and yield qualification.

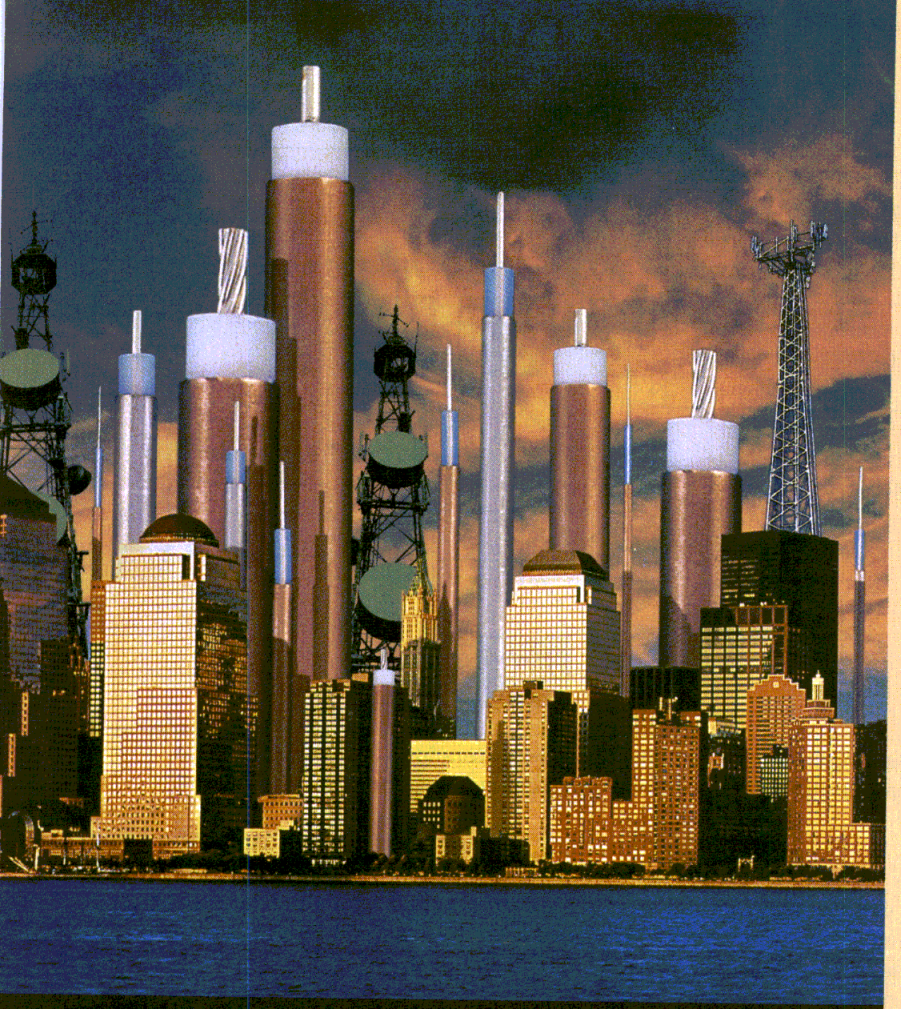
(Fig. 3). The primary goal is to minimize or eliminate time-consuming and very costly design iterations required when failures are revealed at the end of the development process or during field trials.

In a typical design hierarchy, the first level is comprised of the system or sub-system design (in our example, the LMDS transmitter), which gives rise to its own set of specifications. As a rule, designing the system or sub-system represents the most complex and certainly the most expensive step in the process. The individual modules (the second level in the design hierarchy) that make up the system or subsystem typically represent the lowest level of the design effort, and entail working with individual components.

A common problem source is the fact that both levels are designed independently and in parallel, and that each development effort is based an independent set of specifications



4. This design flow effectively reduces system-integration risks. Initially, a detailed system simulation is used to generate MMIC specifications from behavioral models.



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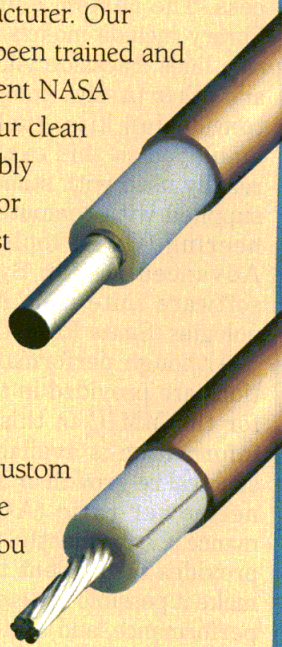

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and requirements. System engineers and circuit designers speak very different languages and lines of communication between the two camps can often be problematic, falling somewhere between "nonexistent" and "minor misunderstanding." While system requirements are expressed in terms such as bit-error rate (BER) and ACPR, circuit specifications focus on things like noise figure and third-order intercept. Specifications translated between the system and component levels are frequently inaccurate or incomplete, and communication of specification changes is frequently an afterthought.

System simulation—a fundamental tool that comes into play right at the start of subsystem and component development—represents an ideal instrument for effectively minimizing discord during the development phase. By simply making use of a shared simulation environment, communication between the various groups participating in the design is vastly improved. This communication enhancement translates into opportunities for early feedback from the circuit designers to the system architects, enabling compromises or changes to be considered at much earlier stages in the development process. Other potential benefits of using a common design tool include:

- Increased likelihood of reasonable specification requirements.
- More complete analysis and verification.
- Better partitioning.
- Enables simulation during system integration phase.
- Provides the ability to share models with customers early in the development.
- Improves the probability for optimizing trade-offs.

Figure 4 illustrates a modified approach to the traditional development process that is optimized to reduce costly and time-consuming design iterations. It should be noted that the success of this methodology is highly dependent on having accurate models. Also, the establishment of a "learning" organization becomes necessary. That is, if the fabricated MMIC does not meet the system

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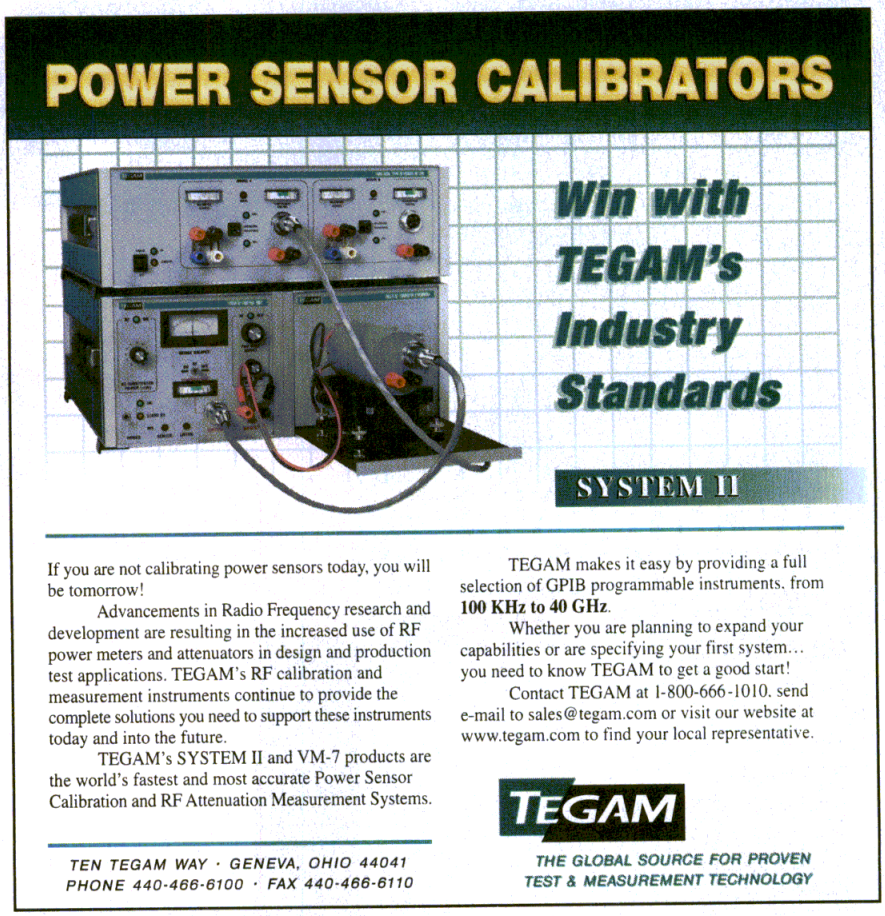

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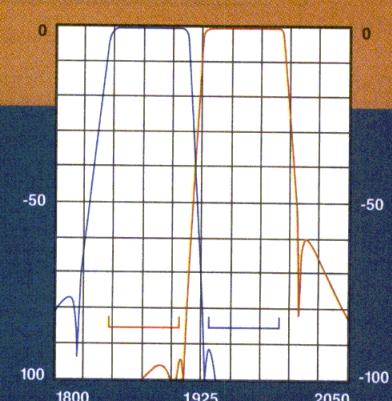

Example Performance:

WD-00007 FULL BAND PCS DIPLEXER

RX Passband Frequency: 1850-1910 MHz
TX Passband Frequency: 1930-1990 MHz

specifications, an engineering decision is forced early on regarding the design of the MMIC or the system specifications.

System simulation provides a fundamental basis for communication between the system and circuit designers. The MMIC designer can use the initial system simulation with behavioral models of components to generate the MMIC specifications (e.g., $P_{1\,dB}$), and the design cannot be finalized until the system simulation meets specifications (e.g., ACPR) using the circuit model. This customized design flow enables immediate feedback to reach system designers that may be considering changes to specs or the design.

For illustration, consider the development of the LMDS transmitter in Fig. 5. The theoretical system includes two major components—a three-port transmitter and a QPSK source. For simulation, the Circuit Envelope Simulator from Agilent Technologies was used to calculate spectral regrowth and adjacent-channel power leakage. This particular simulator provides fast and complete analysis of digitally modulated RF signals. The LMDS transmitter is developed using a hierarchical approach, consisting of MMICs and passive components. The MMICs include a four-stage amplifier, a harmonic mixer, and a filter.

The local-oscillator (LO) arm includes an amplifier to supply the 13.5-GHz signal, which is fed through a bandpass filter before driving the harmonic mixer. The second LO harmonic is then mixed with the intermediate-frequency (IF) signal at 4 GHz. The RF signal is generated at 31 GHz, fed through a bandpass filter, amplified, filtered again, and used to drive the power-stage amplifier. This power-stage amplifier consists of two combined amplifiers, such as the one used as a driver.

It is now necessary to assess the feasibility of using our amplifier in the LMDS transmitter, and then identify the specification that should be assigned to its output power based on linearity requirements. Statistical design techniques invariably result in higher yields, so they must be applied to our MMIC designs.

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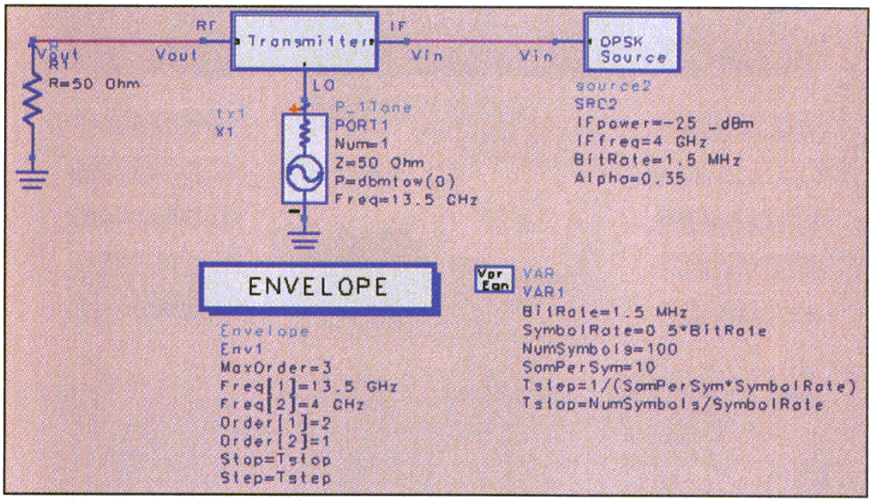


5. In evaluating the system, the transmit LO is set at 13.5 GHz. Then variables describing the QPSK source are defined (as two voltage sources). Two raised-cosine lowpass filters follow the sources and a tuned modulator selects the input harmonic defined by the specified IF and modulates it according to the in-phase (I) and quadrature (Q) modulation inputs.

There are three approaches to applying statistical design techniques. First, one can select the best topology to maximize yields. The design topology's sensitivity to manufacturing process variations should be evaluated early in the development process, not after fabrication.

The second approach is to center parameters to maximize yield. Parameters should be set to minimize performance degradation over the entire range of process variations. The customer is only concerned with worst-case performance.

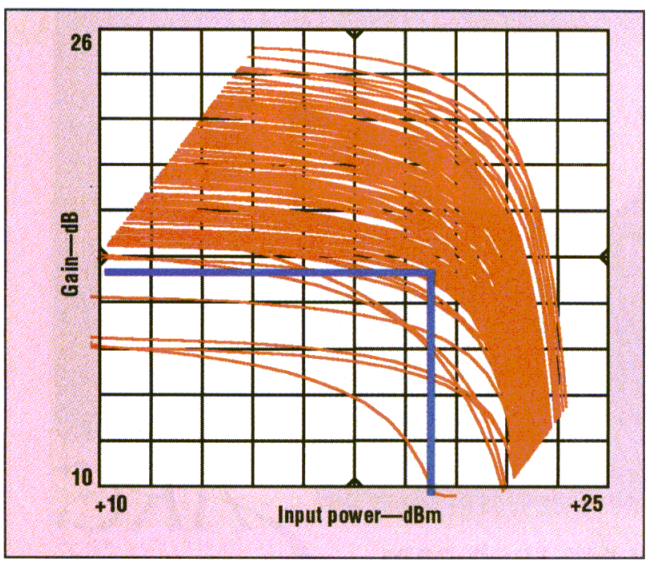
Finally, specifications should be negotiated early. It is easier (and less costly) for system designers to make compromises early in the development process. It is also easier to negotiate specifications when one is armed with statistical data, rather than making guesses about margin.

Behavioral models can be used for system simulation at the lower limit of the distribution as seen in Fig. 6. A behavioral model is defined here as a mathematical representation of the amplifier given specific conditions. In general, behavioral models are more restricted than schematic models, but they provide accurate representations. The key advantage is that they speed up system simulation versus using a full model.

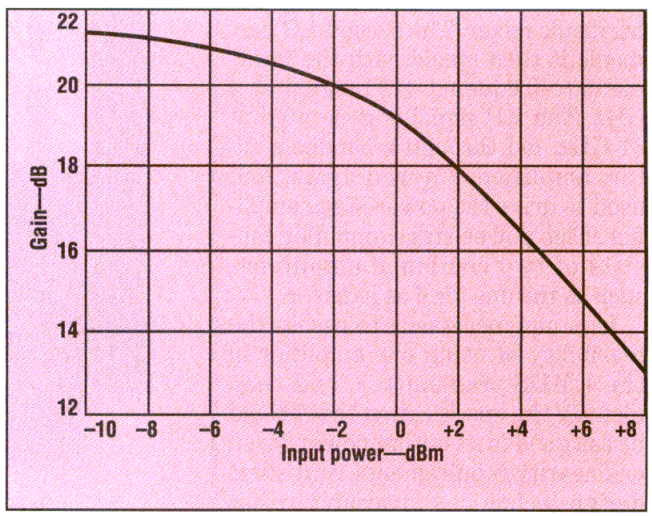
Behavioral models for the MMIC amplifier can be constructed from mathematical equations or with files read by the component. The model defines small-signal performance and gain versus power in (at 31 GHz under recommended bias conditions), where gain is given in real and imaginary terms. Gain must be specified as a function of power input for an amplifier behavioral model. The plot in Fig. 7 shows gain compression versus power input for the MMIC behavioral model.

As with the MMIC amplifier, the transmitter is then created using behavioral models and the mixer and amplifier models. Using behavioral models to simulate a system provides flexibility when choosing limits and conditions, and also speeds up the simulation process—especially for large systems. Behavioral models also help protect intellectual property and are easily updated.

Next, spectral regrowth for the input and output signals from the transmitter is simulated. Filters in the source control the shape of the input signal. The output signal reveals a spectral regrowth of approximately 40 dB. Upon simulating ACPR versus power output, it is found that the transmitter cannot be



6. Knowing the limits of the statistical distribution allows the designer to make decisions based on those limits.



7. This behavioral model of the MMIC amplifier shows gain compression versus power input.

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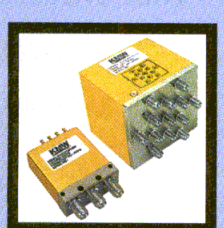

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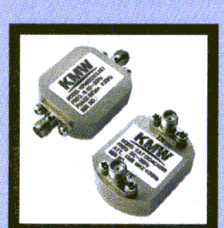

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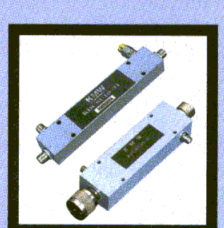

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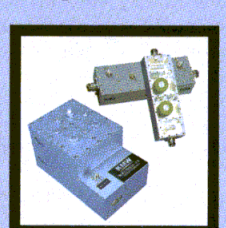

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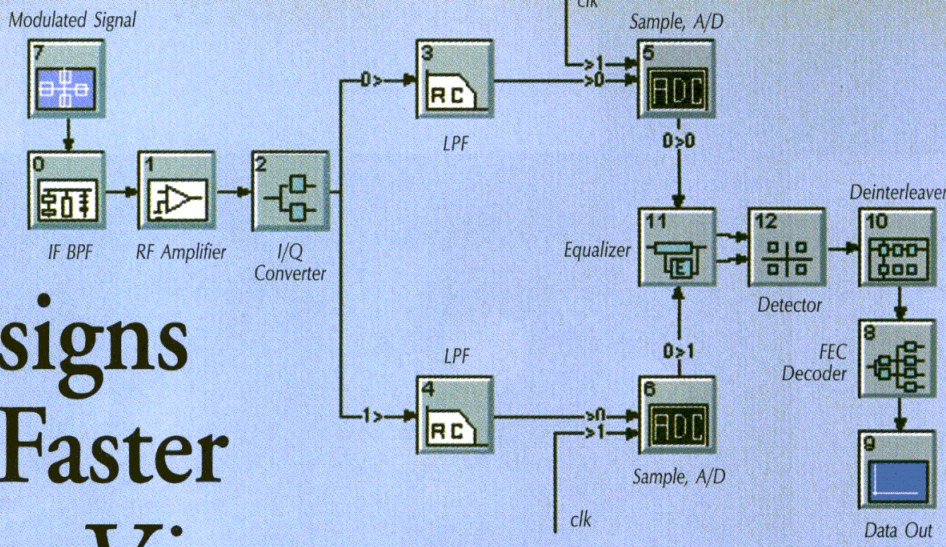

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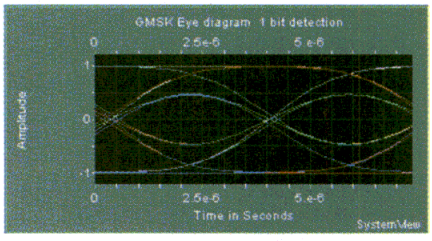
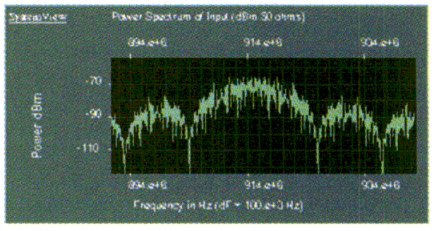
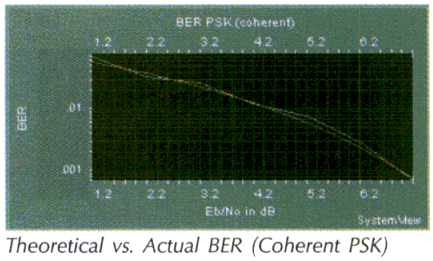


GMSK Eye Diagram (1-bit Detection)



Input Power Spectrum



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operated with output power higher than +20.5 dBm (well below the desired value of +22 dBm) and still remain within specification. Since the transmitter uses two MMIC amplifiers in parallel at the output, the +20.5-dBm limit represents the combined output power for both.

When the power performance for a single MMIC amplifier is plotted, with maximum power defined by the spectral-regrowth specification, it is found that the amplifier must operate approximately 1.5 dB below the $P_{1\,dB}$ level of +19 dBm. Thus, in order to use this part and maintain high yield, the power must remain below this value. Based on simulations, it is known that the power from the transmitter is too low to maintain a high yield for system manufacturing and that additional work will be required by the system designer to maintain a timely product release.

There are a number of options at this point. For example, another amplifier can be added to the output stage to attain the required performance at high manufacturing yields, but the system designer needs to make modifications to accommodate this change.

It is also feasible that only prime chips can be purchased to obtain a higher system yield, but this is likely to be difficult and costly. The MMIC manufacturer may not even be willing to screen devices.

Another option may be to relax the system specifications. The downside is the potential negative impact this could have on the system's ultimate success in the marketplace. The advantage is that fast time to market is maintained.

Finally, a new MMIC can be designed. While this can be done relatively quickly, it negatively impacts time to market, and the risks are quite high.

The potential cost savings from optimizing the design of a MMICs into a system is substantial. The benefits of this optimization extends to the MMIC design because the designer that employs system simulation can directly impact MMIC costs through better product definitions and yields, and even impact device size or design productivity.••

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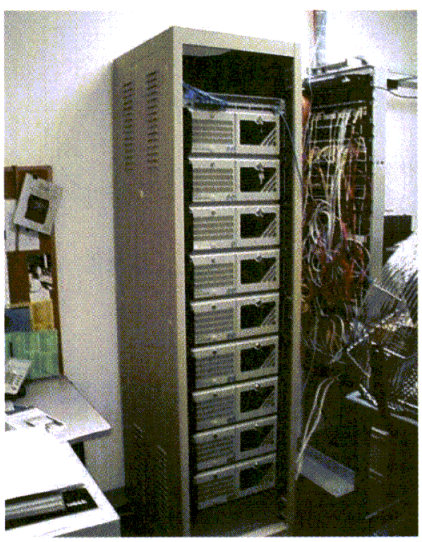

1. An assembled slave-node rack can be seen.

BEOWULF is a computer system design concept pioneered by NASA in 1994. It uses standard, off-the-shelf personal-computer (PC) components and the free Linux operating system to produce low-cost, massively parallel supercomputing systems. Beowulf clusters can serve as a valuable design tool for engineers, by enabling the simulation of a population of designs simultaneously, and dramatically reducing the amount of time required for optimization of design parameters.

EndWave Corp., formerly Endgate (Sunnyvale, CA), was one of the first companies to build a Beowulf system for commercial development purposes. Their Beowulf computer was recently leveraged to design the broadband antenna for Toshiba's Internet-access systems in Japan—in record time. Using the Beowulf computer, along with a customized "genetic algorithm," EndWave was able to evaluate a large number of antenna-parameter scenarios in parallel, thereby reducing the optimization time for the Toshiba antenna from two months to two days. This article will discuss the benefits, applications, and process of building a Beowulf class computer.

A Beowulf class computer cluster is a powerful, inexpensive, and easy-to-build alternative to more-expensive commercial supercomputers. Using low-cost and off-the-shelf PC components, Beowulf can provide significant reduction in design-cycle time. The Beowulf architecture enables engineers to respond rapidly to unique customer requirements, as well as improve design quality to enable the design of more-aggressive product specifications than those previously attempted.

The Beowulf cluster, named after the hero of the epic poem, was first implemented in 1994 at NASA's Goddard Space Flight Center. Thomas Sterling and Don Becker, working under contract to NASA, built a computer system from 16 DX4 (80486 Pentium hybrid) processors to determine the applicability of massively parallel computers to the problems faced by the Earth and space-sciences community. The novelty of Beowulf was not that it used multiple processors, but that it used commodity-off-the-shelf (COTS) parts instead of more-expensive custom hardware. The immediate success of the Beowulf model was made possible by the advent of inexpensive, powerful computing and networking hardware, coupled with the development of standard, publicly available parallel programming tools and platforms, in particular the Linux operating system. In addition to the robust networking and cost advantages of Linux, the open-source model made the development of the required high-speed networking kernel drivers possible. Since 1994, Moore's Law and the bloom of interest in the Linux operating system have made Beowulf clusters a more attractive and powerful option for high-performance computing applications, such as three-dimensional (3D) modeling and design.

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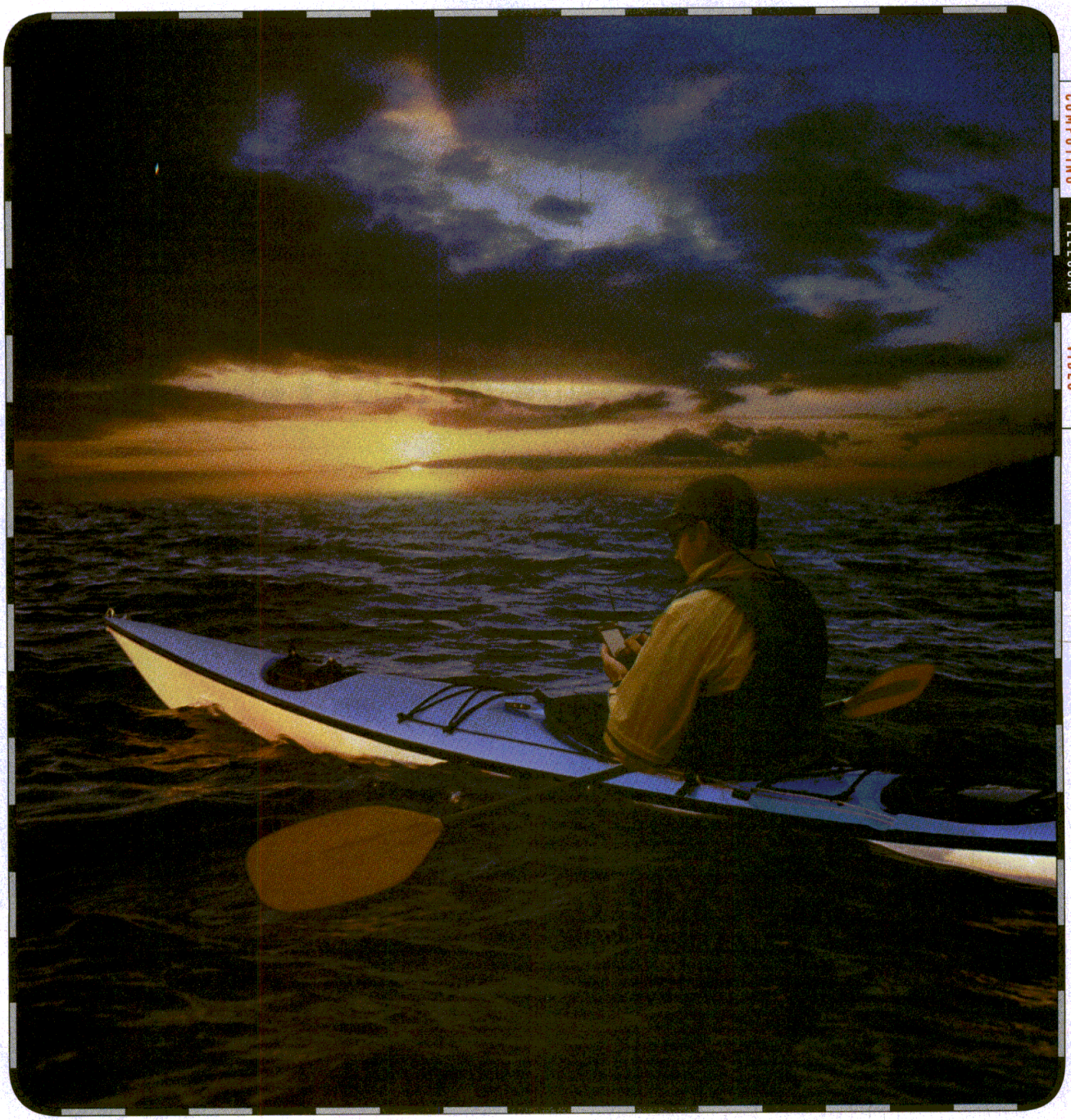

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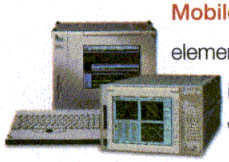

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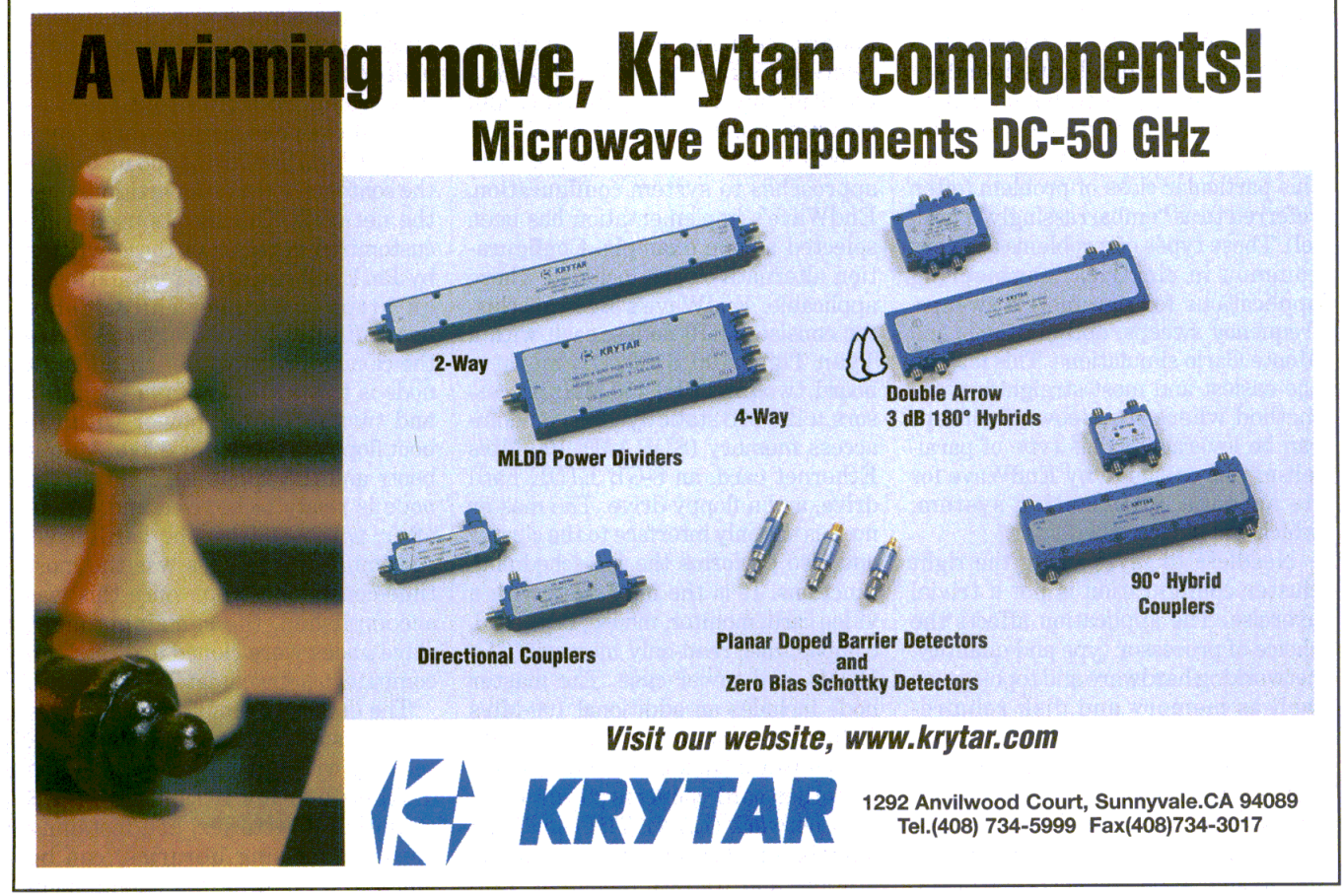
nities were the first to take advantage of these systems, using Beowulf clusters as inexpensive platforms—for high-performance computing and for teaching parallel programming techniques. More than 60 universities and research groups are currently using Beowulf clusters for everything from computational biology to economic forecasting. In the engineering community, Beowulf clusters are making inroads into the realm of commercial-design applications as the unbeatable price/performance ratio and an increasing number of turnkey system and support vendors lure businesses. Companies, such as oil giant Amerada Hess, have saved millions of dollars by replacing expensive leased supercomputers while high-tech startups, such as EndWave Corp., can now harness the kind of computer power once reserved for only the most well funded research groups.

A Beowulf cluster can be defined as a collection of COTS PCs connected by readily available networking hardware running an open-source Unix-like operating system (commonly Linux, but possibly FreeBSD or other operating system). These computers are part of a dedicated cluster, usually on a private network accessed by a single node. The software that is used to implement the parallelism on the cluster usually consists of a set of message-passing libraries for C or FORTRAN, such as MPI or PVM, but can also be something as simple as a script for queuing batch jobs for achieving process-level parallelism.

Applications are the real driver for the design of the cluster. The application must be able to run or be compiled to run on the cluster operating system. Additionally, the application must benefit either from the dividing tasks within the program between multiple processors, or from running multiple iterations of the program simultaneously.

In the first case, where individual tasks within the program are divided between multiple processors, the parallel hooks must be programmed into the application, and the granularity of the parallelism will affect the cluster design. Granularity refers to the amount of computation that can be performed before communication between parallel processes is required. Since the nodes of a Beowulf cluster do not share the same bus or memory as in a traditional supercomputer, latencies can be high. If the parallelism is very "fine grained,"—requiring frequent communication between parallel processes—then it may not be well-suited to a Beowulf, or may require more advanced and expensive network technologies to reduce the network latency (i.e., Myrinet, Scali, Fibrechannel, and Gigabit Ethernet). Regardless of the granularity, the application may also only be scalable to a certain degree, meaning that after a certain number of processors, the improvements in execution time with additional processors diminish to a negligible amount. For the same cost, the less-scalable



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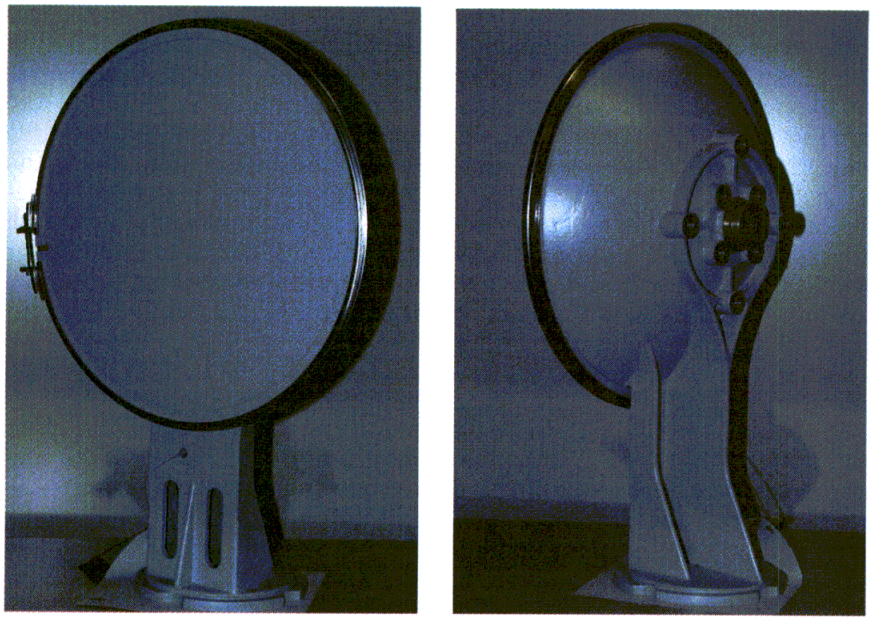

2. The measured performance of the final antenna compares closely with the simulated data generated by the Beowulf computer in a fraction of the time of a conventional PC (left). This millimeter-wave antenna is used with customer premises equipment for wireless Internet access throughout Japan (right).

applications benefit from fewer and faster nodes, while the more scalable applications benefit from a larger quantity of slower nodes.

In the second case, the parallelism is achieved by multiple independent calls to the program, which can be spread over multiple processors. Beowulf clusters are well-suited for this particular class of problem (often referred to as "embarrassingly" parallel). These types of problems are very common in electrical engineering applications for parametric sweeps, frequency sweeps, optimizations, and Monte Carlo simulations. This may be the easiest and most straightforward method where the Beowulf concept can be leveraged. This type of parallelism was exploited by EndWave for its antenna optimization system, which will be described later.

Needless to say, picking the right cluster configuration is not a trivial exercise. The application affects the choice of processor type and quantity, networking hardware and topology, as well as memory and disk requirements. Fortunately, with a Beowulf cluster, it is easy to start small and test the software in order to extrapolate the performance to additional processors. Also, practical help and advice are available on the Internet in the form of how tos, FAQs, and mailing lists. Recommended Internet resources include http://www.beowulf.org and http://www.beowulfunderground.org. Another excellent reference is *How to Build a Beowulf*, by Sterling, Salmon, Becker, and Savarese, published by the MIT Press.

While there are many possible approaches to system configuration, EndWave's implementation has been selected as one example. Configuration alternatives are included where applicable. EndWave's Beowulf cluster consists of 10 nodes, each with a Tyan Tiger 100 S1832DL motherboard, two Intel PIII 450-MHz processors, a 128-MB static dynamic random-access memory (SDRAM), 100-Mb/s Ethernet card, an 8-GB EIDE hard drive, and a floppy drive. The master node is the only interface to the cluster and also performs the job-scheduling functions. It is the only node with a video card, monitor, mouse, keyboard, compact-disc read-only memory (CD-ROM), and tower case. The master node includes an additional 100-Mb/s Ethernet card to provide the only access to the cluster network, while the other card connects to the local-area network (LAN). The remaining nine slave nodes have identical configurations and are rack mounted. Diskless nodes, which boot from a single boot server, are also popular in Beowulf configurations and can reduce cost, but local disks can be chosen as an alternative to provide swap and local scratch space. A 24-port fast Ethernet switching hub is used to connect the nodes. An assembled slave-node rack is shown in Fig. 1. This system costs approximately $20,000. In the least-expensive configuration, a slave node of the type previously described, but without a hard disk and including a mini-tower case, would run at approximately $1000.

In this example, the slave nodes were assembled and hooked up to a video card and monitor to test the basic input/output system (BIOS) and configure the node to boot without a keyboard present. Red Hat Linux Version 6.0 was used as the operating system for the cluster, and a full installation was performed on the master. Any recent Linux distribution will work, but the outdated Extreme Linux CD should be avoided. The operating system was installed on one of the slaves and then "cloned" to the rest of the slave nodes. The cloning process involves booting the new slave node from a floppy, which then mounts its working directories from the master and executes scripts which partition the local hard drive and unarchive the contents of the disk partitions over the network. The scripts used were customized versions of those written by Jan Lindheim of CalTech and found on http://www.cacr.caltech.edu/beowulf/tutorial/building.html. Once the cloning process is set up, adding a node is just a matter of plugging it in and turning on the power with the boot floppy in the floppy drive. After it boots and the floppy is removed, the node is ready to be used and, thereafter, will boot from the hard drive. Assembling the nodes was the most time-consuming step; once this was accomplished, the remaining eight slave nodes were cloned and ready to compute in approximately 10 minutes.

The details of configuring the operating system (relaxing the network security within the cluster, setting up diskless booting if required, NFS mounting directories, and installing message passing libraries) can be found in the how-to documents men-

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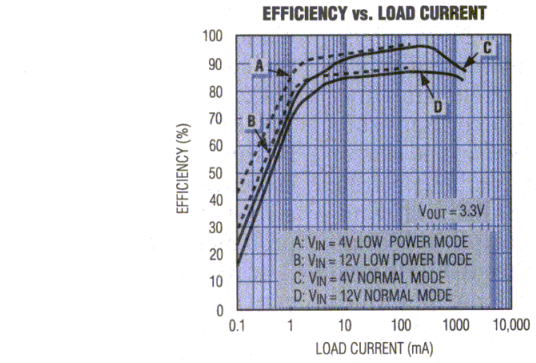

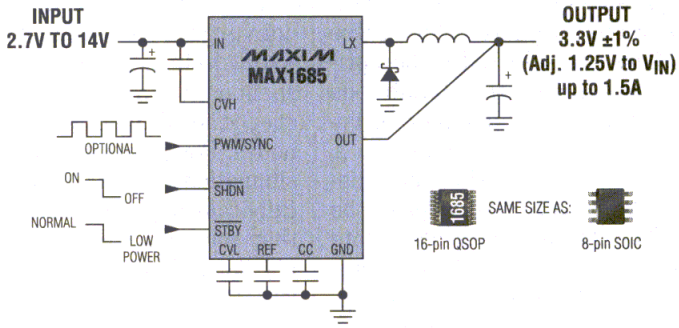

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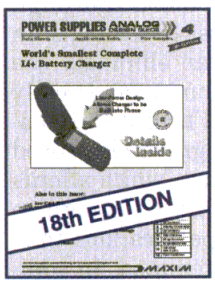

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tioned previously. While these steps are not difficult, they do require a level of familiarity and proficiency with Linux. An alternative to building a Beowulf cluster is to purchase a turnkey system from one of a growing number of vendors, including Paralogic, Inc.; VA Linux Systems; DCG; Alta Technology; and Aspen Systems. This is not a substitute for having someone in-house who understands the system and the operating system (OS), however.

Antenna-design problems typically make use of a number of electromagnetic (EM) simulation engines, including method of moments (MoM), physical or geometric optics, and finite-element or finite-difference time-domain solvers. However, any software, which can be run on the cluster OS in a batch mode, can benefit from process-level parallelism. In EndWave's implementation, a Matlab graphical user interface (GUI) is used to load the application input files which describe the problem geometry, and define parameters based on physical aspects of the structure. When these parameters are selected to be optimized or swept over, the Matlab code writes out a set of input files corresponding to the new geometries. A batch-job queuing routine then distributes the runs to the nodes of the cluster. The Matlab code sends remote shell (rsh) commands to the OS, each one specifying the node and the corresponding input file for the application. The process identifiers (IDs) of these rsh commands are monitored, so that when a job finishes, a new job can be assigned to that cluster node. This process is repeated until all of the jobs have been completed. Then the output from each job can be loaded in order to plot the results or calculate the fitness of optimization cases. While the software described here was developed at EndWave, free (Distributed Queuing System) or commercial (Clustor) alternatives can provide some of the same functionality.

A genetic-algorithm (GA) optimization system was EndWave's largest motivation for building a Beowulf cluster. A GA optimizes a population of designs which can be simulated independently from one another. This property makes the GA ideally suited to take advantage of a Beowulf system. The design parameters of the problem to be optimized are encoded (using a binary or other encoding) as genes, which are then concatenated into a single string known as the chromosome. The process begins by generating a random population of chromosomes that spans the range of permissible parameter values. Every iteration, or generation, of the optimization involves the following steps:

1. Simulating the new members of the population and assigning a fitness value to each one.

2. Selecting a percentage of the population based on fitness to carry over to the next generation (steady-state GA).

3. Filling the remainder of the population with children formed by crossover operations on the chromosomes of the parents (selected with a probability based on their fitness).

4. Randomly mutating a small percentage of the genes of the child population.

These steps are repeated until the convergence criteria are met. The overwhelming majority of time is spent on step number one, which is the step distributed over the cluster nodes. For antenna optimizations, the simulation of each individual can typically take anywhere from 5 s to more than minute for each frequency sample, depending on the size of the problem. Run serially, the simulation time required for a large problem to converge with a population size of a couple hundred can easily run approximately two weeks. This same optimization on a 20-processor Beowulf cluster becomes an overnight run.

EndWave is one of the first companies to develop its own Beowulf system for commercial development, and uses the Beowulf cluster in the design and optimization of millimeter-wave antennas customized for wireless broadband Internet-access systems that link carriers, Internet service providers (ISPs), and business users. As the design process has become less expensive in terms of man hours and time to completion, EndWave has been able to quickly develop custom solutions tailored to the customer's specific needs.

Last year, Toshiba's Telecommunication Systems Division had a need for a small, millimeter-wave customer premise antenna for use in its wireless Internet-access systems throughout Japan. The antenna was limited to a 9-in. (22.86-cm) diameter, with specific requirements for radiation-pattern sidelobes and gain. Existing EndWave feed designs were applied to a reflector of this size, but they failed to meet the tough sidelobe suppression requirements near the mainbeam.

EndWave attacked the problem with the GA optimization system on the Beowulf cluster, using a combined fitness measure that included gain, sidelobe margin with respect to the radiation pattern envelope, and return loss. Several optimizations with different starting feed-geometry seeds were performed. The final design was chosen from among the best individuals from each of these runs based on performance and manufacturability. Within one week, a prototype was designed, optimized, manufactured, and tested. Figure 2 (left) compares the measured performance of the final design with the simulated performance and the Toshiba sidelobe envelope. The production version of the antenna is shown in Fig. 2 (right). By processing a large number of antenna parameter scenarios in parallel, the Beowulf computer was able to reduce the optimization time for the Toshiba antenna from two months to two days.

In conclusion, this innovative and robust modeling methodology can provide design engineers with the capability for rapid response to unique customer requirements. The use of a Beowulf cluster significantly reduces design time, which enables the engineer to attempt a more thorough investigation of possible solutions. In many cases, this can eliminate the need for time-consuming and costly empirical iteration. Further, the quality of optimized designs are improved over other techniques, which affords designers the ability to design to more-aggressive specifications than were previously attempted. In addition, Beowulf clusters have been used in a number of industries and applications, such as computational biology, economic and weather forecasting, circuit synthesis, as well as surface and image processing.●●

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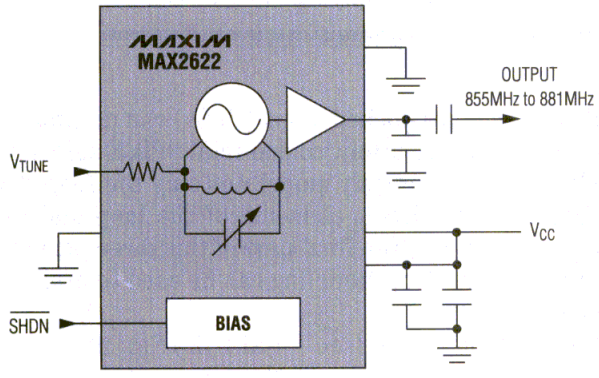

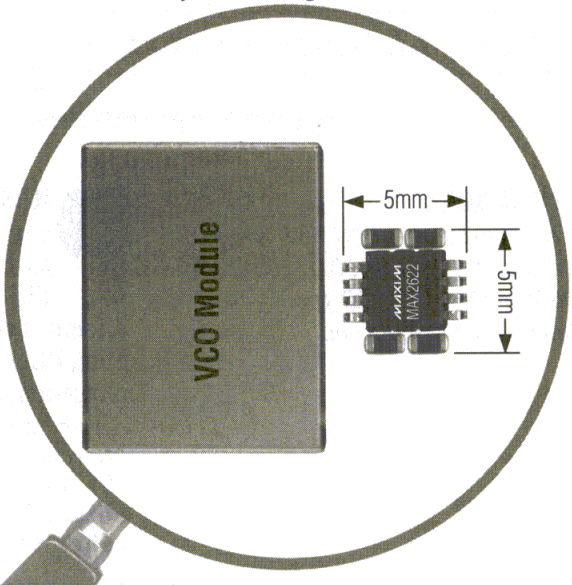

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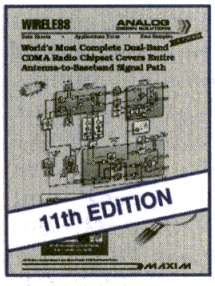
MAX2623	885 to 950	US ISM Band, DECT
MAX2624	947 to 998	US ISM Band



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CIRCLE NO. 209

Model PLL Dynamics And Phase-Noise Performance

Commercial SPICE and math programs can be used to combine noise-source models to develop a composite plot of oscillator phase noise.

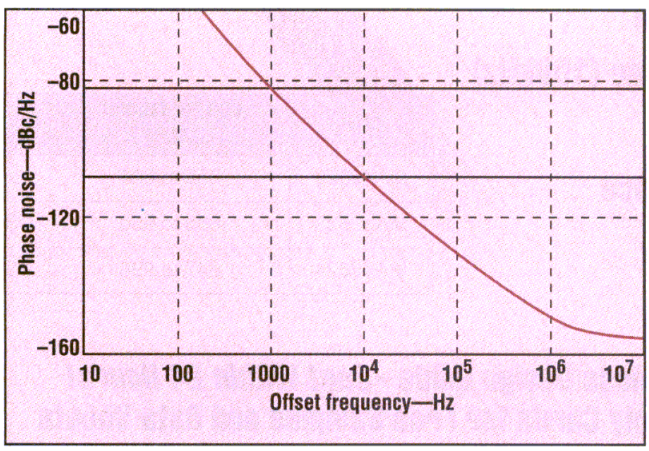
Model PLL Dynamics, Part 3

Eric Drucker

PLL Consultants, 7701 56th Ave. NE, Seattle, WA 98115-6301; (206) 525-0674, FAX: (425) 290-1600, e-mail: linerc@sprintmail.com.

PHASE noise in phase-locked loops (PLLs) can originate from a number of sources. Fortunately, commercial SPICE software tools, such as PSPICE, can effectively model not only the loop dynamics of a PLL, but also the circuit elements of the loop and the combined effects of the noise sources. The final part of this three-part series on modeling PLLs will show how loop modeling can be used to predict synthesizer phase-noise performance.

Part 2 of this article series (see *Microwaves & RF*, February 2000, p. 73) reviewed the basics of phase noise. It introduced a simple schematic-circuit diagram in PSPICE (Fig. 19 of Part 2) to model the phase noise of a voltage-controlled oscillator (VCO). The model assumed three noise sources in the loop that contribute to the output phase noise of the oscillator. Noise equations for use in MathCAD were developed for these sources; they were also modeled using PSPICE. In the final installment of this PLL article series, the individual noise sources will be combined, in MathCAD and PSPICE, to obtain a composite phase-noise plot. This final installment will also show how the noise can be improved by changing the loop parameters.



20. Noise values predicted by PSPICE are within 0.1 dB of the values calculated by the MathCAD noise equation.

Referring back to that schematic diagram (Fig. 19 of Part 2), several resistors were used as sources of noise. For example, resists R_k0_vco1 and R_k0_vco2 generate flat noise. Integrating the noise of resistors R_k2_vco1 and R_k2_vco2 produces $1/f^2$ noise. Since PSPICE requires two connections at every node, two resistors in parallel are needed for VCO

Table 1: Comparing noise levels

Loop parameters	Residual FM (Hz) (300 Hz to 3 kHz)	Residual PM (Radians) (10 Hz to 100 kHz)
1-MHz reference, 1.0-kHz loop BW	4.252	4.812×10^{-3}
1-MHz reference, 2.5-kHz loop BW	3.629	2.856×10^{-3}
10-MHz reference, 6.6-kHz loop BW	1.150	1.356×10^{-3}

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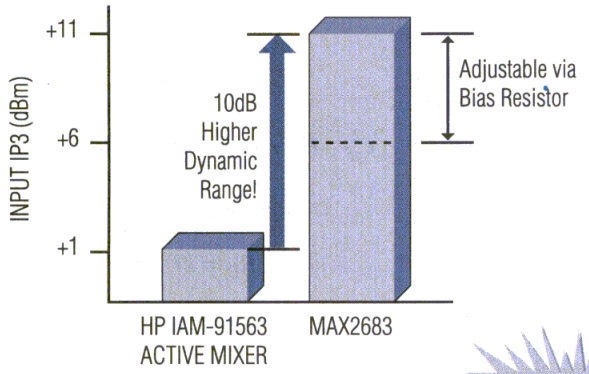

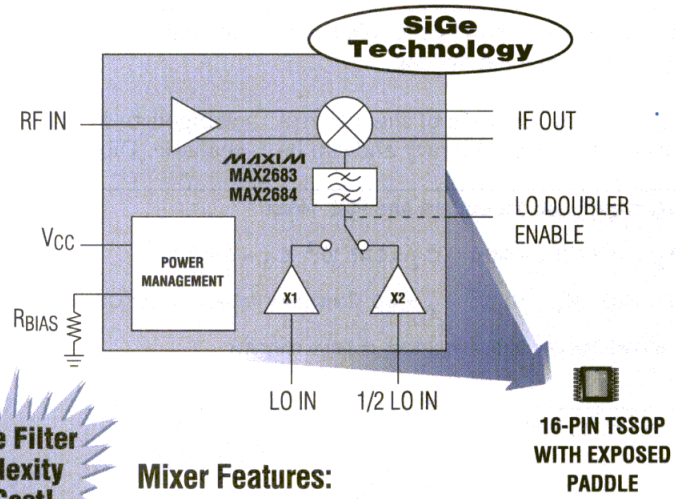

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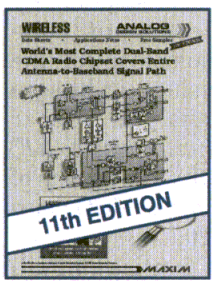

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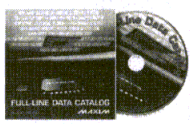

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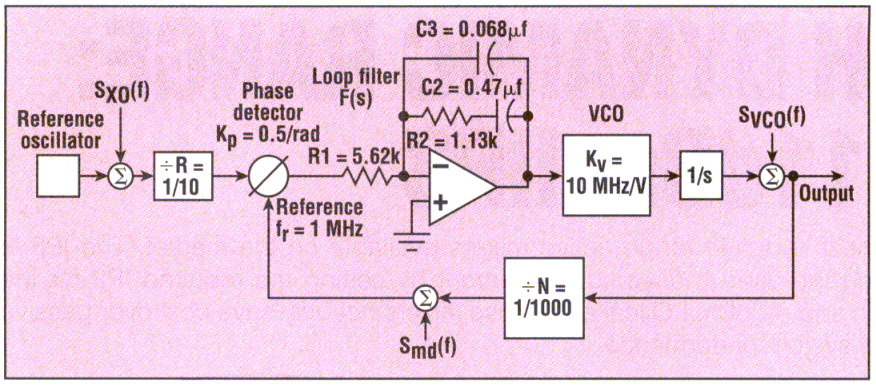
CIRCLE NO. 210



21. A simple PLL can be modeled with three noise sources—the VCO, reference oscillator, and main divider.

noise model. The gain of the integrator is somewhat arbitrary and can be changed to keep the resistor values realistic. Diode Dn1, which is biased by current source In1, produces the 1/f noise. Voltage source Vn1 offsets the diode voltage. Based on this schematic diagram, the PSPICE noise-model equations are as follows (refer to the sidebar, "Calculate PSPICE values for noise models," from *Microwaves & RF*, February 2000, p. 78):

$$e_{n0}^2 = 4kTR$$

$$e_{n1}^2 = R_D^2 i_n^2 = (kT/qI_D)^2(k_f I_D/f)$$
$$= (kT/q)^2(1/I_D)(k_f/f)$$

$$e_{n2}^2 = (4kTR)(k_{i2}^2/f^2)$$

$$e_{n3}^2 = (kT/q)(1/I_D)(k_f/f)(k_{i2}^2/f^2)$$

$$i_n^2 = 2qI_D + (k_f I_D/f)$$

$$R_D = kT/qI_D$$

$$V_o/V_i = k_i/f$$

where:

e^2_{n0} = the k_0 resistor term in the noise equation,

e^2_{n1} = the k_1 diode voltage term in the noise equation,

e^2_{n2} = the k_2 voltage term in the noise equation,

e^2_{n3} = the k_3 voltage term in the noise equation,

i^2_n = the SPICE diode model (ignoring the $2qI_D$ term),

R_D = the diode dynamic resistance (approximately 26 Ω/mA), and

V_o/V_I = the transfer function for the integrator.

Once the diode current is set, only the K_f term in the PSPICE diode model (DNOISE1) sets the value of the 1/f term in the noise equation. By integrating the diode noise, $1/f^3$ noise is produced. The solutions for the resistor and the K_f values can be seen in the MathCAD analysis sidebar from Part 1 (see *Microwaves & RF*, November 1999, p. 70). The results of a PSPICE analysis of the VCO noise can be seen in Fig. 20.

The effects of VCO noise on PLL performance can be best understood by referring to a loop with the assumption of three noise sources (VCO, reference, and main divider) as shown in Fig. 21. The transfer function from the VCO noise to the

Calculate Loop Paramters and Plot Noise

$R_2 := 1130$ $R_1 := 5620$ $C_2 := 0.47 \cdot 10^{-6}$ $C_3 := 0.068 \cdot 10^{-6}$

$N := 1000$ $K_v := 10 \cdot 10^6$ $K_p := 0.5$ Component Values and Loop Constants

Calculate Pole and Zero Frequencies of Active Integrator

$f_z := \frac{1}{2 \cdot \pi \cdot (R_2 \cdot C_2)}$ $f_z = 299.67$ $f_p := \frac{1}{2 \cdot \pi \cdot \left(R_2 \cdot \frac{C_2 \cdot C_3}{C_2 + C_3}\right)}$ $f_p = 2.371 \cdot 10^3$ Zero & Pole Frequency

$F(s) := \frac{R_2}{R_1}\left(\frac{C_2}{C_2 + C_3}\right)\cdot\frac{(s + f_z)}{s}\cdot\frac{f_p}{s + f_p}$ Loop Filter (Active Integrator 3rd Order)

Open Loop Transfer Function & Plot (-120 dB Offset)

$GH(s) := \frac{K_p \cdot K_v \cdot F(s)}{s \cdot N}$ $dB_{open\_loop_i} := 20 \cdot \log(|GH(s_i)|) - 120$ $f_s := 500$

$f_c := root(|GH(j \cdot f_s)| - 1, f_s)$ $f_c = 871.676$ Loop Gain Crossover

$\phi_m := -180 - \frac{180}{\pi} \cdot \arg(GH(j \cdot f_c))$ $\phi_m = -50.842$ Phase Margin

Outside the Loop Bandwidth (VCO) Noise

VCO NoiseTransfer Function (High Pass) & Plot (-80 dB Offset)

$\Theta_{vco}(s) := \frac{1}{1 + GH(s)}$ $dB_{loop\_hp_i} := 20 \cdot \log(|\Theta_{vco}(s_i)|) - 80$

VCO Noise Modified by the Loop & Plot

$S_{\phi\_vco\_loop}(f) := S_{\phi\_vco}(f) \cdot |\Theta_{vco}(j \cdot f)|^2$ $dB_{vco\_loop_i} := 10 \cdot \log(S_{\phi\_vco\_loop}(f_i))$

Inside the Loop Bandwidth Noise

Reference NoiseTransfer Function (Low Pass) & Plot (-80 dB Offset)

$\Theta_{ref}(s) := \frac{GH(s)}{1 + GH(s)}$ $dB_{loop\_lp_i} := 20 \cdot \log(|\Theta_{ref}(s_i)|) - 80$

$R := 10$ Reference Division

Multiplied Sum of Reference and Divider Noise & Plot

$S_{\phi\_mult}(f) := N^2 \cdot \left(\frac{S_{\phi\_ref}(f)}{R^2} + S_{\phi\_md}(f)\right)$ $dB_{mult_i} := 10 \cdot \log(S_{\phi\_mult}(f_i))$

Multiplied Reference and Divider Noise Modified by the Loop & Plot

$S_{\phi\_mult\_loop}(f) := S_{\phi\_mult}(f) \cdot |\Theta_{ref}(j \cdot f)|^2$ $dB_{mult\_loop_i} := 10 \cdot \log(S_{\phi\_mult\_loop}(f_i))$

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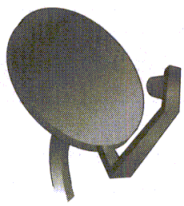

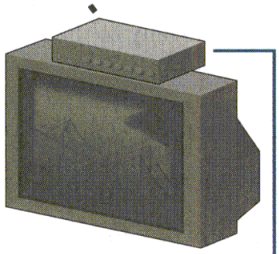

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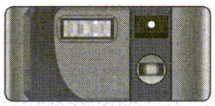

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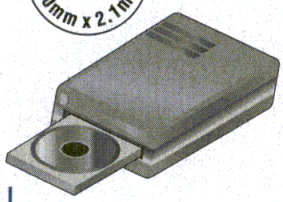

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MAX4014	1	2	+3.15 to +11	30	200	600	0.02/0.02	0.06 to 4.94	No	5-pin SOT23	0.80
MAX4016/4451	2	1	+3.15 to +11	30/55	150/200	600/485	0.02/0.02	0.06 to 4.94	No	8-pin SO/µMAX/SOT23	0.95/0.80
MAX4017	2	2	+3.15 to +11	30	150	600	0.02/0.02	0.06 to 4.94	No	8-pin SO/µMAX	1.09
MAX4018/4020	3/4	1	+3.15 to +11	30	150	600	0.02/0.02	0.06 to 4.94	Yes	14-pin SO, 16-pin QSOP	1.53/1.80
MAX4019/4022	3/4	2	+3.15 to +11	30	150	600	0.02/0.02	0.06 to 4.94	Yes	14-pin SO, 16-pin QSOP	1.66/1.98

(1) V_{CC} = +5V, V_{EE} = 0V. (2) 50,000-up, FOB USA.
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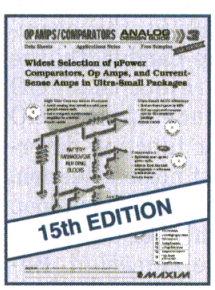

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CIRCLE NO. 211

output, using the control-theory rule (forward gain/(1 + loop gain), where the loop gain is GH(s), is:

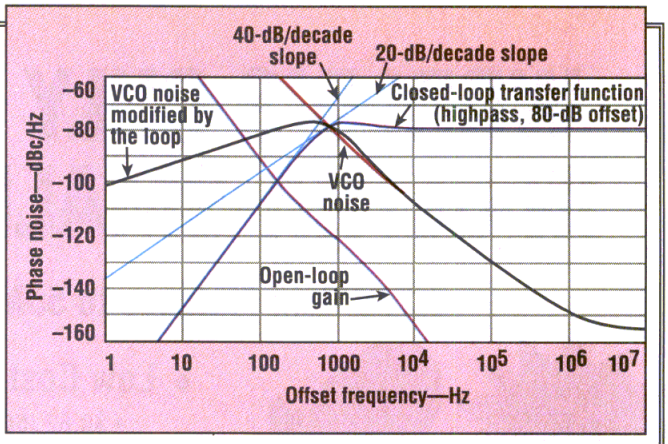
$$\Theta_{VCO}(s) = 1/1+(1/s)[K_v K_p F(s)/N]$$

The phase-noise equation is in terms of noise power, so it is necessary to square the transfer function, which is in terms of voltage ratio. By omitting the 2π term in the transfer-function frequency variable, the loop constants and frequency variable in the noise equation are assumed to be in Hertz. The equation for the VCO noise and output noise due only to the VCO noise is:

$$S_{\phi\_VCO\_Out}(f) = S_{\phi\_VCO}(f)\,|\Theta_{VCO}(jf)|^2 \qquad (15)$$

which can also be written as:

$$S_{\phi\_VCO}(f) = K_{0\_VCO} + K_{2\_VCO}/f^2 + K_{3\_VCO}/f^3 \qquad (16)$$



22. This plot of various transfer functions, including "raw" VCO noise and VCO noise modified by the loop, was generated with the help of MathCAD.

The PLL output-noise contribution from the VCO is "raw" VCO noise modified by the VCO, or highpass, closed-loop transfer function (Fig. 22). The raw VCO noise, VCO (highpass) closed-loop transfer function –80 = 0 dB), and open-loop transfer function –120 = 0 dB) are also shown on this plot as functions of offset frequency. The MathCAD development used to generate these plots is shown in the accompanying sidebar.

This VCO noise is commonly said to be outside the loop noise because the VCO noise is unchanged beyond, or outside, the loop bandwidth (approximately 1 kHz for this example). Inside the loop bandwidth, the loop attempts to reject the VCO noise. The magnitude of the VCO closed-loop transfer function determines the amount of noise rejection. Also, an examination of the open-loop gain is another method for determining the VCO noise rejection. Between DC and the gain-crossover frequency, (|GH(jf) = 1, or 0 dB), the distance between 0 dB and the open-loop gain curve is the rejection. This means the loop can reject VCO noise by 20 dB/decade from gain crossover

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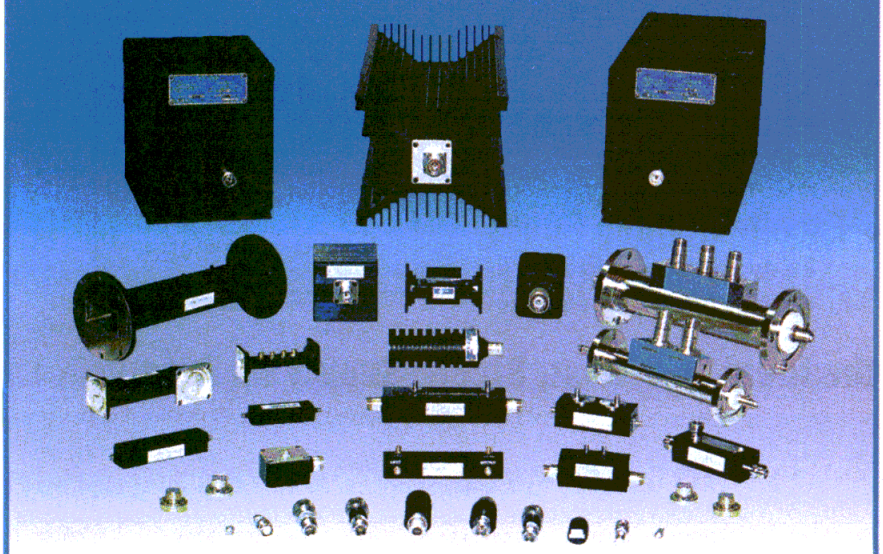

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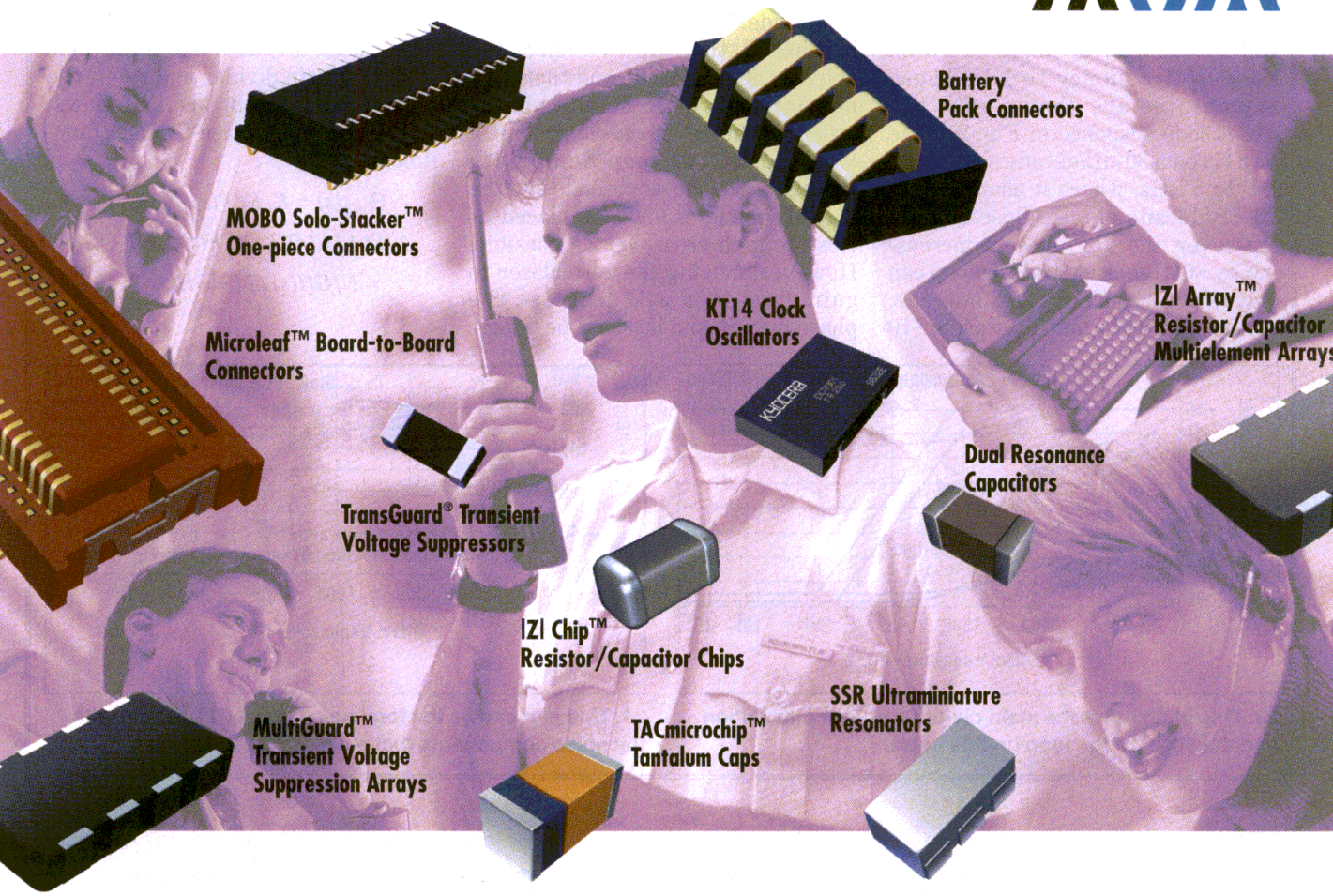

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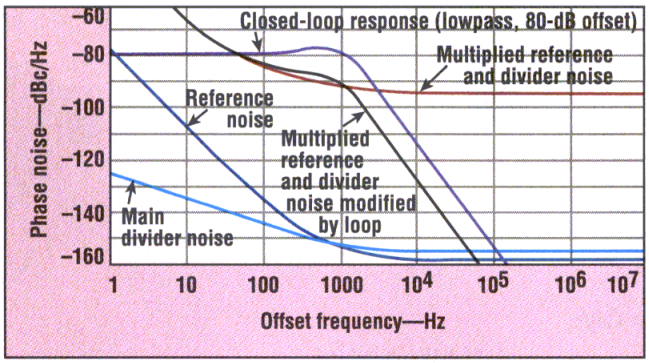
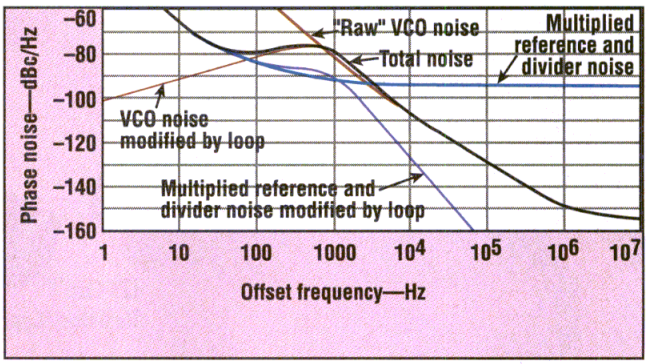
~1 kHz) to the zero frequency (~300 Hz) and then by 40 dB/decade from the zero frequency to DC. These asymptotes are shown with the closed-loop response in Fig. 21. The VCO is on a 30-dB/decade slope from DC to 5 kHz, where it changes to a 20-dB/decade slope. Close to gain crossover, the VCO noise is increasing by 30 dB/decade but the loop can only reject the noise by 20 dB/decade, so there is still a net increase of 10 dB/decade. Once the zero frequency is reached, the loop can reject the noise by 40 dB/decade and there is a net decrease of 10 dB/decade. In order to achieve the most rejection of VCO noise, the loop bandwidth should be as wide as possible and the zero frequency should be placed as close to gain crossover as possible. However, moving the zero closer to gain crossover will decrease the phase margin and increase peaking.

The transfer functions, noise equations, and output-noise equations from the main divider and reference noise to the output are as follows. The main divider noise-transfer function is:

$$\Theta_{md}(s) = K_v K_p F(s)(1/s)/\{1+(1/s)[K_v K_p F(s)/N]\} = N\{GH(s)/[1+GH(s)]\} \quad (17)$$



23. This plot shows the main divider's noise contribution and the effect of the loop on this noise.



24. This plot shows the total noise of the PLL, along with the noise components that make up the total noise.

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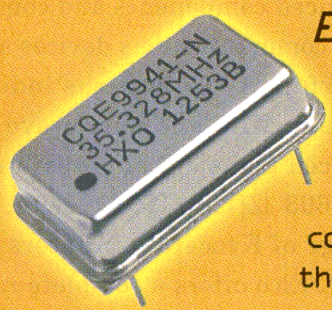

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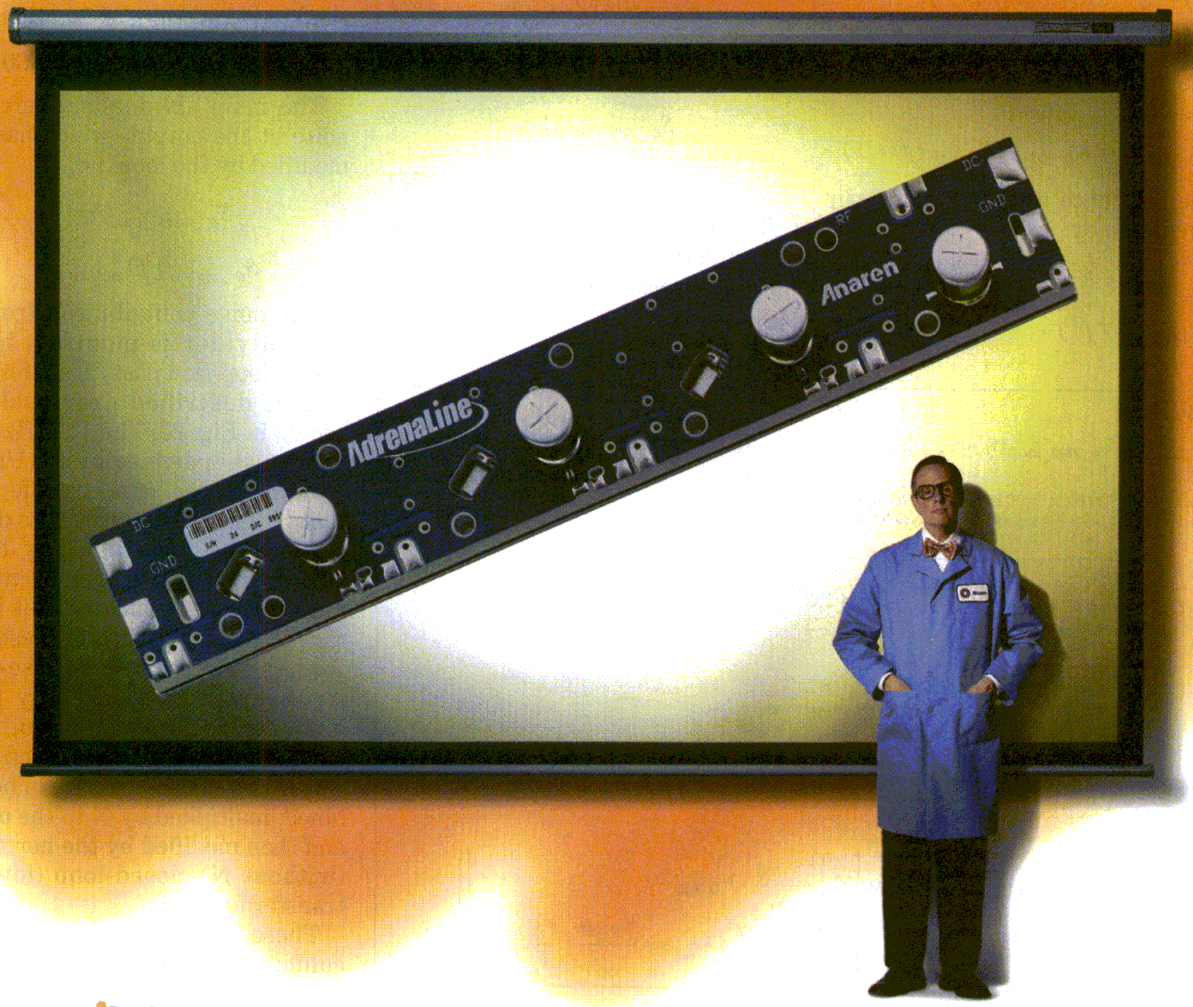

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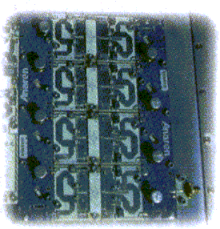

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The reference-oscillator noise-transfer function is:

$$\Theta_{ref}(s) = (1/R)K_v K_p F(s)(1/s) / \{1+(1/s)[K_v K_p F(s)/N]\} = (1/R)N\{GH(s)/[1+GH(s)]\} \quad (18)$$

The main divider noise can be calculated from:

$$S_{\phi\_md}(f) = K_{0\_md} + (K_{1\_md}/f) \quad (19)$$

The reference oscillator noise can be predicted by:

$$S_{\phi\_ref}(f) = K_{0\_ref} + (K_{1\_ref}/f) + (K_{2\_ref}/f^2) + (K_{3\_ref}/f^3) \quad (20)$$

The main divider noise at the output modified by the loop is:

$$S_{\phi\_md\_out}(f) = S_{\phi\_md}(f)\,|\Theta_{md}(jf)|^2 \quad (21)$$

The reference oscillator noise at the output modified by the loop is:

$$S_{\phi\_ref\_out}(f) = S_{\phi\_ref}(f)\,|\Theta_{ref}(jf)|^2 \quad (22)$$

The multiplied sum of the reference and the divider noise at the output (which has not been modified by the loop) is:

$$S_{\phi\_mult}(f) = N^2 S_{\phi\_md}(f) + [S_{\phi\_ref}(f)/R^2] \quad (23)$$

The normalized (without N division) lowpass transfer function is:

$$\Theta_{x0mult}(s) = GH(s)/[1+GH(s)] \quad (24)$$

The main divider and reference noise at the output (which has been modified by the loop) is:

$$S_{\phi\_mult\_out}(f) = S_{\phi\_mult}(f)\,|\Theta_{mult}(jf)|^2 \quad (25)$$

The noise contribution from the main divider is modified by the closed-loop (lowpass) transfer function as described previously and shown in Fig. 23. The 10-MHz reference noise is divided by 10 (1/R) and then obeys the same transfer function. The fact that a noiseless divider reduces the phase noise by 1/(division ratio) can be proven from small-angle modulation theory. It is important to note that both of these noise sources, referred to the phase-detector input, are multiplied by N, the main divider division ratio, to the output, within the loop bandwidth. These two noise sources can be combined, multiplied by N to the output, and then modified by the normalized (without N) closed-loop (lowpass) transfer function.

These summed noise sources are commonly called "inside-the-loop noise" because this noise is unchanged inside the loop bandwidth. Phase-detector noise can also be included. Outside the loop bandwidth, the loop attempts to reject this noise.

The magnitude of the reference closed-loop transfer function determines the amount of noise rejection. Once again, the open-loop transfer function can also be used to determine the rejection (the open-loop gain is shown in Fig. 23). Beyond the gain-crossover frequency, the distance between 0 dB and the open-loop gain curve is the amount of the rejection. This means that the loop can reject this noise by 20 dB/decade from gain crossover (~ 1 kHz) to the pole frequency (~ 2.3 kHz) and then by 40 dB/decade.

The divider and reference noise

Total Noise

$S_{\phi\_total}(f) := S_{\phi\_mult\_loop}(f) + S_{\phi\_vco\_loop}(f)$ $\quad dB_{total_i} := 10\cdot\log(S_{\phi\_total}(f_i))$

Integrate from 300 Hz to 3.00 KHz to Get Residual FM Noise and from10 Hz to 100 KHz to Get Residual PM Noise (Radians)

$$\sigma_{fm} := \sqrt{\int_{300}^{3000} (S_{\phi\_total}(f))\cdot f^2\, df} \qquad \sigma_{fm} = 4.252$$

$$\sigma_{pm} := \sqrt{\int_{10}^{10^5} S_{\phi\_total}(f)\, df} \qquad \sigma_{pm} = 4.812\cdot 10^{-3}$$

Calculate Optimum Loop BW, New Loop Component Values, and Plot Noise

$f_{op} := 2500$ Given

$$10\cdot\log\left(\frac{k_{2\_vco}}{f_{op}^2} + k_{0\_vco} + \frac{k_{3\_vco}}{f_{op}^3}\right) = 10\cdot\log\left[N^2\cdot\left[\frac{k_{1\_md}}{f_{op}} + k_{0\_md} + \frac{\left(\frac{k_{1\_ref}}{f_{op}} + k_{0\_ref} + \frac{k_{2\_ref}}{f_{op}^2} + \frac{k_{3\_ref}}{f_{op}^3}\right)}{R^2}\right]\right]$$

$x := \text{Find}(f_{op})$ $\quad x = 2.582\cdot 10^3$

$R_{21} := 2940$ $\quad R_{11} := 5620$ $\quad C_{21} := 0.047\cdot 10^{-6}$ $\quad C_{31} := 0.0068\cdot 10^{-6}$ New Component Values for 2.5 KHz Loop BW

$$f_{z1} := \frac{1}{2\cdot\pi\cdot R_{21}\cdot C_{21}} \qquad f_{z1} = 1.152\cdot 10^3$$

$$f_{p1} := \frac{1}{2\cdot\pi\cdot\left(R_{21}\frac{C_{21}\cdot C_{31}}{C_{21}+C_{31}}\right)} \qquad f_{p1} = 9.113\cdot 10^3$$

New Zero, Pole & Gain Crossover Frequency

$$f_{xo1} := \frac{K_v\cdot K_p\cdot\frac{R_{21}}{R_{11}}}{N} \qquad f_{xo1} = 2.616\cdot 10^3$$

$$F_1(s) := \frac{R_{21}}{R_{11}}\cdot\left(\frac{C_{21}}{C_{21}+C_{31}}\right)\cdot\frac{(s+f_{z1})}{s}\cdot\frac{f_{p1}}{s+f_{p1}} \qquad GH_1(s) := \frac{K_p\cdot K_v\cdot F_1(s)}{s\cdot N}$$

Active Integrator and Open Loop Gain

$$\Theta_{vco\_1}(s) := \frac{1}{1+GH_1(s)} \qquad \Theta_{ref\_1}(s) := \frac{GH_1(s)}{1+GH_1(s)}$$

VCO & Reference Close Loop Transfer Functions

$f_{c1} := \text{root}(|GH_1(j\cdot f_s)| - 1, f_s)$ $\quad f_{c1} = 2.441\cdot 10^3$

Confirm Gain Crossover and Check Phase Margin

$$\phi_{m1} := -180 - \frac{180}{\pi}\cdot\arg(GH_1(j\cdot f_{c1})) \qquad \phi_{m1} = -49.743$$

$$dB_{total1_i} := 10\cdot\log\left[\left[S_{\phi\_mult}(f_i)\cdot(|\Theta_{ref\_1}(s_i)|)^2\right] + S_{\phi\_vco}(f_i)\cdot(|\Theta_{vco\_1}(s_i)|)^2\right]$$

Total Noise

$\sigma_{fm1} = 3.629$ $\quad \sigma_{pm1} = 2.856\cdot 10^{-3}$ Calculate Residual FM and PM (Equations Not Shown)

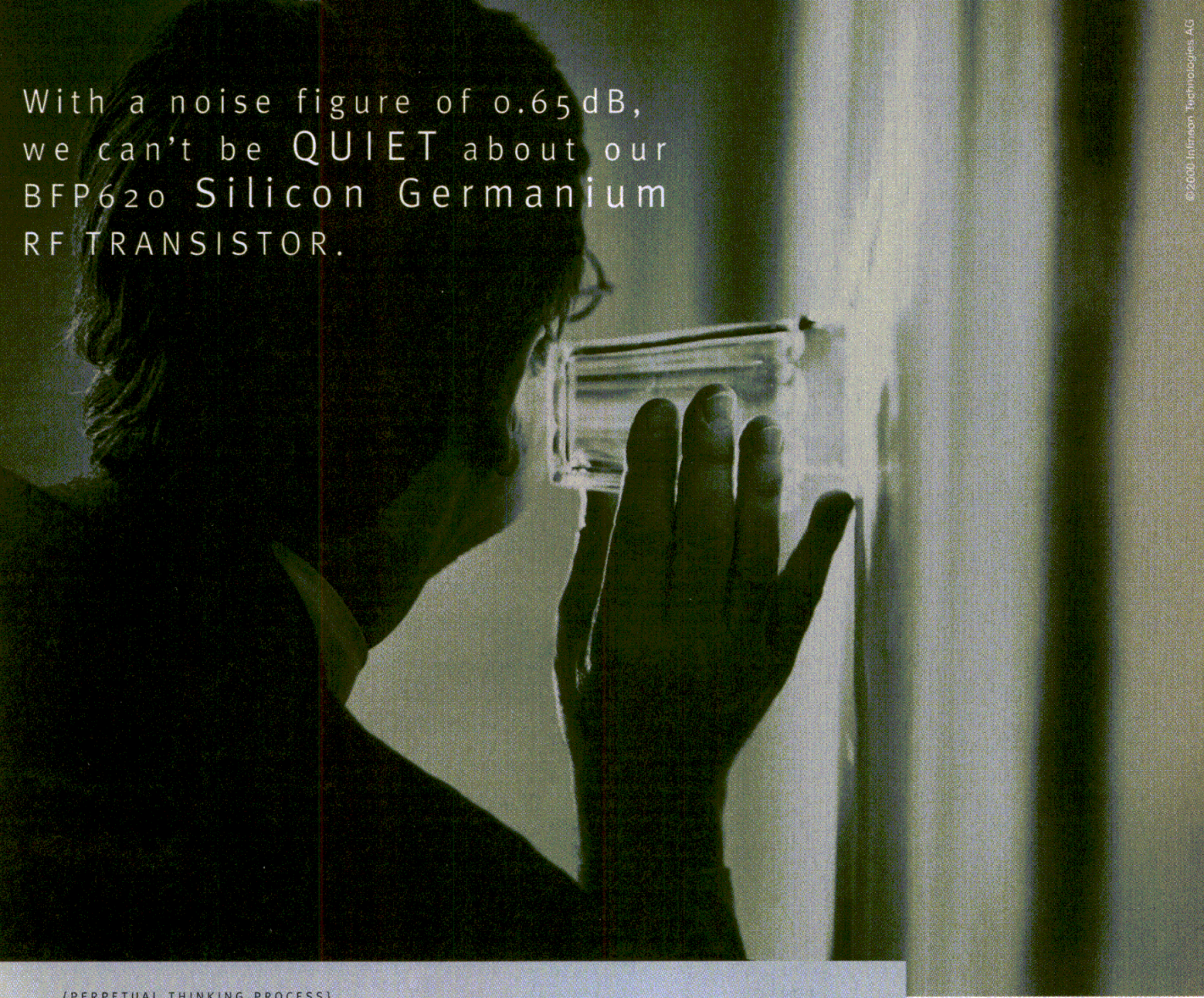

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Table 2: Comparing MathCAD and PSPICE phase-noise results

Offset frequency	MathCAD	PSPICE	PSPICE Noiseless R's
10 Hz	−67.02	−67.01	−67.02
100 Hz	−79.51	−79.28	−79.50
1 kHz	−79.94	−79.03	−79.92
10 kHz	−108.02	−107.70	−107.95
100 kHz	−129.77	−129.68	−129.68

multiplied to the output frequency, without the loop, is essentially flat from gain crossover and beyond. Within the loop, the output noise will decrease by 20 dB/decade until reaching the pole frequency and then decreasing by 40 dB/decade. Making the loop bandwidth narrower will increase the rejection of noise sources inside the loop. Moving the pole closer to gain crossover will also help reject noise within the loop but, once again, decrease the phase margin and increase peaking.

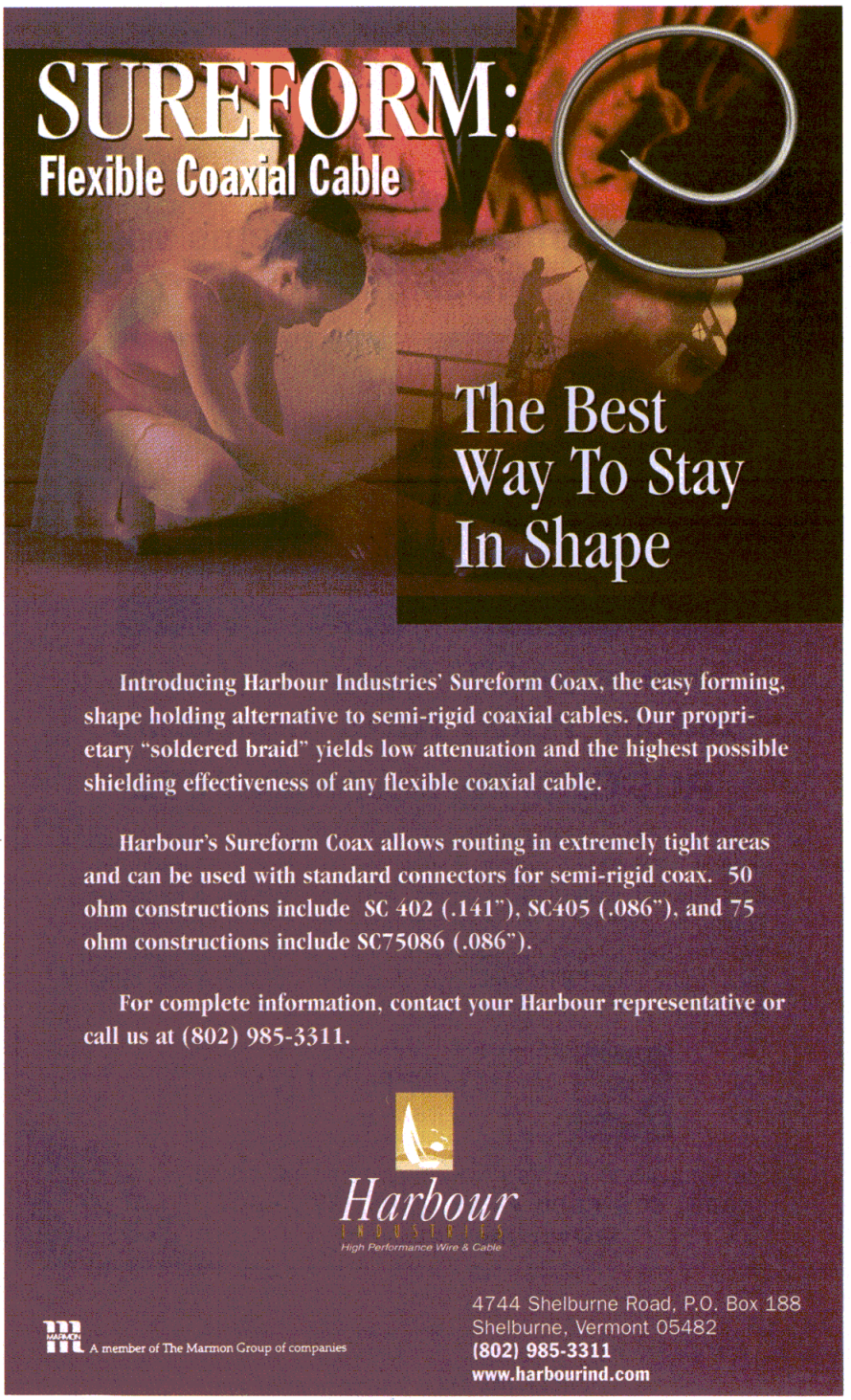
Taking the VCO noise and the multiplied divider plus the divided reference noise scaled by the appropriate transfer function yields the total noise of the PLL. This is shown in Fig. 24, along with the various components that make up the total noise. The total noise of the PLL can be found from:

$$S_{\phi\_total}(f) = S_{\phi\_ref\_out}(f) + S_{\phi\_vco\_out}(f) \quad (26)$$

Equation 26 can be integrated to find the residual phase and frequency noise in a particular bandwidth. This is shown in the MathCAD sidebar.

Looking at the total phase-noise curve, there is a large "bump" between ~100 Hz and 2 kHz. In this region, the multiplied reference noise is approximately 10 dB lower than the final phase-noise curve and the raw VCO noise is significantly higher than the final phase-noise curve. This implies that the original loop bandwidth of 1 kHz is not optimum. A rule of thumb for locating the optimum loop bandwidth for noise is where the raw VCO noise curve intersects with the multiplied sum (N) of the reference, divider, and phase-detector noise. Using MathCAD, it is possible to solve for this frequency (~2.6 kHz), recompute the loop-component values, and replot the noise (Fig. 25). The noise in the 100-Hz-to-2-kHz region has been decreased by approximately 6 dB. In addition, the residual frequency modulation (FM) and phase modulation (PM) are lower. However, the noise in the 2-to-20-kHz region is slightly higher because, due to the wider loop bandwidth, less of the reference noise is being rejected in this region.

Another experiment that can be performed with MathCAD (equations for this part are not shown in the sidebar) is to observe what would



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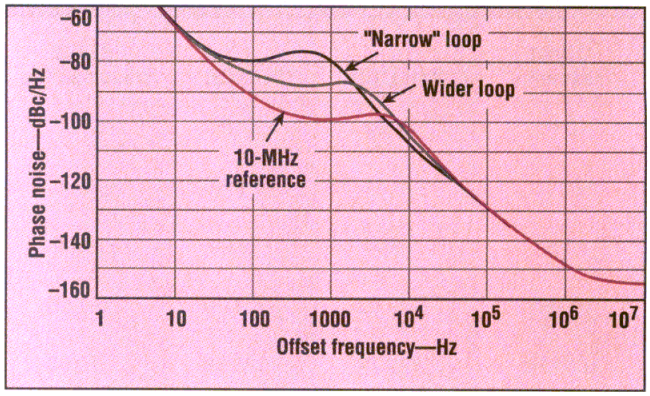
happen to the noise if the phase-detector frequency is 10 MHz instead of 1 MHz (changing the frequency resolution from 1 to 10 MHz). This changes the main division from 1000 to 100 and the reference oscillator goes directly into the phase detector. This is assumed due to operation at 10 MHz—the main divider noise floor is increased by 10 dB from –155 dBc/Hz to –145 dBc/Hz. The optimum loop bandwidth is computed to be ~6.6 kHz (results are shown in Fig. 26. The overall phase noise is improved with the 10-MHz reference.



25. Finding an optimum loop bandwidth involves resolving for loop component values and replotting the noise curves.

For best noise performance, it is preferable to keep the division as low as possible. This reduces the noise inside the loop (phase detector, divider, reference) with respect to the output. This explains the appeal of fractional-N division where one can achieve good frequency resolution with a higher reference frequency and, consequently, lower N numbers, assuming 10-kHz resolution at 1 GHz. With a conventional divider, the multiplication is 100,000 or 100 dB. If one used a 1-MHz reference with 10-kHz fractional N steps, the multiplication would only be 1000, or 60 dB, a 40-dB improvement. A summary of the residual FM and PM for the various cases is shown in Table 1.

PSPICE NOISE MODELS

From Part 2, it was shown that one could model the various noise sources using PSPICE elements, including resistors, diodes, and integrators. The details of the development of the model for the VCO noise were shown along with the MathCAD calculations of the PSPICE elements. Figure 27 shows the complete PSPICE model of the PLL with the three noise sources. The loop is in the upper-right-hand corner. The VCO noise is injected at the output of the VCO block (VCO gain plus the virtual integrator). The main divider noise is injected at the output of the main divider gain block (1/1000). The reference [temperature-compensated-crystal-oscillator (TCXO)] noise is first divided by 10 (0.1), to take into account the noiseless reference divider before being injected at the

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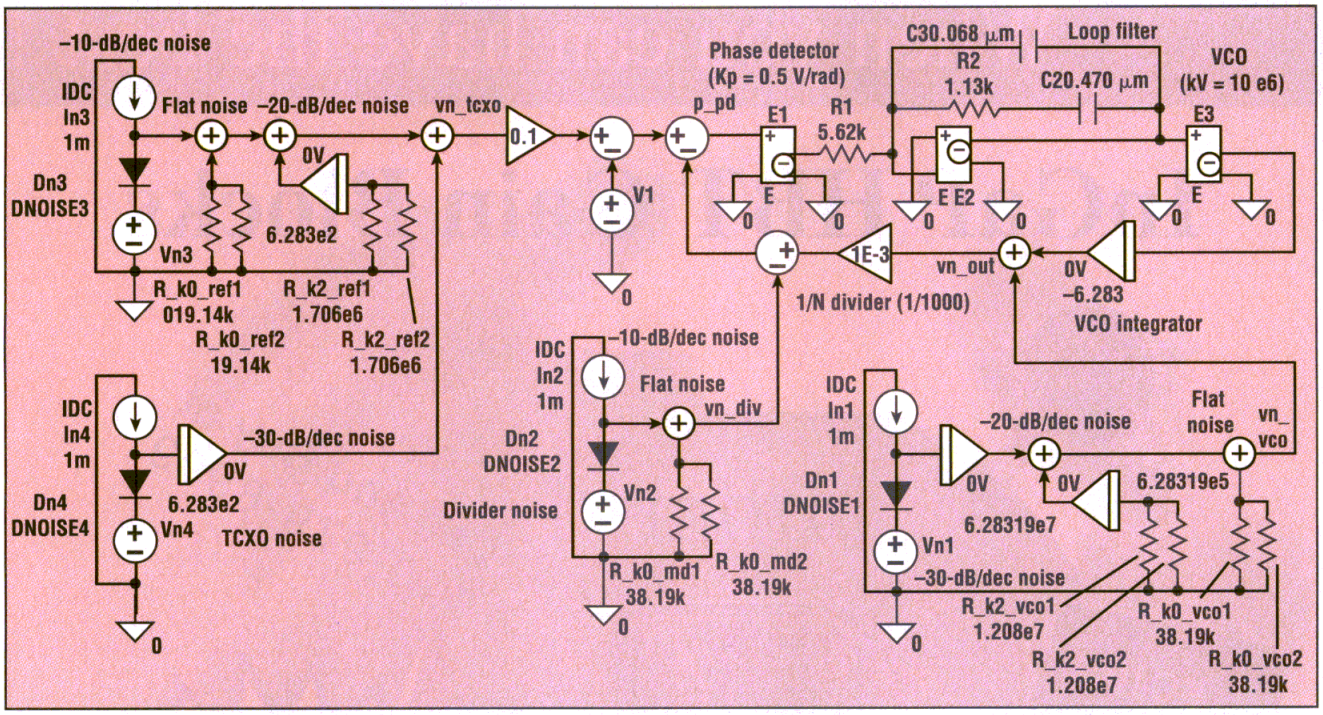
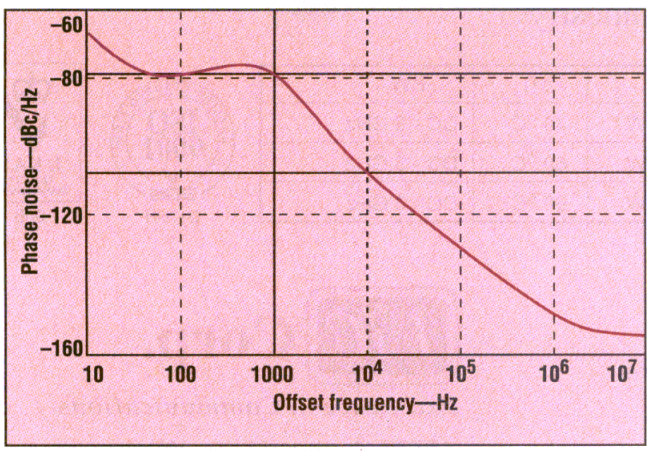


26. This is the complete PSPICE model of the PLL with the three noise sources.

phase detector. The voltage source V1 is used to check the reference closed-loop gain and does not relate to the noise model. The phase-noise output of the loop appears after the VCO phase-noise injection point (vn_out), shown in Fig. 27. Table 2 shows the MathCAD and PSPICE results for the phase noise at various offset frequencies. This is the double-sideband (DSB) phase noise. To convert to single-sideband (SSB) noise, a gain block (0.7071) can be inserted after vn_out.

An advantage in using PSPICE to model the phase noise is that PSPICE automatically takes into account the thermal noise of the resistors in the circuit (integrator input and feedback resistors in the example). Due to this, DSB phase-noise models must be used for the individual noise sources. This thermal noise explains the small discrepancy between the MathCAD and PSPICE results. Noiseless resistors can be simulated in PSPICE using ideal conductance, which shows almost perfect agreement with MathCAD. To show an extreme case, a 50-kΩ resistor is inserted between the opamp and the VCO. The thermal noise from the 50-kΩ resistor will modulate the control line to the varactor in the VCO and "pollute" VCO noise. It is possible to calculate the pollution from the small-signal modulation model. The noise voltage is transformed to FM with a gain of Kv (Hz/V), and then the SSB noise level can be computed.



27. This plot showing the phase-noise output of the loop assumes a probe placed after the VCO phase-noise injection point.

$$e_{n\_series} = (4kTBR)^{0.5} \qquad (27)$$

$$dB_{SSB\_noise} = 20 \cdot \log\left(e_{n\_series} K_v / \sqrt{2} f_m\right) \qquad (28)$$

The DSB phase noise (which is 3 dB greater than this equation) is –90.8 dBc/Hz at 10-kHz offset. Remembering that the "raw" VCO noise is ~–108 dBc/Hz, putting this 50-kΩ resistor in series with the control line degrades the VCO noise performance. PSPICE predicts the noise with the 50-kΩ series resistor to be –90.56 dBc/Hz. This simple experiment indicates why a large unbypassed series resistor should not be connected directly to the input of a VCO. ••

For Further Reading

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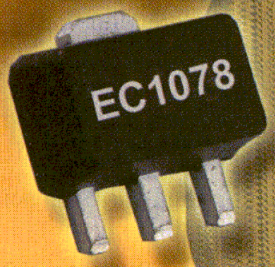

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EC 1078	19.5 dB	21 dBm	37 dBm	120 °C/W	60 °C	DC - 3 GHz
EC-1119	14.8 dB	18.6 dBm	36 dBm	150 °C/W	60 °C	DC - 3 GHz

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Simulate And Build A Ku-Band DRO

Learn how to construct a dielectric resonator oscillator for high-frequency communications systems based on just a few dollars worth of parts.

Ku-Band DRO, Part 1

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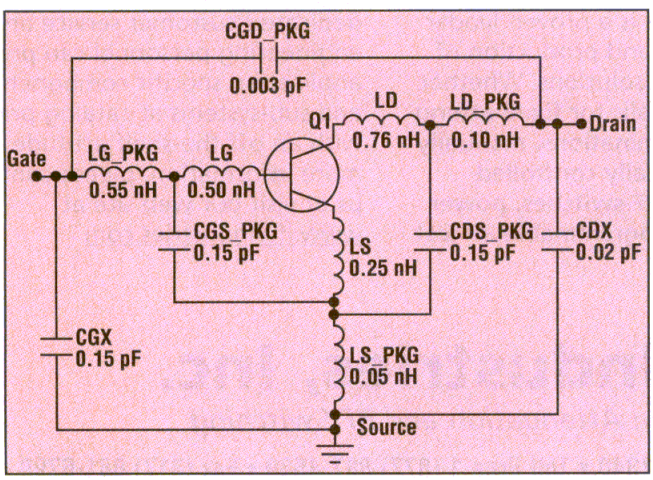
DIELECTRIC resonator oscillators (DROs) are widely used in digital microwave radios and other high-frequency systems. They deliver low phase noise and stable output frequencies with temperature, and represent a good compromise of cost, size, and performance. What follows is part 1 of a design strategy for a 11.25-GHz DRO based on a low-cost plastic-packaged gallium-arsenide (GaAs) metal-semiconductor field-effect transistor (MESFET). The DRO is ideal for use as the local oscillator (LO) of a Ku-band low-noise-block converter (LNB) in a Digital Direct Broadcast System (DDBS).

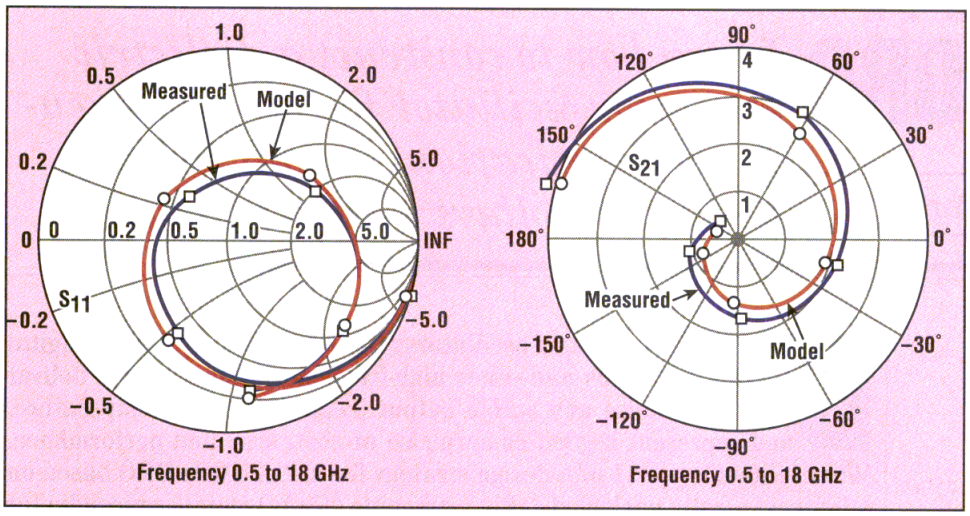
In a DDBS system, the DRO must exhibit low phase noise in order to meet the digital-modulation scheme and bit-error-rate (BER) requirements. It also must have minimal frequency drift over temperature to keep the receiver locked into the selected channel. It should provide enough output power to directly drive the mixer downconverter (usually a diode ring or an active GaAs FET mixer). Because the DC bias (+8 VDC) is usually supplied through the intermediate-frequency (IF) feed from an indoor unit, the voltage requirements are usually not a constraint. However, current draw remains a limitation and should be set to the lowest value possible while still meeting the output-power specifications. An oscillator for DDBS should be compact and light, since the LNB will be a small outdoor unit located at the focal point of an antenna through a light supportive pole. These last needs were met by using 0603 (60-mils-by-30-mils components) surface-mount-technology (SMT) components and by laying out the components within a tightly enclosed cavity. These choices resulted in a 0.950 × 0.750 × 0.500-in. (2.413 × 1.905 × 1.27-cm) design that also includes the metal cavity, tuning screw, and output connectors. Table 1 provides a summary of design goals and simulated/measured performance for the Ku-band DRO.

In choosing a transistor for a DRO, the device should provide low noise figure and enough loop gain at the maximum operating junction temperature and under large-signal conditions. For stable oscillation, a good rule of thumb is to use a transistor with a transition frequency (f_T) at least two to three times the operating frequency. For Ku-band, the GaAs MESFET provides the best compromise high (f_T), good loop gain, and output power.[1]



1. This schematic circuit diagram represents a nonlinear (large-signal) model of the plastic-packaged NEC NE72218 MESFET device.

NEC developed the NE72218 MESFET device for Ku-band oscillator designers. This epitaxial-grown, recessed-gate transistor provides high performance and low phase noise for oscillators to 14 GHz. The transistor is housed in a single four-pin SOT-343 surface-mount plastic package measuring 1.25 × 2.00 mm. With high saturated drain current (I_{DSS}), the NE72218 can also provide enough output power under different bias conditions to drive most Ku-band mixers, even where a resistive buffer has been added. The device is fabricated with optimized ion-implantation technology to minimize flicker noise and provide the lowest 1/f noise performance currently available with GaAs devices (for low phase noise).



2. These plots show the measured and modeled data for the S_{11} and S_{21} responses for the NE72218 MESFET.

The NE72218 features an 0.8-μm recessed gate and triple epitaxial technology. The gate width is 400 μm. In amplifier circuits, it achieves typical gain of 5 dB at 12 GHz when biased with 30 mA and +3 VDC. The output power at 1-dB compression under those same bias conditions is typically −90-dBc/Hz offset 10 kHz from an 11-GHz carrier. The phase noise is typically −110-dBc/Hz offset 100 kHz from an 11-GHz carrier. The device is rated for typical saturated drain current of 60 mA and typical gate-to-source cutoff voltage of −2 VDC. The maximum gate-to-

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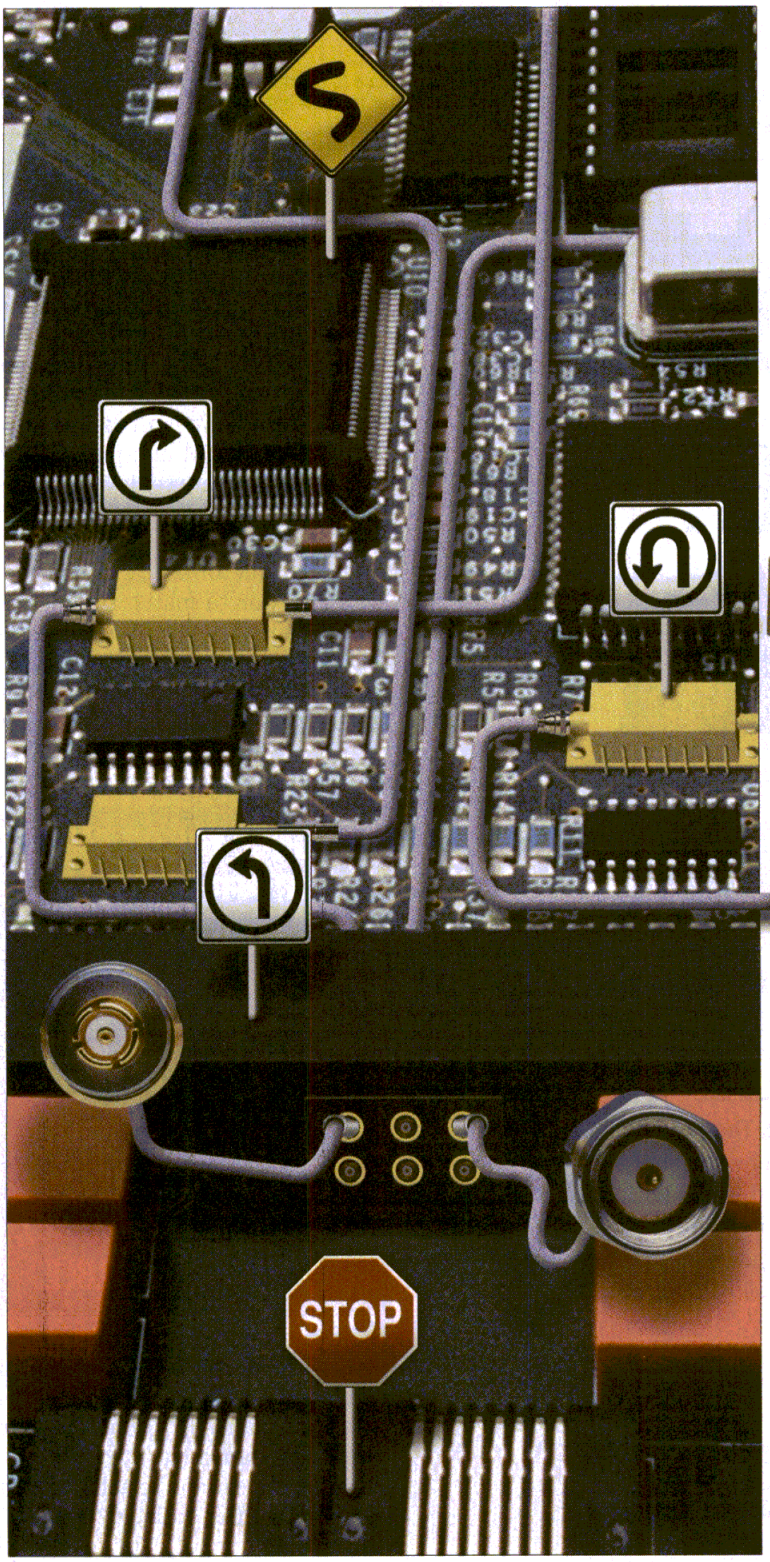

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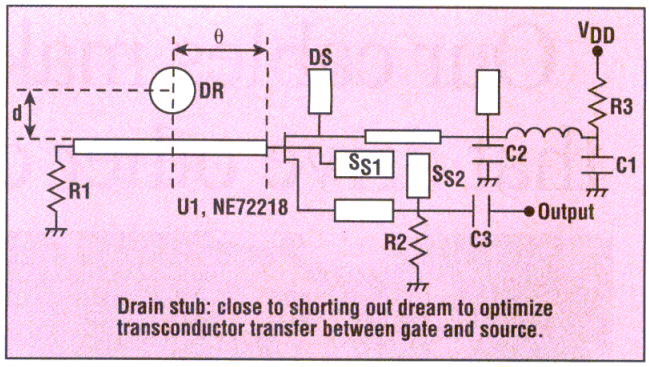
source voltage rating is −5 VDC and the maximum drain-to-source voltage rating is +5 VDC.

The choice of a FET nonlinear model is determined by evaluating the device's measured DC characteristics and comparing these to the characteristics of available nonlinear models. Various models implement a device's DC current-voltage = I-V characteristics differently.[2] The best model for the NE72218 was found to be Triquint's Own Model (TOM) from TriQuint Semiconductor (Hillsdale, OR)



3. The NE72218 transistor and a dielectric resonator (DR) are combined in this circuit to form a 11.25-GHz dielectric resonator oscillator (DRO) for DBS applications.

since the MESFET shows an almost linear increase in drain current with increasing drain voltage at lower gate voltages and an approximately constant drain current with respect to increasing drain voltage at higher gate voltages.

DC-parameter extraction is necessary in order to match the model to the measured I-V curves. The main DC parameters affecting the I-V curves are the nonscalable portion of the threshold voltage (VTO), the current saturation parameter (α), the transconductance coefficient (β), the DC drain-pull coefficient (γ_{DC}), the power-law exponent (Q), the output feedback coefficient (δ), the gate ohmic resistance (R_G), and the source ohmic resistance (R_S). A good fit to the AC data cannot be achieved until a good DC fit is obtained. When the model accurately predicts the device's DC characterization, the AC parameters can then be adjusted.

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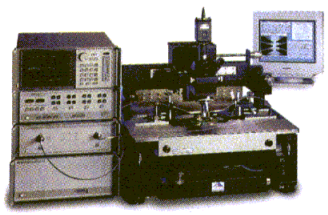

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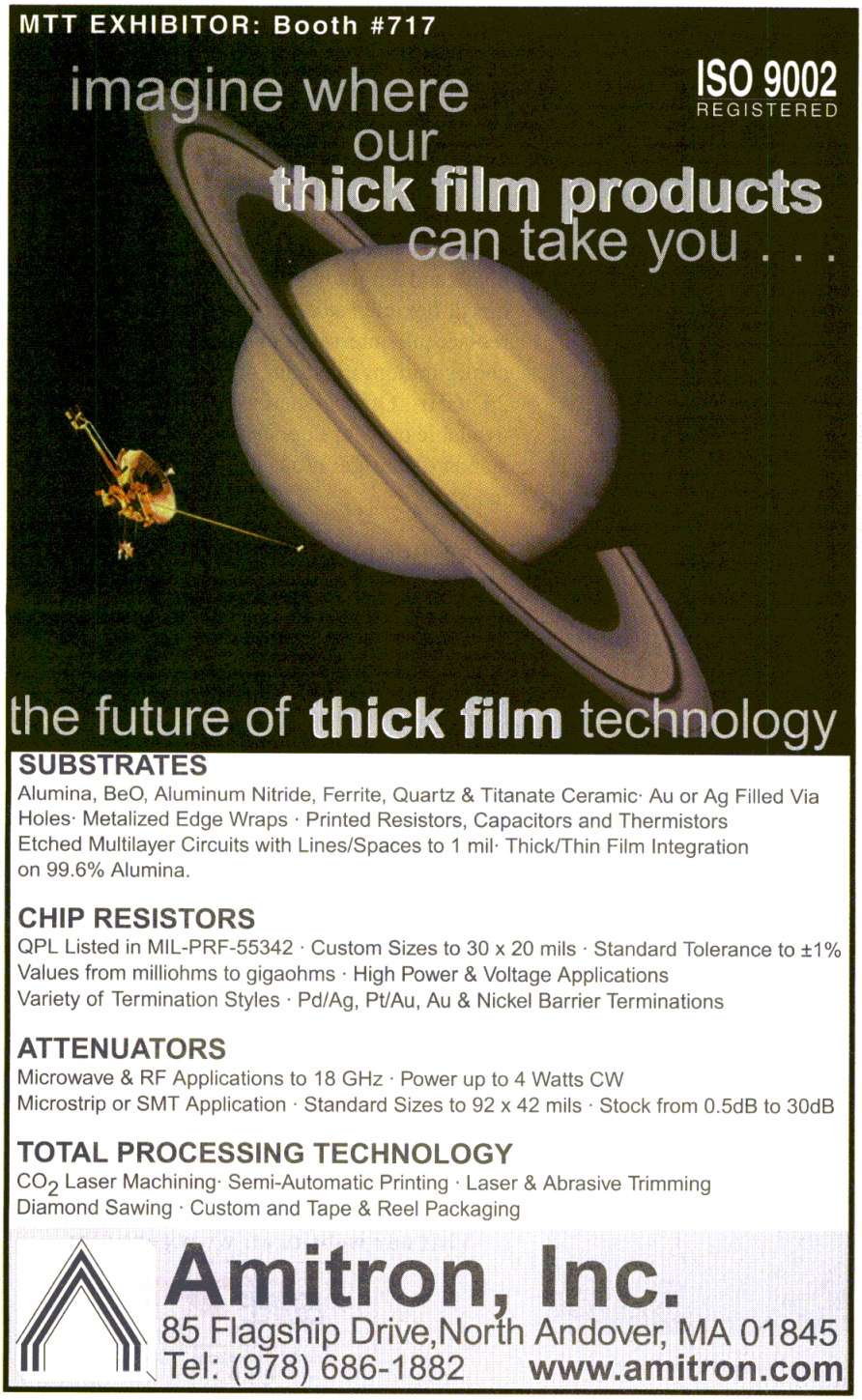
The TOM model parameters that most affect the AC prediction of the model are the transit time under the gate (τ), the AC drain-pull coefficient (γ), the drain-source capacitance (C_{DS}), the zero-bias gate-source capacitance (C_{GS0}), the zero-bias drain-source capacitance (C_{DS0}), R_G, R_S, and the parasitic elements of the transistor package. Once the DC and AC performance of the model is satisfactory, the model can be optimized to fit measured power and noise data (including 1/f noise), where applicable.

Model parameters typically affect more than one type of simulation response. The value of a parameter that results in the model providing the best S-parameter fit may not provide the best fit to measured noise data across a wide range of biases and frequencies. At California Eastern Laboratories, nonlinear models are developed to fit the widest range of biases, frequencies, and applications as possible.[3] There is usually a trade-off in device-model performance when developing this type of model. In general, the DC and AC parameter prediction is approximately equivalent. Then, depending on the targeted application of the device, either the power or the noise performance of the device model is optimized. Sometimes the AC performance of the model is slightly degraded to improve the power or noise prediction of the model. However, the parameters AF and KF are the only model parameters which affect 1/f noise prediction and no compromises to the AC performance need to be made.

THE TWO MOST CHALLENGING ASPECTS OF THE DESIGN WILL BE TO MEET THE LOW PHASE-NOISE SPECIFICATIONS AND THE GOOD FREQUENCY STABILITY OVER TEMPERATURE.

The device model for the NE72218 was extracted over the following ranges: V^{DS} = 0 to +5 VDC and V_{GS} = 0 to –1.4 VDC for DC; V_{DS} = +2 to +4 VDC and I_D = 10 to 40 mA for AC; a frequency range of 0.5 to 18 GHz; and 1/f noise measurements at V_{DS} = +3 VDC and I_D = 30 mA as well as at V_{DS} = +3 VDC and I_D = 40 mA. Figure 1 presents the final device model while Fig. 2 compared the results of the extracted device model to measured data at the desired DRO bias of V_{DS} = +4 VDC and I_{DS} = 20 mA.

The choice of a DRO topology can be made from four different classes—reaction-type, transmission, parallel-feedback, and reflection-type DROs. A reaction-type DRO is a free-running oscillator with enough feedback



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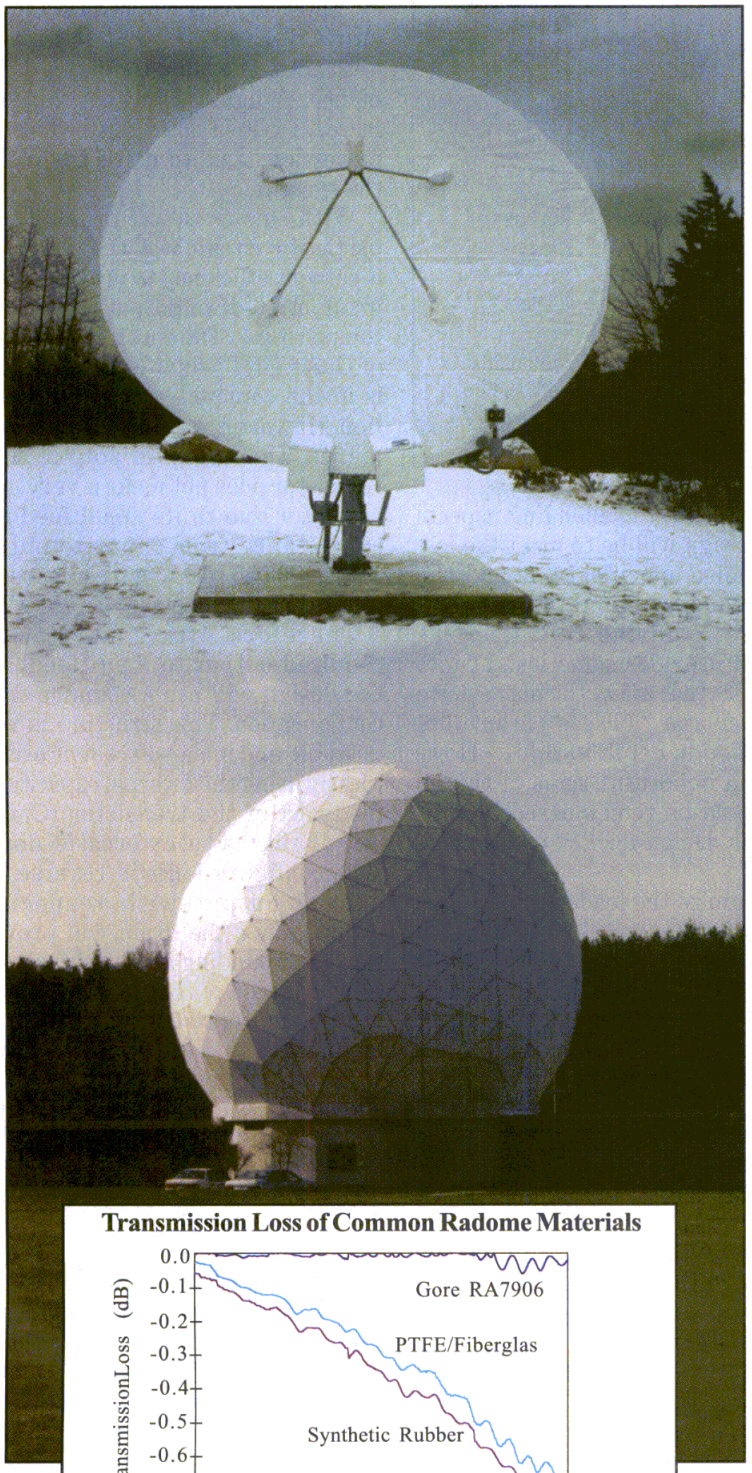

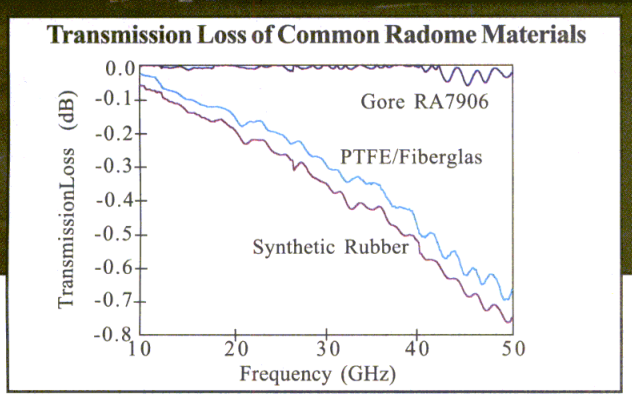

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Table 1: Design goals for 11.25-GHz DRO with simulated and measured results

Parameter	Specification	Simulation	Measured
Supply voltage (V)	8.0 ± 0.25	6.0	6.0 ± 0.25
Supply current (mA)	20 ± 2.5	18	18
Operating frequency range (GHz)	11.25	11.25	11.25
Output power (dBm) at 50 Ω	+7.0 ± 2.5	7.2	6.5
Phase noise (dBc/Hz)			
Offset 1 kHz	–58	–61.3	–62.8
Offset 10 kHz	–80	–91.3	–89.4
Offset 30 kHz	–90		–97.6
Offset 100 kHz	–100	–120.4	–112.3
Offset 1 MHz	–120	–131.2	–130.0
Frequency pushing (kHz)	±800 (max)	Not simulated	±200
Frequency pulling (MHz)	±2.0	Not simulated	±1.0
Harmonics (dBc)	–40 (min)	–20 (min)	–50 (min)
Spurious content (dBc)	–80 (min)	N/A	–80 (min)
Operating temperature (°C)	–55 to +80	Not simulated	–55 to +80
Temperature stability (MHz)	±2.0	Not simulated	±0.7

to oscillate in the desired frequency range. The frequency of oscillation is then stabilized with a dielectric resonator (DR) on the output. Due to the design method, this oscillator usually has a high spurious content and does not provide low phase noise.[4]

The parallel-feedback and the transmission DRO uses the DR between two transmission lines to provide the frequency-selective loop feedback between the input and the output of an amplifier design. Usually, these two configurations do not support much adjustment during on-the-bench tuning and are generally complex to model with a simulator. Due to the tight enclosure required in LNB designs, most DROs display some level of feedback within the cavity.[5] In most cases, however, that effect is undesired and rarely simulated.

Finally, the reflection-type DRO uses the concept of negative resistance where the resonator is placed near a terminated microstrip line connected to the input port of an unstable amplifier. Near its resonant frequency, the dielectric resonator reflects power back to the amplifier, causing an oscillation buildup between the two components that can be tapped into. In this configuration, the coupling between the resonator and the transmission line is easier to model and spurious oscillations can be readily avoided (Fig. 3).

The two most challenging aspects of the design will be to meet the low phase-noise specifications and the good frequency stability over temperature. Studying D.B. Leeson's equation provides some insight into the factors that affect the phase noise of an oscillator.[6] (These parameters are studied in depth in ref. 7.) However, the important rules of thumb that should be remembered to optimize this design for low phase noise are:

- Maximize the loaded quality factor (Q_L) of the tuned circuit. This goal will be achieved with a very high-Q unloaded DR that will only be lightly coupled to the circuit to limit loading effects.
- Choose a device with a low flicker noise. The 1/f noise of the NE72218 makes this device a prime choice for this application.
- Maximize the available source power (P_{avs}) power at the input of the oscillator. A light coupling of the DR will ensure that most of the circuit's available power is stored in the DR and available at the FET's gate.

In addition, the phase noise is also dominated by the signal-to-noise ratio (SNR) at the input (SNR_I), which depends on the noise figure of the active device and on P_{avs}. Consequently, design rules that make good low-noise amplifiers (LNAs) also apply to low phase-noise oscillators.[8] Usually, a typical oscillator runs at approximately 20-percent efficiency. However, this achievement also depends on how much output power is tapped out of the circuit. Higher output power means higher efficiency, however, this will reduce the circuit's loaded Q, which in turn degrades the phase-noise performance. A light output coupling will increase phase noise but reduce the power available to drive the rest of the system.

With this trade-off in mind, this particular circuit was set to achieve 5-percent efficiency to provide a minimum of +6-dBm output power over temperature. The most common use of GaAs FET amplifiers at Ku-band is in the common-source configuration. However, without feedback elements, the common-source FET transistor does not make a very good oscillator due to its small feedback capacitance from input to output (C_{GD0} = 0.055 pF) when compared to other capacitance values in the FET.

Therefore, to generate the required output to input feedback, the design will use a common-drain configuration. This structure is very unstable and makes excellent oscillators by using the internal capacitance feedback of the transistor (C_{GS0} = 0.85 pF) instead of external feedback. This configuration will reverse the normal output with respect to ground since the drain will provide the RF grounding and all signals will be referenced to that port. By choosing the appropriate drain open stub length (DS in Fig. 3), the designer will determine the frequency at which the series negative resistance will be generated on the gate's reflection port. That port should be set to a quarter wavelength at the desired frequency of oscillation.

SELECTING REACTANCE

Selecting the correct reactance at the source (Ss1) maximizes the magnitude of the reflection coefficient at the gate terminal. Adjusting these two parameters will provide the required amount of negative resistance at the needed frequency. Adjusting the output matching network and the amount of output coupling (C3) will drive the output power and loaded Q (and, therefore,

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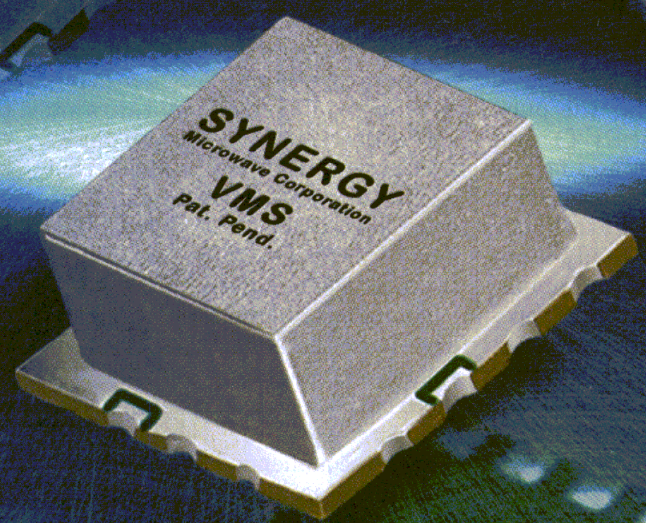

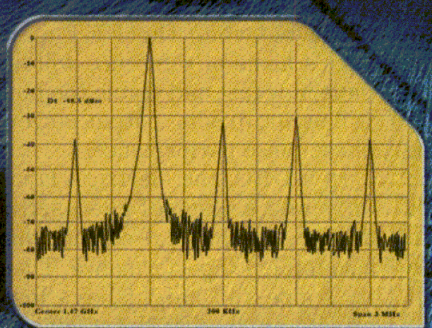

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the phase noise) of the oscillator. As the amplitude of the oscillation increases, the active devices start saturating, and magnitude of the negative resistance decreases until it is equal to the equivalent resistance presented by the DR at the resonant frequency. For a steady-state oscillation to occur, the following condition needs to be met in the input-reference plane of the active device:

$$\Gamma_{in} \times \Gamma_R = 1$$

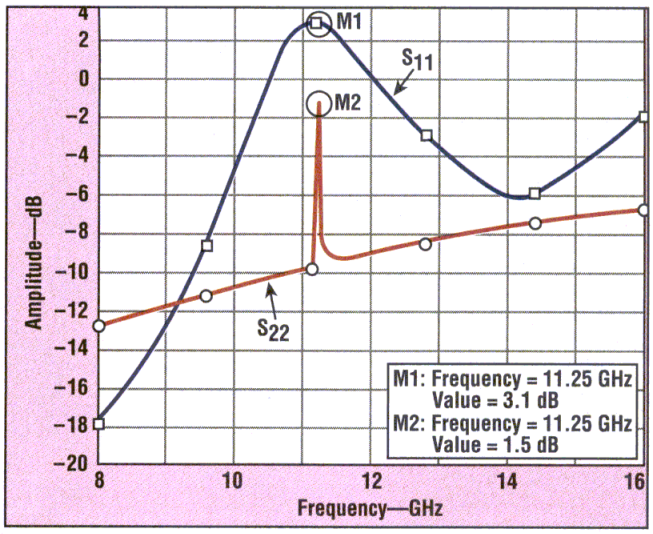
where:

Γ_{in} = the return loss provided by the puck coupled with the transmission line, and

Γ_R = the reflection coefficient (<1) exhibited by the negative resistance amplifier.

Consequently, for a particular amplifier circuit, the designer will need to adjust the puck coupling in terms of phase (how far the puck should be from the FET) and strength of the coupling (how far the puck should be from the transmission line). Eventually, the oscillator should achieve under steady-state oscillation, a loop phase and amplitude of 0 and 1, respectively. Therefore, the next task is to synthesize the gate load by coupling the resonating structure.

As mentioned earlier, the DR is key to the performance of the oscillator in that it defines the Q of the circuit and locks the frequency. The high unloaded Q (Q_0) results in the super-low-noise performance and is defined by dielectric loss tangent of the material and the environmental losses. Recent developments in ceramic material technology have resulted in performance improvements including Q_0 up to 12,000 at 12 GHz and small controllable temperature coefficients. With proper temperature compensation, nearly constant frequency over temperature can be achieved. This parameter also depends on the characteristics of the FET's S-parameters over temperature and the cavity's mechanical-expansion coefficient. Finally, the last important parameter defining the DR is the dielectric constant which ultimately determines the resonator dimensions as well as the cavity (and circuit-design) dimensions. At present, commercially available temperature-stable DR materials exhibit dielectric constants of approximately 36 to 40. These dielectric resonators also come in different forms and modes, however, the cylindrical shape transverse-electric (TE) mode has been widely accepted as the most advantageous one.



4. Linear circuit simulation was used to determine the S_{11} and S_{21} transmission losses from the FET and DR.

Numerous references are available describing the advantages of different ratio in height (H) and diameter (D) from dielectric resonator manufacturers.[7] However, a choice of H/D = 0.4 is recommended to avoid spurious modes oscillations and achieve and optimal Q_0.

Choosing the right DR for the application is key to meeting the DRO's specifications. Due to the series negative resistance of the FET in its feedback circuit, the puck is coupled in series to the circuit through a 50-Ω line and used in a band-reject filter mode. At the desired frequency, the puck will reflect any incoming power back to the FET, producing a buildup between the active device and the DR. Coupling between the microstrip and the resonator is accomplished by orienting the resonator's magnetic momentum perpendicular to the microstrip plane at a distance d (Fig. 3).

The position of the resonator relative to the transmission line determines the oscillator's stability, output power, and phase noise. Optimum positioning can be tricky but is greatly aided by linear and nonlinear simulations. Adjusting d increases or decreases the amount of coupling. A higher coupling provides more output power and robustness of oscillation build-up, however, it reduces the loaded Q and, therefore, the phase-noise performance. A lower coupling will improve phase noise but reduces the output power, and under certain circumstances, the oscillator could fail to start oscillating. Therefore, when designing a low-noise oscillator, the compromise will be to set the distance d small enough that the oscillator will always start (under quick and slow turn-on and at all temperatures) and provides enough power, but large enough to get high loaded Q and low phase noise. Finally, as more energy is stored in the DR, the temperature characteristics of the DRO more closely follow that of the DR. Consequently, a lighter coupling will also provide more control of the LO drift over temperature. The phase relationship of the puck to the active device is as equally critical to efficiently creating an oscillation buildup. The electrical length (θ) [Fig. 3] representing the physical distance between the FET and the puck's plane of reference will determine how fast and stable the buildup will occur, driving output power and phase-noise performance.

In a DRO, the electrical layout is only one aspect of the oscillator design. Mechanical considerations also influence performance. The cavity's size and height have loading effects on the oscillator which can reduce the phase-noise performance and create an unwanted frequency drift over temperature. Under best conditions, the DR would be free to resonate in free space, but due to obvious real-estate consideration, the oscillator must be constrained

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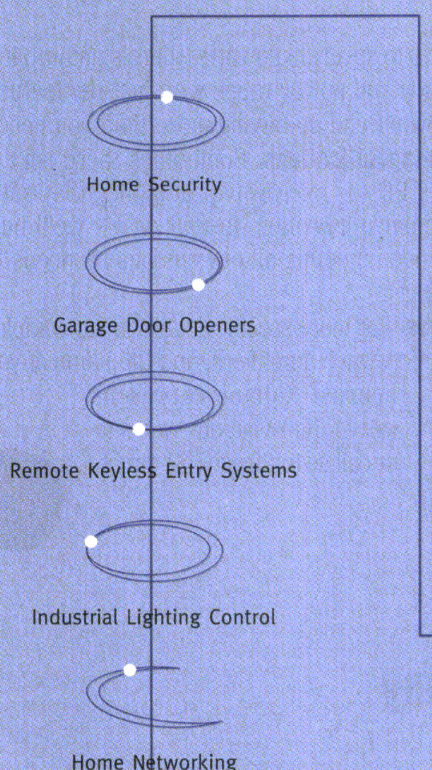

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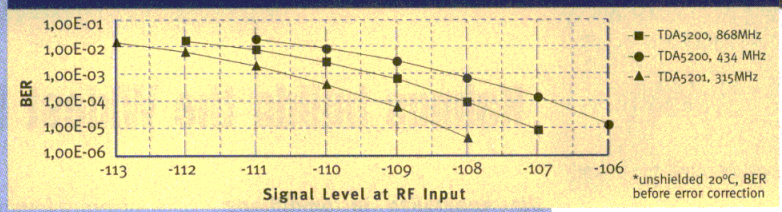

*unshielded 20°C, BER before error correction

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(**50 Ohm source impedance, BER=2E-3 (before error correction), average power level, Manchester encoded data rate=4kbit/s, 280 kHz IF filter bandwidth)

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within a shielded cavity. References 4 and 7 analyze in depth the effects of enclosed cavities, and rules-of-thumb dictate that in order for the cavity to have a reasonable thermal and loading effects, the cavity should be at least three pucks high and three puck's diameter wide. This height requirement is one reason most DRO designers prefer to set their DR on a standoff, so that the housing or printed-circuit board (PCB) on which the DR usually rests does not affect the resonator's performance. The PCB material's mechanical integrity also needs careful consideration due to drift over temperature and long-term aging effects, especially if the cavity is resting on the PCB.

For example, Rogers 4003 material is stable enough to support only minimum frequency aging while providing a low tangent loss at 12 GHz. This low insertion loss provides for low noise-figure performance as well as a high Q for the circuit. However, other low-loss materials such as Teflon mesh compound do not have the same mechanical integrity and the PCB thickness has a tendency to change over time. Although this has little bearing in an amplifier design, in an oscillator, the cavity and ground-plane relations will slowly change over time, providing slow oscillator drift or aging effects that could obsolete the LNB after a few years.

Finally, the fine tuning and adjustment of the DRO will be set through a tuning screw that will increase the DR's resonant frequency as it closes the electrical field above the puck. This should provide as much 80 MHz of tuning range. However, it is important to notice that tuning the frequency with a tuning screw is achieved at the cost of reduction in unloaded Q and temperature stability. This increase in temperature stability is due to the increasing slope of the tuning curve as the metal plate gets closer to the DR surface. Therefore, if no more than approximately 10 MHz of loading can be achieved, the design will provide a good compromise between adjustment for manufacturing variations and high performance. Next month, this two-part article will conclude with the use of linear simulation to create a first-approximation DRO design (Fig. 4). A prototype circuit will be constructed and tested, and a final design and list of required components will be presented. ●●

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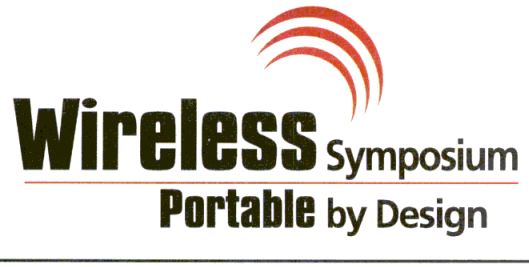

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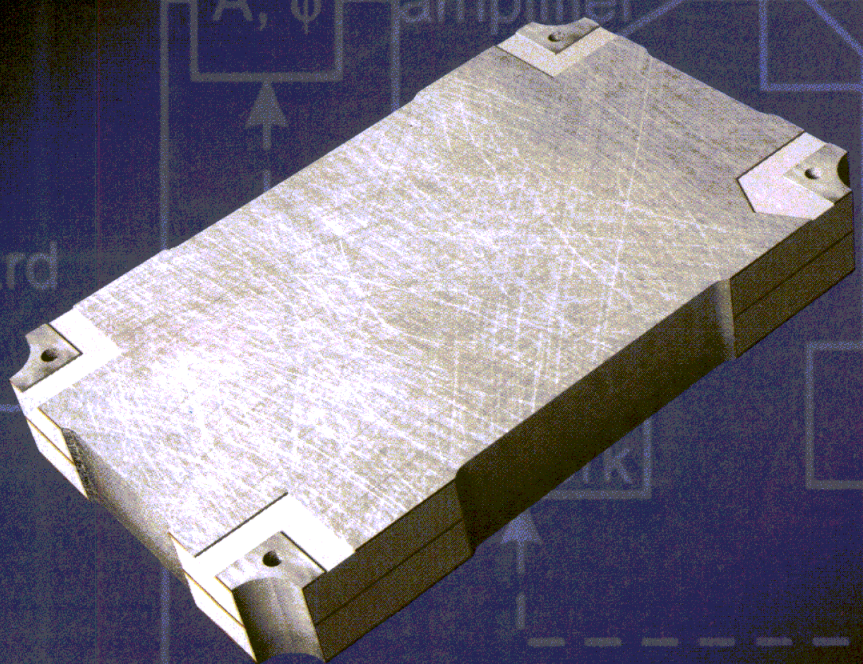

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Update The Design Of Image-Parameter Filters

Computer-aided design has breathed new life into this old-fashioned method for designing filters.

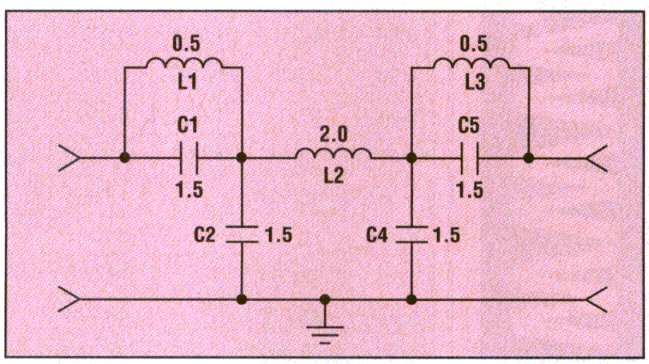
Richard M. Kurzrok
Professional Engineer
RMK Consultants, 82-34 210th St., Queens Village, NY 11427-1310; (718) 776-6343, FAX: (718) 776-6087, e-mail: rmkconsulting@aol.com.

ORIGINAL LC filter designs used the classic method of image parameters. This design method originated at Bell Telephone Laboratories when it was part of American Telegraph and Telephone Company (AT&T). The method was in vogue for more than four decades[1,2] before it was replaced by modern network synthesis. But the advent of computer-aided design (CAD) and analysis has substantially mitigated the shortcomings of image-parameter designs, facilitating the design of useful filters.

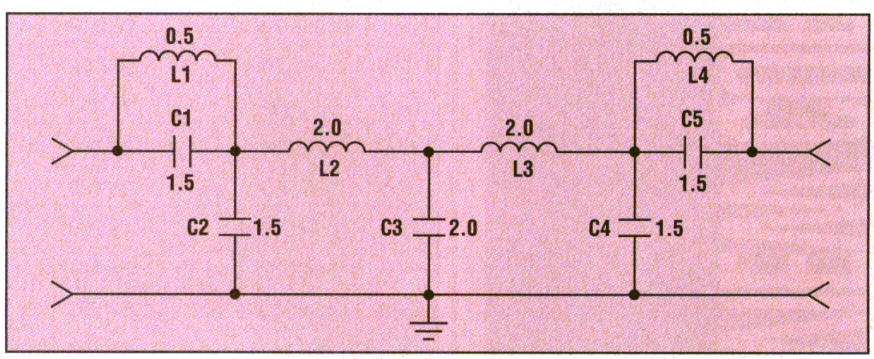
Image-parameter designs suffer from three significant problems:

1. Matching source and load impedance to image impedance is not practical.
2. Image attenuation and phase-shift values are not correct when source and load impedances are resistive.
3. Filter-response-shape degradation due to incidental dissipation is not readily predictable.

A very practical use of image-parameter design is the composite



1. This schematic shows a lowpass filter using one interior constant-K full section.



2. A lowpass filter using two interior constant-K full sections can be seen.

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Table 1: Computed responses versus inductor unloaded Q with m = 0.5 for terminating half-sections and one interior constant-K full section

	QUL = 1 × 10⁶	QUL = 1 × 10⁶	QUL = 50	QUL = 50	QUL = 25	QUL = 25	QUL = 10	QUL = 10
X	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR
0	0	1.0	0.257	1.060	0.506	1.120	1.214	1.300
0.05	0	1.0	0.257	1.060	0.507	1.120	1.216	1.299
0.10	0	1.001	0.258	1.059	0.509	1.119	1.222	1.296
0.15	0	1.005	0.260	1.059	0.512	1.117	1.232	1.292
0.20	0	1.012	0.262	1.059	0.517	1.115	1.245	1.286
0.25	0.001	1.022	0.265	1.060	0.523	1.114	1.260	1.279
0.30	0.001	1.037	0.268	1.066	0.529	1.115	1.279	1.271
0.35	0.003	1.056	0.273	1.077	0.537	1.119	1.299	1.265
0.40	0.006	1.080	0.279	1.094	0.547	1.128	1.321	1.260
0.45	0.011	1.108	0.287	1.116	0.558	1.142	1.346	1.258
0.50	0.018	1.136	0.297	1.141	0.572	1.160	1.373	1.258
0.55	0.025	1.162	0.307	1.164	0.586	1.178	1.403	1.258
0.60	0.030	1.181	0.317	1.180	0.601	1.191	1.436	1.256
0.65	0.031	1.185	0.324	1.183	0.615	1.191	1.471	1.247
0.70	0.025	1.165	0.328	1.164	0.627	1.172	1.511	1.225
0.75	0.013	1.115	0.328	1.117	0.641	1.129	1.565	1.192
0.80	0.001	1.032	0.337	1.049	0.671	1.079	1.651	1.173
0.85	0.005	1.071	0.377	1.083	0.744	1.113	1.818	1.221
0.90	0.020	1.145	0.465	1.156	0.901	1.194	2.154	1.347
0.95	0.005	1.069	0.639	1.140	1.242	1.250	2.900	1.574
1.00	1.938	4.000	2.731	3.500	3.485	3.259	5.523	3.181
1.05	13.367		13.563		13.795		14.660	
1.10	30.723		30.751		30.798		31.047	
1.15	78.374		78.379		78.396		78.508	
1.20	43.266		43.274		43.290		43.380	
1.25	34.067		34.083		34.105		34.205	
1.30	30.069		30.094		30.124		30.242	
1.35	27.954		27.988		28.026		28.165	
1.40	26.770		26.813		26.858		27.018	
1.45	26.120		26.169		26.223		26.401	
1.50	25.802		25.859		25.919		26.115	
1.55	25.704		25.768		25.834		26.047	
2.00	28.389		28.493		28.597		28.917	
4.00	43.813		43.968		44.121		44.576	
6.00	54.010		54.174		54.337		54.819	
8.00	61.383		61.551		61.718		62.209	
10	67.142		67.312		67.480		67.976	

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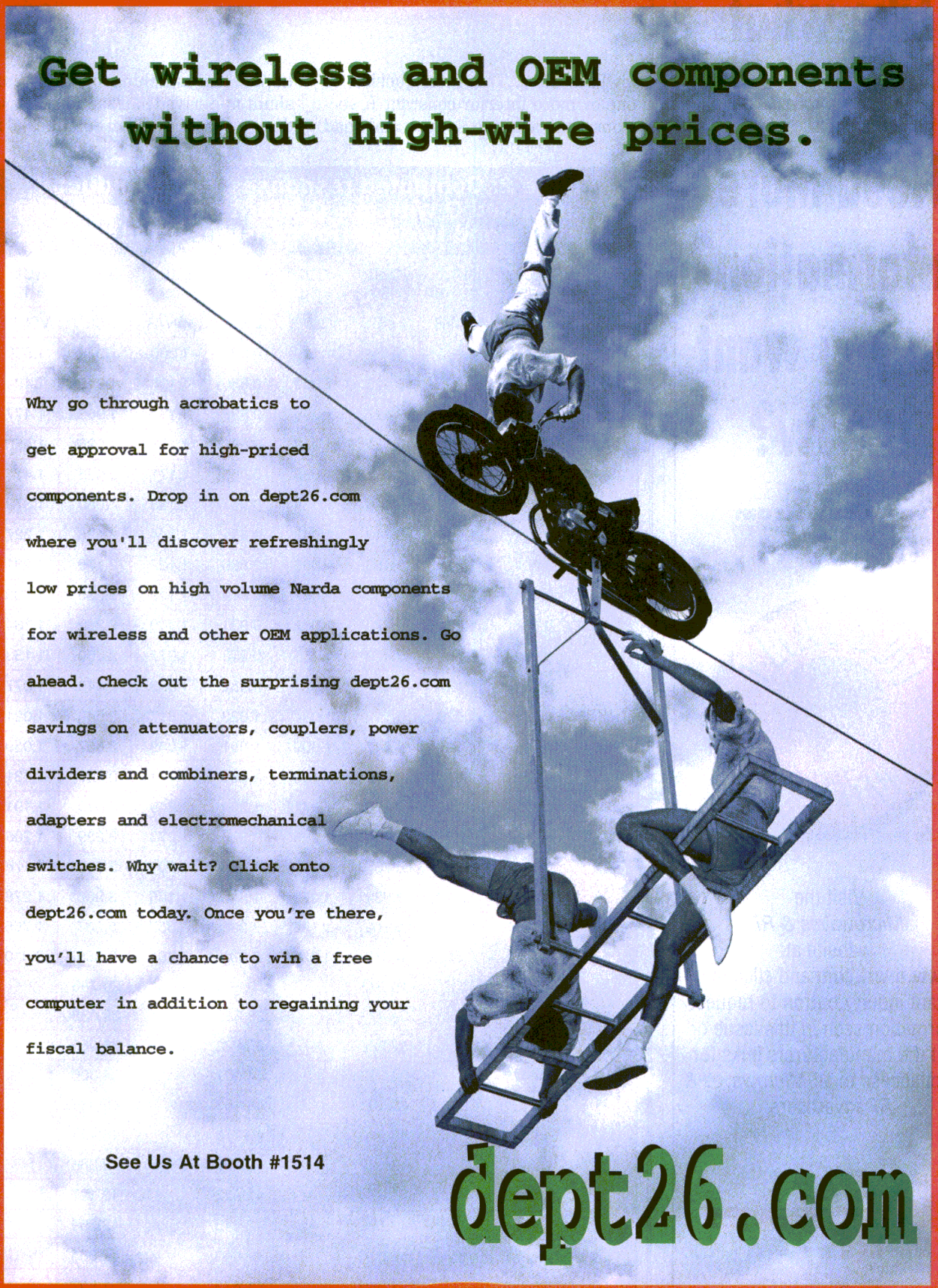

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lowpass filter. This filter consists of one or more interior constant-K sections that achieve source and load impedance matching by using mid-shunt m-derived half sections. Classic design practice sets m equal to 0.6 for

Table 2: Computed responses versus inductor unloaded Q with m = 0.5 for terminating half-sections and two interior constant-K full sections

	QUL = 1 × 10⁶	QUL = 1 × 10⁶	QUL = 50	QUL = 50	QUL = 25	QUL = 25	QUL = 10	QUL = 10
X	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR
0	0	1.0	0.424	1.100	0.828	1.200	1.938	1.500
0.05	0	1.0	0.425	1.099	0.830	1.198	1.945	1.494
0.10	0	1.002	0.427	1.096	0.835	1.192	1.965	1.477
0.15	0	1.008	0.431	1.092	0.844	1.183	1.996	1.450
0.20	0	1.017	0.435	1.088	0.855	1.171	2.035	1.414
0.25	0.001	1.030	0.440	1.084	0.867	1.158	2.080	1.373
0.30	0.002	1.046	0.446	1.083	0.881	1.144	2.128	1.327
0.35	0.004	1.062	0.453	1.085	0.896	1.132	2.179	1.281
0.40	0.006	1.074	0.461	1.087	0.912	1.119	2.233	1.233
0.45	0.006	1.075	0.470	1.082	0.931	1.103	2.292	1.184
0.50	0.004	1.061	0.479	1.065	0.953	1.077	2.360	1.133
0.55	0.001	1.027	0.492	1.030	0.982	1.038	2.440	1.077
0.60	0.001	1.028	0.512	1.024	1.023	1.017	2.540	1.021
0.65	0.011	1.107	0.547	1.099	1.081	1.089	2.667	1.059
0.70	0.038	1.204	0.602	1.189	1.164	1.174	2.827	1.131
0.75	0.072	1.295	0.670	1.271	1.264	1.249	3.019	1.191
0.80	0.086	1.326	0.728	1.298	1.365	1.273	3.239	1.208
0.85	0.049	1.236	0.757	1.217	1.458	1.201	3.511	1.176
0.90	0.002	1.043	0.832	1.079	1.646	1.135	3.996	1.278
0.95	0.002	1.046	1.185	1.192	2.293	1.357	5.267	1.755
1.00	3.010	5.828	4.512	4.222	5.889	3.761	9.385	3.810
1.05	18.796		19.045		19.385		20.786	
1.10	38.435		38.476		38.559		39.038	
1.15	87.774		87.786		87.822		88.069	
1.20	54.079		54.092		54.121		54.306	
1.25	46.111		46.131		46.163		46.335	
1.30	43.211		43.240		43.278		43.454	
1.35	42.095		42.132		42.177		42.364	
1.40	41.830		41.875		41.927		42.127	
1.45	42.034		42.086		42.145		42.359	
1.50	42.517		42.576		42.640		42.868	
2.00	51.264		51.369		51.475		51.807	
4.00	79.659		79.814		79.968		80.425	
6.00	97.055		97.220		97.383		97.866	

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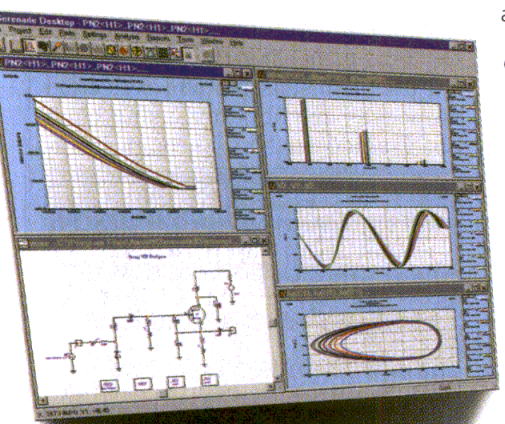

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the terminating half sections. This achieves optimum impedance matching. With m equal to 0.5, impedance matching is no longer optimum, but this is satisfactory for some applications. The filters described in this article use the m value of 0.5. This has the advantage of requiring no more than two different values of inductance and two different values of capacitance.[3,4]

Table 3: Computed responses versus inductor unloaded Q with m = 0.5 for terminating half-sections and three interior constant-K full sections

	QUL = 1 × 10⁶	QUL = 1 × 10⁶	QUL = 50	QUL = 50	QUL = 25	QUL = 25	QUL = 10	QUL = 10
X	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR	L-dB	VSWR
0	0	1.0	0.588	1.140	1.138	1.280	2.607	1.700
0.05	0	1.0	0.589	1.137	1.143	1.274	2.625	1.681
0.10	0	1.003	0.594	1.129	1.155	1.256	2.678	1.627
0.15	0	1.010	0.601	1.117	1.174	1.229	2.755	1.548
0.20	0	1.020	0.608	1.101	1.196	1.195	2.845	1.453
0.25	0.001	1.031	0.617	1.086	1.218	1.158	2.939	1.353
0.30	0.002	1.039	0.626	1.070	1.242	1.119	3.030	1.256
0.35	0.001	1.037	0.636	1.051	1.267	1.079	3.119	1.168
0.40	0	1.020	0.649	1.025	1.296	1.038	3.207	1.096
0.45	0	1.016	0.667	1.014	1.331	1.015	3.299	1.063
0.50	0.005	1.070	0.691	1.067	1.375	1.067	3.399	1.097
0.55	0.018	1.138	0.725	1.129	1.428	1.128	3.512	1.150
0.60	0.036	1.199	0.764	1.185	1.488	1.180	3.638	1.193
0.65	0.044	1.222	0.797	1.206	1.547	1.198	3.783	1.204
0.70	0.028	1.174	0.818	1.161	1.607	1.156	3.963	1.167
0.75	0.002	1.045	0.850	1.045	1.698	1.050	4.222	1.083
0.80	0.017	1.135	0.949	1.117	1.877	1.096	4.625	1.031
0.85	0.072	1.293	1.124	1.250	2.167	1.203	5.221	1.069
0.90	0.038	1.206	1.292	1.166	2.527	1.118	6.074	1.068
0.95	0	1.005	1.742	1.176	3.390	1.318	7.815	1.623
1.00	4.087	8.127	6.338	4.601	8.349	3.967	13.270	4.001
1.05	24.255		24.539		24.969		26.861	
1.10	46.142		46.195		46.313		47.017	
1.15	97.174		97.193		97.249		97.632	
1.20	64.891		64.909		64.952		65.232	
1.25	58.153		58.176		58.219		58.463	
1.30	56.352		56.383		56.430		56.664	
1.35	56.236		56.275		56.327		56.562	
1.40	56.892		56.939		56.997		57.237	
1.45	57.952		58.006		58.069		58.318	
1.50	59.236		59.296		59.365		59.623	
2.0	74.143		74.247		74.355		74.700	
3.0	98.131		98.272		98.413		98.840	

Three different composite lowpass filters are considered in this article. The schematics for these filters—using one, two, or three interior constant-K full sections—are shown in Figs. 1, 2, and 3. The circuit elements are normalized for the source/load impedance and the cutoff frequency of a single constant-K section.

Each normalized circuit element can be denormalized by multiplying all normalized element values by a normalizing inductance or capacitance:

The normalizing inductance, in μH, is:

$$L_0 = Z_0 / (2 \times PI \times f_c) \qquad (1)$$

The normalizing capacitance, in pF, is:

$$C_0 = 1 \times 10^6 / (2 \times PI \times f_c) \times Z_0) \qquad (2)$$

where:

Z_0 = the source/load impedance in ohms, and

F_c = the cutoff frequency (usually the 3-dB point) in MHz.

Filter transmission and reflection responses are obtained using computer-aided analysis through iterative ABCD matrix multiplications. Image-parameter, lowpass-filter insertion loss and VSWR versus normalized frequency variable x (where x = the actual frequency/cutoff frequency) have been computed for three composite filters with one, two, or three interior constant-K sections and with m = 0.5, m-derived, termi-

THE FILTERS THAT ARE DESCRIBED IN THIS ARTICLE USE THE M VALUE OF 0.5. THIS HAS THE ADVANTAGE OF REQUIRING NO MORE THAN TWO DIFFERENT VALUES OF INDUCTANCE AND TWO DIFFERENT VALUES OF CAPACITANCE.

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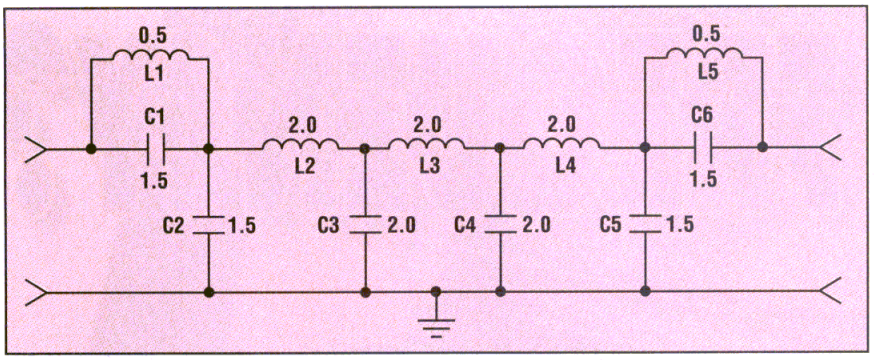

3. This schematic shows a lowpass filter using three interior constant-K full sections.

Table 4: Parts for illustrative composite lowpass filter

Circuit element	Value	Realization	Quantity
C1, C2, C4, C5	150 pF	P-series polypropylene	4
C3	200 pF	CD-15 series dipped mica	1
L1, L4	0.125 μH	Micro metals T25-10 5T-#26	2
L2, L3	0.500 μH	Micro metals T25-10 14T-#26	2
In/out coaxial connectors	BNC female	Pomona 2447 panel receptacle	2
Enclosure	Al box	Hammond 1590A or Bud Cu-123	1
PCB	Cut by hand	Vector board 169P44C1	1
Standoffs	Male/female	Amatom 9794-SS-0440	4

nating half sections. These responses are shown in Tables 1, 2, and 3 for unloaded Q's of 1×10^6, 50, 25, and 10.

It should be noted that x = 1 does not necessarily correspond to the 3-dB normalized cutoff frequency of the composite filter. For a lossless filter with one interior constant-K section, x = 1.0086 at the 3-dB cutoff. For a lossless filter with two interior constant-K sections, x = 1.00 at the 3-dB cutoff. For a lossless filter with three interior constant-K sections, x = 0.9965 at the 3-dB cutoff. For lossy filters, the relative normalized frequencies for the 3-dB cutoff will be different.

Table 5: Measured insertion loss versus frequency for illustrative lowpass filter

Frequency (MHz)	Insertion loss (dB)	Frequency (MHz)	Insertion loss (dB)
10	0.2	32	6.0
15	0.2	33	10.0
20	0.2	34	16.5
23	0.3	35	20.0
26	0.5	36	>35
28	0.8	38	>35
30	1.8	40	>35
31	3.2	50	>35

The author has designed, constructed, and tested an engineering model of a composite L-C lowpass filter. The filter exhibits a nominal 3-dB cutoff frequency of 31.831 MHz and a source/load impedance of 50 Ω. Figure 2 shows a normalized schematic of the filter, with two interior constant-K full sections and an m = 0.5 m-derived terminating half section. Table 4 lists the parts list for this filter.

Table 5 shows the data measured for the lowpass filter. Reasonable agreement between theory and experiment was obtained using a limited dynamic-range test setup. The effective unloaded Q's for the toroidal inductors were approximately 100.

Image-parameter filter-design methods can be used to obtain composite lowpass filters with satisfactory transmission and reflection responses. These filters can provide useful performance for inductor unloaded Q's that are as low as 10. They require a small number of different component values. These characteristics are compatible with the high-volume, low cost, surface-mount manufacturing techniques for wireless and other commercial applications.

Image-parameter, lowpass filters provide design simplicity, circuit symmetry, reasonable passband behavior, and sharp stopband selectivity. In some LC filter situations, the old-fashioned design method of image parameters is still a viable alternative to modern network synthesis. Image parameters are also used at microwave frequencies for transmission-line filters of non-commensurate line lengths.[5] ●●

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4. R.M. Kurzrok, correction, *Applied Microwave & Wireless*, p. 12, May 1999.
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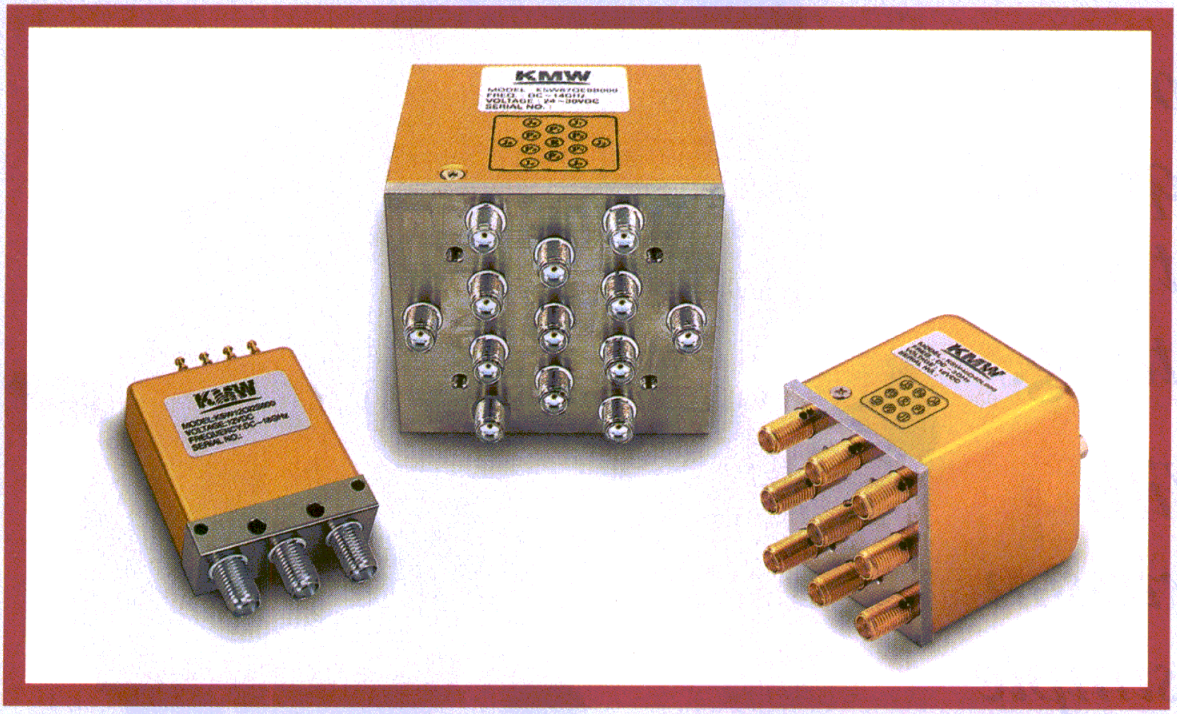

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PTF 10136	1000	6	18.0	28	20244	N
PTF 10147	1000	10	15.0	26	20244	N
PTF 10137	1000	12	13.0	28	20244	N
PTF 10007	1000	35	12.0	28	20222	N
PTF 10052	1000	35	12.0	28	20235	N
PTF 10015	1000	50	12.0	28	20235	N
PTF 10031	1000	50	12.0	28	20222	N
PTF 10139*	1000	60	12.0	28	20235	N
PTF 10138*	1000	60	12.0	28	20222	N
PTF 10009	1000	85	12.0	28	20230	N
PTF 10049	470–860	85	12.0	32	20240	I
PTF 10159	470–860	120	12.0	32/28	20240	I
PTF 10019	860–900	70	13.0	28	20237	I
PTF 10133	860–900	85	13.0	28	20237	I
PTF 10100	860–900	165	12.0	28	20250	I
PTF 10162	860–960	18	14.0	26	20222	N
PTF 10036	860–960	85	11.0	28	20240	I
PTF 10160*	860–960	85	15.0	26	20248	I/O
PTF 10020	860–960	125	11.0	28	20240	I
PTF 10149	921–960	70	15.0	26	20252	I
1.0–2.2 GHz – GOLDMOS FET						
PTF 10111	1500	6	15.0	28	20222	N
PTF 10107	2000	5	11.0	26	20244	N
PTF 10135	2000	5	11.0	26	20249	N
PTF 10041*	2000	12	10.0	26	20249	N
PTF 10053	2000	12	10.0	26	20244	N
PTF 10021	1400–1600	30	11.0	28	20237	I/O
PTF 10125	1400–1600	135	11.5	28	20250	I/O
PTF 10045	1600–1650	30	10.0	28	20222	N
PTF 10112	1800–2000	60	11.0	28	20248	I/O
PTF 10153*	1800–2000	60	12.5	28	20248	I/O
PTF 10120	1800–2000	120	10.0	28	20250	I/O
PTF 10043	1900–2000	12	11.0	26	20222	I
PTF 10035	1900–2000	30	11.0	28	20237	I/O
PTF 10123*	2100–2200	5	11.0	28	20244	N
PTF 10119	2100–2200	12	10.0	28	20222	I
PTF 10048	2100–2200	30	10.0	28	20237	I/O
PTF 10122	2100–2200	50	10.0	28	20248	I/O
PTF 10134*	2100–2200	100	10.0	28	20250	I/O

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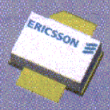

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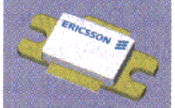

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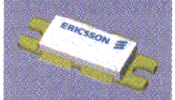

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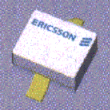

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Predict Total Losses In Amplifier Matching Networks

Realistic estimations of total circuit losses can be obtained by considering the losses imposed by impedance-matching networks on high-power transistor amplifiers.

Oleksandr Gorbachov
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IMPEDANCE-MATCHING networks are commonly used in transistor power amplifiers (PAs). Many of the design strategies assume these networks to be ideal and without losses. Such an assumption can result in an incorrect choice of impedance values due to the high-reflection coefficients of matching lines needed to match high-power transistors with inherently low-characteristic impedances. This article will attempt to present a choice of lossy-matching networks for PAs to help designers of real-world amplifiers.

In an energy transmission from a generator to a load, the total loss in a homogeneous matching line can be defined by eq. 1:

$$Loss\ (in\ db) = 10\ log\ \{(1-|\Gamma|^2 e^{-4\beta l})/[e^{-2\beta l}(1-|\Gamma|^2)]\} \qquad (1)$$

where:

l = the length of the line between the generator and load,

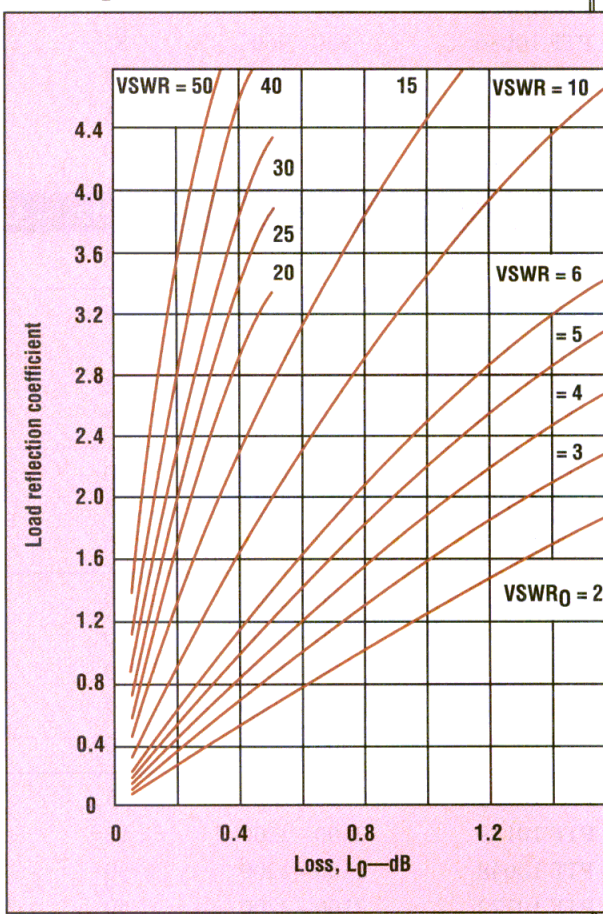
β= the attenuation factor of the line for the wave type considered, and

Γ = the reflection-coefficient of the load (which is assumed to be a well-matched generator).

For the case of a transistor-based PA, losses can be represented by S-parameters in the following way:

$$Loss\ (in\ db) = 10\ log\ \{[1-(S_{21}S_{12}\Gamma)^2]/[S_{21}S_{12}(1-|\Gamma|^2)]\} \qquad (2)$$

Figure 1 presents results calculated from eqs. 1 and 2. In Fig. 1, L refers to loss (in decibels), $L_o = e^{\beta 1}$, or $L_o = S_{21} = S_{13}$ (in decibels). These represent losses for a reciprocal line well-matched with a load. It can be seen how high values of the load-reflection coeffi-



1. These losses are typical characteristics for a transmission line terminated with a reflective load.

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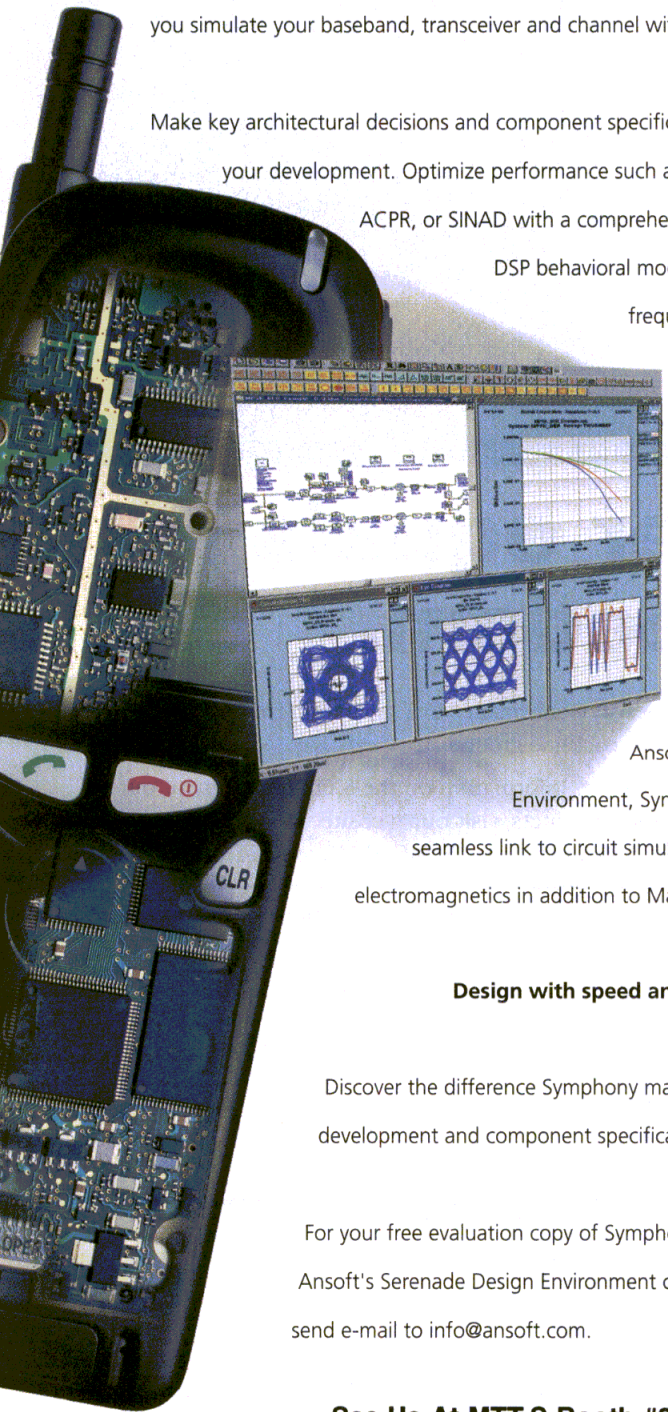

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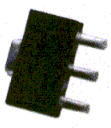

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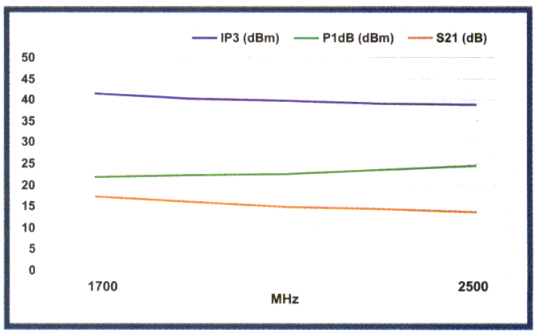

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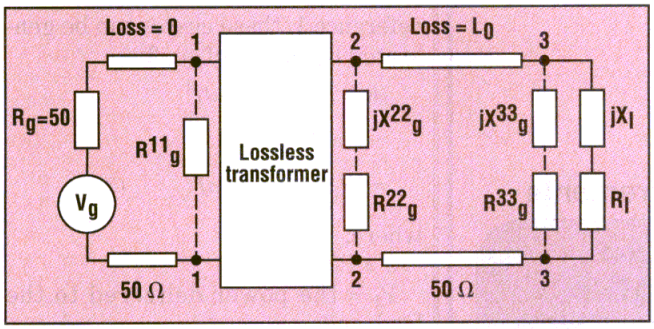
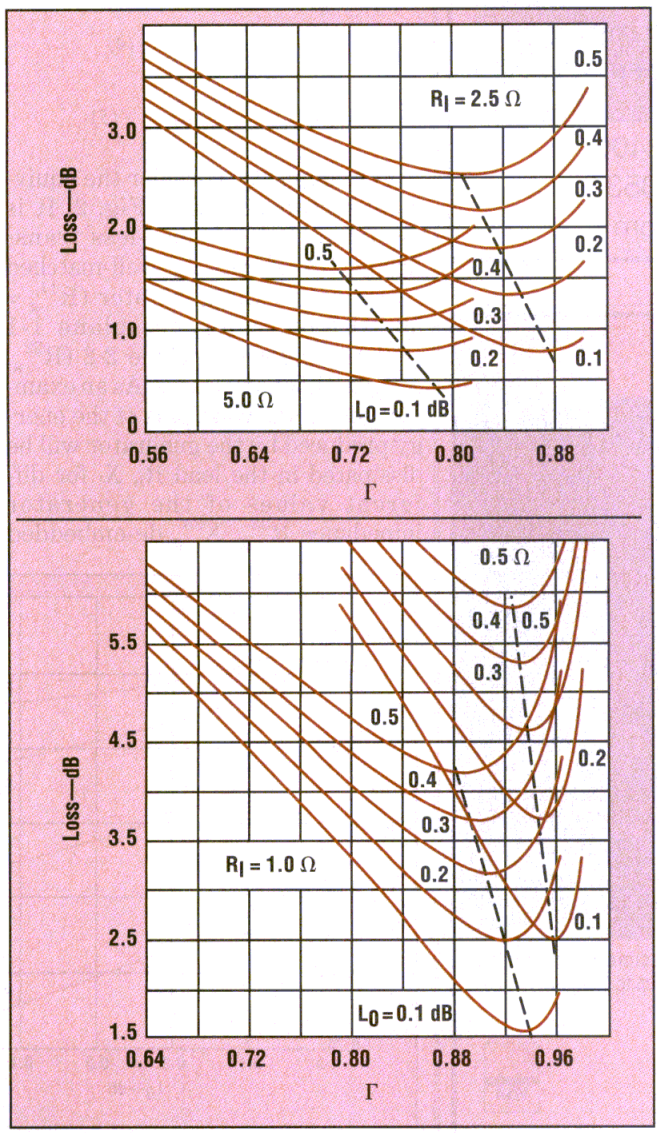


2. This equivalent circuit can be applied to more realistic lossy matching conditions.

cient or voltage-standing-wave ratio (VSWR) result in high total losses. It should be obvious that to achieve lower values of the load-reflection coefficient, designers must choose matching network materials with small inherent losses, choose a minimal distance between matched ports, and choose a matching line impedance as close as possible to the load impedance. Usually, it is not possible to satisfy all of these requirements simultaneously. It is well-known that for maximum delivery of energy from the generator to the load, the conjugate-matching condition must be satisfied. Losses attributed to an unmatched condition between the generator and the load can easily be found from the equivalent-circuit model presented in Fig. 2. In Fig. 2, R_g is the resistance of the generator, R_l is the resistance of the load, X_g is the reactance of the generator, X_g is the reactance of the load, V_g is the voltage developed by the generator, and I is the common current for the generator and the load. Constant-gain circles on a Smith Chart usually represent these losses. For the rectangular coordinates R and X and from



3. The total losses from the generator to the load are plotted as a function of the reflection coefficient of the load, Ω, for various conditions of initial line loss and load resistance.

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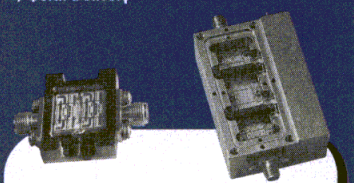

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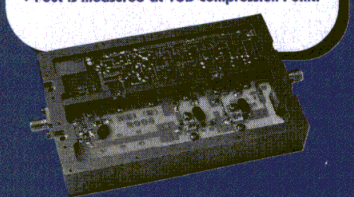
Freq.:	2.0-6.0GHz	4.0-8.0GHz
Gain:	32.0dB min.	26.0dB min.
N.F.:	3.5dB max. 2.0 dB Typ.	6.0 dB max.
Pout:	+12.0dBm min.	+20.0dBm min.
VSWR:	2.0:1/2.0:1	2.0:1/2.0:1
DC Pwr.:	+12VDC@150mA	+12 VDC@150mA

Model:	MSH-6343408-DI	MSH-7264401-DI
Freq.:	8.0-12.0GHz	8.0-18.0GHz
Gain:	23.0dB min.	16.0dB min.
N.F.:	3.0dB max. 2.0dB Typ.	5.0dB max. 3.5dB Typ.
Pout:	+18.0dBm min.	+16.0dBm min.
VSWR:	2.0:1/2.0:1	2.0:1/2.0:1
DC Pwr.:	+12 VDC@150mA	+12VDC@130mA

Model:	MSH-6546602	MSD-3488601-TTL
Freq.:	8.0-12.0GHz	.05-3.0GHz
Gain:	34.0dB min.	30.0dB min.
N.F.:	4.0dB max. 3.0dB Typ.	10.0dB max.
Pout:	+30.0dBm min.	+30.0dBm min.
VSWR:	2.0:1/2.0:1	2.0:1/2.0:1
DC Pwr.:	+15 VDC@1.13A	+15 VDC@1.0A

⇒ Pout is measured at 1dB Compression Point.



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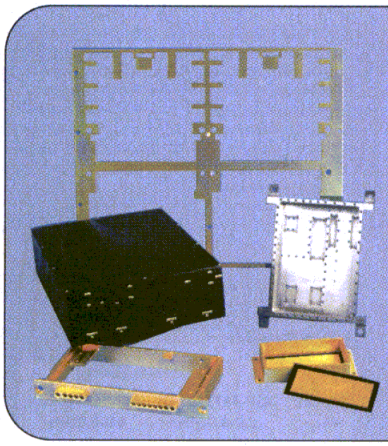

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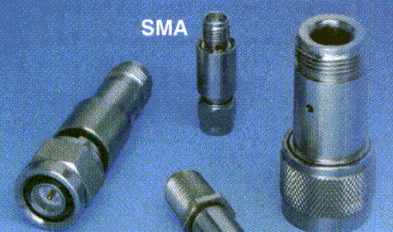

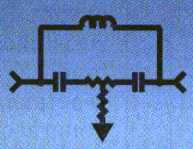

Model	Frequency	Impedance
9093-N	500 MHz - 2 GHz	50Ω
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reference 1, these circles can be generated by means of eq. 3:

$$\{R_l - R_g[(2-a)/a]\}^2 + (X_l + X_g)^2 = 2R_g[(1-a)^{0.5}/a]^2 \quad (3)$$

where:

$a = P_l/P_g$,

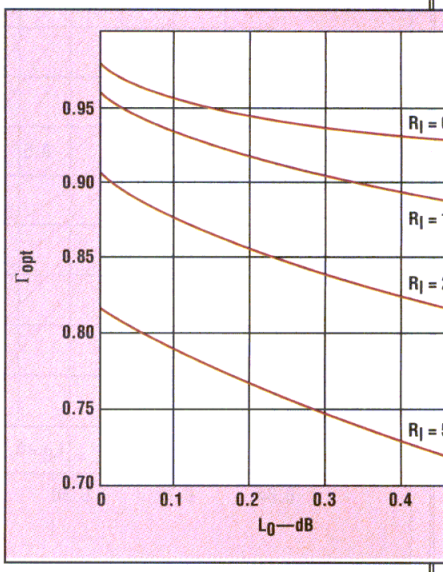
P_l = the power delivered to the load, and

P_g = the power produced by the generator. The coordinates R^0_l and X^0_l of the centers of the circles from eq. 3 and their radius R_O can be calculated[2] according to:

$$R^0_l = R_g[(2-a)/a]; \quad X^0_l = -X_g \quad (4)$$

$$R_0 = 2R_g[(1-a)^{0.5}/a] \quad (5)$$

For simplicity, consider the equivalent circuit shown in Fig. 2. It is assumed that the loss-less transformer transforms the well-matched resistance of the generator (R^{11}_g = 50 Ω) from reference plane 1-1 (R^{11}_g) to reference plane 2-2 (R^{22}_g, X^{22}_g) without any losses. As an example, determine what part of the energy produced by the generator will be dissipated in the load R_l, X_l for different values of the generator impedance R^{33}_g, X^{33}_g de-embedded



4. The optimum reflection coefficient of the generator is plotted as a function of the initial loss for different values of load resistance.

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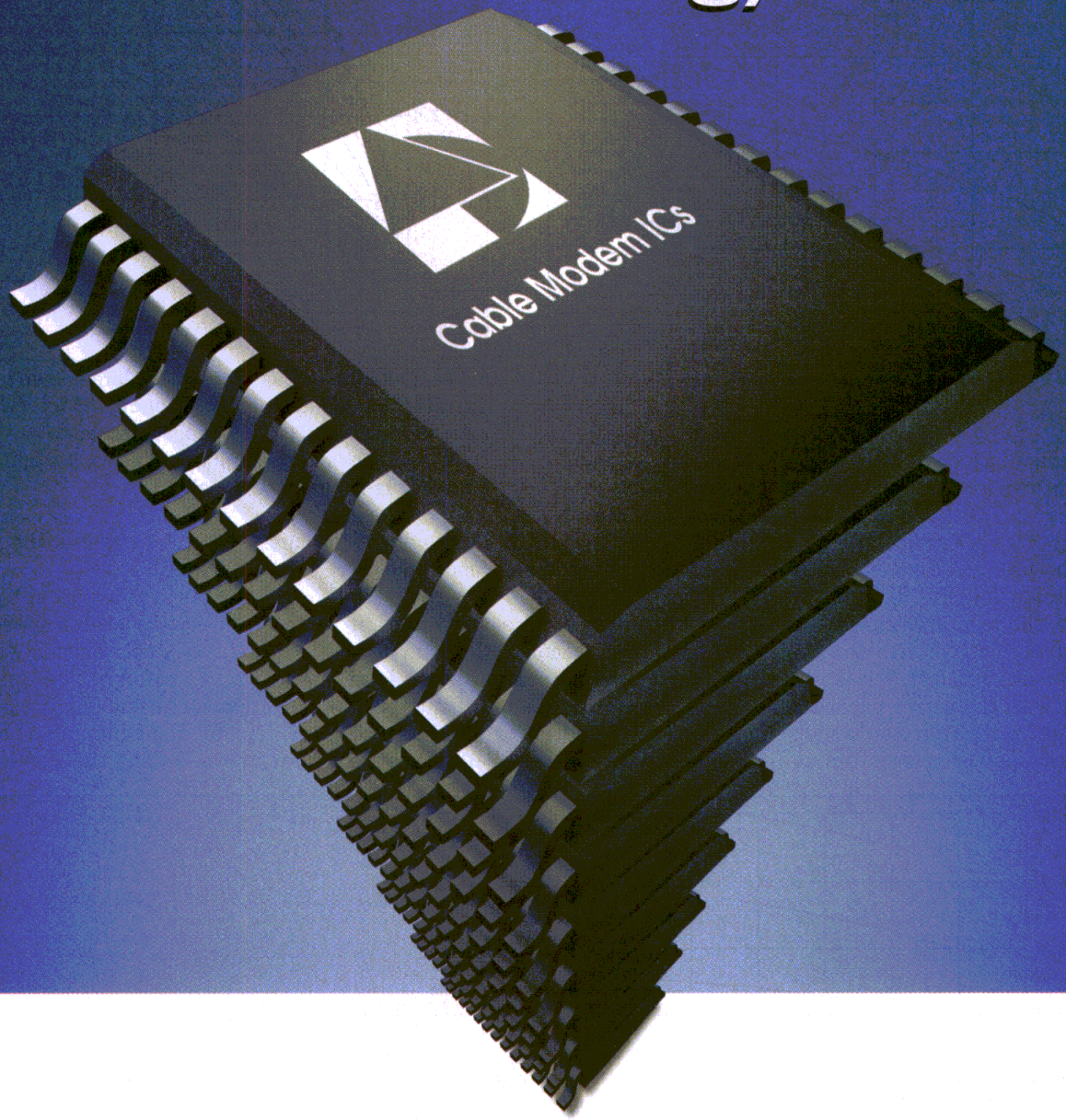

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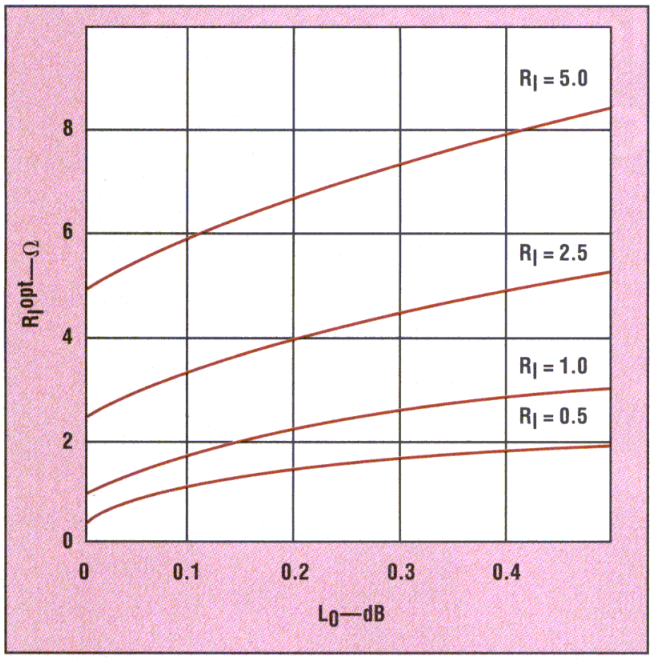
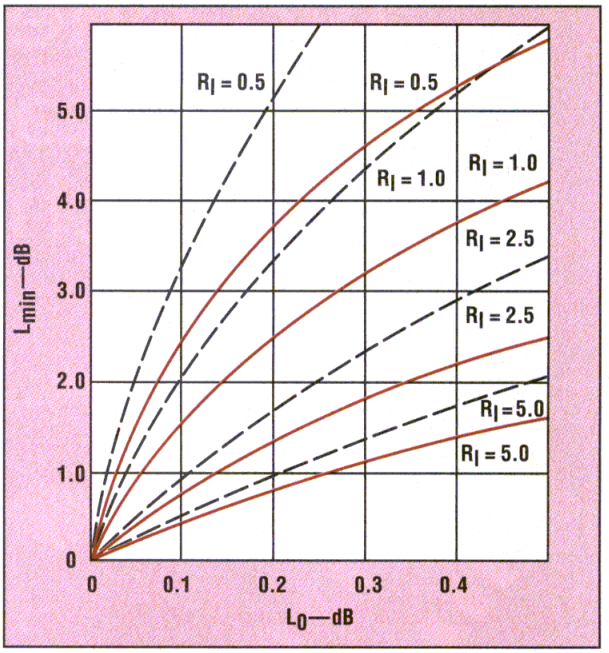


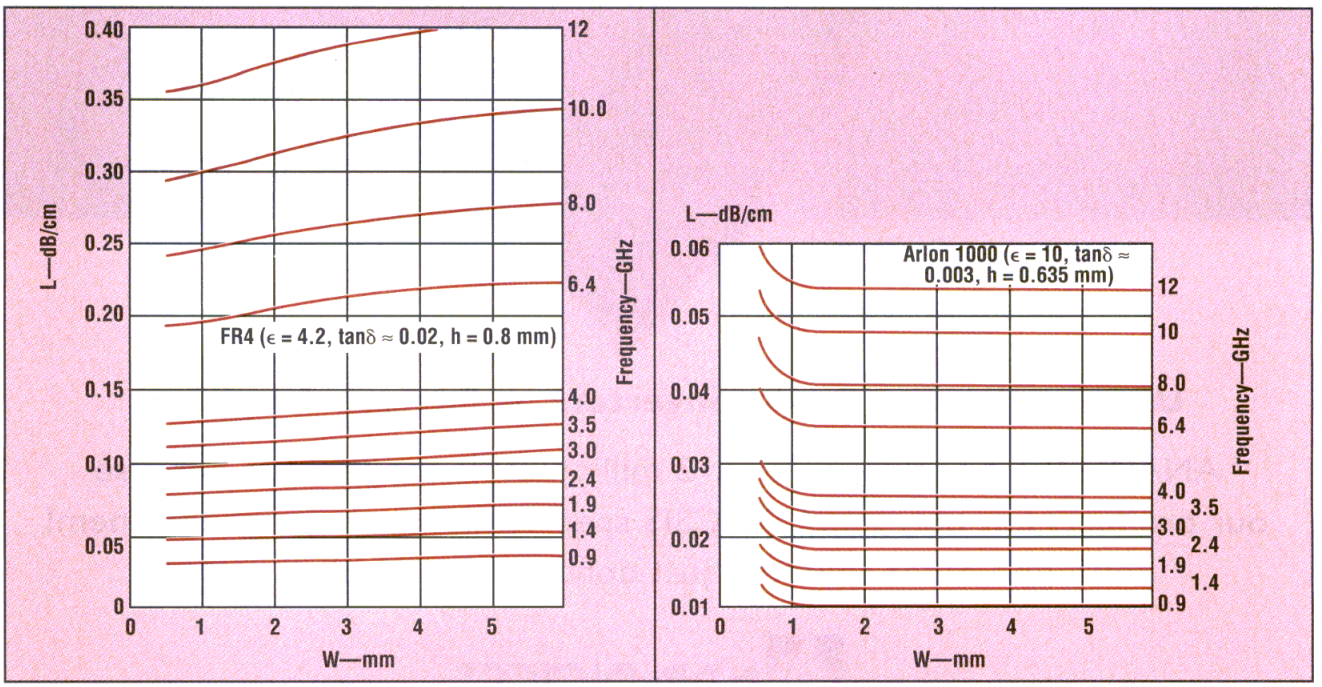
5. The optimum load resistance is plotted as a function of the initial loss for different values of load resistance.



6. The minimum values of loss, L_{min}, are shown for optimum impedance-matching conditions for the generator and load (solid lines) compared to calculations based on original values for the optimum load resistance (dashed lines).

to the load reference plane 3-3 through the loss (L_O) of the 50-Ω transmission line (Fig. 3). In the obvious case, the load impedance is chosen entirely active with the resistance, R_l, as 0.5...5.0 Ω. These values coincide with the active parts of the input and output impedances of microwave power transistors at L- and C-band frequencies. The generator's impedance at reference place 3-3 is also assumed to be active. By combining the two loss conditions, it is possible to calculate the total losses from the generator to the load (Fig. 4) for a wide range of initial line losses, L_O, load resistance, R_l, and generator resistance at reference plane 3-3 represented as the reflection coefficient, Γ.

In all cases, it is apparent that the choice of a low generator impedance, $R^{33}_g = R_l$, does not coincide with the minimum total loss. The optimum reflection coefficient for the generator, Γ_{opt}, for the minimum amount of loss, "moves" along the curve of the



7. Two typical circuit-board materials, FR4 and Arlon 1000, were used as reference points to plot total losses in conductors and dielectric as a function of line width, W.

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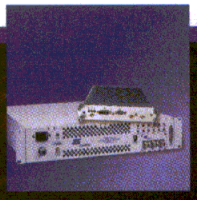

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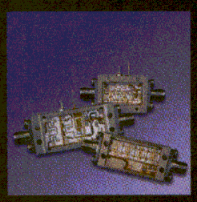

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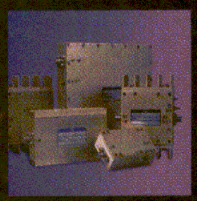

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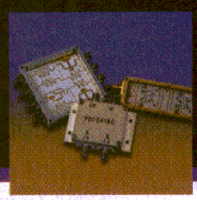

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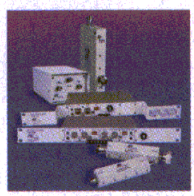

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dashed lines (Fig. 4) to lower values of Γ by increasing the initial loss, L_O. Figure 5 provides the optimum values of the reflection coefficient for the generator, H_{opt}, versus initial loss, L_O; these values can be used for an evaluation of a matching condition not only for a resistive load but also for a complex (reactive) load. The results shown in Fig. 5 indicate that for the maximum transfer of energy from the generator to the load through a lossy transmission line, a principally "unmatched" condition must be chosen in accordance with the optimum reflection coefficient for the generator, Ω_{opt}.

Figure 6 shows the results of calculations for a purely resistive load for the optimum load resistance, R^{opt}_l, versus the initial loss, L_O. As can be seen, the optimum load resistance increases dramatically in comparison with its initial values especially for the small resistance values typically required for high-power microwave transistors.

Figure 6 presents calculations of the minimum losses in a matching line for the optimum matching of the generator and load (solid lines). This figure also presents the losses (dashed lines) resulting from choosing the optimum-load resistance equal to the original value. A significant difference can be seen in the two sets of losses, especially for those cases of smaller load resistances and higher values of initial line loss.

In referring back to Fig. 4, it can be seen that, for high values of the load resistance, tuning of a impedance-matching line for minimum losses is not particularly crucial for small values of resistance and for higher values of the initial line loss. This can be explained by a reduction in the quality factor (Q), one factor of the transistor resonant circuit represented by the matching line.

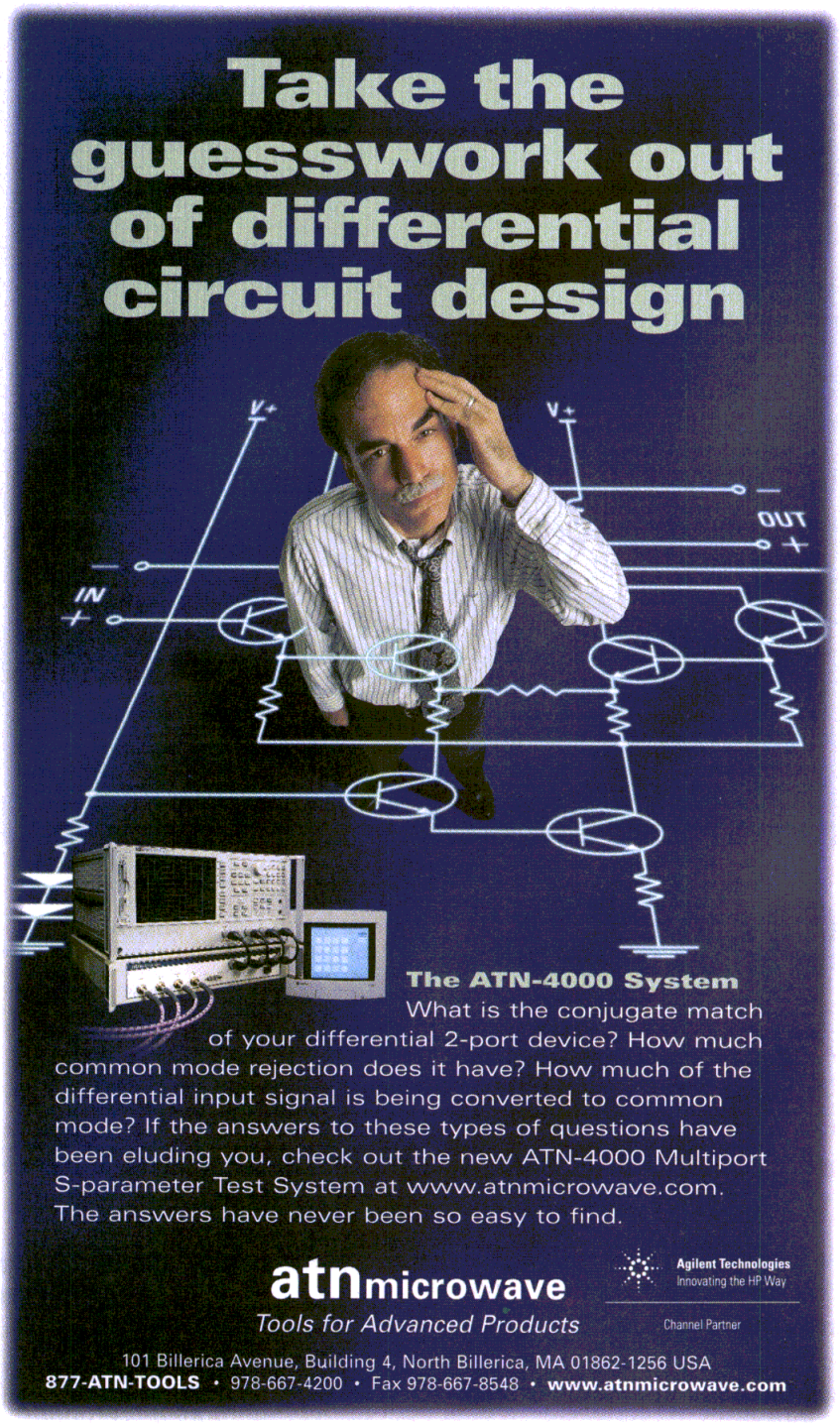
PREDICTING LOSSES

By using the calculated values presented in Figs. 1 through 6, it should be easy to predict the total losses in a matching line and choose the optimum load and generator resistances for a wide range of cases. Although this article has presented calculations for cases based on 50-Ω matching lines, similar results are possible by scaling the impedance to other values.

For reference purposes, Fig. 7 offers the calculated results of the total losses in the metal conductors and dielectric material for a typical microstrip circuit board, using common circuit-board materials such as FR4 and 1000 material from Arlon and with the copper (Cu) metallization at different line widths (where W is the line width in millimeters).[3] ●●

References

1. I.V. Lebedev, *EHF Technology and Devices*, Vol.1, Moscow, 1961.
2. R.S. Carson, *High-Frequency Amplifiers*, Wiley, New York, 1975.
3. O. Gorbachov, "Choice of Matching Networks for High-Precision RF/Microwave Measurements," *Applied Microwave & Wireless*, to be published.



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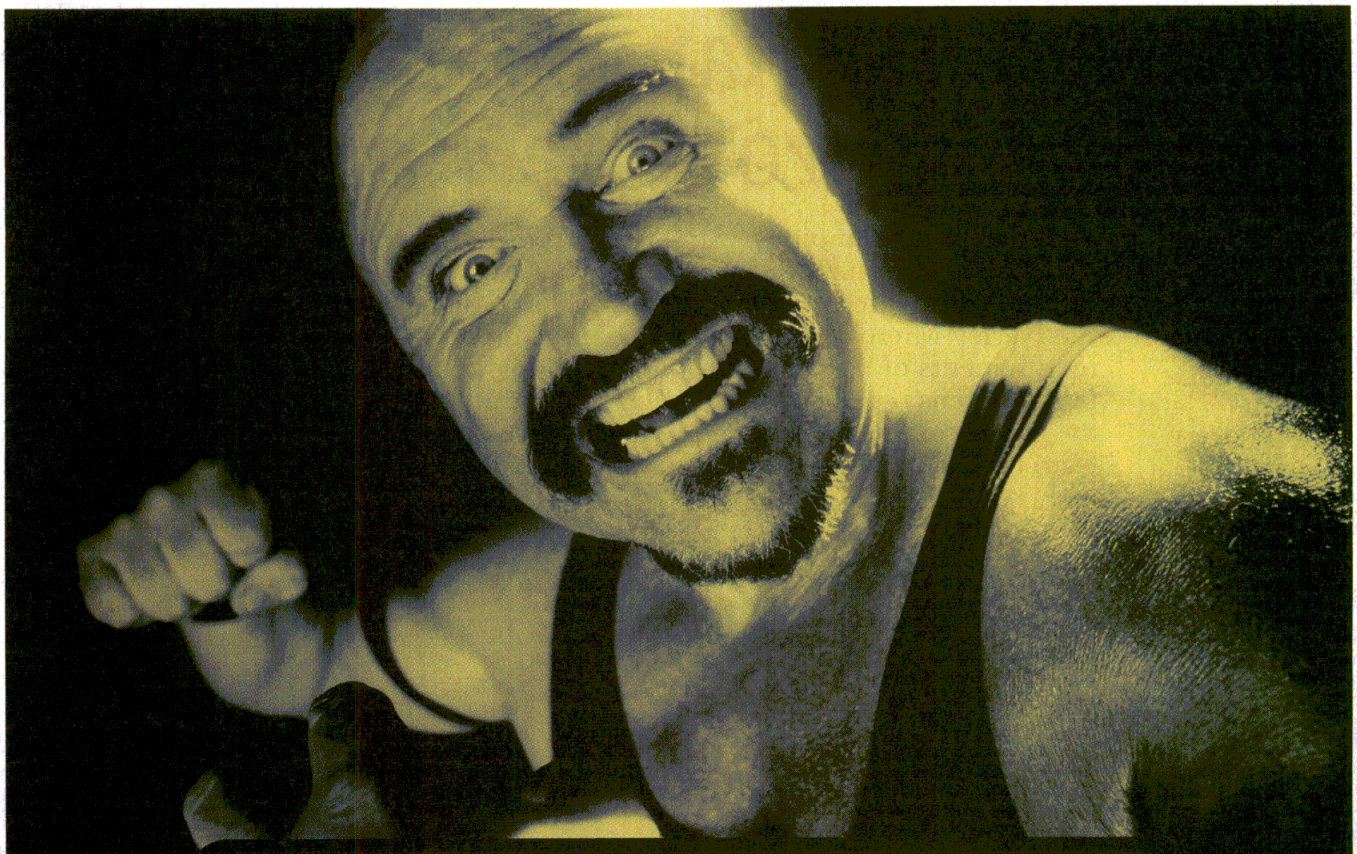

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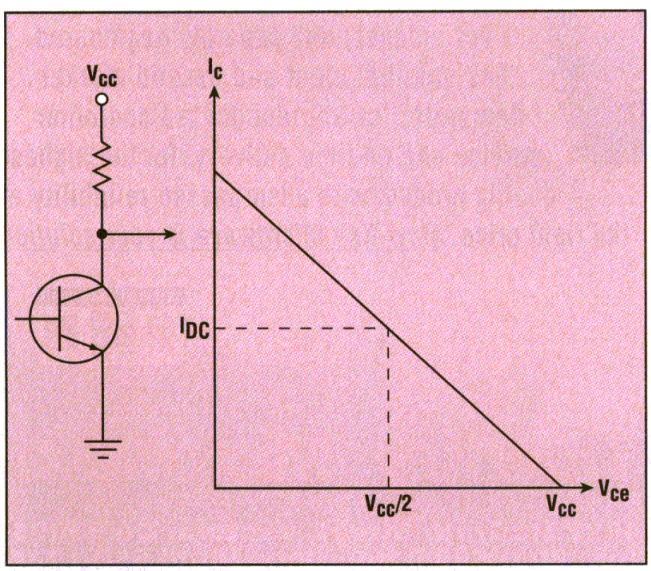
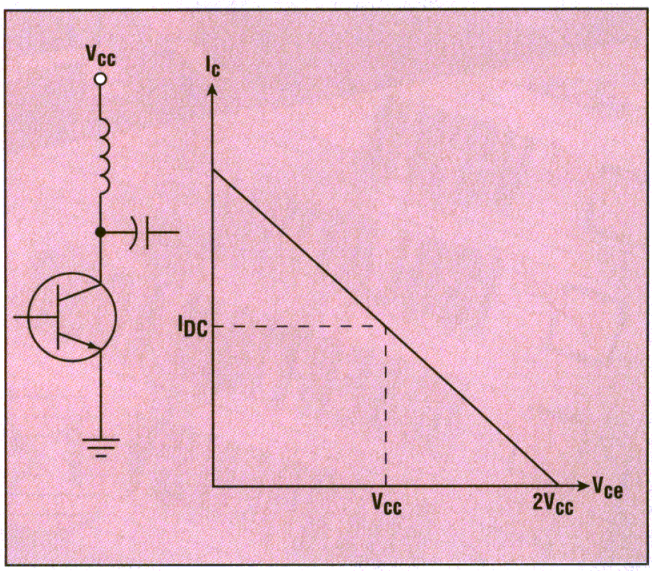
INTERCEPT points have been used for many years to predict spurious responses.[1,2] In this technique, two signals with frequencies, F_1 and F_2, are injected into the input of an amplifier. The nonlinearity of the amplifier produces intermodulation (IM) products at the output. The second-order IM terms are $F_1 \pm F_2$ and the third-order IM terms are $2F_1 \pm F_2$ and $2F_2 \pm F_1$. The intersection of the fundamental and third-order responses on a log-log scale is known as the third-order intercept point (IP3), and the intersection of the fundamental and second-order responses is called the second-order intercept point (IP2). To find those intercept points, measurements are made in the linear portion the amplifier's response—well below its 1-dB compression point.

One piece of literature published on this subject states that, empirically,[1] the second-order plot intercepts the fundamental plot at the same point as the third-order plot, unless some technique is used that suppresses even-order spurious responses (such as a push-pull operation). Another statement is that the 1-dB compression point is 10 dB less than the IP3. But an analysis performed on a single-stage transistor amplifier shows that those statements are inaccurate.



1. A Class A transistor-amplifier schematic and load line are shown here.



2. An inductive-coupled, Class A transistor-amplifier schematic and load line can be seen.

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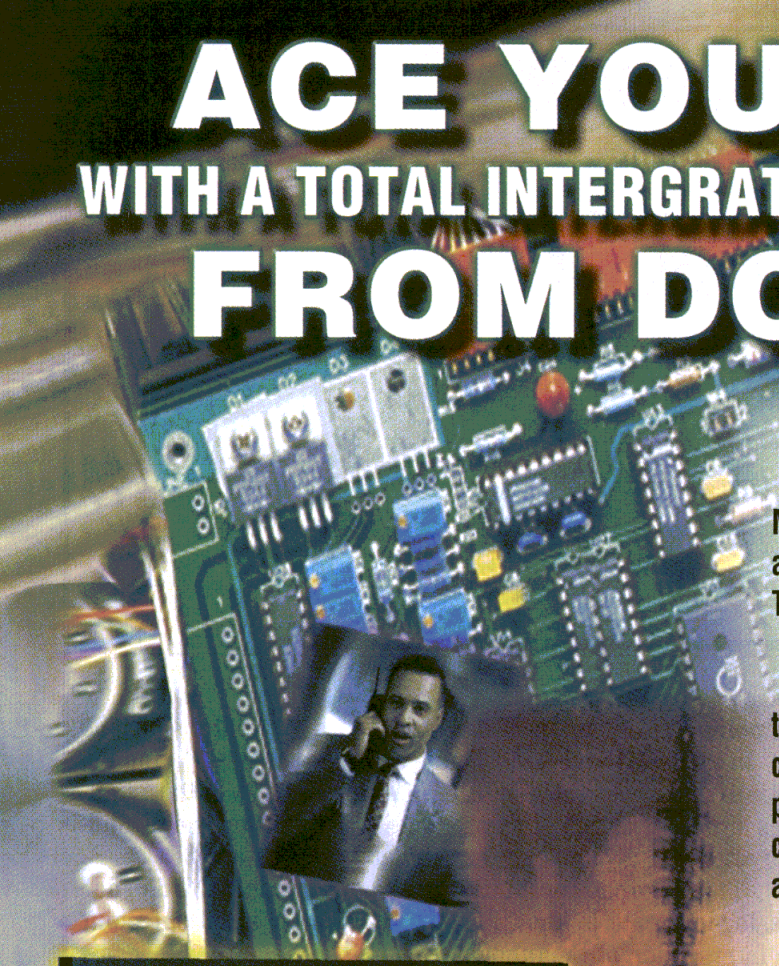

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This article discusses the analysis performed on a single-stage transistor amplifier to determine the IP2 and IP3, as well as the 1-dB compression point, as a function of the DC current of the transistor. The results of the analysis are provided in a set of simple equations.

INTERCEPT-POINT ANALYSIS

The emitter current of a transistor can be expressed as:[3]

$$i_E(t) = I_{ES} exp(v_{BE} / V_T)$$

where:

V_T = kT/q,

I_{ES} = the emitter saturation current,

v_{BE} = the base-emitter voltage,

k = Boltzmann's constant,

T = the temperature in K, and

q = the electronic charge.

The input base-emitter voltage is a combination of DC and AC components.

$$v_{BE} = V_{DC} + v(t)$$

And so the emitter current becomes:

$$i_E(t) = I_{ES} exp[(V_{DC} + v(t)) / V_T] = I_{ES} exp(V_{DC} / V_T) exp(v(t) / V_T)$$

The collector current is:

$$i_0(t) = \alpha I_E(t)$$

if:

$$I_Q = \alpha I_{ES} exp(V_{DC} / V_T)$$

Then the collector current becomes:

$$i_0(t) = I_Q exp(v(t) / V_T) \qquad (1)$$

For small-signal analysis:

$$I_Q = I_{DC}$$

Using the expansion of exp(x):

$$exp(x) = 1 + x + x^2/2! + x^3/3! + \ldots$$

then:

$$V_0(t) = i_0(t) R_0 = I_{DC} R_0 [1 + (1/V_T) v(t) + (1/2V_T^2) v^2(t) + (1/6V_T^3) v^3(t)]$$

From intercept-point theory, $V_0(t)$ is assumed to follow the Maclaurin series:

$$V_0(t) = k_0 + k_1 v(t) + k_2 v^2(t) + k_3 v^3(t) + \ldots$$

therefore:

$$k_1 = I_{DC} R_0 / V_T \qquad (2a)$$

$$k_2 = I_{DC} R_0 / (2V_T^2) \qquad (2b)$$

$$k_3 = I_{DC} R_0 / (6V_T^3) \qquad (2c)$$

When two equal-amplitude signals are applied to the input of an amplifier:

$$v(t) = a \sin \omega_1 t + a \sin \omega_2 t$$

Then, from intercept-point theory, the fundamental amplitude is $k_1 a$, the second-order intermodulation is $k_2 a^2$, and the third-order IM is $3k_3 a^3/4$. The IP2 can be found from the intersection of the fundamental and the second-order IM.

$$k_1 a = k_2 a^2 \text{ or } a_{2IP} = k_1 / k_2$$

$$V_{2IP}(out) = k_1 a_{2IP} = k_1^2 / k_2$$

Substituting k_1 and k_2 from equation 2:

$$V_{2IP}(out) = 2 I_{DC} R_0 \qquad (3a)$$

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

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

The IP3 can be found from the intersection of the fundamental and the third-order IM.

$$k_1 a = 3k_3 a^3 / 4 \text{ or } a^2_{3IP} = 4k_1 / (3k_3)$$

$$V^2_{3IP}(out) = (k_1 a_{3IP})^2 = 4k_1^3 / (3k_3)$$

Substituting k_1 and k_3 from equation 2 is:

$$V^2_{3IP}(out) = 8 I_{DC}^2 R_0^2 \qquad (4a)$$

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

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$$P_{3IP}(out) = 8\, I_{DC}^2 R_0 \qquad (4c)$$

From equations 3 and 4, it can be seen that the IP3 is 3 dB higher than the IP2.

COMPRESSION POINT

Using equation 1 and defining v(t) as:

$$v(t) = V_1 \cos \omega t$$

and

$$x = V_1 / V_T \qquad (5)$$

then:

$$i_0(t) = I_Q exp(x \cos \omega t)$$

And from a known Fourier series expansion:

$$exp(x \cos \omega t) =$$

$$I_0(x) + 2\sum_{n=1}^{\infty} I_n(x) \cos(n\omega t)$$

where:

$I_n(x)$ = a modified Bessel function of the first kind, of order n, and argument x.

$$i_0(t) = I_Q I_0(x)$$

$$\left[1 + 2\sum_{n=1}^{\infty} [I_n(x)/I_0(x)] \cos(n\, \omega\, t)\right]$$

The DC current is:

$$I_{DC} = I_Q I_0(x)$$

On a small signal, or x ~ 0, $I_0(0) = 1$ and IDC = IQ. But on a large signal, or x >> 0, the DC current shifts and is no longer IQ.

The amplitude of the fundamental component of $i_0(t)$ is:

$$I_{01} = I_{DC}[2\, I_1(x) / I_0(x)] \qquad (6)$$

IT SHOULD BE NOTED THAT SINCE LARGE-SIGNAL ANALYSIS IS BEING DEALT WITH, AN OPTIMUM R_0 SHOULD BE CHOSEN IN ORDER TO USE THE MAXIMUM VOLTAGE AND CURRENT SWING.

The fundamental output power is:

$$P_0 = I_{01}^2 R_0 / 2 \qquad (7a)$$

$$P_0 = I_{DC}^2 [2\, I_1(x) / I_0(x)]^2 R_0 / 2 \qquad (7b)$$

And by using equation 5, the input power is:

$$P_{in} = V_1^2 / (2R_{in}) = x^2 V_T^2 / (2R_{in}) \qquad (8)$$

Defining conductance as:

$$g_m = I_{DC} / V_T$$

The gain is:

$$G_p = P_0 / P_{in} =$$

$$g_m^2 [2\, I_1(x) / (x\, I_0(x))]^2 R_{in} R_0 \qquad (9)$$

The gain is small signal when x → 0. And, from Bessel functions:

$$[2\, I_1(x) / (x\, I_0(x))] \rightarrow 1$$

The 1-dB compression point is where the gain, G_p, decreases by 1 dB.

$$10 \log[2\, I_1(x) / (x\, I_0(x))]^2 = -1$$

$$2\, I_1(x) / (x\, I_0(x)) = 0.891$$

From Bessel tables, it can be found that this complies when x =1, which yields:

$$2\, I_1(x) / I_0(x) = 0.891$$

Substituting back in equation 7, the 1-dB compression point is:

$$P_{1dB} = 0.891^2\, I_{DC}^2 R_0 / 2 \qquad (10a)$$

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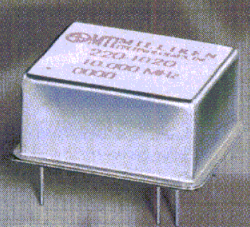

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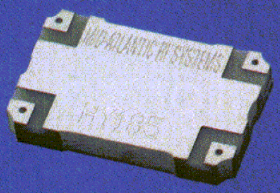

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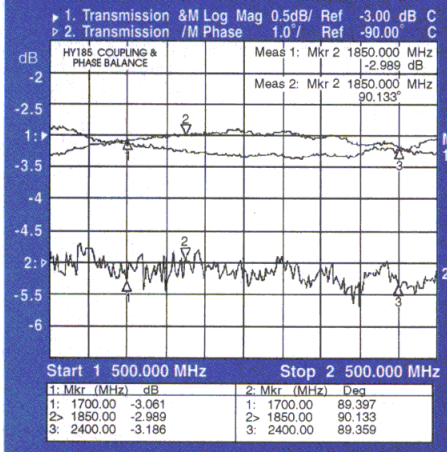

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Intercept Point

$$P_{1dB} = 0.4\, I_{DC}^2 R_0 \qquad (10b)$$

It is interesting to note that the compression point is where x = 1 (i.e., the input voltage, V_1, equals V_T).

It should be noted that since large-signal analysis is being dealt with, an optimum R_0 should be chosen to use the maximum voltage and current swing. For Class A amplifiers, the schematics and the $V_{CE} - I_C$ graph are shown in Fig. 1.

The optimum RL is $V_{CC} / (2_{IDC})$ and the 1-dB compression point is:

$$P_{1dB} = 0.2\, V_{CC} I_{DC} = 0.2\, P_{DC} \quad (11)$$

Compare this to the generally used approximation of $0.25P_{DC}$ for maximum power. For inductive-coupled Class A amplifiers, the schematics and the $V_{CE} - I_C$ graph are shown in Fig. 2.

The optimum RL is V_{CC}/I_{DC}, and the 1-dB compression point is:

$$P_{1dB} = 0.4\, V_{CC} I_{DC} = 0.4\, P_{DC} \quad (12)$$

Compare this to the approximation of $0.5P_{DC}$ that is generally used in literature.

From equations 4 and 10, the ratio of intercept point to 1-dB compression point is:

$$P_{3IP}(out) / P_{1dB} = 8 / 0.4 = 20 = 13\ dB \qquad (13)$$

This equation assumes that the same R_0 is taken in both cases. But since the intercept point is measured at small signals and is extrapolated, a higher R_0 than the optimum load resistance can be taken in equation 4, and the ratio provided in equation 13 for a different load resistance can be much higher.

This analysis has generated equations for IP2, IP3, as well as the 1-dB compression point. These equations show that all terms are related to the load impedance and the square of the DC current, and differ only by a constant.••

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1. Franz C. McVay, "Don't Guess the Spurious Level," *Electronic Design 3*, February 1, 1967, pp. 70-73.
2. Harold B.Goldberg, "Predict intermodulation distortion," *Electronic Design 10*, May 10, 1970, pp. 76-78.
3. K.K. Clarke and D.T. Hess, *Communication Circuits: Analysis and Design*, Addison-Wesley Publishing Co., 1971.

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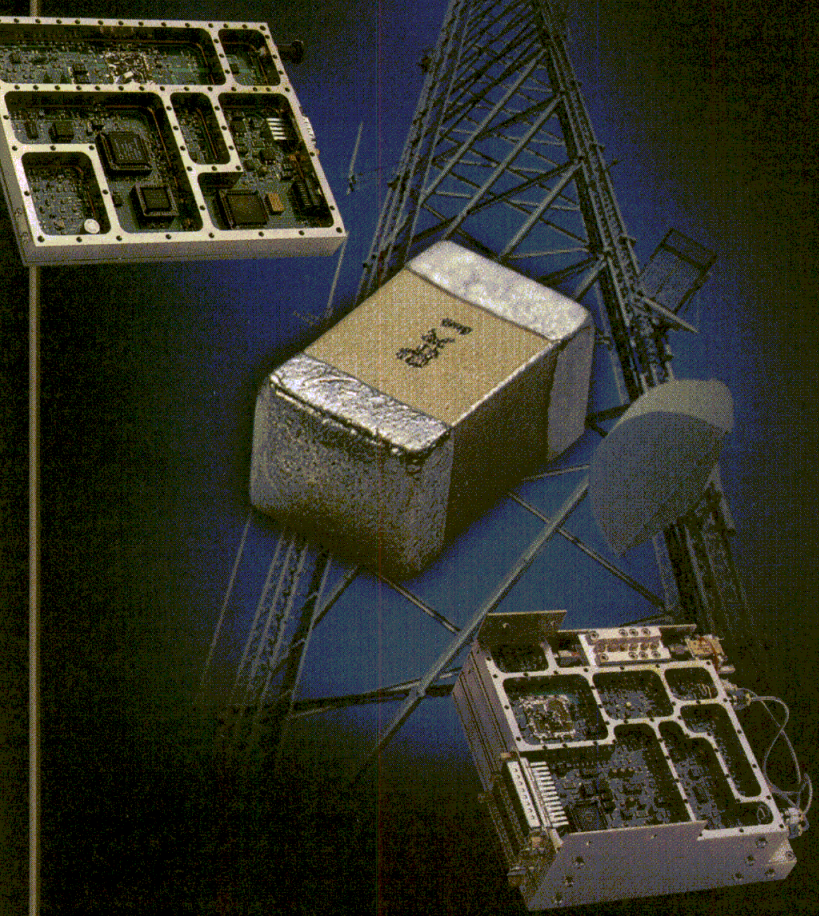

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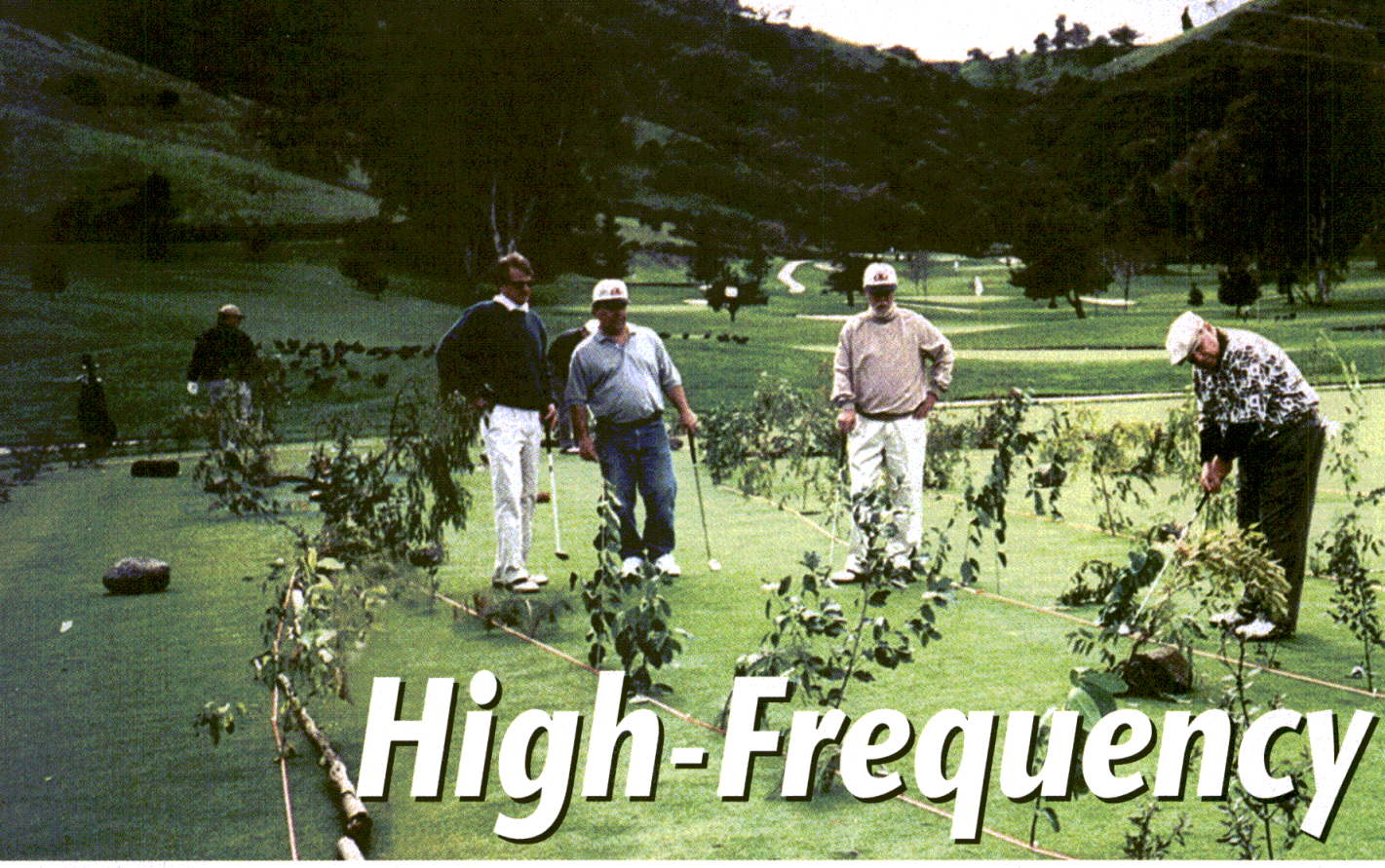

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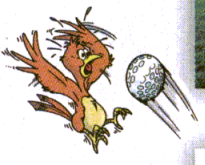
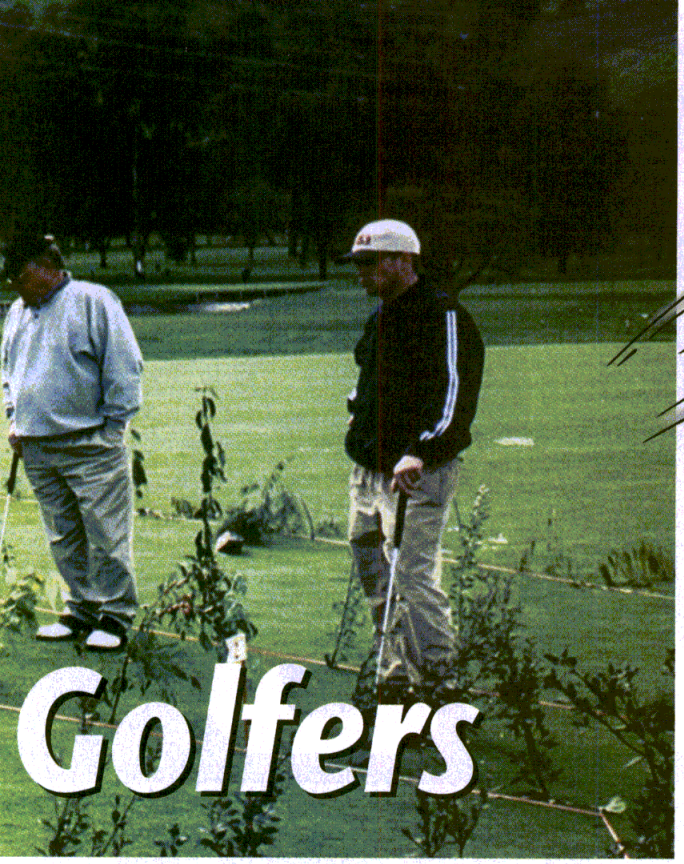
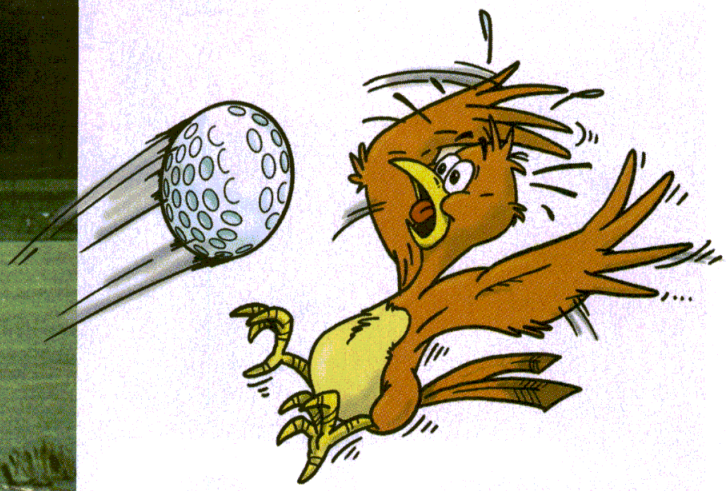


GOLF AFFICIONADOS CONVENED AT the close of the Eighth Annual *Wireless Symposium/Portable By Design Conference & Exhibition* in search of the perfect stroke. More than 150 representatives from leading wireless and high-frequency electronics manufacturers competed on the links at the Bay Area's picturesque Santa Teresa Golf Club in San Jose. The Gene Roberts Invitational Tournament (GRIT), named after *Microwaves & RF*'s West Coast Regional Sales Manager, has become a regular part of the annual Wireless Symposium.





Golfers

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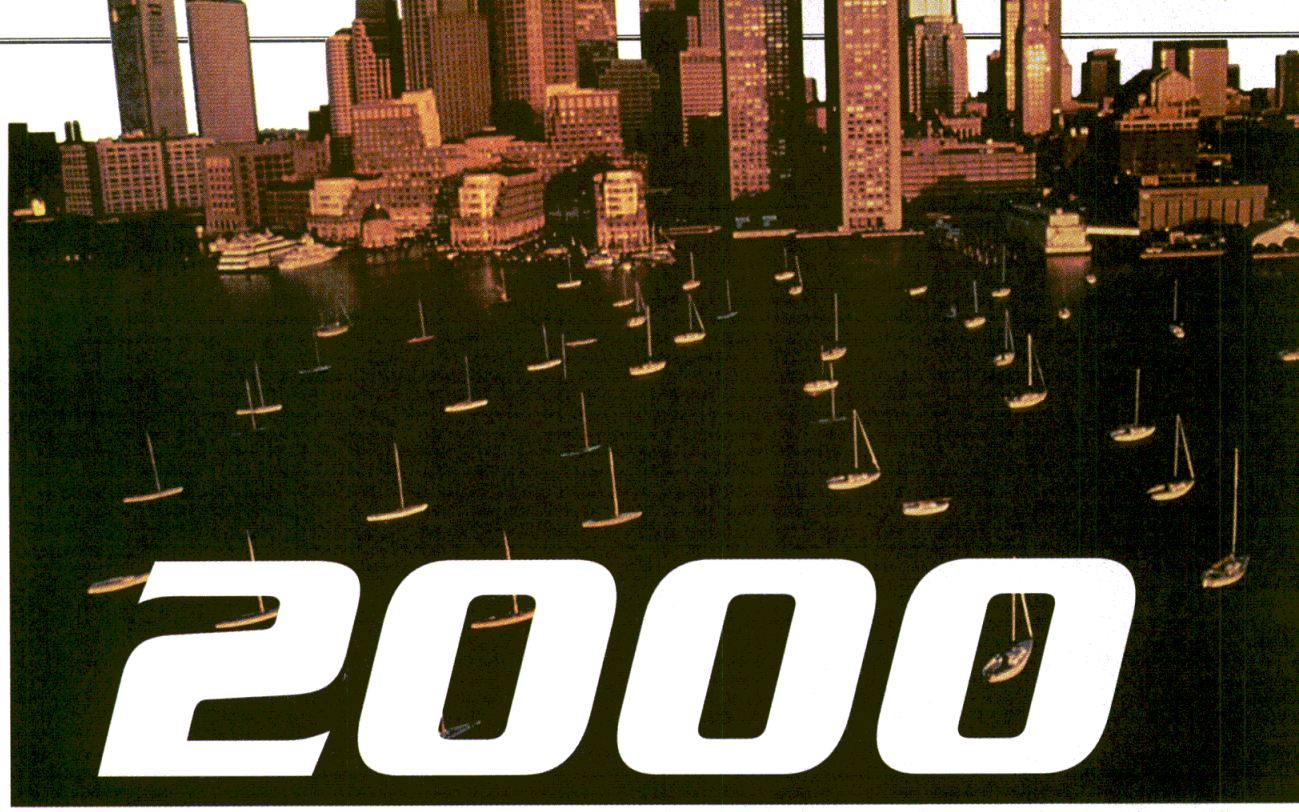
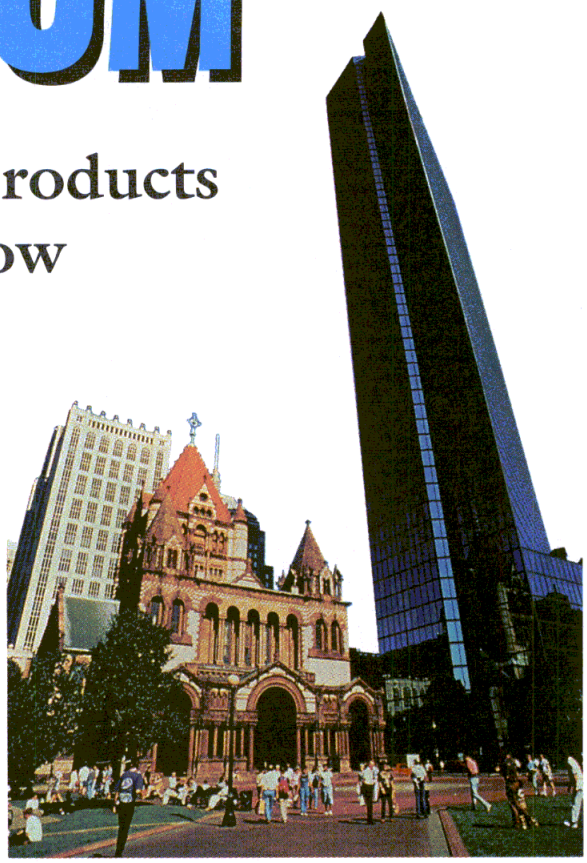


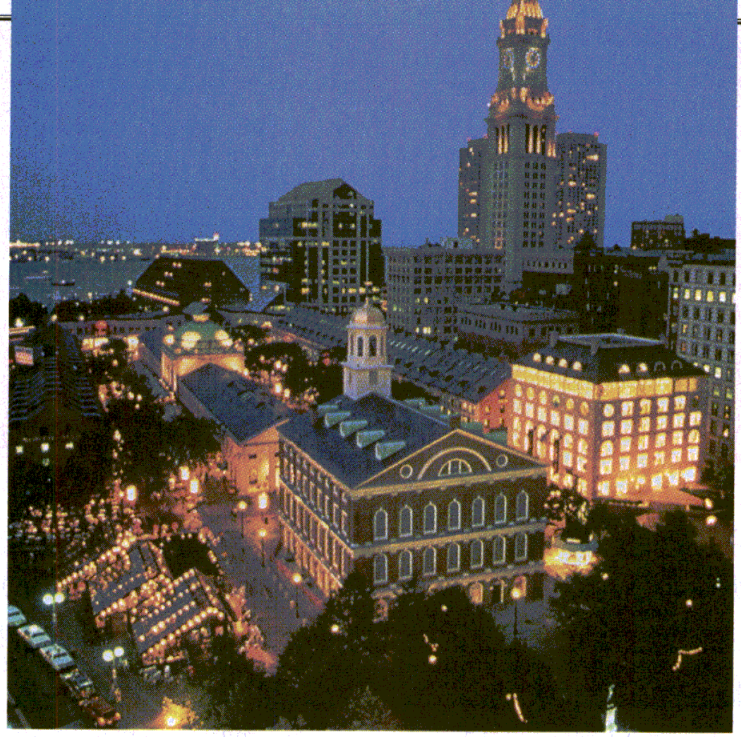
2000 INTERNATIONAL SYMPOSIUM

The Key Papers and Products at this Year's Show

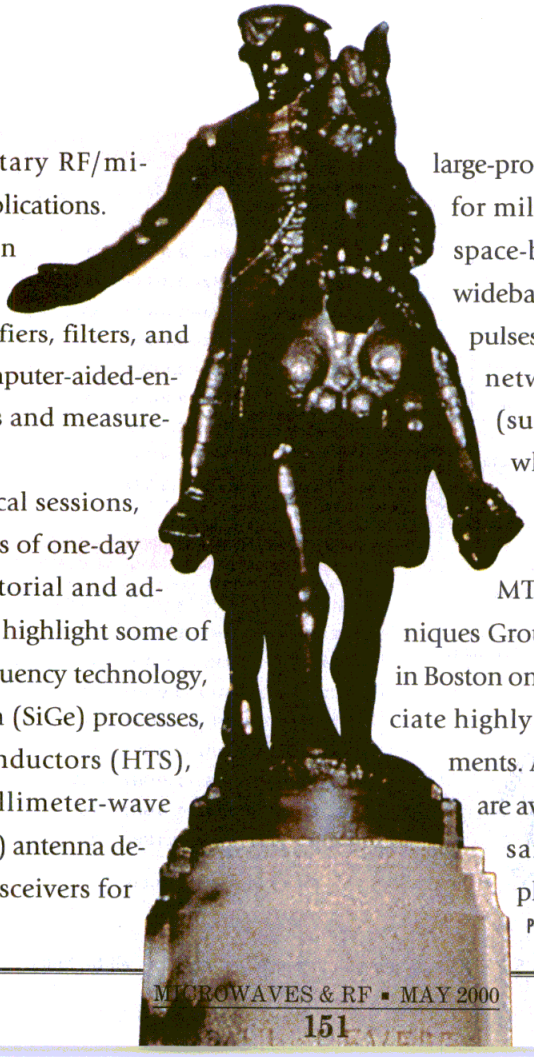
Microwave engineers generally possess a diversified set of skills and interests. The wide range of topics at the upcoming 2000 IEEE/International Microwave Theory & Techniques Symposium (MTT-S) next month is testimony to the versatility of the modern microwave engineer, with topics that cover everything from automotive sensors to wireless communications.

Scheduled for June 11-16, 2000 at the John B. Hynes Veterans Memorial Convention Center (Boston, MA), the 2000 IEEE/MTT-S features more than 400 technical papers





MICROWAVE



on commercial and military RF/microwave technology and applications. Technical sessions focus on component-related topics, such as the design of amplifiers, filters, and multipliers, as well as on computer-aided-engineering (CAE) techniques and measurement methods.

In addition to the technical sessions, the conference boasts a series of one-day workshops that provide tutorial and advanced content. Workshops highlight some of the latest trends in high-frequency technology, including silicon-germanium (SiGe) processes, high-temperature superconductors (HTS), Bluetooth technology, millimeter-wave technology, advanced (smart) antenna designs, millimeter-wave transceivers for large-production applications, digital receivers for military and commercial applications, space-based links for radar systems, ultra-wideband systems based on extremely short pulses, and applications of artificial neural networks to RF and microwave design (supporting model development even when established mathematical relationships are not available).

Finally, in conjunction with the MTT-S, the 53rd Automatic RF Techniques Group (ARFTG) Conference will be held in Boston on Friday, June 16, for those who appreciate highly accurate high-frequency measurements. Also, a variety of additional programs are available for spouses of attendees. For a sampling of the technical sessions, please see the pages that follow.

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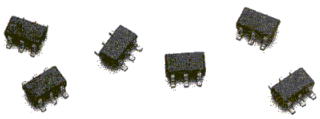

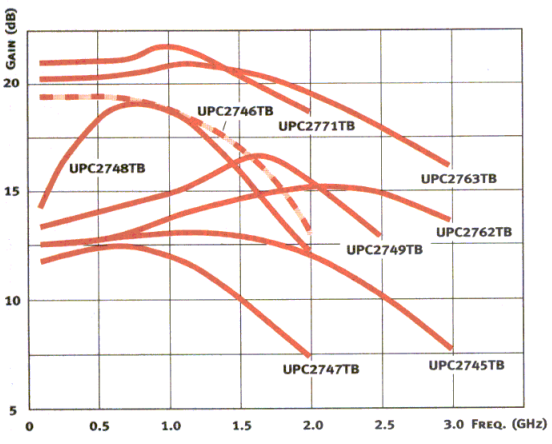

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UPC2710TB	50MHz–1.0GHz	33	3.5	+10.8	22	500MHz
UPC2711TB	50MHz–2.9GHz	13	5	-2.6	12	1.0GHz
UPC2712TB	50MHz–2.6GHz	20	4.5	-0.4	12	1.0GHz
UPC2713T	50MHz–1.2GHz	29	3.2	+0.3	12	500MHz
UPC2776TB	50MHz–2.7GHz	23	6.0	+6	25	1.0GHz
UPC2791TB	50MHz–1.9GHz	12	5.5	+1	17	500MHz
UPC2792TB	50MHz–1.2GHz	20	3.5	0	19	500MHz

3V WIDEBAND RFIC AMPS — FROM 35¢

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UPC2746TB	50MHz–1.5GHz	19	4	-3.7	7.5	500MHz
UPC2747TB	100MHz–1.8GHz	12	3.3	-11	5	900MHz
UPC2748TB	200MHz–1.5GHz	19	2.8	-8.5	6	900MHz
UPC2749TB	100MHz–2.9GHz	16	4	-12.5	6	1.9GHz
UPC2762TB	100MHz–2.9GHz	14.5	7	7	27	1.9GHz
UPC2763TB	100MHz–2.4GHz	20	5.5	6.5	27	1.9GHz
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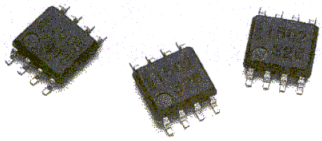

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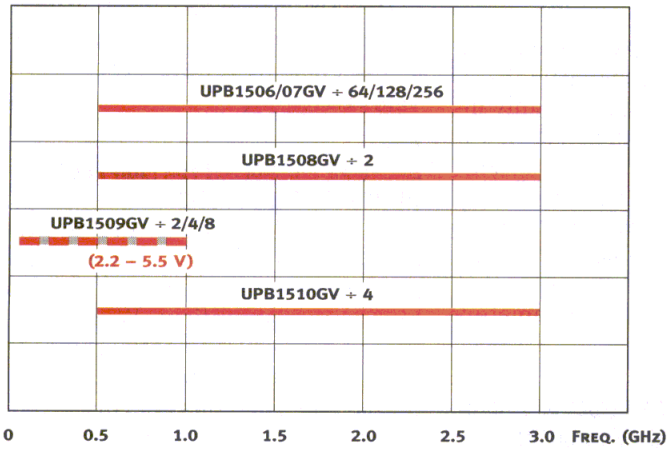
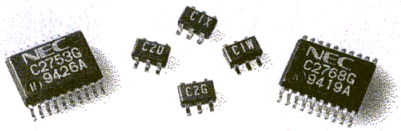


Frequency Range



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UPC2758TB[1]	100 - 2000	11	17	+6
UPC2768GR[1]	10 - 450	7	80	-17
UPC8106TB[2]	100 - 2000	9	9	+1
UPC8112TB[1]	800 - 2000	8.5	13	-10
UPC8116GR[3]	100 - 500	4.1	6.5	—

1. Downconverter 2. Upconverter 3. AM/ASK Receiver IC

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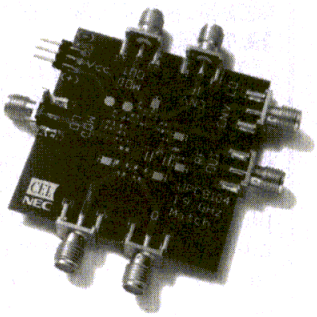

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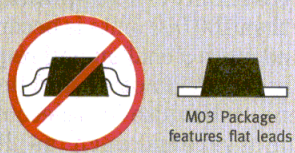

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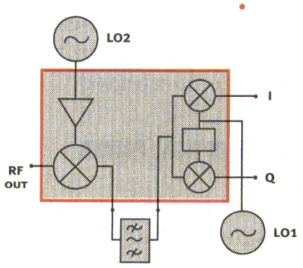

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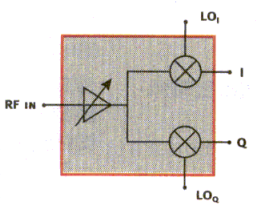
- 1.9GHz RF Output



UPC2766GR

IQ Demodulator

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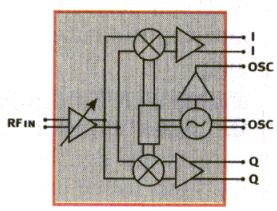
- IF BW – DC to 100MHz
- 35dB typ AGC dynamic range
- 30dBc typ distortion



UPC3205GR

IQ Demodulator

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CIRCLE NO. 244

Sunday, June 11, 2000

8:00 AM-5:00 PM

WSA

SILICON/SILICON-GERMANIUM BiCMOS PROCESSES AND CIRCUIT TECHNIQUES FOR RF ICs

RF IC BiCMOS processes are coupled with analog design techniques that are often unfamiliar to the RF circuit designer. These circuit techniques are employed in addition to familiar RF design techniques to take full advantage of process characteristics. The second half of the workshop will concentrate on RF IC BiCMOS circuit techniques employed by several successful RF IC designers.

WSB

EFFICIENCY AND LINEARITY ENHANCEMENT METHODS FOR PORTABLE RF/MW POWER AMPLIFIERS

The digital modulation formats used in present second-generation cellular systems, proposed 3G cellular systems, and many other wireless applications require linear power amplifiers in the transmitter unit of the mobile radio. The workshop will be tutorial and advanced in nature. The operation of various emerging amplifier architectures will be reviewed and their performance potential, cost, and size attributes examined.

WSC

MICROWAVE FILTER SYNTHESIS AND EQUIVALENT CIRCUIT EXTRACTIONS

Network synthesis using modern methods based on transformed variables will be presented. Examples to be presented include cross-coupled filters, physically asymmetric filters (such as singly terminated structures) and other new realizations. The process of equivalent circuit-element extraction using electromagnetic simulations also will be described.

WSD

BLUETOOTH TECHNOLOGY FROM CONCEPT TO IMPLEMENTATION

This workshop will present an overview of Bluetooth technology. The history and current status of Bluetooth will be introduced. Potential applications will be presented. System-level specifications and considerations will be discussed. Bluetooth baseband and radio architecture will be addressed. Finally, circuit implementations of Bluetooth on different technologies (CMOS, BiCMOS, and SOI) will be presented.

WSE

In this workshop, participants will obtain a clear understanding of the issues involved in designing highly integrated packages, an in-depth understanding of the current manufacturing techniques that the packaging industry is using, and an overview of the development of novel, millimeter-wave packaging technologies that will impact the future of millimeter-wave packaging.

WSF

AN INTRODUCTION TO THE THEORY AND PRACTICE OF NUMERICAL ELECTROMAGNETICS

The purpose of this tutorial is to provide insight into the operating principles of electromagnetic simulators and show how these principles translate into their properties as engineering tools. The characteristics of frequency and time-domain simulators based on finite-element, finite-difference, and method-of-moment formulations will be presented in terms accessible to the practitioner, and the resulting performance profiles of simulators based on these methods will be demonstrated.

WSG

SPATIAL AND QUASI-OPTICAL POWER COMBINING ARRAYS

In this workshop, successful circuit-design approaches as well as CAE tools employed for circuit-level as well as electromagnetic modeling of quasi-optical circuits and systems will be presented. Deliberations in the workshop will emphasize circuit/electromagnetic-field interaction and modeling, 3D interconnects, the design of efficient coupling systems, electrothermal issues and circuit/device tolerances.

WSH

ANTENNAS AND PROPAGATION FOR WIRELESS COMMUNICATION

This one-day tutorial/workshop will provide a fundamental and broad introduction to antenna properties, antenna-design considerations and RF propagation issues. The workshop will begin with the basic concepts and definitions used in the antenna and propagation industry. It will conclude with some practical examples of the concepts presented and demonstrations of actual antenna design using some commercially available antenna-design software.

1:00 PM-5:00 PM

WSI

EMERGING TECHNOLOGY FOR HIGH-POWER > 100-W POWER AMPLIFIERS

This workshop will include an overview of the current state-of-the-art techniques applied in production high-power amplifiers. Wide bandgap semiconductor developments then will be presented and their potential effects on this field discussed. Advanced cooling technologies and their potential impact also will be covered.

WSJ

NEW DEVELOPMENTS IN MIXED-MODE TECHNIQUES FOR NONLINEAR CIRCUIT DESIGN

This workshop will present in-depth tutorial discussions as well as new developments in mixed-mode techniques for analyzing nonlinear circuits. It will also provide discussions on problems of consistent time-domain modeling and new ideas for circuit-simulation algorithms. Participants are strongly encouraged to bring transparencies and to explain their point of view.

WSK

MICRO-/MILLIMETER-WAVE TRANSCEIVERS FOR MASS PRODUCTION DESIGN, TECHNOLOGY, IMPLEMENTATION

Production of micro-/millimeter-wave transceivers in quantities higher than 100,000 units is becoming more and more a challenging request not only for automotive and industrial sensors but also for telecommunication systems such as PTP, PTMP/LMDS, and WLAN. In the workshop, technical and industrial experts will present and discuss the implementation and production processes of several micro-/millimeter-wave transceivers.

MONDAY, June 12, 2000

8:00 AM-NOON

WMI

DIGITAL RECEIVERS FOR MILITARY AND COM-

HIGH IP3 MIXERS

When your wireless communications system calls for very low intermodulation distortion and enhanced dynamic range, look into **Synergy's** new line of ***HIGH IP3 MIXERS***. Standard models are available in specialized frequency bandwidths covering UHF, Cellular, PCS and ISM bands. Additional features are low conversion loss and high interport isolation. Most models operate at +17 dBm of local oscillator drive level and exceed +30 dBm of input third order intercept point. Higher L.O. drive level models with higher third order intercept points are also available.

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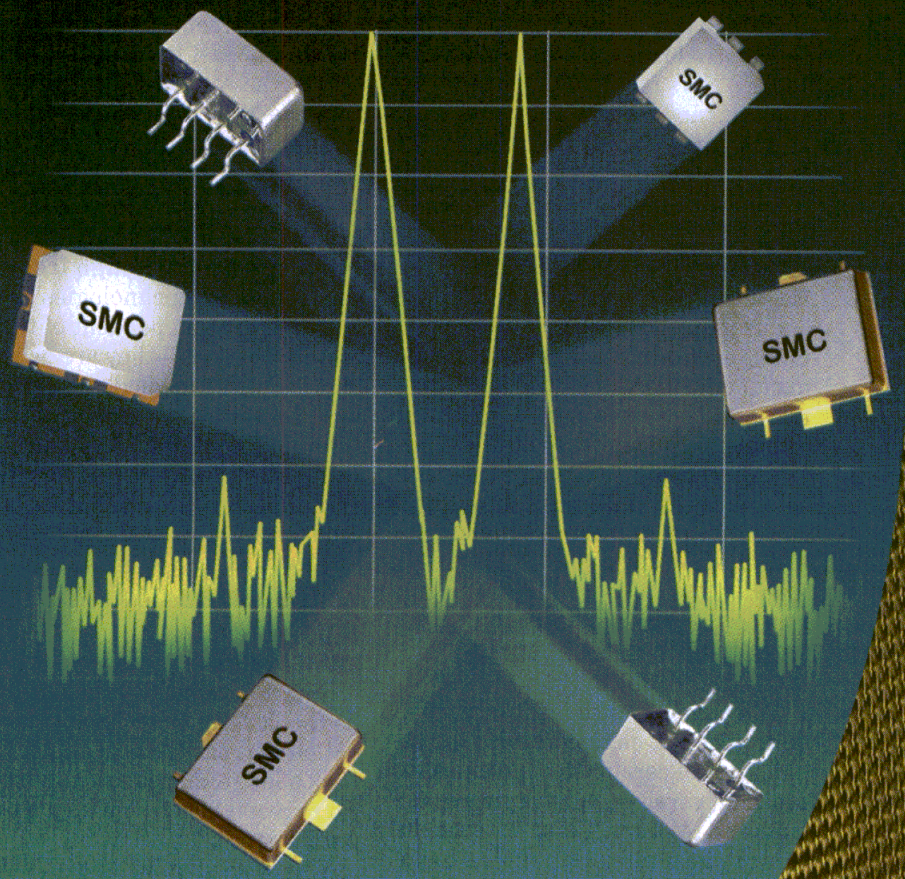

For additional information, contact Synergy's sales and application team:

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201 McLean Boulevard, Paterson, NJ 07504
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E-mail: sales@synergymwave.com
Web site: www.synergymwave.com

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See Us At MTT-S Booth #506 CIRCLE NO. 305

MERCIAL APPLICATIONS

Several wideband and narrowband digital receiver systems have been developed for both military and commercial applications during the past few years. This workshop will bring together leading digital-receiver experts. It will begin with tutorial talks on folding and interpolating (for wideband) and sigma-delta over-sampling (for large dynamic range) ADCs to provide a basic understanding of the crucial elements of digital receivers. It is aimed at providing basic digital receiver concepts and as the latest developments in both the military and commercial world.

NOON-1:15 PM

PMA

WILL SiGe STEP ON THE GaAs?

Research and development (R&D) of silicon-germanium (SiGe) technology has been ongoing for the last 12 years. Repeated promises of performance superior to GaAs at reduced cost have not materialized in that time frame. A number of questions that need to be answered before the success of this technology can be fully assessed are both technical and financial in nature.

8:00 AM-5:00 PM

WMA

RF POWER AMPLIFIERS, CLASSES A THROUGH S—HOW THEY OPERATE, HOW TO DESIGN THEM, AND WHEN TO USE EACH

This full-day course is for beginner through advanced design engineers and their supervisors who are concerned with designing any type of RF power-amplifier product. Learn how all types of linear and nonlinear RF PA circuits operate, how to design them and when to use each, and be able to design manufacturable RF PAs that work satisfactorily under all foreseen normal and abnormal operating conditions.

WMB

WAVELETS FOR EM, DEVICE, AND CIRCUIT MODELING

The use of wavelets in scientific computations is an area that has been exploding in significant results and applications, much faster than written scientific documentation can keep up. The application of wavelets for the solution of problems in electromagnetics and RF circuits has been a very recent and exciting activity.

The goals of this workshop are to introduce computational EM professionals to a rapidly progressing new subfield in the electronics disciplines, to acquaint RF professionals with the major impacts that this technical area will have on RF applications and to provide practical "how to" information for researchers.

WMC

APPLICATIONS OF ARTIFICIAL NEURAL NETWORKS TO RF AND MICROWAVE DESIGN

The recent introduction of artificial neural networks (ANN) to the microwave field marks the birth of an unconventional alternative to modeling and design optimization problems in RF and microwave CAD. This tutorial introduces the fundamentals of using ANNs for RF and microwave modeling and design. The course will also bring the audience to the forefront of this emerging field with state-of-the-art research results.

WMD

FUTURE TRENDS IN FERRITE DEVICES AND TECHNOLOGY

Future trends in ferrite technology include challenges in applications, manufacturing techniques and materials, and cost is a critical factor in commercial and defense applications. Novel applications are anticipated in several areas including solitons, gyroelectric media, and frequency-selective MSW devices.

WME

ADVANCES IN CERAMIC INTERCONNECT TECHNOLOGIES FOR WIRELESS, RF, AND MICROWAVE APPLICATIONS

The purpose of the workshop will be to acquaint engineers with the capabilities of ceramic technologies for realizing wireless and microwave circuits and systems. Ceramic technologies are generally less familiar to design engineers and yet offer significant benefits in terms of circuit performance, weight, and reliability.

WMF

EM-BASED CAD OF PRINTED-CIRCUIT COMPONENTS, WAVEGUIDE NETWORKS AND ANTENNAS—STATE-OF-THE-ART AND PROMISING TECHNIQUES FOR THE NEXT DECADE

The focus of this workshop will be the presentation of the state-of-the-art and recent progress in the development of advanced EM techniques and their use for performance assessments and the design of key components relevant to communication systems for the 21st century. Examples will be given from the application of such CAD capability to a variety of practical component design problems, such as printed-circuit antennas and arrays, microwave circuits, planar circuit interconnects, and waveguide filters.

WMG

SPACE-BASED RADAR MICROWAVE LINKS

This workshop will consist of tutorial sessions on the basic AF Space-based-radar (SBR) program, SBR microwave/photonics requirements and some system trade study results, a review of past microwave photonics efforts, the Air Force Research Lab's in-house Microwave Links program, and a basic review of microwave-link engineering for SBR applications.

WMH

RF FRONT-END ARCHITECTURES

Developments of high frequency semiconductor technologies are to a large degree independent of the system applications—they are driven by improvements in yield, materials, device, and circuit design. This workshop shall contribute to an architecture bridge between technology and systems in order to facilitate the introduction of revolutionary system applications.

1:00 PM-5:00 PM

WMJ

SMART ANTENNAS

Smart antennas are considered to be the last major technological innovation that has the potential for leading to large increases in wireless-communication-systems performance.

This workshop will begin with tutorial presentations that introduce the concept of smart antennas and the role they play, particularly in commercial wireless communication.

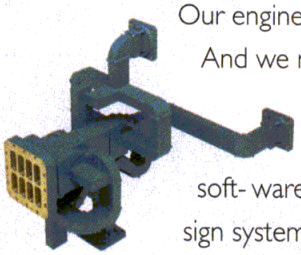
WMK

ULTRA WIDEBAND SYSTEMS AND APPLICATIONS

Ultrawideband (UWB) signals oc-

The fastest tools help us engineer the finest waveguide components.

THE SHORTEST DISTANCE BETWEEN *DESIGN* AND *DONE*.



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- WAVEGUIDE TO COAX ADAPTERS
- WAVEGUIDE PRESSURE WINDOWS
- COMMERCIAL WAVEGUIDE ASSEMBLIES

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Harmonics:	-40 dBc
Frequency Stability:	±5 ppm
Optionally:	±2.5 ppm
Phase Noise at 12.3 GHz (Typ.)	
100 Hz	-83 dBc/Hz
1 k	-100 dBc/Hz
10 k	-110 dBc/Hz
100 k	-113 dBc/Hz
1 M	-128 dBc/Hz

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CIRCLE NO. 340

cupy more than 100-percent instantaneous bandwidth and are generated using intense, short, video-like pulses. This workshop will discuss the history, communication, radar, and future applications of this technology. A demonstration of a UWB communication system will be conducted.

TUESDAY, June 13, 2000

8:00 AM-10:00 AM

10:10 AM-11:50 AM

TU2A

LOW-NOISE TECHNOLOGY

JOINT IMS-RFIC SESSION

Chair: L. Boglione, Co-chair: E. Niehenke

TU2A-1

HBT Low-noise Performance in an 0.18-mm SiGe BiCMOS Technology

TU2A-2

Design of GaAs MMIC Transistors for the Low-Power, Low-Noise Applications

TU2A-3

High-Gain-density K-band PHEMT LNA MMIC for LMDS and Satellite Communication

TU2A-4

High-Performance 60-GHz Coplanar MMIC LNA Using InP Heterojunction FETs with AlAs/InAs Superlattice Layer

TU2A-5

Design and Performance of Wideband, Low-noise, Millimeter-wave Amplifiers for Microwave Anisotropy Probe Radiometers

TU2A-6

Design and Characterization of MMIC Active Cold Loads

TU2B

MILLIMETER-WAVE SIGNAL SOURCES

Chair: R. Alm, Co-chair: J.F. Luy

TU2B-1

Low-Phase-Noise 58-GHz SiGe HBT Push-push Oscillator with Simultaneous 29-GHz Output

TU2B-2

A 94-GHz HEMT Oscillator Using High Order Subharmonic Synchronization

TU2B-3

Low-Phase-Noise Signal-Generation Circuits for 60-GHz Wireless Broadband System

TU2B-4

A Low Phase-Noise 38-GHz HBT MMIC Oscillator Utilizing a Novel Transmission-Line Resonator

K. Hosoya, S. Tanaka, Y. Amamiya, NEC Corp., Japan

TU2B-5

A Power Combined W-band HBT Oscillator

TU2C

MILLIMETER-WAVE PACKAGING

Chair: R.W. Jackson, Co-chair: J. Pavio

TU2C-1

Low-Cost Multilayer Ceramic Package for Flip-chip MMIC up to W-band

TU2C-2

Millimeter-wave Ceramic Package for a Surface Mount

TU2C-3

Development of a 36-GHz Millimeter-wave BGA Package

TU2C-4

Optimized Flip-chip Interconnect

HARMONIC (COMB) GENERATORS 0.1 - 50 GHz

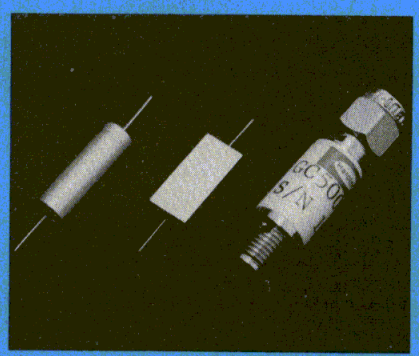
NEW - *COAXIAL TO 50 GHz*

GC Series Comb Generators

GC Typical Specifications

Model	Input Freq (MHz)	Minimum Output (dBm) @ GHz				
		12.4	18	26	40	50
GC100	100	-30	-40	—	—	—
GC500	500	-15	-20	—	—	—
GC1000	1000	-10	-15	-35	—	—
GC2026	2000	0	-10	-20	—	—
GC1040A	1000		-15	-30	-45	—
GC1540A	1500		-10	-25	-40	—
GC2040A	2000		-5	-15	-30	—
GC1050A	1000		-15	-30	-45	-50
GC1550A	1500		-10	-25	-40	-50
GC2050A	2000		-5	-15	-30	-40

- Broadband Output to 50 GHz
- No Bias Required
- Drop-In Modules or with Connectors
- Input Frequencies 10 MHz to 2 GHz
- Specified at 1/2 Watt Drive
- Surface Mount in 0.25" SQ with formed leads
- 50 GHz units in 1.32" long coaxial



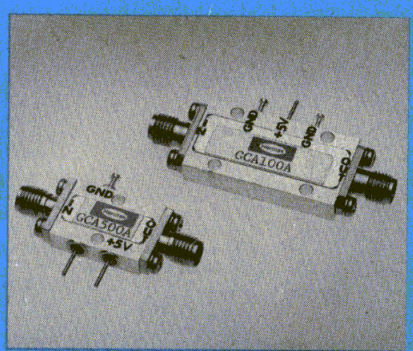
GC Series

GCA Series Comb Generators with Integral Preamplifiers

- 0 dBm or +10 dBm Input
- Small Size
- Input Frequencies 30 MHz to 2 GHz
- +5 VDC Power Supply

GCA Typical Specifications

Model	Input Freq (MHz)	Minimum Output (dBm) @ GHz		
		12.4	18	26
GCA0526	500	-15	-20	-40
GCA1026	1000	-10	-15	-35
GCA1526	1500	-5	-10	-25
GCA2026	2000	0	-10	-20



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TU2C-5
RF W-band Wafer-to-wafer Transition

TU2D
ADVANCED TOPICS IN NONLINEAR CAD
Chair: P. Draxler, Co-chair: T. Brazil

TU2D-1
Accurate RF and Microwave System Level Modeling of Wide Band Nonlinear Circuits

TU2D-2
Circuit Reduction Technique for Finding the Steady State Solution of Nonlinear Circuits

TU2D-3
Determining the Steady-state Responses in RF Circuits Using GMRES, CGS, and BICGSTAB Solutions in sSPICE for Linux

TU2D-4
Rigorous RF and Microwave Oscillator Phase Noise Calculation by Envelope Transient Technique

TU2D-5
Techniques for Oscillator Nonlinear Optimization and Phase-noise Analysis Using Commercial Harmonic-balance Software

TU2D-6
Large-displacement Modeling and Simulation of Micromechanical Electrostatically Driven Resonators Using the Harmonic Balance Method

TU2E
ADVANCED CONCEPTS IN FREQUENCY-DOMAIN NUMERICAL TECHNIQUES
Chair: M.Salazar-Palma, Co-chair: H.Y.D. Yang

TU2E-1
Efficient MM/FE GSMS Technique for the CAD of Broadband Lateral Coax Feeds in Rectangular Waveguide

TU2E-2
Modeling of Arbitrary Shaped Radiating Structures by the Wave Concept Iterative Process

TU2E-3
Macro-elements for Efficient FEM Simulation of Small Geometric Features in Waveguide Components

TU2E-4
The Perfectly Matched Layer as Lateral Boundary in Finite-difference Transmission-Line Analysis

TU2E-5
Analysis of Planar Microwave and Millimeter-wave Circuits with Anisotropic Layers Based on Generalized Transmission-Line Equations and on the Method of Lines

TU2E-6
Applying Wavelets to Electromagnetic Field Simulation: The Method of Lines

NOON-1:15 PM

PTB
HIGH-DATA-RATE COMMUNICATION—PREPARING THE FUTURE

This panel will address the actual preparations for high-data-rate communication expected to become a reality during the first decade of the next century.

1:20 PM-3:00 PM

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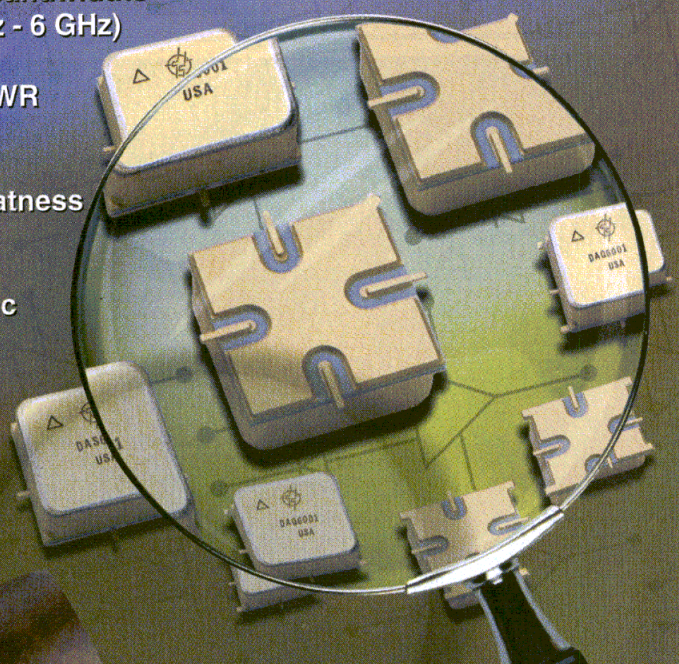

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CIRCLE NO. 243

TU3A
WIRELESS CIRCUIT TECHNIQUES
JOINT IMS-RFIC SESSION
Chair: N. Camilleri, Co-chair: K.B. Ashby

TU3A-1
(Invited) FM Demodulation Using an Injection-locked Oscillator

TU3A-2
(Invited) (1:45 PM) A Fast Response, Programmable PA Regulator Subsystem for Dual Mode CDMA/AMPS Handsets

TU3A-3
A Comparison Study of ESD Protection for RF ICs: Better Performance and Less Parasitic

TU3A-4
(2:35 PM) Effects of Offsets on Bipolar Integrated-Circuit Mixer Even-Order Distortion Terms

TU3B
MEMS SWITCHES/PHASE SHIFTERS
FOCUSED SESSION
Chair: B.E. Sigmon, Co-chair: D. Peterson

TU3B-1
MEMS Microswitch Arrays for Reconfigurable Distributed Microwave Components

TU3B-2
A Low-Voltage Actuated Micromachined Microwave Switch Using Torsion Springs and Leverage

TU3B-3
One- and Two-bit Low-loss Cascadable MEMS Distributed X-band Phase Shifters

TU3B-4
Design of Low Actuation Voltage RF MEMS Switches

TU3B-5
High-isolation Inductively-tuned X-band MEMS Shunt Switches

TU3B-6
Characterization of Source-to-drain Capacitance (CDS) Effect of GaAs PHEMT for Microwave and Millimeter-wave Switch

TU3C
FILTERS (1)
Chair: A.E. Atia, Co-chair: M. Guglielmi

TU3C-1
A Tunable X-band Active Notch Filter with Low-distortion Passband Response

TU3C-2
Quasi Dual-mode Resonators

TU3C-3
A First Practical Model of Very Small and Low Insertion Loss Laminated Duplexer Using LTCC Suitable for WCDMA Portable Telephones

TU3C-4
Low-profile TM-mode Dielectric Disk Resonator BPF for WCDMA

TU3C-5
A New Circuit Configuration to Obtain Large Attenuation with a Coupled-resonator Band Elimination Filter Using Laminated LTCC

TU3C-6
A Miniaturized Bandpass Filter with Double Surface Electrodes

TU3D
MONOLITHIC AND SEMICONDUCTOR TECHNOLOGY
Chair: Z. Barda, Co-chair: T. Lee

TU3D-1
Magnetic RF Integrated Thin-film Inductors

TU3D-2
Fabrication of High-Frequency Passives on BiCMOS Silicon Substrates

TU3D-3
A Compact and Wideband GaAs PHEMT Distributed Amplifier IC Based on a Micro-machined CPW

TU3D-4
A Compact 5-b Phase Shifter MMIC for K-band Satellite Communication Systems

TU3D-5
RF Loss and Cross Talk on Extremely High Resistivity (10K-1M Ω-cm) Si Fabricated by Ion Implantation

TU3D-6
Design Guide of Coupling between Inductors and Its Effect on Reverse Isolation of CMOS LNA

TU3E
ADVANCES IN TIME-DOMAIN MODELING
Chair: A. Cangellans, Co-chair: A. Beyer

TU3E-1
Unconditionally Stable FDTD Algorithm for Solving Three-dimensional Maxwell's Equations

TU3E-2
Finite Element Time-Domain-Based Diakoptic Method for Microwave Circuit Analysis

TU3E-3
Application of Wavelet-Galerkin Method to Electrically-large Optical Waveguide Problems

TU3E-4
Multi-resolution 2D-TLM Technique Using Haar Wavelets

TU3E-5
An Extension of the MRTD Technique to the Fast Computation of the Radiation from Planar Open Structures

TU3E-6
Characterization of Micromachined Transmission Lines Using MRTD (Multiresolution Time-domain Technique)

TU3E-7
Improved Autoregressive (AR) Signal Modeling for FDTD Resonance Estimation

3:30 PM-5:10 PM

TU4A
INTEGRATED TRANSCEIVERS
JOINT IMS-RF IC SESSION
Chair: S. Kiaei, Co-chair: D. Lovelace

TU4A-1
(Invited) SiGe Bipolar Technologies for Low-Phase-Noise RF and Microwave Applications

TU4A-2
A Highly Integrated Dual-Band Tri-mode Transceiver Chip set for CDMA, TIA/EIA-95, and AMPS Applications

TU4A-3
A Highly Integrated Radio for High-Performance, Low-Cost 2.4-GHz ISM Cordless Applications

TU4A-4
An RF Transceiver for WDCT in a 25-GHz Silicon Bipolar Technology

TU4B
MICROMACHINING AND MEMS TECHNOLOGY
FOCUSED SESSION
Chair: G. Rebeiz, Co-chair: H.A. Hung

TU4B-1
Fabrication of High-performance On-chip Suspended Spiral Inductors by Micromachining and Electroless Copper Plating

TU4B-2
Post-processed RF Micromechanical Switches on Commercial RF ICs

TU4B-3
Fully Integrated Micromachined Capacitive Switches for RF Applications

TU4B-4
X-band and K-band Lumped Wilkinson Power Dividers with a Micromachined Technology

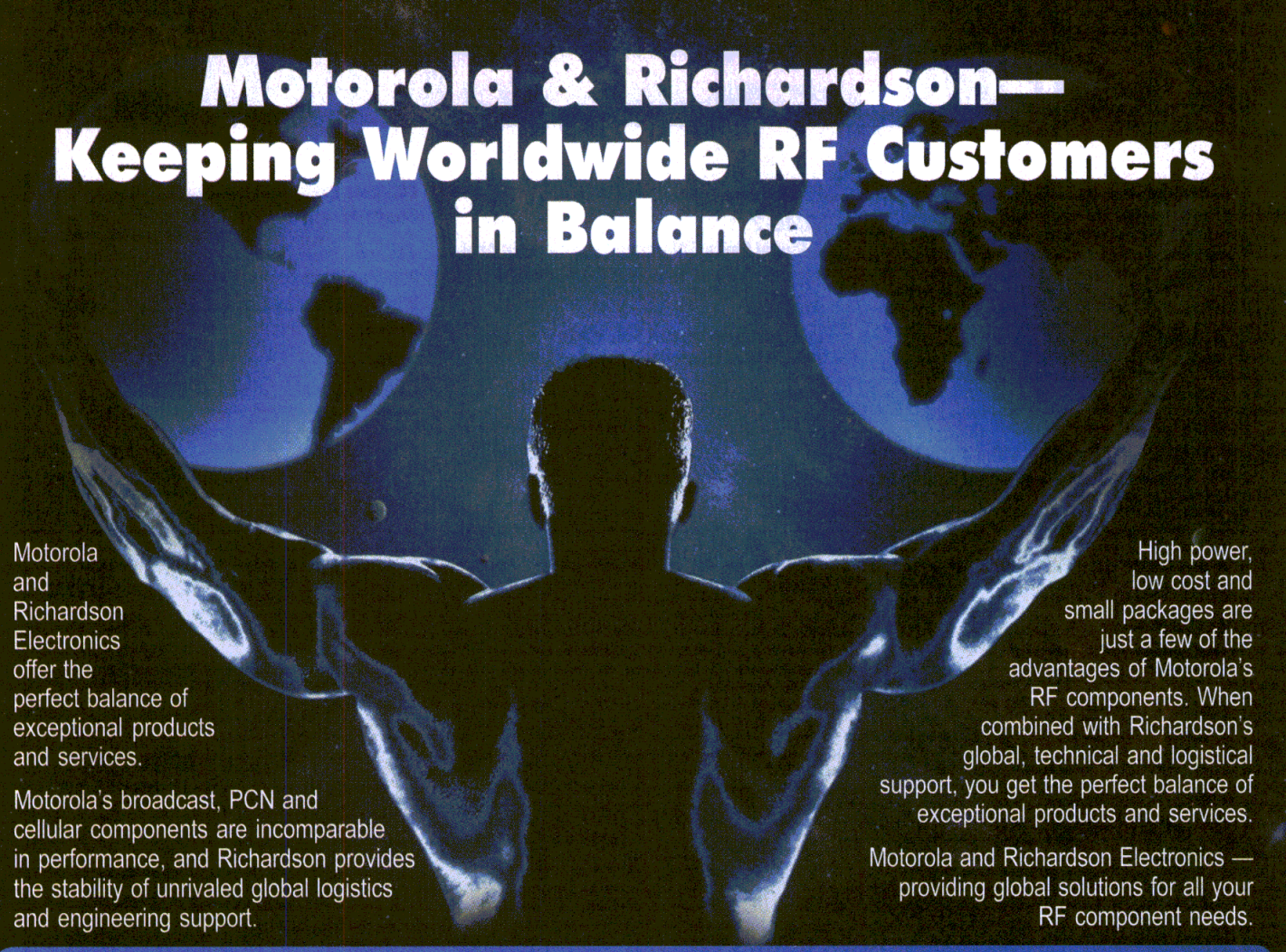
TU4B-5
Micromachined Branch-Line Coupler in CMOS Technology

TU4B-6
Micromachined Circuit-Combining Networks for W-band Applications

TU4B-7
A New Micromachined Overlay CPW Structure with Low Loss over Wide Impedance Ranges

TU4B-8
Multilevel Finite Ground Coplanar-Line Tran-

Motorola & Richardson— Keeping Worldwide RF Customers in Balance



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Motorola and Richardson Electronics — providing global solutions for all your RF component needs.

PCN PRODUCT

Device	Frequency	Voltage	Oper. Gain (Typ.)	Output Power
MRF18060A/AS	1805-1880 MHz	26 Volts	13.0 dB	60 Watts CW
MRF18060B/BS	1805-1880 MHz	26 Volts	13.0 dB	60 Watts CW
MRF18090A/AS	1930-1990 MHz	26 Volts	13.5 dB	90 Watts CW
MRF18090B/BS	1930-1990 MHz	26 Volts	13.5 dB	90 Watts CW
MRF19030/S	1930-1990 MHz	26 Volts	13.0 dB	30 Watts PEP
MRF19045/S	1930-1990 MHz	26 Volts	14.0 dB	45 Watts PEP
MRF19060/S	1930-1990 MHz	26 Volts	12.5 dB	60 Watts PEP
MRF19085/S	1930-1990 MHz	26 Volts	12.5 dB	90 Watts PEP
MRF19125/S	1930-1990 MHz	26 Volts	12.5 dB	125 Watts PEP
MRF21125/S	1930-1990 MHz	28 Volts	12.0 dB	125 Watts PEP
MRF21180/S	1930-1990 MHz	28 Volts	11.3 dB	160 Watts PEP

CELLULAR

Device	Frequency	Voltage	Oper. Gain (Typ.)	Output Power
MRF9180	880 MHz	26 Volts	17.0 dB	180 Watts PEP
MRF9085/S	880 MHz	26 Volts	17.0 dB	85 Watts PEP
MRF9045/S	945 MHz	28 Volts	18.0 dB	45 Watts PEP
MRF9045M	945 MHz	28 Volts	16.0 dB	45 Watts PEP

BROADCAST

Device	Frequency	Voltage	Oper. Gain (Typ.)	Output Power
MRF372	470-860 MHz	28 Volts	14.0 dB	180 Watts PEP
MRF373A/AS*	470-860 MHz	28 Volts	11.2 dB	100 Watts PEP
MRF374A*	470-860 MHz	28 Volts	12.0 dB	100 Watts PEP

** Under development.*

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CIRCLE NO. 333

sitions for High-density Packaging Using Silicon Micromachining

TU4C
FILTERS (2)
Chair: H.C. Bel, Co-chair: S.J. Fiedziuszko

TU4C-1
Compact Multi-level Folded-line Bandpass Filters

TU4C-2
Aperture Compensation Technique for Innovative Design of Ultra-broadband Microstrip Bandpass Filter

TU4C-3
Transmission-Line Filters with Advanced Filtering Characteristics

TU4C-4
Third Integrated Narrowband Filters for Millimeter-wave Wireless Applications

TU4C-5
A Design of the Novel Coupled-Line Bandpass Filter Using Defected Ground Structure

TU4C-6
Microstrip Triangular Patch Resonator Filters

TU4C-7
Edge-coupled Coplanar Waveguide Bandpass Filter Design

TU4C-8
Development of Integrated Three-Dimensional Bluetooth Image-Reject Filter

TU4D
MICROWAVE DIGITAL CIRCUITS
Chair: K.C. Wang, Co-chair: A. Oki

TU4D-1
A 40-Gb/s Monolithic Digital OEIC Module Composed of Uni-traveling-carrier Photodiode and InP HEMT Decision Circuit

TU4D-2
Compact Differential InP-based HBT VCOs with a Wide Tuning Range at W-band

TU4D-3
A 50-GHz Monolithic RSFQ Digital Phase Locked Loop

TU4D-4
A 13-GHz Phase-Locked-Loop Development for Micro- and Millimeter-wave Applications

TU4D-5
Prospects of Silicon-Germanium-Based Technology for Very High-speed Circuits

TU4E
APPLICATIONS OF TIME-DOMAIN MODELING
Chair: R. Vahldieck, Co-chair: L. Roselli

TU4E-1
Time-Domain Global Modeling of EM Propagation in Semiconductor Using Irregular Grids

TU4E-2
A Global Finite Difference Time-Domain Analysis of a Silicon Nonlinear Transmission Line

TU4E-3
Interpolating Wavelet Scheme toward Global Modeling of Microwave Circuits

TU4E-4
Direct Synthesis of Microwave Filters Using Inverse Scattering TLM (Transmission-Line Matrix) Method

TU4E-5
Extension of the Lumped Network (LN)-FDTD Approach to the Modeling of Nonlinear Elements

TU4E-6
Methods for the Modeling of Thin Wire Structures with the TLM Method

TU4E-7
The Inclusion of Fringing Capacitance and Inductance in FDTD for the Robust Accurate Treatment of Material Discontinuities

2:30 PM-5:00 PM

TUIF
INTERACTIVE FORUM
Chair: D. Swanson, Co-chair: D. Teeter

TUIF-1
Low-power Active Mixer for Ku-band Application Using SiGe HBT MMIC Technology

TUIF-2
A mm-wave Monolithic Gilbert Cell Mixer

TUIF-3
Approximate Analytical Modeling of Current Crowding Effects in Multi-turn Spiral Inductors

TUIF-4
Novel Alternative Design Methods Based upon Combined Amplification, Matching, and Filtering Approaches in MMIC Technology

TUIF-5
Monolithic Upconversion and Reference Downconverter IC for Power Amplifier Linearization Using GaAs HBTs

TUIF-6
A Unified Approach of Near- and Far-carrier PM Noise Calculation in Large Multitone Autonomous and Forced Circuits at RF and Microwaves

TUIF-7
Analysis of mm-wave Nonlinear Circuits by Combining Genetic Algorithm and Harmonic Balance Technique

TUIF-8
The Estimation of Volterra Transfer Functions with Applications to RF Power-Amplifier Behavior Evaluation for CDMA Digital Communication

TUIF-9
Fast Prediction of the Performance of Wireless Links by Simulation-trained Neural Networks

TUIF-10
Linearity Considerations of WCDMA Front-ends for UMTS

TUIF-11
Optimal Filters for Nonlinear Power-Amplifier Modeling and Equalization

TUIF-12
Computer-aided Optimization of Adjacent Channel Power in Nonlinear Communications Amplifiers

TUIF-13
Two-tone IMD Asymmetry in Microwave Power Amplifiers

TUIF-14
A New Method in Characterizing the Nonlinear Current Model of MESFETs Using Single-tone Excitation

TUIF-15
Generalized Distributed Nonlinear Device Modeling for Krylov-subspace-Based Microwave Circuit Analysis

TUIF-16
Evaluation of Non-linear Modeling Techniques Based on Vectorial Large-signal Measurements as Applied to MOSFETs Coupled Electro-thermal Simulations and Measurements of Power FETs

TUIF-18
A Computational Load-pull System for Evaluating RF and Microwave Power-Amplifier Technologies and Load Impedance on the Intermodulation Products of GaAs FETs

TUIF-20
Additional Conditions to Formation of the Maximally Flat Waveforms in a Class F Power Amplifier

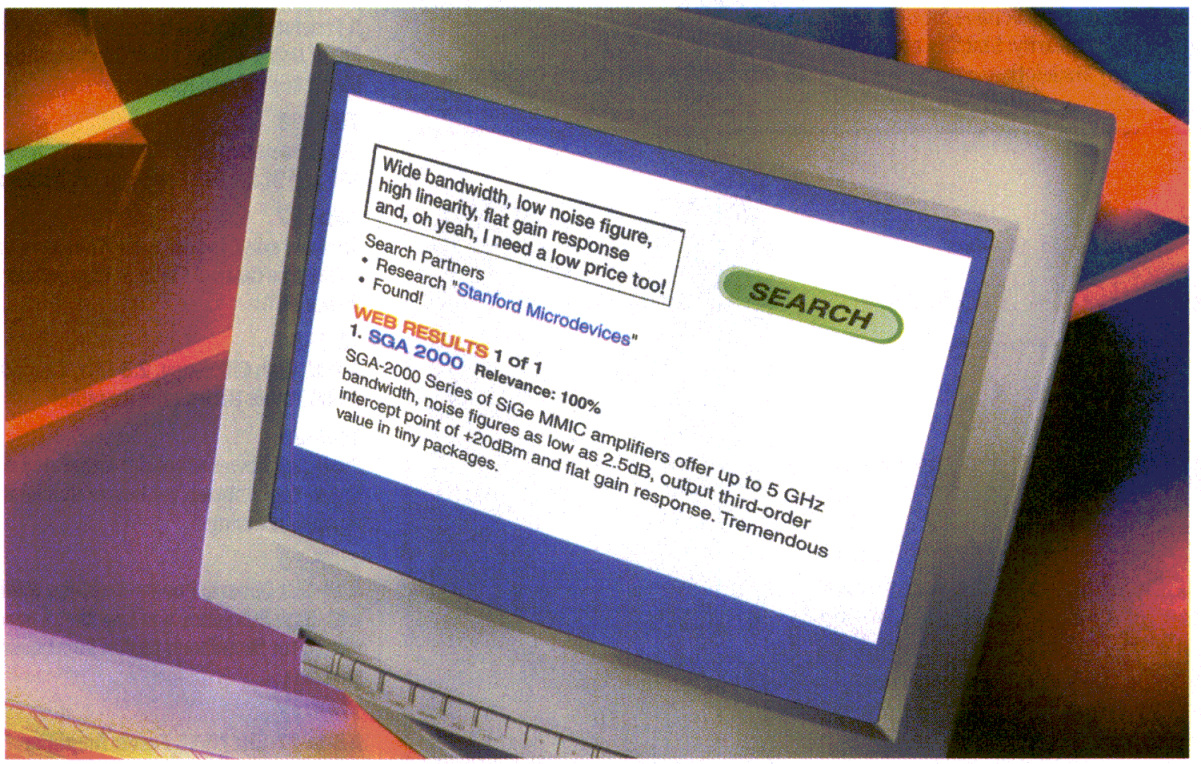
TUIF-21
On Robust Suppression of Third-order Intermodulation Terms in Small-signal Bipolar Amplifiers

TUIF-22
Performance Improvement of Feedforward Power Amplifiers by Using a Novel Injection Combining Technique

TUIF-23
Adaptive Algorithms for Feedforward Power-Amplifier Linearizer

TUIF-24
Adaptive Baseband/RF Predistorter for

Still searching for the right amplifier for your latest design?



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SPECIFICATION MATRIX

	SGA-2163 SGA-2186	SGA-2263 SGA-2286	SGA-2363 SGA-2386	SGA-2463 SGA-2486
Frequency (GHz)	DC-5.0	DC-3.5	DC-2.8	DC-2.0
Gain (dB)	10.5	15.0	17.4	19.6
TOIP (dBm)	20.0	20.0	20.0	20.0
P1dB (dBm)	7.0	7.0	7.0	7.0
N.F. (dB)	4.1	3.2	2.9	2.5
Supply Voltage (Vdc)	2.2	2.2	2.7	2.7
Supply Current (mA)	20	20	20	20

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

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CIRCLE NO. 267

Power Amplifiers through Instantaneous AM-AM and AM-PM Characterization Using Digital Receiver

TUIF-25

Microwave Power Amplifiers with Digitally-controlled Power-Supply Voltage for High Efficiency and High Linearity: Stability and Incidental Frequency Modulation

TUIF-27

A Programmable High-isolation Monolithic Millimeter-wave SPST Switch

TUIF-28

1-to-26-GHz High Power p-i-n Diode Switch

TUIF-29

38-GHz MMIC PHEMT-based Tripler with Low-Phase-noise Properties

TUIF-30

Broadband, Planar, Doubly Balanced Star Mixers

TUIF-31

A Passive-type Even Harmonic Quadrature Mixer Using Simple Filter Configuration for Direct Conversion Receiver

TUIF-32

CMOS Resistive Ring Mixer MMICs for GSM-900 and DCS-1800 Base-Station Applications

TUIF-33

Analysis and Experimental Waveform Study on Inverse Class-F Mode of Microwave Power FETs

TUIF-34

A 2-W, High Efficiency, 2.8-GHz, Cascode HBT MMIC Power Distributed Amplifier

TUIF-35

Push-pull Power Amplifier Integrated with Quasi-Yagi Antenna for Power Combining and Harmonic Tuning

TUIF-36

A 20-GHz Doherty Power-Amplifier MMIC with High Efficiency and Low Distortion Designed for Broadband Digital Communication Systems

TUIF-37

A 5-to-27-GHz MMIC Power Amplifier

TUIF-38

RF Power-Amplifier Integration in CMOS Technology

TUIF-39

Demonstration of a High-Efficiency Nonuniform Monolithic Gallium Nitride Distributed Amplifier

TUIF-40

A Ka-band HBT MMIC Power Amplifier

TUIF-41

An S-band High-power Broadband Transmitter

TUIF-42

Ka-band, 30-W Solid-State Power Amplifier

TUIF-43

Designing Commercially Viable mm-wave Modules

TUIF-44

Submillimeter Quasioptical Semiconductor Isolators

TUIF-45

A 230-GHz Low-Noise Subharmonically Pumped SIS Mixer

TUIF-46

Low-Cost Millimeter-wave Aperture Coupled Antenna Array on Polymer Membrane Sub-

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CIRCLE NO. 462

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PHILIPS

Philips RF Wideband Transistors

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

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CIRCLE NO. 393

strate

TUIF-47
Improved Photonic Bandgap Cavity and Metal Rod Lattices for Microwave and Millimeter-Wave Applications

TUIF-48
Rectangular Phased-Array Antennas Fed by Single-crystal YIG Phase Shifters

TUIF-49
Reconfigurable Transmission-type Beam-former

TUIF-50
Characterization of a Multi-mode Microstrip Aperture for Phased-Array Applications

TUIF-51
Measurement of the Phase and Amplitude Distributions of Coupled Oscillator Arrays

TUIF-52
Range Demonstration of an Ultra-wideband, Continuous, Time-Steered Array Using a Fiber-optic, Cascaded Grating Prism

Wednesday, June 14, 2000
8:00 AM-9:40 AM

WE1A
NEW TRANSMISSION-LINE ELEMENTS
Chair: C.K.C. Tzuang, Co-chair: M. Dydyk

WE1A-1
Engraved NRD Guide for Millimeter-wave Integrated Circuits

WE1A-2
Characterization of Asymmetric Coupled CMOS Lines

WE1A-3
A Novel Low Impedance Line for MMIC Using Air-gap Stacked Structure

WE1A-4
A Broadband CPW-to-waveguide Transition Using Quasi-Yagi Antenna
N. Kaneda, Y. Qian, T. Itoh

WE1A-5
Novel Quadrature Branch-line Coupler Using CPW-to-microstrip Transitions

WE1A-6
Excitation of the Microstrip Higher-Order Leaky Modes by Aperture-coupling Method

WE1B
MIXER TECHNIQUES
Chair: B. Nelson, Co-chair: L. Reynolds

WE1B-1
A Novel DC-offset Cancellation Technique for Even-harmonic Direct Conversion Receivers

WE1B-2
A Novel Uniplanar Balanced Subharmonically Pumped Mixer for Low-cost Broadband Millimeter-wave Transceiver Design

WE1B-3
A Ku-band Frequency Converter Using 0.25-mm PHEMT Technology

WE1B-4
A New Millimeter-wave MMIC Mixer for Sensor Applications

WE1B-5
Ultra-small MMIC Mixers for K- and Ka-band Communications

WE1C
SUPERCONDUCTING COMPONENTS AND TECHNOLOGY
Chair: G. Lyons, Co-chair: R. Mansour

WE1C-1
A Narrowband HTS Bandpass Filter at 18.5 MHz

WE1C-2
Superconductors and Cryotechnology for Space Communications—Adaptation of a New Technology for Applications

WE1C-3
Highly Selective HTS Bandpass Filter with Multiple Resonator Cross-Couplings

WE1C-4
High-Temperature Superconducting Nonlinear Transmission Lines

WE1C-5
Superconducting Filter for IMT-2000 Band

WE1C-6
Superconducting Microstrip Line Bandpass Filter for Mobile Applications

WE1C-7
An HTS Microwave Narrowband Filter Using Dual-mode Multi-zigzag Microstrip Loop Resonators

WE1D
MICROWAVE PHOTONICS: DEVICES AND SYSTEMS
Chair: D. Jager, Co-chair: C. Gee

WE1D-1
High-power Waveguide Integrated Photodiode with Distributed Absorption

WE1D-2
40-GHz Broadband Optical Receiver Combining a Multimode Waveguide Photodiode Flip-chip Mounted on a GaAs-based HEMT Distributed Amplifier

WE1D-3
Bias-Free Optical Control of Microwave Circuits and Antennas Using Improved Optically Variable Capacitors

WE1D-4
Wide-band Analog-digital Photonic Link with Third-order Linearization

WE1D-5
Design and Characterization of a Fiber-optic RF Remoting Link

WE1D-6
Integrated Optic RF Phase Shifter for Continuous Beam Steering at 1-to-1200-MHz

WE1E
X-BAND T/R MODULES: SEMICONDUCTORS, PACKAGING, AND ASSEMBLY ISSUES
SPECIAL SESSION
Chair: B. Kopp

WE1E-1
(Invited) X-band Transmit/Receive Module Overview

WE1E-2
(Invited) Wideband Semiconductors for X-band Amplifiers

WE1E-3
(Invited) X-band GaAs MMIC Size Reduction and Integration

WE1E-4
(Invited) Microwave Flip Chip and B&A Technology

WE1E-5
(Invited) Advanced Technology for Future Space-Based Antennas

10:10 AM-11:50 AM

WE2A
ADVANCES IN MICROWAVE SIGNAL GENERATION
Chair: A. Boudiaf, Co-chair: S. Wetenkamp

WE2A-1
Low Phase Noise, Low-Power IC VCOs for 5-to-7-GHz Wireless Applications

WE2A-2
SiGe MMICs Beyond 20 GHz on a Commercial Technology

WE2A-3
A Low-Power 20-GHz SiGe Dual-modulus Prescaler

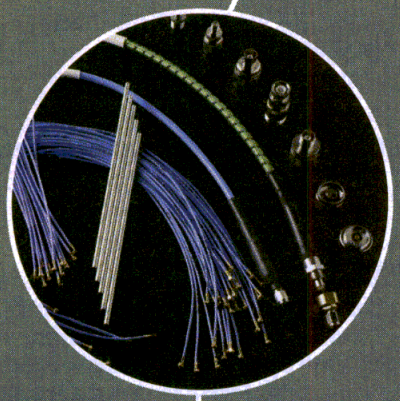
WE2A-4
A Low-voltage 6-GHz-band CMOS Monolithic LC-tank VCO Using a Tuning-range Switching Technique

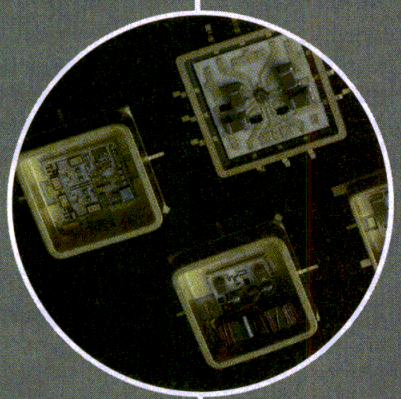
WE2A-5
Building of a Highly Precise and Compact Synthesizer in E-band by Means of Gunn Oscillator and MMIC Chip set

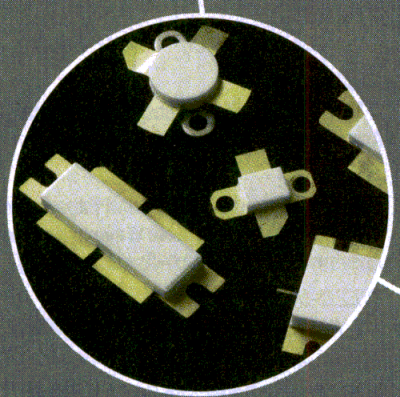
WE2B
NONLINEAR TRANSISTOR MODELING
Chair: J. Atherton, Co-chair: W. Curtice

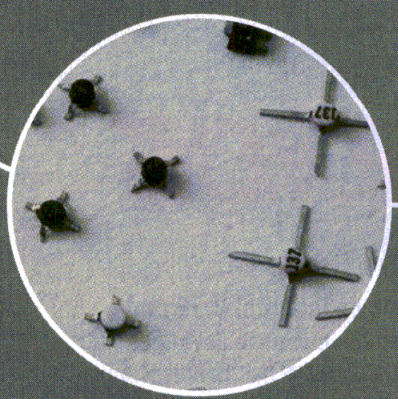
WE2B-1
An Improved Deep Sub-micron MOSFET RF Nonlinear Model with New Breakdown Current Model and Drain to Substrate Nonlinear

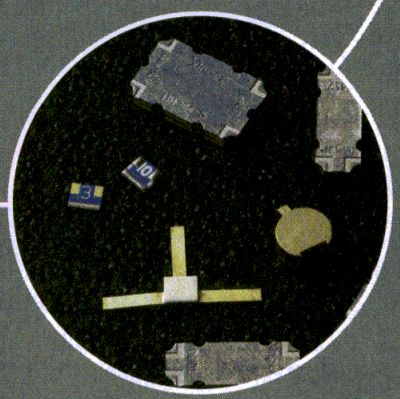
WHY GO ANY FURTHER?











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Coupling

WE2B-2
Accurate Large-signal Modeling of SiGe HBTs

WE2B-3
Scalable GaInP/GaAs HBT Large-signal Model

WE2B-4
Influence of Collector Design on InGaP HBT

Linearity

WE2B-5
Validation of an Analytical Large-Signal Model for AlGaN/GaN HEMTs

WE2B-6
Characterization and Modeling of Nonlinear Trapping Effects in Power SiC MESFET

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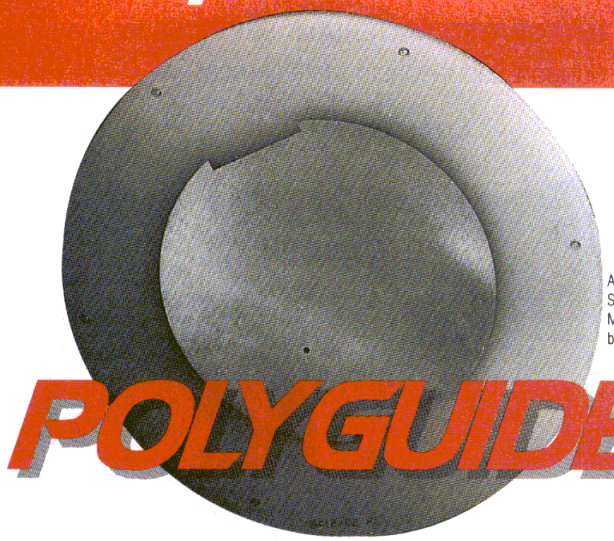

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CIRCLE NO. 407 or visit www.mwrf.com

WE2C
HF/VHF/UHF POWER AMPLIFIERS AND LINEARIZERS
Chair: F. Raab, Co-chair: J. Modelski

WE2C-1
Circuit-Design Technique for High Efficiency Class F Amplifiers

WE2C-2
Analysis of Class-F and Inverse Class-F

WE2C-3
Class-E Switching-mode High-efficiency Tuned RF/Microwave Power Amplifier: Improved Design Equations Include Effects of Load-network Finite-Loaded Q and Resulting Harmonic Currents in Load Network

WE2C-4
Class E RF/Microwave Power Amplifier: Linear Equivalent of Transistor's Nonlinear Output Capacitance, Normalized Design, and Maximum Operating Frequency versus Output Capacitance

WE2C-5
Stabilizing Circuit for Class-E Amplifier

WE2C-6
A Simple Method to Cut Down Configuration of Feedforward Power Amplifier

WE2D
ACTIVE QUASI-OPTICS AND SPATIAL POWER COMBINING
Chair: D. Rutledge, Co-chair: M. DeLisio

WE2D-1
A 25-W and 50-W Ka-band Quasi-optical Amplifier

WE2D-2
Electromagnetic Modeling and Experimental Verification of a Complete Waveguide-based Aperture-coupled Patch-Amplifier Array

WE2D-3
A 5-W, 37-GHz Monolithic Grid Amplifier

WE2D-4
A 16-element Reflection Grid Amplifier

WE2D-5
A Three-dimensional Retrodirective Self-oscillating Mixer Array

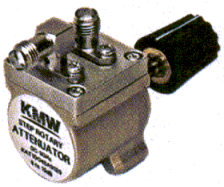
WE2D-6
A 1.6-W Power-Amplifier Module at 24 GHz Using New Waveguide-based Power Combining Structures

WE2D-7
K-band Spatial Combiner Using Active Array Modules in an Oversized Rectangular Waveguide

WE2E
PHASED ARRAYS
Chair: K. Tomiyasu, Co-chair: D. Parker

a Wide Range of
Quality Attenuators
Step-Rotary / Continuously Variable / Fixed Coaxial

Designed by KMW for consistency and stability, these Attenuators will be your first choice for building wireless communication system applications



KAT2DO4SA000
KAT1DO4SA002

Step-Rotary Attenuators

Product Code No.	KAT2DO4SA000	KAT1DO4SA002
Operating Type	Make-Before-Break	
Frequency Range	DC ~ 3GHz	DC ~ 3GHz
Insertion Loss (Max.)	0.2dB	0.2dB
VSWR (Max.)	1.15:1	1.15:1
Incremental Attenuation Range (dB)	0 ~ 1	0 ~ 10
Attenuation Step (dB)	0.2	1
Nominal Impedance	50 ohm	
I/O Port Connector	SMA(F) / Right Angle SMA(F)	
Average Power Handling	1W @ 2GHz	
Temperature Range	-30°C ~ +80°C	
Dimension (inch)	1.925*1.567*2.224	

KAT13O4CA000

KAT13O4CA001

Continuously Variable Attenuators

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 (Min.)	13dB @ 2GHz		
Nominal Impedance	50 ohm		
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*0.457 B type : 1.224*1.102*0.457		

Fixed Coaxial Attenuators are available

N-type, SMA-type Connectors

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CIRCLE NO. 372

KMW
RF & Microwave Products

Celebrating 25 Years

OMNIYIG, INC. 25th ANNIVERSARY 1973 1998

Since 1973 OMNIYIG has been at the forefront of advanced YIG technology, manufacturing state-of-the-art YIG components, Detectors, Limiters, Comb Generators, and integrated subsystems. We are continually designing and manufacturing new state-of-the-art YIG components for Airborne Systems/Space/Ship Board Commercial/EW/ECM/Jammers/Radar/ Communication.

OMNIYIG's products are available for commercial use, yet are built under stringent military specifications (MIL-E-5400) Class II/ MIL-STD-883 (Typ). Our advanced products are designed into some of the world's most sophisticated systems, including commercial, EW, and ECM programs such as the ALQ-99, ALQ-117, ALR-172, ALR-56C, ALR-62, ALR-64, ALR-67, ALR-69, APR-39, WLR-8, Rapport III and we are on board the F-15, F-16, F-18, EA6B, EF111, B-1, B2 Stealth, L-130 and more. We maintain and continue technology improvement.

YIG BAND PASS AND BAND REJECT FILTERS

- Frequency Range: 0.4 GHz to 40 GHz
- Octave and Multioctave Frequency Ranges
- Sizes: 1.2 x 1.2 x 1.2 inch
  1.0 x 1.4 x 1.4 inch
  1.4 x 1.4 x 1.4 inch
  1.7 x 1.7 x 1.7 inch

YIG MULTIPLIERS

- Frequency Range: 0.4 GHz to 40 GHz
- Octave and Multioctave Frequency Ranges
- Input Frequency: 100 MHz to 2000 MHz; fixed or variable
- Output Frequency: 100 MHz to 40 GHz
- Sizes: 1.4 x 1.4 x 1.4 inch
  1.7 x 1.7 x 1.7 inch

YIG OSCILLATORS

- Frequency Range: 0.4 GHz to 20 GHz
- Octave and Multioctave Frequency Ranges
- Sizes: 0.5 x 1.0 x 1.0", 1.0 x 1.0 x 1.0 ", 1.2 x 1.2 x 1.2", 1.0 x 1.4 x 1.4", 1.7 x 1.7 x 1.7", 1.7 x 1.7 x 1.7"

LIMITERS, COMB GENERATORS and DETECTORS

DRIVERS are available as integrated components with the YIG devices. Analog drivers with 0–10 volts and 12-bit digital drivers.

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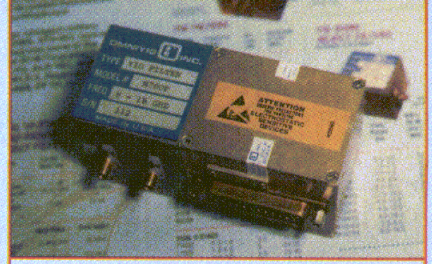
Web: www.OMNIYIG.com

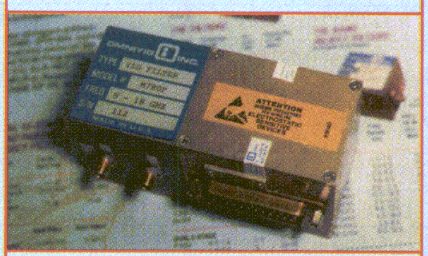
OMNIYIG INC.

WE HAVE OVER 1000 DESIGNS AVAILABLE
WE ALSO DO CUSTOM DESIGNS



YIG BAND REJECT FILTERS

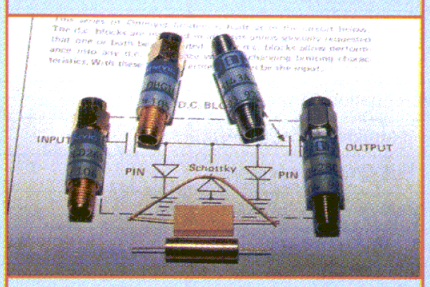
Omniyig Model No.	Freq. Range (GHz)	Ins. Loss (max.) (dB)	Bandwidth @ 40 dB (min.) (MHz)
6- , 8- and 12-STAGES			
P106RX	0.5 – 1.0	1.5	10
L106RX	1.0 – 2.0	1.5	10
C105RX	2.0 – 8.0	1.5	10
X106RX	8.0 – 12.4	1.5	20
Ku106RX	12.0 – 18.0	1.8	20
M102RX	4.0 – 12.4	1.5	8
M103RX	4.0 – 12.4	1.5	10
M104RX	4.0 – 18.0	1.5	10–60



YIG BAND PASS FILTERS

- Frequency Range: 0.4 GHz to 40 GHz
- 3-Stages, Dual 3-Stages
- 4-Stages, Dual 4-Stages
- 6-Stages
- 7-Stages

The filters can be integrated with analog drivers or 12-bit digital drivers built to commercial or military specifications.



LIMITERS, COMB GENERATORS and DETECTORS

- Frequency Range: 0.01 GHz to 40 GHz

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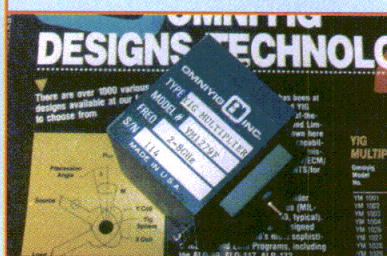
THIN FILM YIG OSCILLATORS

Omniyig Model No.	Freq. Range (GHz)	RF Pwr. Output (mW)	2 Ha (dB
YOM1517	0.5 – 2.0	20–60	1
YOM1518	1.0 – 4.0	20–60	1
YOM1514	4.0 – 12.0	10	1
YOM1513	4.0 – 10.0	10	1
YOM83	2.0 – 6.0	20	1
YOM1948	3.5 – 10.5	15	1
YOM1317	2.0 – 8.0	20	1
YOM818	8.0 – 18.0	20	1
YOM1516	6.0 – 18.0	20	1
YOM2320	2.0 – 10.0	13	1
YOM2321	5.0 – 18.0	13	

We offer other models with 2nd Harmon –60 dBc. Oscillators integrated with 2-sta filters are available.



YIG MULTIPLIERS

STANDARD UNITS
Output Computer Tested

Omniyig Model No.	Input Freq.	Output (GHz)	Outp Pw (dB
YM 1001	1–2 GHz	2 – 13	6
YM 1002	100 MHz	1 – 12	–3
YM 1003	200 MHz	1 – 12	–2
YM 1004	500 MHz	1 – 12	–1
YM 1026	1–2 GHz	2 – 18	2
YM 1027	100 MHz	1 – 18	–4
YM 1028	200 MHz	1 – 18	–3
YM 1029	500 MHz	1 – 18	–2
YM 1087	.1–.2 GHz	1 – 12	–2

* RF input power on all models 0.5 to 1 watts.

Fast switching, MIL-Spec Hi-Reliability a units integrated with drivers, oscillators amplifiers also available.

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YIG FILTERS

Omniyig Model No.	Freq. Range (GHz)	Ins. Loss (dB)	Bandwidth @ 3 dB (MHz)
6-STAGE			
P106	0.5 – 1.0	6.5	12–30
L106	1.0 – 2.0	5.5	20–35
S106	2.0 – 4.0	5.0	20–40
C106	4.0 – 8.0	4.5	25–40
X106	8.0 – 12.4	4.5	25–40
Ku106	12.4 – 18.0	4.5	28–45
3-STAGE			
P103	0.5 – 1.0	5.0	14–25
L103	1.0 – 2.0	3.5	20–35
S103	2.0 – 4.0	3.0	20–35
C103	4.0 – 8.0	3.0	25–40
X103	8.0 – 12.4	3.0	25–40
Ku103	12.4 – 18.0	3.5	30–45
4-STAGE			
P104	0.5 – 1.0	6.0	12–23
L104	1.0 – 2.0	4.5	20–35
S104	2.0 – 4.0	4.0	20–35
C104	4.0 – 8.0	4.0	25–40
X104	8.0 – 12.4	4.0	25–40
Ku104	12.4 – 18.0	4.0	28–45
DUAL 2-STAGE			
P1022	0.5 – 1.0	3.5	17–30
L1022	1.0 – 2.0	3.0	24–35
S1022	2.0 – 4.0	2.5	25–40
C1022	4.0 – 8.0	2.5	25–40
X1022	8.0 – 12.4	2.5	25–40
Ku1022	12.4 – 18.0	2.5	30–45
MULTIOCTAVE BANDS			
M1611/2	1.0 – 18.0	5.5	25–65
M1612/4	2.0 – 18.0	6.5	25–75
M102/2[5]	1.0 – 12.4	5.0	25–60
M1613/2	1.0 – 12.4	5.5	25–60
M1048/4	4.0 – 18.0	6.0	25–60
M203/4[5]	1.0 – 18.0	6.5	25–70

Other Multioctave YIG Filters are available. Analog and 12-bit digital drivers are available for all YIG devices.

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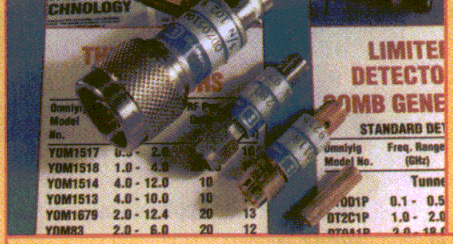

DETECTORS, LIMITERS and COMB GENERATORS

Since 1973 OMNIYIG has been designing and improving the performance of detectors, limiters, and comb generators. We have many new designs in many sizes. We also integrate input power amplifiers with the comb generators. Some of our standard designs are shown below. Send for our complete catalog.

ZERO BIAS SCHOTTKY DETECTORS

Omniyig Model No.	Freq. Range (GHz)	k Factor (mv/mw)	TSS (dBm)
Zero Bias			
ODZ0004A	0.1 – 18	1200	–51
ODZ0510A	0.5 – 4	1750	–53
ODZ0518A	1.0 – 12	1250	–52
ODZ0527A	2.0 – 12	1250	–52
ODZ0328A	2.0 – 18	1250	–52
Tunnel Planar			
ODT0004A	0.1 – 18	750	–50
ODT0510A	0.5 – 4	800	–50
ODT0527A	2.0 – 12	800	–50
ODT0328A	2.0 – 18	700	–50
ODT0240A	6.0 – 18	700	–50

COMB GENERATORS

Omniyig Model No.	Input Freq. (MHz)	Output Freq. Range (GHz)	Output Pwr. (dBm)
OHG10140	100	0.1 – 4	–10
OHG10118	100	0.1 – 18	–40
OHG20218	200	0.2 – 18	–34
OHG30318	250	0.25–18	–29
OHG51027	500	0.5 – 18	–20
OHG61027	1000	0.1 – 18	–33
OHG72027	2000	2.0 – 18	–10
OHG61026	1000	1.0 – 26	–35
OHG71026	2000	2.0 – 26	–20

TRACKING YIG FILTERS/ YIG OSCILLATORS

Omniyig Model No.	Freq. Range (GHz)	Filter IL (dB)	Tracking (MHz)
M129YTO	0.5–2	5.5	5
M120YTO	2 – 8	5.0	7
M121YTO	8–18	5.0	8

Oscillator RF power output +13 dBm. Integrated with Analog or 12-bit TTL drivers.

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LIMITERS

Omniyig Model No.	Freq. Range (GHz)	Max. Ins. Loss (dB)	Max. Lkg. Pwr. (dBm)*
PIN Diode			
OLP2801A	0.1 – 0.5	0.5	+20
OLP2817A	1.0 – 4.0	0.5	+19
OLP2726A	2.0 – 8.0	1.2	+19
OLP2640A	6.0–18.0	2.0	+18
OLP2650A	2.0–18.0	2.5	+18
Schottky			
OLD2802A	0.1 – 1.0	0.5	+15
OLD2709A	0.5 – 2.0	0.5	+15
OLD2762A	2.0 – 8.0	1.0	+14
OLD2635A	4.0–18.0	2.5	+14
OLD2650A	2.0–18.0	2.5	+13

*Measured at 1 Watt CW Power

COMB GENERATORS with INTEGRATED INPUT RF AMPLIFIER

Omniyig Model No.	Input Freq. (MHz)	Output Freq. Range (GHz)	Output Pwr. (dBm)
CG252A	100–200	0.1 – 12	–24
CG253A	100	1.0 – 18	–38
CG256A	200	1.0 – 18	–35
CG259A	250	1.0 – 18	–30
CG262A	500	1.0 – 18	–18
CG265A	1000	1.0 – 18	–15
CG266A	1000	2.0 – 26	–33
CG268A	2000	2.0 – 18	–20
CG269A	2000	2.0 – 26	–35

Call Omniyig with your requirements or for a new updated catalog. We offer many other models covering a variety of frequencies.

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WE2E-1
Oscillator Reproducibility Consideration in Coupled Oscillator Phase-steering Arrays

WE2E-2
A Phased-array Antenna Using a Multi-line Phase Shifter Controlled by a Piezoelectric Transducer

WE2E-3
A Beam-steerer Using Reconfigurable PBG Ground Plane

WE2E-4
Demonstration of a Reconfigurable Beam-former for Simplified 2-D, Time-steered Arrays

WE2E-5
Optically Smart Active Antenna Arrays

WE2E-6
A Transmit/Receive Active Antenna with Fast Low-power Optical Switching

NOON-1:15 PM

PWC
RF MEMS FOR TUNABLE APPLICATIONS

MEMS components offer the capabilities of very loss operation and, therefore, very high-Q operation in applications such as tunable filters, tunable oscillators, and tunable matching networks. The panel will present recent advances in high-Q MEMS varactors, tunable filters, and oscillators from 200 MHz to 10 GHz.

PWD
DEVICES FOR HIGH-SPEED FIBER OPTICS—TECHNOLOGY NEEDS AND WHAT'S NEXT?

Fiber optics has often been described as a "killer app" technology. This panel session will examine device design, material systems for discrete devices, material systems for OEICs and other needs for next-generation and future high-speed fiber-optic applications.

1:20 PM-3:00 PM

WE3A
POWER AMPLIFIERS AND HANDSETS
Chair: A. Platzker, Co-chair: D. Teeter

WE3A-1
A Monolithic +2.5-VDC, 1-W Silicon Bipolar Power Amplifier with 55-percent PAE at 1.9 GHz

WE3A-2
A SiGe HBT Power Amplifier with 40-percent PAE for PCS CDMA Application

WE3A-3
An Integrated 900-MHz Push-pull Power Amplifier for Mobile Applications

WE3A-4
An L-band High Efficiency and Low Distortion Power-Amplifier Module Using an HPF/LPF Combined Interstage Matching Circuit

WE3A-5
42-percent High Efficiency Two-stage HBT Power-Amplifier MMIC for WCDMA Cellular Phone System

WE3A-6
High-Efficiency CDMA RF Power Amplifier Using Dynamic Envelope Tracking Technique

WE3B
CAD, MODELING, AND OPTIMIZATION TECHNOLOGY
Chair: K.C. Gupta, Co-chair: A. Pavio

WE3B-1
Neural Frequency-Space Mapping EM Optimization of Microwave Structures

WE3B-2
A New Macromodeling Approach for Nonlinear Microwave Circuits Based on Recurrent Neural Networks

WE3B-3
A Novel Nonlinear Statistical Modeling Technique for Microwave Devices

WE3B-4
Optimization of Microwave Circuits Using Analytically Calculated Gradients in the Finite Element Method

WE3B-5
Efficient Sensitivity Analysis of Transmission-line Networks Using Model Reduction Techniques

WE3C
FERRITE AND SAW DEVICES
Chair: J. Owens, Co-chair: R. Weige

WE3C-1
Spike Leakage and Suppression in Frequency Selective Limiters

WE3C-2
Analysis of Six-port Disk Circulator Using Subdivided Ports

WE3C-3
MMW Monolithic Y-junction Circulator on Single-crystal SC-doped Ba-hexaferrite

WE3C-4
High Q VCO (Voltage Controlled Oscillator) Using Electro-acoustic Device Compatible with CMOS Integrated Circuit Technology

WE3C-5
A Miniature High-Q Grating-mode-type SAW Resonator and a Wide-band 1-GHz SAW-VCO for Mobile Communications

WE3D
BIOLOGICAL EFFECTS AND MEDICAL APPLICATIONS
SPECIAL SESSION
Chair: A. Rosen , Co-chair: Y. Kotsuka

WE3D-1
(Invited) Microwave Treatments for Prostate Disease

WE3D-2
(Invited) Advances in Catheter Ablation for Treatment of Cardiac Arrhythmias

WE3D-3
(Invited) Radiometric Sensing: An Adjuvant to Mammography to Determine Breast Biopsy

WE3D-4
(Invited) Thermoacoustic CT

WE3D-5
(Invited) Cellular Telephones: Hazards or Not?

3:30-5:10 PM

WE4A
POWER AMPLIFIER INTEGRATED CIRCUITS
Chair: J. Schellenberg, Co-chair: H.J. Kuno

WE4A-1
A Compact S-band MMIC High Power Amplifier Module

WE4A-2
A 7.4-to-8.4-GHz High Efficiency PHEMT Three-stage Power Amplifier

WE4A-3
Full W-band MMIC Medium Power Amplifier

WE4A-4
Compact W-band Solid-state MMIC High-Power Sources

WE4A-5
A 3-to-10-GHz LCR-matched Power Amplifier Using Flip-chip Mounted AlGaN/GaN HEMTs

WE4A-6
14-W GaN-based Microwave Power Amplifiers

WE4B
CAD MODELING OF PASSIVE COMPONENTS
Chair: K. Nashadham, Co-chair: A. Weisshaar

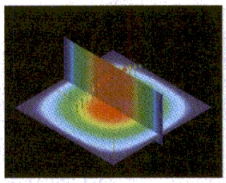
WE4B-1
Broadband Physics-based Modeling of Microwave Passive Devices through Frequency Mapping

WE4B-2
Reduced Order Modeling of Coupled On-chip Interconnects for Silicon-based RF Integrated Circuits

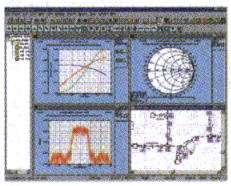
WE4B-3
Development of CAD Formulas of Integrated Circuit Components and Simplified Formula-

287 specs. 23 designers. 4 departments. 1 goal.

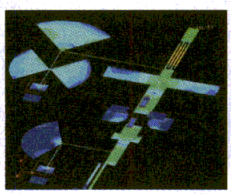
...is everybody on the same page?



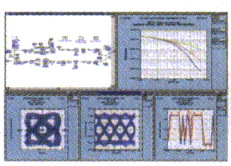
Ansoft HFSS
3D electromagnetic
structure simulation



Harmonica
linear & nonlinear
circuit simulation



Ensemble
planar electromagnetic
field simulator



Symphony
wireless & wired
system simulation

Interoperate

Ansoft's Serenade Design Environment

...delivering system, circuit, electromagnetic simulation, synthesis, and physical design in a single software suite. Streamline your product development process using integrated schematic capture, simulation, layout, IC package modeling, and links to third-party tools from companies such as Cadence®, Mentor Graphics and EEsof.

Handshaking between modules allows engineers to focus on critical components at any stage of development. Each simulator can be used as a powerful, stand-alone tool, or in concert with others for end-to-end high-frequency design. The Serenade Design Environment is built on Ansoft's core electromagnetics technology; and offers accuracy, an easy-to-use Windows® interface, advanced design utilities, and superior value.

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- Connector Choice: SMA, N, 2.92mm

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- Connector Choice: SMA, 2.92m, N
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- Low Intermodulation Design Options
- Connector Choice: SMA, 2.4mm, 2.92m, 3.5mm, N, 7/16, BNC

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- Blind-mate Connector Systems
- PLANAR CROWN® Connector System

Directional Couplers
- dc-4 GHz • Octave & Broadband

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- Plug & Go Programmable Attenuators and Relay Drivers
- IEEE-488 & Standard Serial Interfaces/Controllers
- Bus Controlled Programmable Attenuator Units
- Custom Subsystem Design for Specialized Applications

Terminations & Loads
- Models from dc-40 GHz
- High Power Models up to 1,000 W
- Low Intermodulation Design Options
- Connector Choice: SMA, 2.4mm, 2.92m, 3.5mm, N, 7/16

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CIRCLE NO. 228

tion Followed by Rigorous Derivation

WE4B-4
Unified CAD Model of Microstrip Line with Backside Aperture for Multilayer Integrated Circuit

WE4C
ADVANCES IN MILLIMETER WAVES AND SUBMILLIMETER WAVES
Chair: K. Agarwal, Co-chair: S. Weinreb

WE4C-1
A 183-GHz Low Noise Amplifier Module for the Conical-scanning Imager Sounder (CMIS) Program

WE4C-2
A Coplanar 148-GHz Cascode Amplifier MMIC Using 0.15-mm GaAs PHEMTs

WE4C-3
Low-cost 38-and 77-GHz CPW MMICs Using Ion-implanted GaAs MESFETs

WE4C-4
A 33-GHz Power Amplifier Based on an Extended Resonance Technique

WE4C-5
High Efficiency MMIC Frequency Triplers for Millimeter and Submillimeter Wavelengths

WE4C-6
NbN Hot Electron Bolometric Mixer with Intrinsic Receiver Noise Temperature of Less than Five Times the Quantum Noise Limit

WE4C-7
Experimental Investigation of WDM Channel Spacing in Simultaneous Upconversion Millimeter-wave Fiber Transmission System at 60-GHz Band

WE4D
BIOLOGICAL EFFECTS AND MEDICAL APPLICATIONS
Chair: R. Bansal, Co-chair: K. Carr

WE4D-1
Measurements of Thermal Effects Induced on a Human Head Exposed to 900-MHz Mobile Phones

WE4D-2
Evaluation of the Power Absorbed in a Human Head Model Exposed to Cellular Phones Equipped with Helical Antennas

WE4D-3
A Study of Uncertainties in Modeling the Handset Antenna and Human Head Interaction Using the FDTD Method

WE4D-4
Effect of the Human Tissue on EMI from a Cellular Telephone with a Hearing Aid

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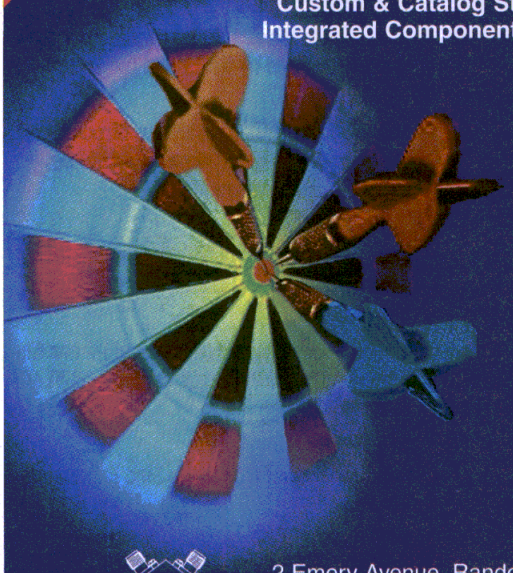

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WE4D-5
Algorithm for Retrieval of Deep Brain Temperature in New-born Infant from Microwave Radiometric Data

WE4D-6
Microwave Breast Cancer Detection

WE4D-7
Characterization of Biological Tissues up to Millimeter Wave: Test Fixtures Design

WE4E
PACKAGING AND INTERCONNECTS
Chair: M. Harris, Co-chair: C.P. Wen

WE4E-1
Interconnects and Transitions in Multilayer LTCC Multichip Modules for 24-GHz ISM-band Applications

WE4E-2
Application of Digital PGA Technology to 20-GHz Microwave Packages

WE4E-3
Multi-layer Spiral Inductors in a High-precision Fully-planar MCM-D Process

WEIF
INTERACTIVE FORUM
Chair: Dan Swanson, Co-chair: Doug Teeter

WEIF-1
Properties of CPW in the Sub-mm Wave Range and Its Potential to Radiate

WEIF-2
New Closed-form Electromagnetic Green's Functions in Layered Media

WEIF-3
Equivalent Network for Rectangular-waveguide H-plane Step Discontinuity—Multi-transmission Line and Multi-port Ideal Transformer

WEIF-4
Full-wave Analysis of a Periodic Array of Rectangular Lossy Conductive Thick Patches

WEIF-5
Leaky Modes of Vertically Stacked Microstrips at Higher Order

WEIF-6
New Efficient Design of Microwave Inhomogeneous Media Filters

WEIF-7
Reconstruction of Permittivity Profiles in Cylindrical Objects Illuminated by Higher Order Modes

WEIF-8
Full-wave Characterization of an Edge-coupled Coplanar-waveguide Structure with Backed Conductor

WEIF-9
Reconstruction of the Non-minimum Phase Function from Amplitude Only Data

WEIF-10
BI-RME Modeling of 3D Waveguide Components Enhanced by the Ewald Technique

WEIF-11
Multiport Analysis of Arbitrary Circular-rod Insets in Rectangular Waveguide by the Generalized Admittance Matrix

WEIF-12
Mix-RWG Current Basis Function and Its Simple Implementation in Method of Moments

WEIF-13
Time Domain Characterization of Planar Microwave Transformers Using the SCN-TLM Method

WEIF-14
A Higher-order FDTD Using Sinc Expansion Function

WEIF-15
Development of a Three-dimensional Unconditionally Stable Finite-difference Time-domain Method

WEIF-16
Visualization of Near-field and Far-field in the Time Domain

WEIF-17
Optimal Shape Design of Passive Device Using FDTD and Design Sensitivity Analysis

WEIF-18
The Application of the Wave Potential Functions to the Analysis of Transient Electromagnetic Fields

WEIF-19
Determination of Optimum Conductivity Profile for PML and PML-D Using Multiple Variables Pade Approximation

WEIF-20
General Design Equations of Three-port Unequal Power Divider Terminated by Arbitrary Impedances

WEIF-21
40-to-90-GHz Impedance-transforming CPW Marchand Balun

WEIF-22
A Very-small-sized Reversed-phase Hybrid Ring

WEIF-23
A Whispering Gallery Mode Sapphire Resonator with Semi Spherical Enclosure

WEIF-24
Low Loss Passive Microwave Components in Standard Silicon Technology

WEIF-25
Accurate Analysis of Silicon, VLSI-technology Compatible Spiral Inductors

WEIF-26
A New Network Model for Miniaturized Hairpin Resonators and Its Applications

WEIF-27
Improvements in Losses and Size of Frequency Tunable Coplanar Filter Structures Using MMIC Negative Resistance Chips for Multistandard Mobile Communication Systems

WEIF-28
Low Temperature Cofired Ceramic (LTCC) Ridge Waveguide Multiplexers

WEIF-29
Application of the Planar I/O Terminal to Dual-Mode Dielectric Waveguide Filters

WEIF-30
Synthesis of Cascaded Quadruplet Filters Involving Complex Transmission Zeros

WEIF-31
The Design of a Half-wavelength Resonator BPF with Attenuation Poles at Desired Frequencies

WEIF-32
Microwave Breakdown in Output Multiplexers Filters

WEIF-33
A Design Procedure for Co-existing Multi-mode Waveguide Bandstop Filters

WEIF-34
Anisotropic Permittivity Extraction from Phase Propagation Measurements Using an Anisotropic Full-wave Green's Function Solver for Coplanar Ferroelectric Thin-Film Devices

WEIF-36
A Novel HTS Microstrip Quasi-Elliptic Function Bandpass Filter Using Pseudo-lumped Element Resonator

WEIF-37
A Novel Triple-Push Oscillator Approach

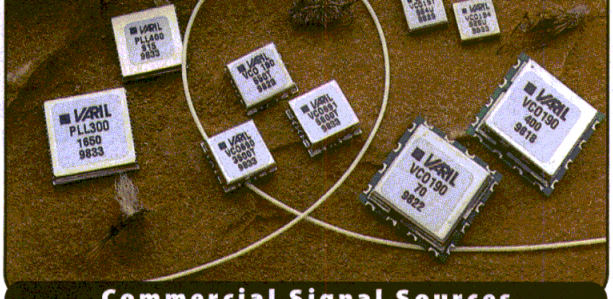
WEIF-38
77-GHz Planar Gunn VCOs On AlN Substrates Using Novel Flip-Chip InP Gunn Diodes

WEIF-39
Impact of 1/f noise in Ka-band InGaP/GaAs HBT Frequency Sources

WEIF-40
A Very High-Q-feedback Oscillator at 33 GHz Using MMIC Amplifiers

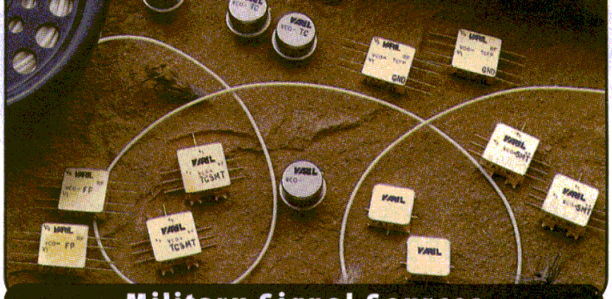
WEIF-41
MEMS Devices for High Isolation Switching and Tunable Filtering

WEIF-42
Modeling of MEMS Capacitive Switches by TLM



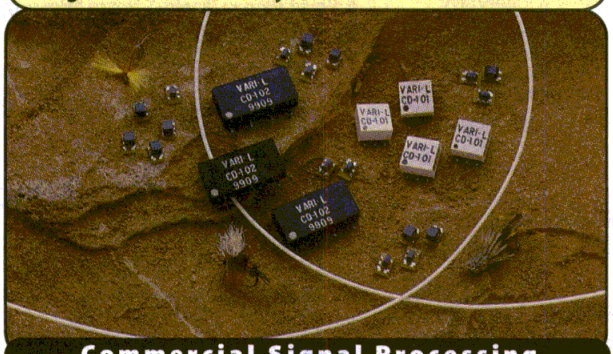
Commercial Signal Sources

- High Performance Voltage Controlled Oscillator Modules
- High Performance PLL Synthesizer Modules



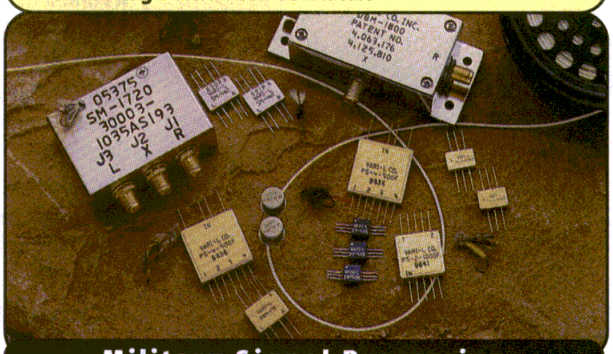
Military Signal Sources

- Ruggedized High Performance Hybrid
- Voltage Controlled Oscillators



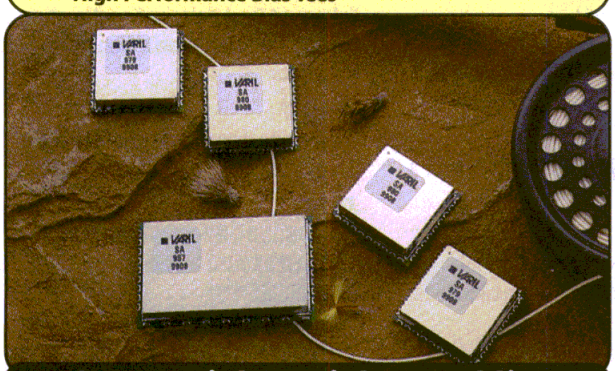
Commercial Signal Processing

- High Performance Wideband RF Transformers
- High Performance Power Dividers and Couplers
- High Performance Double Balanced Mixers
- High Performance RF Chokes
- High Performance Bias Tees



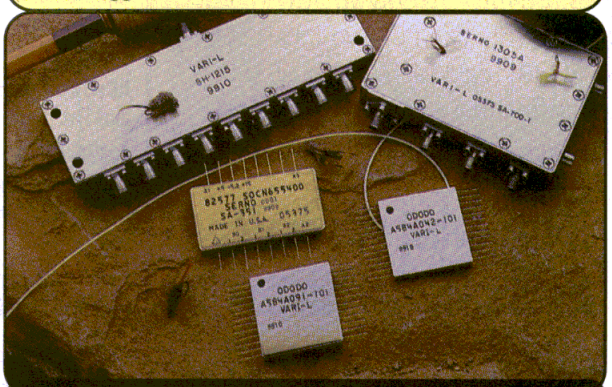
Military Signal Processing

- Ruggedized Double Balanced Mixers
- Ruggedized Wideband RF Transformers
- Ruggedized Power Dividers and Couplers
- Ruggedized I/Q Modulators and Demodulators



Commercial Special Assemblies

- Special Frequency Conversion Modules
- Special Frequency Generation Modules



Military Special Assemblies

- Ruggedized Special Frequency Generation Assemblies
- Ruggedized Special Frequency Conversion Assemblies
- Ruggedized Special RF Distribution Assemblies

Subscriber Signal Sources

- Miniature Voltage Controlled Oscillator Modules
- Miniature PLL Synthesizer Modules

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WEIF-43
Membrane-supported Ka Band Resonator Employing Organic Micromachined Packaging

WEIF-44
A Parametric Model of MEMS Capacitive Switch Operating at Microwave Frequencies

WEIF-45
Noise Performance of Submicron HEMT Channels under Low Power Consumption Operation

WEIF-46
On the Performance of Low Noise, Low DC Power Consumption Cryogenic Amplifiers

WEIF-47
Temperature-dependent Noise Parameters and Modeling of Inp/InAlAs/InGaAs HEMTS

WEIF-48
Interlayer MEMS RF Switch for 3D MMICs

WEIF-49
Simulation of Noise-power Ratio with the Large-signal Code CHRISTINE

WEIF-50
ACeS Communications System Phase Noise and Transient Effects

WEIF-51
Optical Distribution of Local Oscillators for a Rotating Antenna

WEIF-52
Data Transmission Using an Optical Microwave Link for Wireless Communications

THURSDAY, June 15, 2000

8:00 AM-9:40 AM

TH1A
WIRELESS AND CELLULAR COMMUNICATIONS
Chair: R.G. Ranson, Co-chair: J.K. McKinney

TH1A-1
An Integrated RF Front End for Multi-mode Handsets

TH1A-2
5.8-GHz OFDM GaAs MESFET MMIC Chip Set

TH1A-3
A Calibratable 4.8-to-5.8-GHz MMIC Vector Modulator with Low Power Consumption for Smart Antenna Receivers

TH1A-4
A New Antenna Switching Architecture for Mobile Handsets

TH1A-5
Quadrature Direct Conversion Receiver Integrated with Planar Quasi-Yagi Antenna

TH1A-6
A Silicon MMIC Active Balun/Buffer Amplifier with High Linearity and Low Residual Phase Noise

TH1B
COUPLERS
Chair: E.J. Denlinger, Co-chair: J. Taub

TH1B-1
A Novel Planar Magic-T Using Microstrip-to-slotline Transition

TH1B-2
A Microwave Octave-band Lumped Element Quadrature Coupler

TH1B-3
Novel HDI Microwave Chip on Flex (MCOF) Broadside Coupler Designs

TH1B-4
Wideband 0-dB Branch-line Directional Couplers

TH1B-5
A New Waveguide Directional Coupler/Hybrid Type Favorably Suited for Millimeter Wave Applications

TH1B-6
Enhanced Dual Polarization Directional Coupler for Dual Polarization Beam Forming Networks

TH1B-7
Waveguide Branch Couplers for Tight Couplings

TH1C
BROADBAND AND HIGH-EFFICIENCY AMPLIFIERS AND NOVEL HYBRID MODULE TECHNIQUES
Chair: G. Brehm, Co-chair: F. Sullivan

TH1C-1
The Gain Advantages of Four Cascaded Single Stage Distributed Amplifier Configurations

TH1C-2
Multi-octave 200-MHz to 20-GHz Amplifier for Extended Range Digital Log Video Amplifier Application

TH1C-3
A Compact PA MMIC Module for K-band High speed Wireless Systems

TH1C-4
Suppression of Resonant Leakage Effects in Coplanar MMIC Packages using an Si Submount Layer

TH1C-5
Vertical Integrated Transformer Using Bondwires for MMICs

TH1C-6
Low Voltage Tunable Barium Strontium Titanate Thin Film Capacitors for RF and Microwave Applications

TH1D
FERROELECTRIC DEVICES

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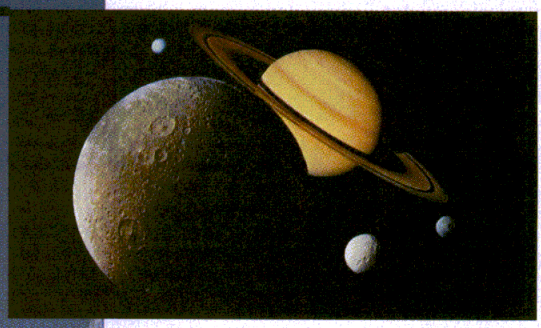

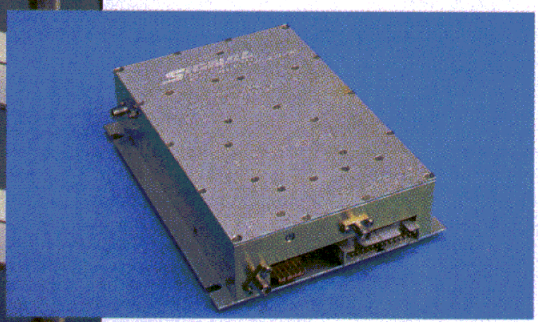

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CIRCLE NO. 392

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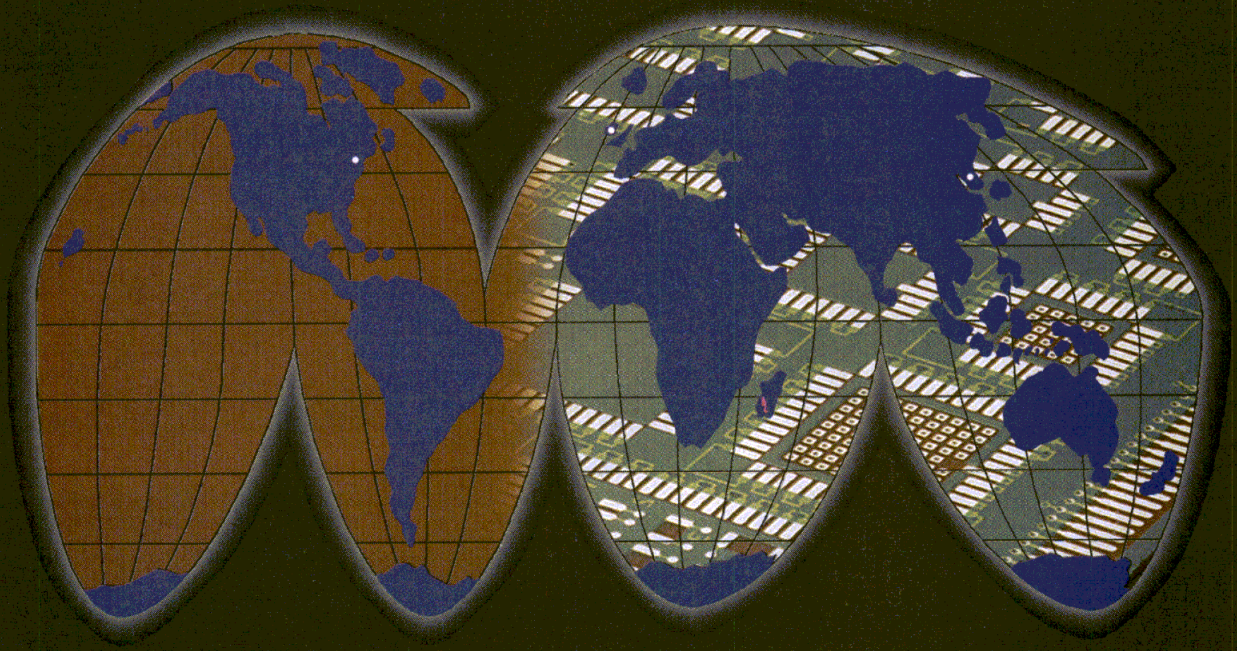

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CIRCLE NO. 281

FOCUSED SESSION
Chair: S. Stitzer, Co-chair: L. Davis

TH1D-1
A K-band Linear Phased Array Based on BaSrTiO Thin Film Phase Shifters

TH1D-2
Application of Ferroelectrics in Phase Shifter Design

TH1D-3
Ferroelectric/Ferrite Tunable Phase Shifters

TH1D-4
A Ferroelectric Tunable Microstrip Lange Coupler for K-band Applications

TH1D-5
Full Wave Analysis of Coplanar Phase Shifter Printed on High Purity Ferroelectric Material

TH1D-6
Limiting Characteristics of a Tunable Resonator with a Ferroelectric Capacitor.

TH1E
LINEAR MODELS OF ACTIVE DEVICES
Chair: K. Chang, Co-chair: D.E. Root

TH1E-1
A 110-GHz Scalable FET Model Based on 50-GHz S-parameter Measurements

TH1E-2
Bias-dependent Linear, Scalable Millimeter-wave FET Model

TH1E-3
Empirical Loadline Capacitance Models for HEMT

TH1E-4
Improved Parameter Extraction of Small-sized FETs for Low-power Applications

TH1E-5
A New Simplified and Reliable HEMT Modeling Approach Using Pinched Cold FET S-parameters

TH1E-6
A New Analytical and Broadband Method for Determining the HBT Small-signal Model Parameters

TH1E-7
Direct Extraction Method for Internal Equivalent Circuit Parameters of HBT Small-signal Hybrid-Pi Model

10:10AM -11:50 AM

TH2A
AUTOMOTIVE SENSORS AND ID SYSTEMS
Chair: S. Kanamluru, Co-chair: H. Kondoh

TH2A-1
A Wireless Pressure Measurement System Using a SAW Hybrid Sensor

TH2A-2
A Realtime Close-range Imaging System with Fixed Antennas

TH2A-3
A Pre-crash Radar Sensor System Based on Pseudo-noise Coding

TH2A-4
Reduction of Interference in Microwave Automotive Radars

TH2A-5
A 76.5-GHz Vector Modulator and Its Application to a Digitally Modulated Automotive Collision Avoidance Radar

TH2A-6
Microwave Backscatter Modulation Systems

TH2A-7
A Novel Active Retrodirective Array for Remote Tagging and Wireless Sensor Applications

TH2B
PASSIVE COMPONENTS
Chair: A. Fathy, Co-chair: K. Zaki

TH2B-1
High Q Two Dimensional Defect Resonators—Measured and Simulated

TH2B-2
Tunable Dielectric Resonators

TH2B-3
An Electronically Tunable Photonic Bandgap Resonator Controlled by Piezoelectric Transducer

TH2B-4
A Design of Novel Grooved Circular Waveguide Polarizers

TH2B-5
Novel Technique of Phase Velocities Equalization for Microstrip Coupled Line Phase Shifters

TH2C
LINEAR AND NONLINEAR NETWORK MEASUREMENTS
Chair: D. Williams, Co-chair: Z. Skvor

TH2C-1
High Power Time Domain Measurement System with Active Harmonic Load-pull for High Efficiency Base-Station Amplifier Design

TH2C-2
An Active Pulsed RF and Pulsed DC Load-pull System for the Characterization of Power Transistors Used in Coherent Radar and Communication Systems

TH2C-3
Measurement Based Nonlinear Modeling of Spectral Regrowth

TH2C-4
A Low Cost AM/PM Sideband Generator for Low Noise Calibration

TH2C-5
Realistic Sampling Circuit Model for a Nose-to-nose Simulation

TH2C-6
A Comprehensive Millimeter-wave Calibration Development and Verification Approach

TH2C-7
A New Method for Characteristic Impedance Determination on Lossy Substrate

TH2D
POWER AMPLIFIER LINEARIZATION
FOCUSED SESSION
Chair: M. Eron, Co-chair: A. Katz

TH2D-1
Digital Controlled Adaptive Feedforward Amplifier for IMT-2000 Band

TH2D-2
Optimization of Feedforward Amplifier Power Efficiency on the Basis of Input Power Statistics

TH2D-3
A 900-MHz 16-QAM Direct Carrier Modulation Transmitter Using Feedforward Linearization

TH2D-4
Input Adaptive Linearizer System

TH2D-5
A Linearized Power Amplifier MMIC for +3.5-VDC Battery Operated Wide-band CDMA Handsets

TH2D-6
An 18-GHz-band MMIC Linearizer Using a Parallel Diode with a Bias Feed Resistance and a Parallel Capacitor

TH2D-7
Analog Predistortion Linearizer for High Power RF Amplifiers

TH2D-8
A Multicarrier Amplifier with Low Third-order Intermodulation Distortion

TH2E
PERIODIC STRUCTURES AND NEW TECHNIQUES
Chair: A.A. Oliner, Co-chair: T.K. Sarkar

TH2E-1
Surface-wave Band-gap and Leaky-waves on Integrated Circuit Structures with Planar Periodic Metallic Elements of Resonance-type Periodic Structures

TH2E-2
Analysis and Application of Coupled Microstrips on Periodically Patterned Ground Plane

TH2E-3
Numerical and Experimental Characterization of Slow-wave Microstrip Line on Periodic

Ground Plane

TH2E-4
Dispersion Characteristics of Microstrip with Periodic Perturbations

TH2E-5
Revisiting Guided-wave and Leakage-wave Modal Properties in Near Cutoff Region with the Concept of Complex Effective Dielectric Constant

TH2E-6
Formulas of Microstrip with Truncated Substrate by Synthetic Asymptotes—A Novel Analysis Technique

NOON-1:15 PM

PTHE
LOW-COST LMDS TERMINALS—TECHNOLOGY VERSUS MARKET VOLUME AND TIMING

LMDS systems for data communications to the home or small business are being rapidly deployed at various frequencies from 23 to 42 GHz. This panel will address the various technologies being used for the RF sections of the terminals and provide views of the cost and availability of these technologies.

1:20 PM-3:00 PM

TH3A
APPLIED SENSOR TECHNOLOGY
SPECIAL SESSION
Chair: P. Hide, Co-chair: T. Weller

TH3A-1
A 76-to-77-GHz Pulsed-Doppler Radar Module for Autonomous Cruise Control Applications

TH3A-2
Theoretical and Experimental Characterization of a Broadband Random Noise Radar

TH3A-3
Low Power FMCW Radar System for Level Gaging

TH3A-4
A Pulse Radar Gauge for Level Measurement and Process Control

TH3A-5
Multiparameter Microwave Sensors for Determining Composition or Condition of Substances

TH3A-6
Distributed, Micropower Wireless Integrated Network Sensors (WINS)

TH3B
TRENDS IN FREQUENCY MULTIPLIERS AND SWITCHES
Chair: D. Peterson, Co-chair: B.E. Sigmon

TH3B-1
Design and Stability Test of a 2-to-40-GHz Frequency Doubler with Active Balun

TH3B-2
Wideband Monolithic Millimeter-wave Phase Shifter with Minimum Insertion Loss Variation

TH3B-3
Millimeter Wave Broadband Frequency Tripler in GaAs/InGaP HBT Technology

TH3B-4
A Ka-band MMIC Super Regenerative Detector

TH3B-5
High Output Power, Broadband 28-to-56-GHz MMIC Frequency Doubler

TH3B-6
L-band Glass-epoxy 5-b Phase Shifter Module Using SW-BANK Chips and Low Cost L, C Chips

TH3B-7
A Novel Method for Nulling the LO Leakage in Resistive FET Mixers

TH3B-8
High-power Millimeter-wave Planar Doublers

TH3C
OPTICAL PROBING AND MILLIMETER-WAVE MEASUREMENTS
JOINT IMS-ARFTG SESSION
Chair: R.D. Pollard, Co-chair: C.M. Weil

TH3C-1
A New, Simple Test Set for On-wafer Characterization of Millimeter-wave Electro-optic Devices

TH3C-2
A Bilateral Lightwave Network Analyzer

TH3C-3
Novel Electromagnetic Field Probe Using Electro/Magneto-optical Crystal Mounted on an Optical-fiber Facet for Microwave Circuit Diagnosis

TH3C-4
Non-contact Internal-MMIC Measurement Using Scanning Force Probing

TH3C-5
Contrast Quantification of Millimeter-wave Scanning Near-field Microscope with a Slit Probe

TH3C-6
A Novel W-band Spectrometer for Dielectric Measurements

TH3C-7
A Near-field Alignment Technique at Millimeter and Sub-millimeter Wavelengths

TH3D
FILTERS (3)
Chair: D. Schmitt, Co-chair: M. Guglelimi

TH3D-1
Tuning & Measurements of Couplings and Resonant Frequencies for Cascaded Resonators

TH3D-2
Computer Diagnosis and Tuning of Microwave Filters Using Model-based Parameter Estimation and Multi-level Optimization

TH3D-3
Filters With an Almost Constant Stopband

TH3D-4
Full Wave Modeling of Electric Coupling Probes in Combline Resonators and Filters

TH3D-5
Mobius Resonators and Filters

TH3D-6
Single-passband, Single-stopband Narrowband Filters

TH3E
NEW PROPAGATION EFFECTS ON CPW AND MICROSTRIP
Chair: P. Lampariello, Co-chair: J. Zehentner

TH3E-1
Dispersion and Radiation Characteristics of Conductor-backed CPW with Finite Ground Width

TH3E-2
The Influence of a Top Cover on the Leakage from Microstrip Line

TH3E-3
Behavioral Feature of the Space-wave Leaky Mode, the Surface-wave Leaky Mode and the Bound Mode

TH3E-4
Mode-coupling Effect between Different Two Types of the Surface-wave Leaky Mode on Conductor-backed Coplanar Strips (CBCPS)

TH3E-5
Low-loss Millimeter-wave Propagation Characteristics of NRD-guide Surface-mounted on Planar Substrate for Hybrid Integrated Circuit

3:30 PM-5:10 PM

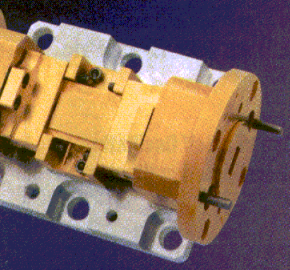
TH4A
NOVEL MICROWAVE SYSTEMS AND COMPONENTS
Chair: R. Gupta, Co-chair: R. Kaul

TH4A-1
LMDS Upconverter and Downconverter MMIC

TH4A-2
A 25-GHz Downconverter Integrated with a

MILLIMETER WAVE COMPONENTS

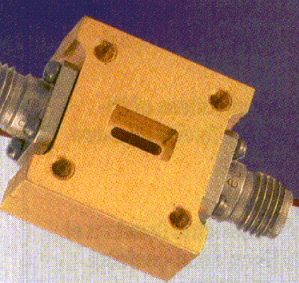
AMPLIFIERS • MIXERS • MULTIPLIERS



AMPLIFIERS

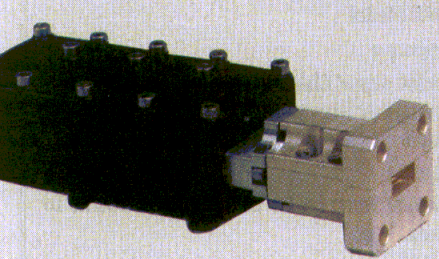
Model Number	Frequency (GHz)	Gain (dB, Min.)	Gain Flatness (±dB, Max.)	Noise Figure (dB, Max.)	I/O VSWR (Max.)	Output Power at 1dB Comp.* (dBm, Typ.)
JSW4-18002600-18-5A	18-26	28	1.0	1.8	2.0:1/2.0:1	5
JSW4-26004000-25-5A	26-40	25	2.5	2.5	2.0:1/2.0:1	5
JSW4-18004000-32-8A	18-40	21	2.0	3.2	2.0:1/2.5:1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

* Higher output power options available



MIXER/CONVERTER PRODUCTS

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



MULTIPLIERS

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

* Higher output power options available

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CIRCLE NO. 290

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TH4A-3
Chip Multilayer Antenna for 2.45-GHz-band Application Using LTCC Technology

TH4A-4
A Novel Broadband Receiver Architecture for Wireless Communication and Radar Systems

TH4A-5
Chaotic Radar Systems

TH4A-6
Digital Communications Using Self-phased Arrays

TH4B
VERY HIGH POWER COMPONENTS
Chair: J. Goel, Co-chair: R. Mallavarpu

TH4B-1
Low Distortion High Power GaAs Pseudomorphic Heterojunction FETs for L/S-band Digital Cellular Base Stations

TH4B-2
100-W L-band GaAs Power FP-HFET Operated at +30 VDC

TH4B-3
A 240-W Push-pull GaAs Power FET for WCDMA Base Stations

TH4B-4
High Power Modular S-band Solid State Transmitters Family for ATC and Naval Applications Radars

TH4B-5
Modeling Single- and Multi-carrier Performance of a Helix TWTA with the 1-D Parametric Code, CHRISTINE

TH4B-6
Multi-megawatt X-band Semiconductor Microwave Switches

TH4C
PHOTONIC SYNTHESIS OF MICROWAVE SIGNALS
SPECIAL SESSION
Chair: P.D. Bieracki, Co-chair: A.J. Seeds

TH4C-1
Fibre-integrated Heterodyne Optical Injection Phase-locked Loop for Optical Generation of Millimetre-wave Carriers

TH4C-2
(Invited) All-optical Heterodyne RF Signal Generation Using a Mode-locked-laser Frequency Comb: Theory and Experiments

TH4C-3
(Invited) A Good Prospect for Broadband Millimeter-wave Fiber-radio Access System—An Approach to Single Optical Component at Antenna Base Station

TH4C-4
Optical Millimeter Wave Generation Utilizing a Subharmonic Reference

TH4C-5
30-GHz Microstrip HEMT Oscillator Using Indirect Optical Injection Locking

TH4D
FILTERS/MULTIPLEXERS (4)
Chair: C. Wang, Co-chair: R. Snyder

TH4D-1
A Wide-band Tunable Filter Technique Based on Double-diplexing and Low-Q Elements

TH4D-2
Triple Mode Filters with Coaxial Excitation

TH4D-3
Accurate Modeling of Narrow-band Filters for Satellite Communications

TH4D-4
A Simplified Analysis for High Power Microwave Bandpass Filter Structures

TH4D-5
Six-pole Triple Mode Filters in Rectangular Waveguide

TH4D-6
Transversal Filter Using Whispering Gallery Quarter Cut Resonators

2:30 PM-5:00 PM

THIF
INTERACTIVE FORUM
Chair: Dan Swanson, Co- Chair: Doug Teeter

THIF-1
Frequency-space Mapping Optimization of Microwave Circuits Exploiting Surrogate Models

THIF-2
Passive Model-reduction of Distributed Networks with Frequency-dependent Parameters

THIF-3
Frequency and Geometry Parameterization for 3D Microwave Circuit Design and Optimization Using Pole Expansion and Derivatives

THIF-4
Modeling Source Inductance in FETs

THIF-5
Microstrip-to-coplanar Waveguide Transition Full-wave Analysis Model

THIF-6
Electromagnetic Modeling of Linear PHEMTs for Millimeter Waves

THIF-7
Integrated Microstrip to NRD-guide Transition Using a Spurious Mode Suppressing Technique

THIF-8
Ultra Low-loss Transmission Lines on Low Resistivity Silicon Substrate

THIF-9
Computation of Causal Characteristic Impedances

THIF-10
Wide Bandwidth Microwave Phase Detection Using Optical Intensity Modulators

THIF-11
Nonlinear Distortion Compensation of Microwave Fiber Optic Links with Asymmetric Adaptive Filters

THIF-12
Optoelectronic Mixing in a Self Oscillating InP/InGaAs Photo-Heterojunction Bipolar Transistor

THIF-13
Linearized Optical Directional-coupler Modulators for Analog RF/Microwave Transmission Systems

THIF-14
A Novel Fiber-optic Link with Remote Heterodyning for Phase Stabilized Transmission of Microwave and Millimeter Wave Signals

THIF-15
General Accuracy Considerations of Microwave On-wafer Silicon Device Measurements

THIF-16
Measurements of the Dielectric Constant of Low-loss materials Using Semi-confocal (Open-resonator) at 55-65 GHz.

THIF-17
Accurate Determination of Dielectric Constant of Substrate Materials Using Modified Wolff Model

THIF-18
Vector Signal Characterization of 38-GHz Power Amplifier with 200 Mb/s QPSK Modulation

THIF-19
A Novel Setup for Co-channel Distortion Ratio Evaluation

THIF-20
An Automatic Harmonic Selection Scheme for Measurements and Calibration with the Nonlinear Vectorial Network Analyzer

THIF-21
Effects of Pin Depth in LPC 3.5-mm, 2.4-mm, and 1.0-mm Connectors

THIF-22
In-circuit Testing of Complex Circuits Using On-wafer Probing and Electromagnetic Coupled Ground Interconnects

THIF-23

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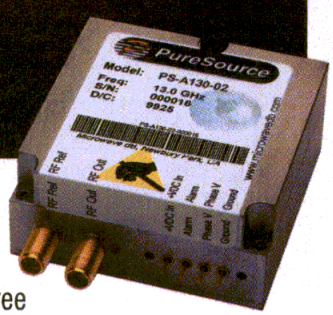

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CIRCLE NO. 253

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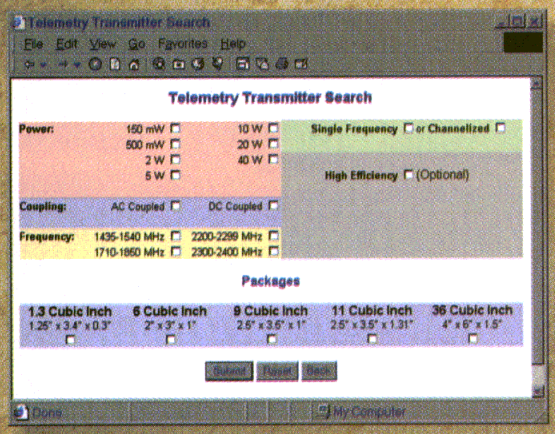

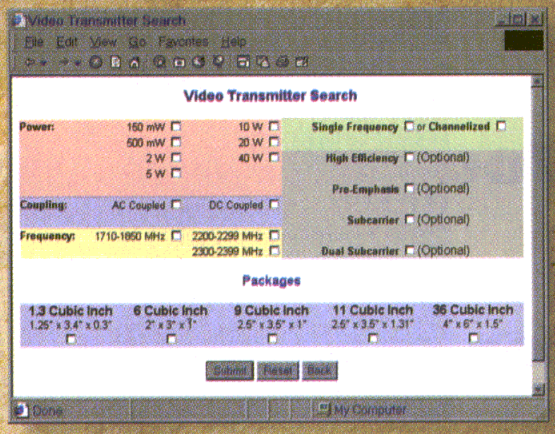

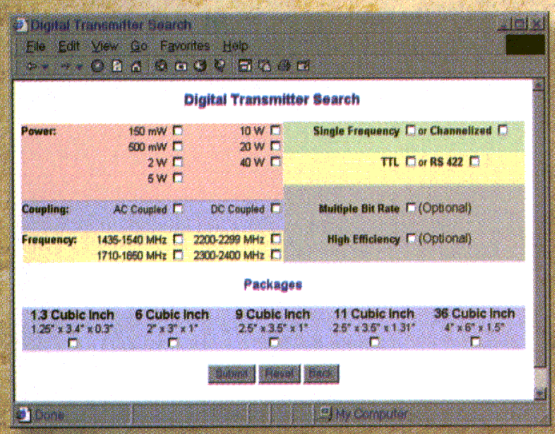

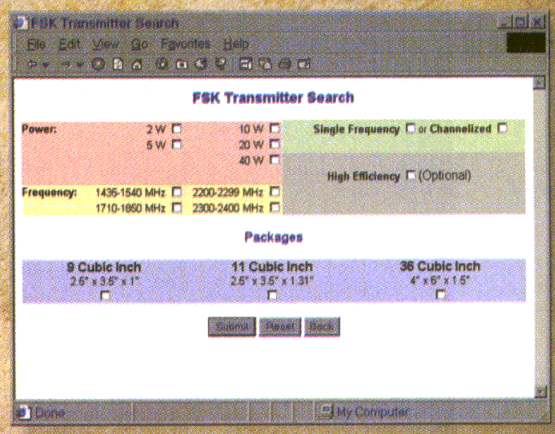

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CIRCLE NO. 273

A Modified Open-ended Coaxial Probe for Concave Surface Coating Materials Testing

THIF-24

Measurement of a DSB Modulated RF Signal by Use of Discriminant Filters and Envelope Detection.

THIF-25

Isolation in Three-dimensional Integrated Circuits

THIF-26

Development of Microwave/Millimeter-wave Integral Passives for Multi-layer Organic MCMs

THIF-27

A New Concept of RF Feed-through Applied to Multi-chip Modules for Space Equipment

THIF-28

Characterization of Surface Mount Packages at Microwave Frequencies Using Wafer Probes

THIF-29

Development of Finite Ground Coplanar (FGC) Waveguide 90-deg. Crossover Junctions with Low Coupling

THIF-30

A Ku-band Frequency-tunable Active Matched Feedback MMIC Amplifier Using Variable-capacitance Elements

THIF-31

Si/SiGe HBT Active Integrated Antenna on High Resistivity Silicon Substrate

THIF-32

GaN Gunn Diodes for THz Signal Generation

THIF-33

Three Dimensional Power Amplifier and Special Elements Design

THIF-34

New Methodology for Microwave/Millimeter-wave MMIC Development

THIF-35

Survivability of InP HEMT Devices and MMICs under High RF Input Drive

THIF-36

Global Electromagnetic Analysis to Help the Conception of an Active Module

THIF-37

A New Thin Film Passive Integration Technology for Miniaturization of Mobile Phone Front End Modules; Integration of a Dual-band Power Amplifier, Switch, and Diplexer for GSM

THIF-38

Surface Integral Equation Modeling Approach to the Handset Antenna and Human Body Interaction

THIF-39

Analysis of a Passive Spatial Combiner Using Tapered Slotline Array in Oversized Coaxial Waveguide

THIF-40

Electro-thermal Modeling and Measurement of Thermal Time Constants and Natural Convection in Spatial Power Combining Grid Arrays

THIF-41

Improved Balanced Crossed Dipole Quasi-Optical Frequency Doubler

THIF-42

Generation of Cs-Xe Discharge Plasma Slabs for Nonlinear Microwave Quasioptical Devices

THIF-43

Influence of Output Impedance on Power Added Efficiency of Si-bipolar Power Transistors.

THIF-44

Low-cost Millimeter-wave Transmitter Architecture Using Software Radio Techniques

THIF-45

Effect of Out-of-band Terminations on Cross Modulation in CDMA Receivers

THIF-46

A 2-GHz Band Even Harmonic Type Direct Conversion Receiver for WCDMA Mobile Terminal Utilization

THIF-47

New Type of High Speed Pulse Radar Based on the NRD Guide Technology at 60-GHz Band

THIF-48

Beam Steering Antenna for MMW Imaging Radar

THIF-49

Monolithic GaAs Surface Acoustic Wave Chemical Microsensor Array

THIF-50

Low-cost Automobile Crash Sensor Using a 61-GHz Active Integrated SIMMWIC Antenna

THIF-51

76-GHz Post-wall Waveguide Fed Parallel Plate Slot Array with 45-deg. Linear Polarization

J. Hirokawa, M. Ando

WFA

INTEGRATED TRANSCEIVER DESIGN USING SILICON-BASED SEMICONDUCTORS

RF transceivers are key elements in the expanding communications area. A critical need exists to provide the RF engineer with the design tools and performance information necessary to incorporate the integrated design in this low-cost medium.

This workshop will focus on silicon-based RF integrated design techniques. CMOS and BiCMOS process characteristics and RF performance will be presented for active (diodes, BJTs and MOSFETs) as well as passive components (resistors, capacitors, and inductors).

WFB

HIGH POWER TRANSMITTER SYSTEMS AND RELATED SUBSYSTEMS FOR WIRELESS BASE STATIONS

Together with the switches, the base stations constitute some of the costliest investments a wireless operator needs to make in order to establish wireless service. It is also an application ground for some of the most rapidly changing and evolving technologies such as DSP, linearization, digital radios, high power devices and components, software, power supplies and cooling.

WFC

MICROWAVE OSCILLATORS—LOOKING BACK AND LOOKING FORWARD (OVERVIEW AND STATE OF OSCILLATOR THEORY AND DESIGN)

This workshop will provide an overview of oscillator design together with the new developments that include nonlinear simulation and measurements.

The workshop will consist of invited papers that will outline oscillator design and provide an overview of best design practices and the newest developments in the field.

WFD

MICROWAVE AND PHOTONIC APPLICATIONS OF MEMS

Microelectromechanical systems (MEMS) has seen tremendous growth recently and is generally considered a revolutionary, enabling technology. The aims of this workshop are to disseminate, update, and discuss ongoing research in the MEMS field relating to device development and systems applications and to stimulate new developments at several levels.

WFE

FERROELECTRIC MATERIALS AND MICROWAVE APPLICATIONS

Advances in materials science have

led to a variety of new materials with strong potential applications to microwave and millimeter-wave components. Materials are available in a thin film that can be integrated into existing MMIC processes.

WFF

AUTOMATED CIRCUIT OPTIMIZATION USING ELECTROMAGNETIC SIMULATORS

Most, if not all, commercial CAE systems for RF, wireless, and microwaves have automatic optimization capabilities. The role of optimization in modeling, simulation, and design is taken for granted by microwave circuit designers. This workshop will emphasize optimization methodologies as a cornerstone in simulation, modeling, design, and manufacturing.

WFG

BIOLOGICAL EFFECTS OF EM FIELDS

Public concerns amount from hazards of non-ionizing EM fields and radiation including electric and magnetic fields and radiation, which arise from sources like power lines, distribution lines, substations, radio transmitters, cellular stations, and other electrical devices. This short course will review the long history of this issue, the controversy concerning possible biological effects and expected health hazards. It will provide details of the engineering research and scientific data that result in the setting of health safety standards.

WFH

PRINTED ANTENNA TECHNOLOGY FOR WIRELESS AND SATELLITE COMMUNICATIONS

This half-day workshop will provide participants with a good understanding of the miniaturization concepts for printed antenna technology for wireless connectivity and satellite communications covering the frequency range of 450 MHz to millimeter waves.

2000 RF IC SYMPOSIUM TECHNICAL PROGRAM

Monday, June 12, 2000

8:00 AM

MON1A

EXCITING APPLICATIONS OF RF ICs

PLENARY SESSION

Chair: F. Ali, Co-chair: D. Lovelace

MON1A-1

(Invited) Communication ICs for the Next Millennium

MON1A-2

(Invited) Bluetooth Radio Architectures

MON1A-3

(Invited) RF and System Requirements for Cable Modem Applications

MON1A-4

(Invited) High Data Rate Wireless LANs

10:30 AM

MON2A

CUTTING-EDGE RF ICs

Chair: D. Lovelace, Co-chair: F. Ali

MON2A-1

(Invited) Broadband Tuner on a Chip for Cable Modem, HDTV and Legacy Analog Standards

MON2A-2

(Invited) Symphony - A Sigma Delta Synthesizer

MON2A-3

(Invited) A Direct Conversion Transceiver for Multi-band GSM Application

MON2B

MILLIMETER-WAVE COMPONENTS AND SUBSYSTEMS

Chair: L. Liu, Co-chair: J.P. Mondal

MON2B-1

67-GHz Static Frequency Divider Using 0.2-mm Self-aligned SiGe HBTs

MON2B-2

Monolithic Resistive Mixers for 60-GHz Direct Conversion Receivers

MON2B-3

Monolithic Power Amplifiers Covering 70 to 113 GHz

MON2B-4

A High-performance 85-to-119-GHz GCPW MMIC Low Noise Amplifier

MON2B-5

Applications of Virtual Modeling to the Development of a 39-GHz MMIC Transceiver on Multilayer Multi Chip Assembly (MCA) Substrate

1:30 PM

MON3A

INTEGRATED VCOs

Chair: T. Tewksbury, Co-chair: J. Moniz

MON3A-1

(Invited) A Novel RF IC for UHF Oscillators

MON3A-2

A Fully-monolithic SiGe Differential Voltage-controlled Oscillator for 5-GHz Wireless Applications

MON3A-3

A Fully Integrated SiGe Bipolar 2.4-GHz Bluetooth Voltage-controlled Oscillator

MON3A-4

A Low Phase Noise Monolithic VCO in SiGe BiCMOS

MON3A-5

A Fully Integrated 5.35-GHz CMOS VCO and a Prescaler

MON3B

INTEGRATED RECEIVERS

Chair: S. Lloyd, Co-chair: K. Ashby

MON3B-1

(Invited) A Highly Integrated Commercial GPS Receiver

MON3B-2

Analog Baseband IC for Use in Direct Conversion WCDMA Receivers

MON3B-3

A Low-IF, Polyphase Receiver for GSM Using Log-domain Signal Processing

MON3B-4

Low Cost 900-MHz Single-chip Cordless Telephone Receiver

MON3C

RF POWER AMPLIFIERS—WATT'S UP

Chair: M. Kumar, Co-chair: F. Ali

MON3C-1

A +3.2-VDC, 45-percent Efficient, Novel Class AB+C CDMA MMIC Power Amplifier Using Quasi Enhancement Mode PHEMTs

MON3C-2

An Ultra Low Current Consumption Two Step Power Control Driver MMIC with Self Current Control for Wide-band-CDMA Handsets

MON3C-3

(Student Paper) A Single Supply Very High Power and Efficiency Integrated PHEMT Amplifier for GSM Applications

MON3C-4

A 3.0-to-4.0-GHz Fully Matched High Linearity GaAs MMIC Power Amplifier for Wireless Communications

MON3C-5

Effect of Bias Scheme on Intermodulation Distortion and Its Use for the Design of PCS TX Driver

3:30 PM

MON4A

TRANSMIT ICs

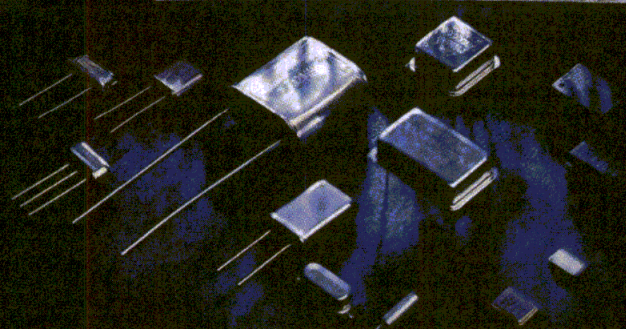
Chair: J. Moniz, Co-chair: T. Tewksbury

MON4A-1

(Invited) The Current Status of Microwave Amplifier Linearity and Intermodulation Im-

(continued on page 242)

The difference is Crystal Clear.



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OSCILLATORS

Oscillator type
- TTL
- HCMOS
- DUAL (HCMOS/TTL)
- TRI-STATE
- VCXO

Frequency Tolerance
- 10 , 25 , 50 & 100ppm

Operating temp. range
- Commercial
- Industrial
- Custom

Size / Style
- Full-size (14 pin)
- Half-size (8 pin)
- Gull-wing (14 /8 pin)
- Surface mount

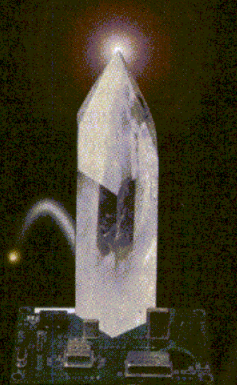
CRYSTALS

Holder styles
- Through-hole
- Surface mount

Calibration tolerance
- 10 , 25 , 50 & 100ppm

Operating temp. range
- Commercial
- Industrial
- Custom

Options
- Gull-wing
- Ground wire



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CIRCLE NO. 394

Avionics transistor pushes 350-W power

Model TAN 350 is a high-power common-base bipolar transistor designed for pulsed avionics systems operating from 960 to 1215 MHz. Geared for +50-VDC operation, the transistor offers at least 350-W pulsed (10-μs pulses) output power with 7.5-dB typical gain and 40-percent typical collector efficiency. **GHz Technology, Inc., 300 Oakmead Village Dr., Santa Clara, CA 95051-0808; (408) 986-8031, FAX: (408) 986-8120, Internet: http://www.ghz.com.**

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Foam helps insulate sensitive components

ECCOSTOCK FPH is a high-temperature rigid foam-in-place urethane material that is ideal for filling cavities and protecting sensitivity electronics components from vibration. The two-part liquid foam system creates a rigid foam material with low dielectric loss, good thermal insulation, and good vibration-damping properties. Kits are available in three sizes: 2, 10, and 70 lbs. (0.9, 4.5, and 31.5 kg). Once cured, the in-place liquid foam can provide densities of 2, 4, 6, and 10 lbs. (0.9, 1.8, 2.7, and 4.5 kg)/ft^3. **Emerson & Cuming Microwave Products, 28 York Ave., Randolph, MA 02368; (781) 961-9600, FAX: (781) 961-2845, Internet: http://www.emerson-cumingmp.com.**

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Four-port circulator handles PCS band

A four-port circulator has been developed for applications in the personal-communications-services (PCS) band covering 1930 to 1990 MHz. The circulator features 0.6-dB maximum insertion loss (with typical performance of 0.4 dB) and 40-dB minimum isolation (with typical performance of 47 dB). The circulator, which handles 100-W continuous-wave (CW) power, features maximum VSWR of 1.25:1. It incorporates a tabbed 20-dB attenuator output port with ±1-dB attenuation flatness and an attenuation port rating of 10 W CW. The circulator is rated for operating temperatures that range from –40 to +90°C. **Alcatel Ferrocom Ferrite Products, 6385 San Ignacio Ave., San Jose, CA 95119; (408) 220-8171, FAX: (408) 229-8506, Internet: http://www.ferrocom.com.**

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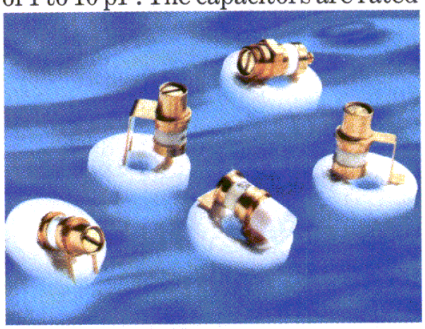
Trimmer capacitors span 1 to 10 pF

Eco-Trim variable capacitors in the 1500 series are low-cost, high-performance air-dielectric capacitors covering a capacitance tuning range of 1 to 10 pF. The capacitors are rated



for voltages to +250 VDC and operating temperatures from –65 to +125°C. The components are available in five mounting styles for printed-circuit and surface-mount applications. P&A: $1.95 (2500 qty.); 4 to 6 wks. **Johanson Manufacturing, Rockaway Valley Rd., Boonton, NJ 07005; (973) 334-2676, FAX: (973) 334-2954, Internet: http://www.johansonmft.com.**

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Tiny termination absorbs 250-W power

Model 83-3994TC is a high-power, flangeless signal-sampling termination designed for used to terminate an RF circuit and provide a low-level signal sample to a detector. The termination, which is designed for use from DC to 1 GHz, has a temperature coefficient of less than 200 PPM over an operating temperature range of –55 to +150°C. It is designed to handle average power of 250 W to +100°C. The termination has nominal impedance of 50 Ω. The termination is ideal for monitoring the signal-power imbalance in hybrid couplers or to monitor the reverse power in a three-port circulator. **Florida RF Labs, Inc., P.O. Box 899, Stuart, FL 34995-0899; (800) 544-5594, (561) 286-9300, FAX: (561) 283-5286, Internet: http://www.rflabs.com.**

CIRCLE NO. 78 or visit www.mwrf.com

MMIC power amp boosts 21 to 23 GHz

Model ES/FMM5811GJ-2 is a three-stage monolithic-microwave-integrated-circuit (MMIC) power amplifier (PA) designed for use from 21.3 to 23.6 GHz. It achieves typical output power at 1-dB compression of +24 dBm (0.25 W) with a typical power-added efficiency (PAE) of 20 percent. The power gain at 1-dB compression is typically 15 dB while the gain flatness across the operating frequency band is typically 1 dB. The gallium-arsenide (GaAs) MMIC device, which is internal matched to 50 Ω at its input and output ports, draws 190-mA drain current at a drain voltage of +6 VDC and 7-mA gate current at a gate voltage of –5 VDC. It is well-suited for point-to-point and point-to-multipoint microwave links. **Fujitsu Compound Semiconductor, Inc., 2355 Zanker Rd., San Jose, CA 95131-1138; (408) 232-9500, FAX: (408) 428-9111, Internet: http://www.fcsi.fujitsu.com.**

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Microwave analyzers scan 1 MHz to 24 GHz

The 6840 series of RF and microwave system analyzers integrates a synthesized signal source, spectrum analyzer, and a scalar network analyzer into a single compact housing. Available in variants from 1 MHz to 24 GHz, the test systems can perform conventional transmission and reflection measurements as well as testing on frequency-conversion components, such as RF/microwave mixers. The system's synthesizer achieves phase noise of –106-dBc/Hz offset 10 kHz from a 1-GHz carrier, with harmonics of less than –55 dBc. The spectrum analyzer offers resolu-

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V.S.W.R. (max)	1.9:1	2.0:1	2.0:1
Insertion loss (max)	0.8dB	0.9dB	0.9dB
Isolation (min)	50dB	50dB	50dB
RF Power	Avg. 10 watts	Avg. 5 watts	Avg. 3 watts
Switching time	15 msec max	15 msec max	15 msec max
Operating mode	Failsafe or Latching	Failsafe or Latching	Normally Open
Environmental	Designed to meet MIL-E-5400 and MIL-S-3928, ISO-9001 Certified		

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tion-bandwidth filters ranging from 1 kHz to 3 MHz, with an instantaneous dynamic range of better than 80 dB. The scalar network analyzer offers AC and DC detector modes and accurate return-loss measurements through autotesters with 40-dB directivity. **IFR Systems, Inc., 10200 West York St., Wichita, KS 67215; (888) IFR-HELP, Internet: http://www.ifrsys.com.**

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Isolators/circulators aim at 3G WCDMA systems

A line of affordable isolators and circulators has been developed for third-generation (3G) wideband code-division-multiple-access (WCDMA) communications systems. Designs include drop-in configurations sized at 0.75 × 0.75 × 0.23 in. (1.905 × 1.905 × 0.5842 cm) and thermally compensated with a typical performance of 20-dB isolation (min), 0.4-dB insertion loss (max), VSWR of 1.22:1, and a frequency range of 1.9 to 2.1 GHz. **Renaissance Electronics Corp., 1300 Massachusetts Ave., Boxborough, MA 01719; (978) 263-4994, FAX: (978) 263-4944, Internet: http://www.rec-usa.com.**

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Chip trimmer caps span 4 to 20 pF

The JZL and JRL series of chip trimmer capacitors maintains ±1-percent capacitance drift at low cost. The capacitors are available in capacitance ranges of 4 to 10 pF and 7 to 20 pF. **Voltronics International Corp., 100 Ford Rd., Denville, NJ 07834; (973) 586-8585, FAX: (973) 586-3404.**

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Generator and tester combine for 3G solution

The model MG3681A signal generator and the MS8608A transmitter tester team to provide test capability for third-generation (3G) wideband-code-division-multiple-access (WCDMA) communications standards. The signal generator features 30-MHz internal as well as external in-phase/quadrature (I/Q) modulation bandwidth and support for chipping rates to 16 Mchips/s. The transmitter tester, which combines a transmitter, power meter, and spectrum analyzer in one cabinet, provides a 20-MHz analysis bandwidth. The transmitter tester covers a frequency range of 300 kHz to 3 GHz. **Anritsu Co., 1155 E. Collins Blvd., Richardson, TX 75081; (800) ANRITSU, (972) 644-1777, FAX: (972) 644-3416, Internet: http://www.anritsu.com.**

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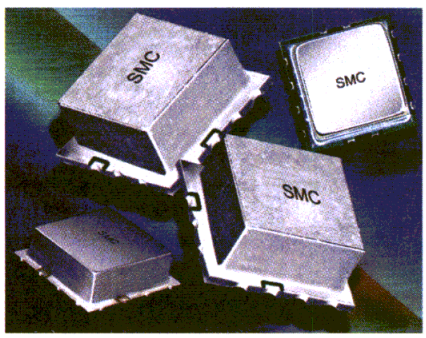
Cavity filters provide low insertion loss

A line of cavity/comb line filters is intended for infrastructure applications in the cellular and personal-communications-services (PCS) frequency bands. The devices offer an



insertion loss of 1 dB (typical) in the passband and more than 70-dB rejection at ±35 MHz in the cellular band. Power-handling capacity is 100 W. **Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8000, FAX: (973) 881-8361, e-mail: sales@synergymwave.com.**

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VCXOs offer less than 1-ps jitter

The CO-600V series of voltage-controlled crystal oscillators (VCXOs) features a low-jitter performance that is better than 1 ps at 155.52 MHz. Available at frequencies spanning the 10-to-160-MHz range, these precision oscillators are compatible with emitter-coupled-logic (ECL) and positive ECL (PECL) signal levels and are available in a variety of package options for use in high-performance communications system applications. **Vectron International, 166 Glover Ave., Norwalk, CT 06856-5160; (88) VECTRON-1, (888) FAX-VECTRON, Internet: http://www.vectron.com.**

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InGaP HBT amp module serves WCDMA systems

Model CHP0230-PM is a two-stage power amplifier (PA) that is developed for multimode wireless applications ranging across the 824-to-849-MHz frequency range. Fabricated with an InGaP heterojunction-bipolar-transistor (HBT) process, the CHP0230 amplifier module achieves 40-percent linear power-added efficiency (PAE) along with 60-percent analog PAE. It features +28.5-dBm output power in wideband- code-division-multiple-access (WCDMA) systems and can generate 30-dB gain when operating from a single-voltage supply as low as +3.2 VDC. **Celeritek, Inc., 3236 Scott Blvd., Santa Clara, CA 95054; (408) 986-5060, FAX: (408) 986-5095, Internet: http://www.celeritek.com.**

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Avionics amplifiers deliver peak power to 6 kW

A line of high-power linear amplifiers based on proprietary laterally-diffused-metal-oxide-silicon (LDMOS) device technology has been developed for pulsed avionics applications. Based on LDMOS transistors that are capable of 50 to 100 W per device at frequencies spanning the range from 100 to 2400 MHz, the devices are combined in modules which, in turn, can be combined in amplifier systems that are capable of providing pulsed output-power levels to 6 kW at avionics frequencies. **Zeta, an Integrated Defense Technologies Co., 2811 Orchard Pkwy., San Jose, CA 95134; (408) 434-3600, FAX: (408) 433-0205, Internet: http://www.zeta-idt.com.**

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Technique measures mixer group delay

Uncertainty regarding the accuracy of mixer group-delay measurements can be minimized through a new test technique. The measurement method, which is detailed in the first issue of *Device Test Forum* from Agilent Technologies (Palo Alto, CA), uses only a single broadband mixer and a vector network analyzer (VNA) with time-domain option.

The new technique, which requires that the mixer's intermediate-frequency (IF) port be terminated with a 50-Ω airline and short, measures the frequency response of the mixer's reflection coefficient. This frequency response is then transformed into an impulse response using the VNA's time-domain capability. Provided that the airline has been previously characterized and its electrical parameters are well-known, the two-way reflections from the short can be examined in the time domain and used to calculate the absolute delay through the mixer.

There are certain factors to consider when using this technique. For one, if the mixer contains any components with isolation, such as an isolator or preamplifier, the reflected signals from the short may be at levels that are too low for the VNA to make accurate measurements. The mixer must also have very broadband characteristics. In addition, the new measurement technique assumes that the device under test (DUT) is reciprocal in nature, with one-way transmission being identical when measured from the RF-to-IF ports and from the IF-to-RF ports. Differences in these responses translate into uncertainties in the mixer group-delay measurements. Copies of the first issue of *Device Test Forum* are available upon request from: **Agilent Technologies, P.O. Box 50637, Palo Alto, CA 94303-9512; Internet: http://www.agilent.com/find/DeviceTest-Forum.**

CIRCLE NO. 194 or visit www.mwrf.com

Selecting a digital storage oscilloscope

Digital storage oscilloscopes (DSOs) are workhorse instruments for designers of high-frequency analog and digital circuitry. With the appropriate bandwidth and sampling speed, a DSO can capture continuous and transient signals for engineers who are involved with troubleshooting, prototype design, and basic research. But selecting the best DSO for a particular application involves a number of choices, and these are outlined in an eight-page application note, "Selecting The Best Digitizing Storage Oscilloscope For Your Application," available from Tektronix, Inc. (Beaverton, OR).

The application note reviews the key specifications of DSOs, including bandwidth, the bandwidth-rise-time relationship, sampling rate, triggering capabilities, and record length. For example, the number of signals to be captured and displayed simultaneously determines the number of oscilloscope channels that are needed.

A DSO's bandwidth specification defines its measurement frequency range. The bandwidth is specified as the frequency where sinusoidal input signal has been attenuated to 70.7 percent of its amplitude at midband, also known as the 3-dB point. Unfortunately, the vertical amplitude at this point has an error of approximately 30 percent, so that performing measurements in this range provides unacceptable accuracy for most applications. For that reason, the "five times rule" is usually applied when determining the necessary bandwidth of a DSO. Using this rule, the bandwidth is calculated by multiplying the highest signal component to be measured by five. So, to measure signals in the 200-MHz range, the DSO measurement bandwidth should be at least 1000 MHz (1 GHz).

In some cases, the bandwidth of a signal to be measured may not be known. However, if the rise time of that signal is known, its bandwidth can be determined. Using a simple formula of 3-dB bandwidth (in Hertz) times the rise time (in seconds) equals a factor K, the rise time can be calculated for most DSOs. The value of K is usually assumed to be approximately 0.40, in which case a signal with typical rise time of 1.5 ns would translate to a signal bandwidth of 230 MHz.

Although written in 1997, the application note is nonetheless current in its advice for DSO specifiers. It provides carefully chosen graphics to illustrate key measurement parameters, along with tables of pertinent specifications. Copies of the application note are available upon request from: **Tektronix, Inc., P.O. Box 500, M/S 50-255, Beaverton, OR 97077-0001, (503) 627-1916, Internet: http://www.tek.com.**

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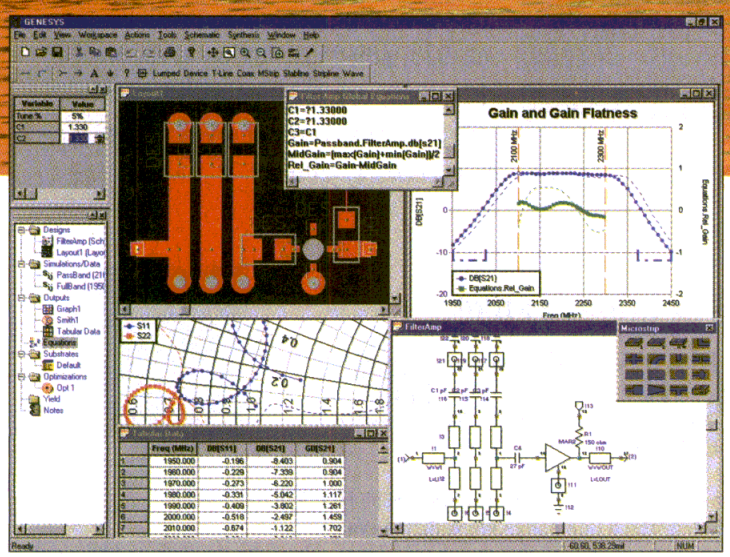

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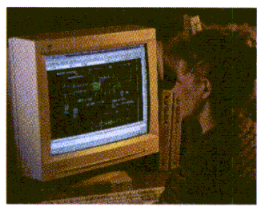

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CIRCLE NO. 301

Analyzer Adds Speed And Accuracy To NF Testing

The successor to the venerable 8970 series of noise-figure meters brings enhanced accuracy and operating speed to noise testing, with adjustable measurement bandwidth to 3 GHz.

Gareth Thomas and Ewan Shepherd

Agilent Technologies, Power and Noise Product Generation Unit, South Queensferry, West Lothian EH30 9TG, Scotland; (44) 1313357415, FAX: (44) 1313357322, Internet: http://www.agilent.com.

NOISE measurements have long been an instrumental part of RF/microwave component and system characterization and performance verification. Even with the move toward higher-order modulation schemes in wired and wireless communications systems, noise figure remains as important today as it was 46 years ago when the term was introduced to designers. However, there is an increasing need for a simplified process. Greater speed and accuracy in the measurement of noise figure are also desired (see sidebar). Agilent Technologies (Queensferry, Scotland) has introduced the NFA series of noise-figure analyzers to satisfy these requirements. The series is the successor to the company's 8970 noise-figure meters, the standard-bearers in noise-figure measurement for decades.

The NFA series of noise-figure analyzers provides up to twice the accuracy and as much as three times the speed of its predecessors, with many new capabilities and a more intuitive, streamlined interface for the user (Fig. 1). The series consists of two models—the Agilent N8972A, a 1.5-GHz "entry-level" successor to the 8970, and the 3.0-GHz Agilent N8973A, which has a higher level of performance and many new features (see table). Both instruments provide real-time display of noise figure and gain versus frequency, input of noise source, excess-noise-ratio (ENR) data from diskette, easy measurement set up, internal data storage, a 3.5-in. DOS-compatible diskette drive, and ports for GPIB, RS-232, printer (parallel), and a GPIB port that is dedicated to local-oscillator (LO) control for mixer measurements.

1. The NFA series of NF analyzers includes models with top frequencies of 1.5 and 3.0 GHz with improved operating speed and measurement accuracy compared to their predecessors.

The 3.0-GHz Agilent

N8973A adds the following features to this list:

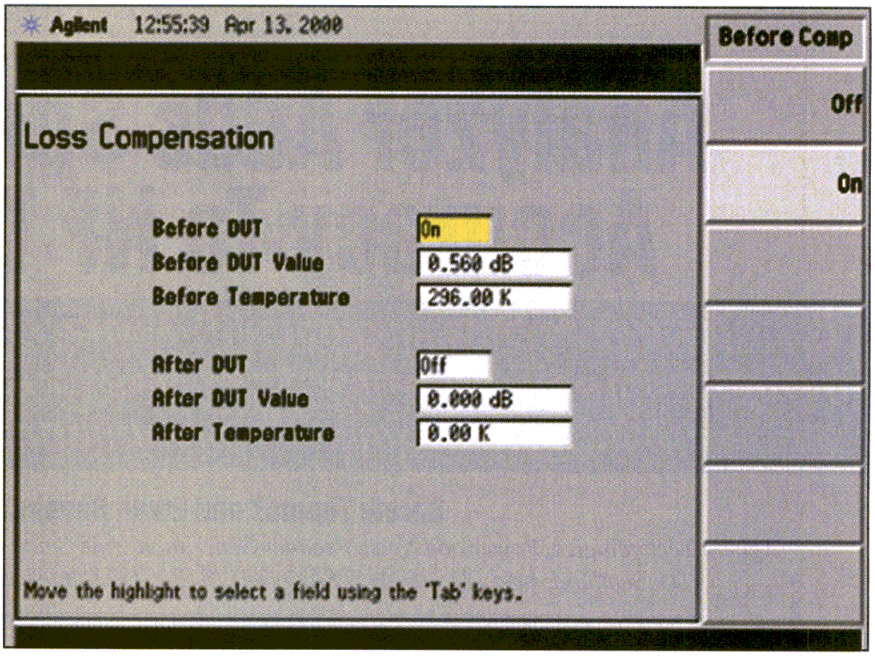
- It cuts the instrument noise-figure uncertainty of the 8970B in half.
- It provides greater repeatability and more accurate characterization of individual system components.
- It offers three times the measurement speed of the 8970B, which makes it better suited for testing devices such as direct-broadcast-satellite (DBS) receiver low-noise block downconverters (LNBs).
- It features measurement bandwidths as narrow as 100 kHz for testing amplifier/filter subsystems.

The features that are embodied in the Agilent's NFA series of instruments are the direct result of feedback from customers faced with noise-figure measurements. These features either directly strengthen performance levels established by the 8970B, or add capabilities required by current and (likely) future applications.

One of the most important changes is in the interface to the operator. Many designers perceive noise-figure measurement to be overly complex, and many of the instruments that are used to measure it do not help alleviate their apprehension. Instruments that measure noise figure, whether they are noise-figure meters or analyzers, spectrum analyzers, or network analyzers, are increasingly operated by nontechnical operators, which places the burden on the instrument designer in order to make operation as intuitive as possible.

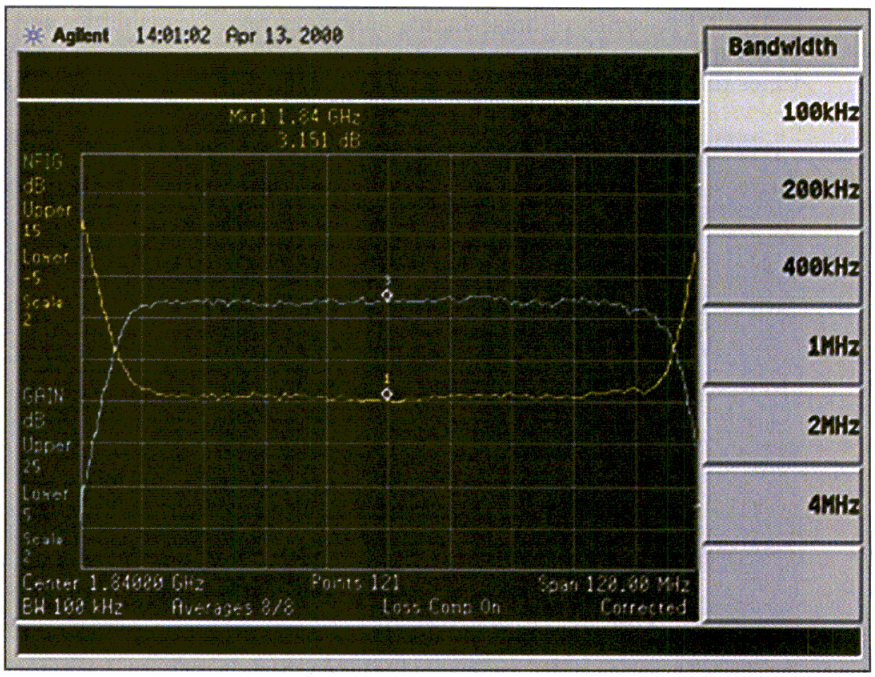
The NFA series achieves noteworthy progress in this area. The overall goal was to make it more difficult for mistakes to be made that affect the integrity of the measurements. Keys with complementary functions are grouped together, and the number of keystrokes required to make measurements is reduced. In addition, the soft-key menus are more intuitive than in previous instruments. An example of these improvements is the way the instruments support loss compensation either before or after the device under test (DUT) is connected (Fig. 2).

The display allows the user to switch on "before-DUT" or "after-DUT" loss compensation, as well to enter the loss compensation value. This means that loss compensation is automatically taken into account during the measurement, and the user can easily see what loss compensation is being applied within the measurement. Programming is also easier thanks to the comprehensive use of a small-computer programmable interface (SCPI), including the commands for each front-panel operation.



2. Entering loss compensation values has become a "fill-in-the-blank" operation with the NFA series instruments.

The color display of the NFA series allows noise figure and gain versus frequency to be displayed, with log or linear scaling. Noise figure and gain can be displayed simul-



3. Noise and gain are displayed with frequency on the liquid-crystal-display (LCD) screen of the NFA series of analyzers.

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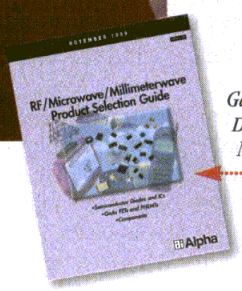

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NECESSITY'S GRANDCHILD (THE ROOTS OF NOISE FIGURE)

IF necessity is the mother of invention, then surely noise figure is the grandchild of that lineage. Until this term was defined, there were several ways to characterize the effect of noise on a system (especially receiver) performance, but not one stood far enough above the rest to become dominant. However, the ubiquitous presence of noise spawned noise figure as the fundamental parameter that quantified how much noise a component or system contributes to signals that pass through them. This parameter reigns essential still.

The importance of noise figure to overall system performance and cost is hard to overestimate. For example, by reducing half the noise figure of a direct-broadcast satellite (DBS) receiver in half from 2 dB to 1 dB has nearly the same effect on performance as doubling the transponder's transmit power. In a satellite receiver production line, noise figure can often be reduced by 1 dB simply by adjusting impedance levels or selecting appropriate transistors. This 1-dB noise figure reduction has approximately the same effect as increasing antenna diameter by as much as 40 percent. For wireless communication applications, a base station with a low receiver noise figure allows the mobile units it communicates with to reduce their transmit power, which has a positive effect on battery life along with the size and corresponding weight of the batteries they use.

Considering its performance implications, it is not surprising that noise has been the bane of receiver designers since there were signals to detect. However, it took Harold Friis, a consultant to Hewlett-Packard Laboratories, to coin the term, its theory, and derivation in 1944. Friis defined it as the signal-to-noise ratio (SNR) at the input port of an RF or microwave device divided by the SNR at the device's output port, expressed in decibels (dB).

As its name implies, SNR is the ratio of signal level to noise level in a particular transmission environment. The greater the number is, the more the signal exceeds the noise, thus making it more easily detected. In contrast, the lower the noise-figure value is the better, since ideally noise created by a microwave component, subsystem, or system would add nothing to the signal as it passes through. This is never the case, since all electrical devices add some noise to a signal, but devices that contribute the least are best, and have the lowest noise figures.

Thanks to Friis, there was now a convenient measure of noise performance, and the first instrument to make the measurement was the Model 340 from HP, which was introduced in 1958. It was primarily used in radar testing. Making proficient measurements required racks of equipment and a maze of cable. The tests back then were mainly carrier-to-noise, and noise-figure measurements gave designers greater confidence that their designs would meet their performance goals.

In 1981, HP introduced the HP 8970A noise-figure meter, which added automation to the measurement process for the first time. It was followed by the HP 8970B, which when it was introduced in 1987, offered optional frequency coverage to 2 GHz, and increased speed. Companies serving aerospace and defense markets remained the primary users of the instruments, although the satellite-communications and microwave-radio markets were increasingly interested in the instruments.

Now, almost a decade-and-a-half later, companies that manufacture or install wireless communications systems are the primary customers for noise-figure meters. While complex modulation schemes abound in today's wireless systems, and tests such as bit-error ratio (BER) and the ratio of effective bits to noise (E_b/N_o) are the staples, noise figure continues to be a fundamental parameter by which receiver performance is judged. Harold Friis would be proud.

taneously, either overlaid on the same graph or separately (Fig. 3). The instruments also allow spot measurements to be made and displayed in a large, highly legible font, using the meter display. Multiple points can be displayed in a table format. Gain is always shown corrected, and noise figure can be displayed in either corrected or uncorrected formats. The graphical display also provides markers and limit lines for checking measurements against predetermined limits in pass/fail testing environments. Users can select up to 401 measurement points, with the key trade-off being a penalty in measurement speed with the increasing number of measurement points.

Another issue addressed in the NFA series is electromagnetic interference (EMI), which is often present in environments where noise figure is measured. The source of this radiation may be nearby base stations, other instruments, or even fluorescent lights. A frequency-list mode in the NFA series allows operators to select measurement frequencies that are known to be free of EMI.

NARROW BANDWIDTHS

The 8970B noise-figure meter performs tests across a 4-MHz measurement bandwidth. But designers working with amplifier/filter combinations and measuring out-to-band edges often require a more flexible (narrower) measurement-bandwidth scheme. The Agilent N8973A has the ability to make measurements as narrow as 100 kHz, which makes it possible to evaluate filters used in Global System for Mobile Communications (GSM) equipment using a 200-kHz bandwidth. In addition, base-station subassemblies can now be evaluated directly at the band edges.

The need to provide noise-figure measurement data to others is important to many designers, whether transporting it to someone else in the company or to a customer. The NFA series addresses this by allowing more than one setup routine to reside in the instrument. Internal storage of traces, states, and ENR tables are provided, which allows existing information to be recalled

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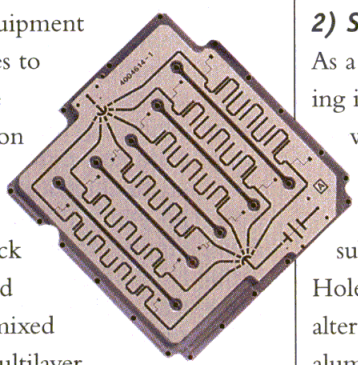

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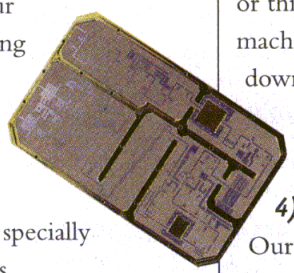

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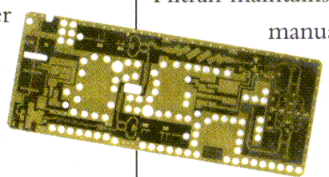

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Our Sodium Etchant formulation for PTH and edge plating gives us tight control of processing, resulting in reliable, high-performance circuits and excellent adhesion of copper to PTFE.

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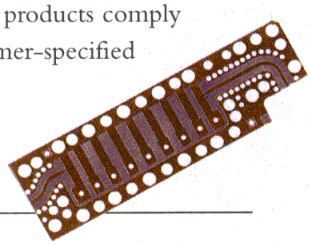
Our stringent program of in-process inspection results in high yields on the most complex circuits. We strictly monitor all production processes, and our upfront engineering, SPC and final inspection all help to ensure that products comply with customer-specified standards.



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CIRCLE NO. 221

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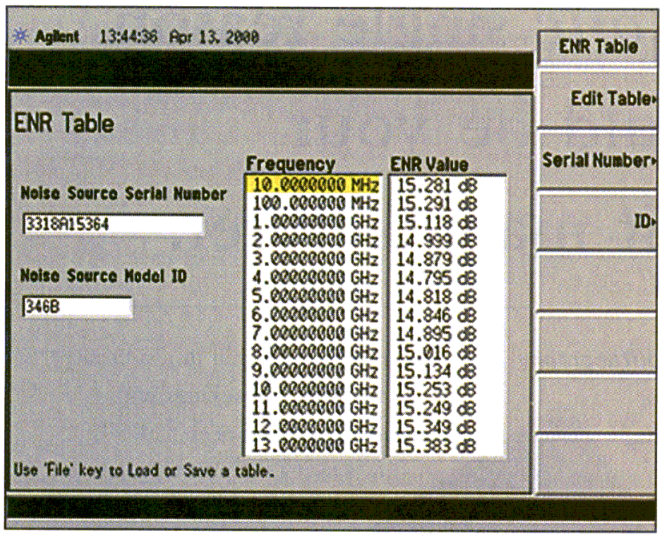


4. ENR data can easily be recalled for appropriate noise sources, and no longer must be entered manually.

consuming setup as well as recalibration.

As many as 30 sets of data can be stored on a diskette, depending on how complex the measurement set up is. The instruments' printer port is compatible with PCL3 and PCL5 parallel printers, and trace data can be saved in standard formats for inclusion in spreadsheet programs or reports. In addition to GPIB, the instruments can be controlled via the RS-232 serial port. An external video-graphics-array (VGA) display can be employed as well.

Entering ENR data into the instrument is substantially easier now because, for all new noise sources, Agilent supplies calibration data on diskette. Once entered, an ENR table for a particular HP or Agilent noise source can be recalled easily, showing model and serial number, providing all frequencies entered, along with the ENR data at these frequencies (Fig. 4). All Agilent 346 and 347 noise sources are now supplied with the ENR data on diskette.

Agilent has made increased education an integral part of its introduction, in order to provide a better understanding of the noise-figure measurement process. In addition to the standard technical specifications and configuration guide, several other documents have been released. An application note that has been provided to thousands of designers over the years ("Fundamentals of Noise Figure Measurements," Application Note 57-1) has been thoroughly updated with current information.

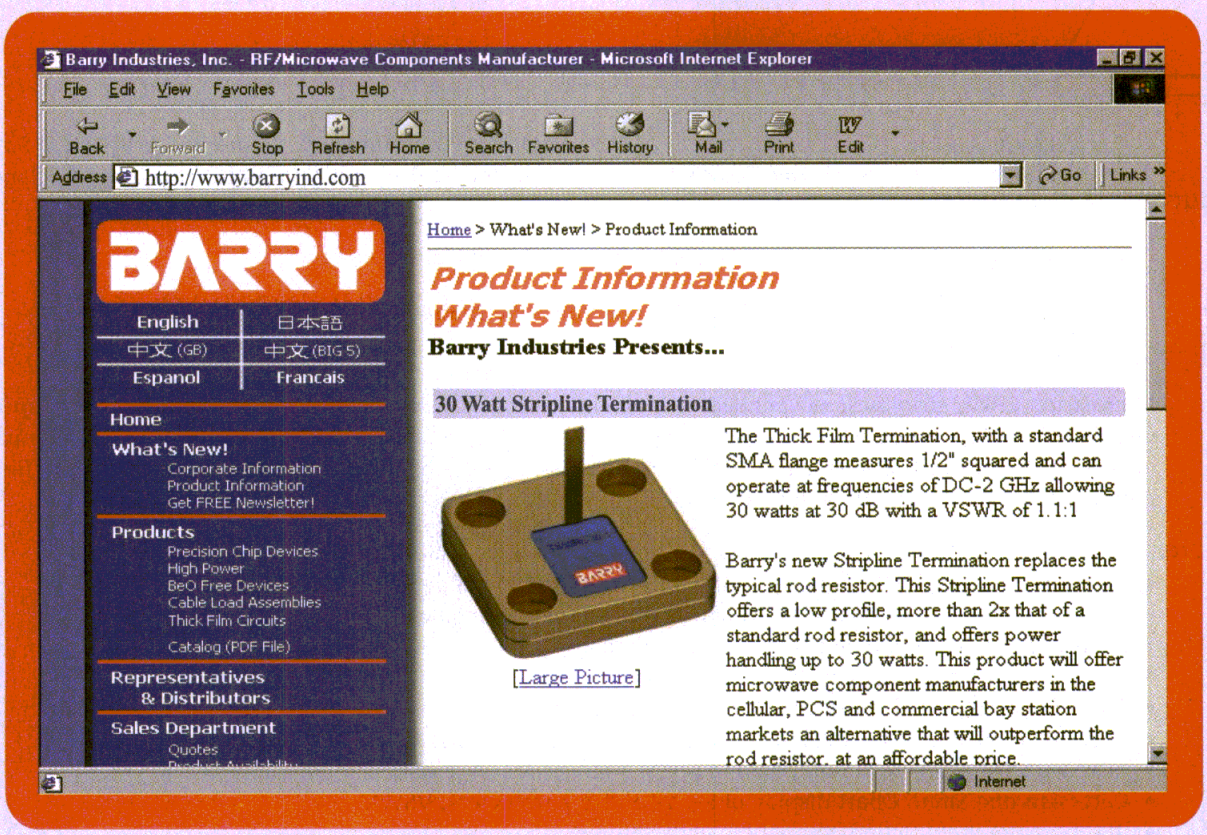
In addition, a publication called *10 Hints For Making Successful Noise Figure Measurements* addresses more specific issues. To find out more about the NFA, visit the Agilent website at http://www.agilent.com/find/nfa. The site provides links to application notes and tutorials that cover noise figure theory, measurement and applications. An uncertainty calculator is available at the website as well. It provides users with an indication of the factors that affect measurement uncertainty and lets them focus on reducing measurement uncertainty. The 1.5-GHz N8972A is priced at $16,750 and the 3-GHz N8973A is priced at $22,750. The NFA series is available now. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844 ext. 7021, Internet: http://www.agilent.com.**

CIRCLE NO. 51 or visit www.mwrf.com

The Agilent NFA series at a glance

	N8972A	N8973A
Measurement frequency range	10 MHz to 1.5 GHz	10 MHz to 3.0 GHz
Measurement bandwidths	4 MHz	4 MHz, 2 MHz, 1 MHz, 400 kHz, 200 kHz, 100 kHz
Sweep Number of points Trigger	2 to 401 or fixed frequency Continuous or single	
Instrument noise-figure uncertainty	+/−0.10 dB	+/−0.05 dB
Measurement speed 64 averages	<80 ms/sample	<42 ms/sample
Display Channels Markers Display units	17 cm (6 3/4-in.) color LCD 2 4 Noise figure (F dB or as ratio), gain (G dB or as ratio) y factor (y dB or as ratio), effective input noise temperature (K, °C, °F), relative power density (P_{hot}, P_{cold}, dB or as ratio)	
Interfaces	GPIB, LO GPIB, 9-pin RS-232, parallel, VGA, external reference	
Data storage Internal drive Diskette drive	30 traces, states, or ENR tables	
Dimensions without handle	222 × 410 × 375 mm	
Weight	<34.2 lb	
Calibration interval	1 year	
Warranty period	3 year standard	

High Power Resistors



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CIRCLE NO. 229

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- Mixed electromagnetic simulation and nodal analysis for large scale simulation
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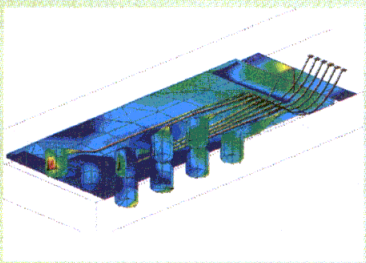
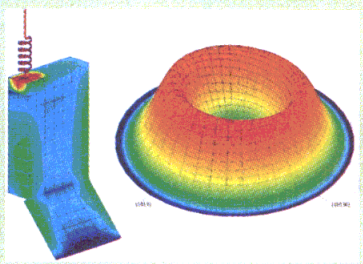
- RLC parameter extraction compatible with SPICE
- 2D and 3D display of current distribution, radiation patterns and near field
- Calculation of antenna and scattering parameters including directivity, efficiency and RCS
- Current and near field animation

IE3D Simulation Examples and Display

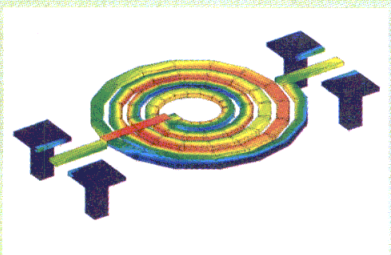
The current distribution on an AMKOR SuperBGA model at 1GHz created by the IE3D simulator



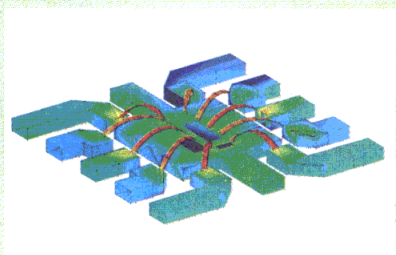
The current distribution and radiation pattern of a handset antenna modeled on IE3D



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

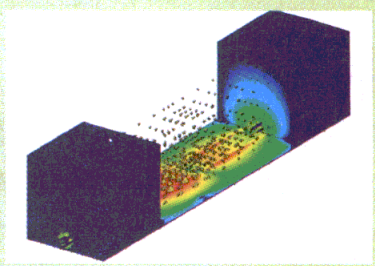


IE3D modeling of an IC Packaging with Leads and Wire Bonds

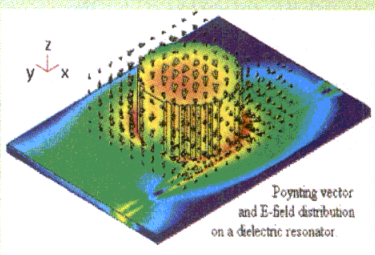


FIDELITY Examples

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



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



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CIRCLE NO. 319

RF Notch Filters Are Programmable

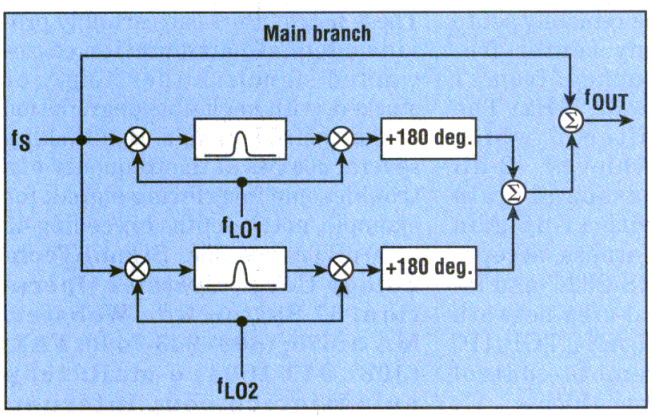
These remote-controllable notch filters offer selectable center frequencies and notch widths to improve the sensitivity of HF and UHF receivers.

Theodore A. Coppola

Signal Technology Corp., Systems Operation,
37 Sutton Rd., Webster, MA 01570; (508) 943-7440, FAX: (508) 949-1804,
e-mail: tcoppola@sigtech.com, Internet: http://www.sigtech.com.

NOTCH filters are ideal for reducing or removing the effects of a strong signal in close proximity to a weaker signal. In communications and test systems, for example, notch filters support the capture of weak signals amidst interfering signals. But notch filters are generally not variable in their tuned frequencies or filter characteristics, with the exception of a line of programmable notch filters from the Systems Operation of Signal Technology Corp. (Webster, MA). These filters are ideal for improving the dynamic range of a receiver and minimizing the effects of receiver desensitization in the presence of high-level signals. Filters are currently available for high-frequency (HF) and ultra-high-frequency (UHF) applications.

Notch filters have traditionally been fixed in frequency and notch characteristics, and usually custom designed according to the needs of a particular signal scheme. In general, notch filters are designed for placement in series with the signal path to remove a single unwanted interference signal. Because of this series configuration, with multiple filters cascaded from input to output, the use of several notch filters to remove multiple unwanted interference signals can cause excessive signal loss and distortion.



1. The programmable notch filters create separate signal paths for desired and undesired signals, eliminating the unwanted signals by means of cancellation techniques.

A conventional multiple-notch filter network consists of a notch filter for each notch frequency, with the filter responses cascaded together. Because of this, incoming signals pass through each of the notch filters, with each notch removing its corresponding signal. Unfortunately, because the passband responses are cascaded (in series), the desired signals are somewhat attenuated and degraded in terms of phase linearity, intermodulation (IM) performance, noise figure, and group delay as they pass through the successive notch filters.

PATENTED TECHNOLOGY

The patented Signal Technology notch filters were developed to provide a means of removing multiple undesired signals from wide operating bandwidths at minimal loss and degradation of desired signals.[1] Rather than filter incoming signals through series filters, the programmable notch filters create separate notch-filter paths for each notch frequency (Fig. 1). Each notch-filter path produces an output spectrum that is equal in magnitude but 180-deg. out of phase (inverted) with respect to the undesired input spectrum. The separate outputs of each notch-filter path are then combined in parallel with the RF input signal to produce and RF output signal with the desired spectra relatively unchanged and unattenuated, with all undesired signal spectra attenuated.

In order to add programmable notch-filter capability to the parallel notch-filter architecture, frequency translation to an intermediate frequency (IF) is employed. Frequency translation is achieved by a basic heterodyne configuration, using mixers

Table 1: Summarizing a programmable HF notch filter

Model number	5301-012B-1/5
Notch center frequency	1 to 32 MHz
Frequency resolution	1 kHz
Notch shape factor	8:1
Notch depth	25 dB
Passband gain	2 dB ±1 dB
VSWR	1.50:1
Output third-order intercept point	+75 dBm
Output second-order intercept point	+35 dBm

Table 2: Summarizing a programmable VHF notch filter

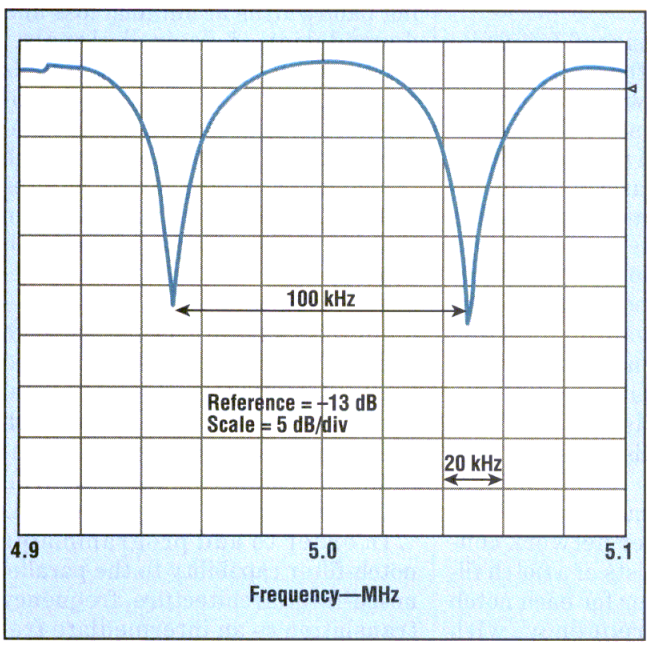
Model number	5550-001N-1
Signal passband	600 to 1200 MHz
Notch center frequency	800 to 1000 MHz
Frequency resolution	6.25 kHz
Notch shape factor	8:1
Notch depth	20 dB
Passband gain	–0.5 dB
VSWR	1.50:1
Output third-order intercept point	+45 dBm

with a tunable local oscillator (LO) to downconvert incoming RF signals. In this way, the center frequency of a bandpass filter can be selected by changing the frequency of the tunable LO. As a result of the heterodyne architecture, fixed bandpass filters in the programmable notch filter are replaced by selectable bandpass filters.

DIGITAL ADJUSTMENTS

To achieve the precise 180-deg. phase balance needed for signal inversion, the filters employ a digitally programmable fine-tuning phase shifter in the IF path. Each bandpass filter section is precharacterized for residual phase errors, and the reciprocals of those errors are stored digitally. These reciprocal values are then applied to the programmable phase shifter to maintain an exact 180-deg. difference between signals to be attenuated and their mirrored versions. A similar approach maintains equal magnitude of mirrored signals through a digitally programmable fine-tuning attenuator.

So far, this patented notch-filter design approach has been applied to two series of programmable filters, for use at HF and UHF. The HF filters include a model with tunable notch center frequencies of 1 to 32 MHz (Table 1) and notch depths of 25 dB (Fig. 2). The filter is designed for insertion in a signal path at any point between the receiver and the antenna. It is designed in a three-space-high (3U) 19-in. (48.26-cm) rack-mount chassis that can accept up to 12 notch-filter modules as well as power-supply and control modules. The tunable-filter modules each have four selectable notch widths—5, 10, 25, and 50 kHz. Each of the tunable-filter modules can be remotely set to any center frequency from 1 to 32 MHz. The filter, which achieves 2-dB passband gain with ±1-dB gain flatness, accepts RS-232C and local-area-network (LAN)[TCP/IP] remote-control connections.



2. These close adjacent notches show a depth of better than 25 dB for forward-transmission measurements made at a center frequency of 5 MHz.

The UHF filters include the model 5550 notch filter for cellular and other applications at notch center frequencies ranging from 800 to 1000 MHz (Table 2). With tunable frequency resolution of 6.25 kHz and a notch-filter depth of 20 dB, this programmable tuner is suitable for enhancing cellular-receiver sensitivity by removing interference from adjacent cell sites. The filter, which is supplied in a 1U 19-in. (48.26-cm) rack-mount enclosure, minimizes passband insertion loss to a mere 0.5 dB. It also achieves impressive dynamic-range performance, with an output third-order intercept point of +45 dBm.

The Signal Technology programmable notch filters do not suffer from the effects of cascading as with traditional notch filters. In fact, the technology allows for the stacking of notches to increase the notch depth. Because of the growing use of digital components, such as analog-to-digital converters (ADCs) and digital signal processors (DSPs) in radio receivers, removal of interfering signals becomes mandatory in modern communications systems, in order to prevent over-ranging of the ADCs. These notch filters can certainly provide adequate attenuation of unwanted signals, either singly or stacked, with negligible degradation of the desired signals. By stacking several notches at the frequency of a troublesome interfering signal, for example, notch depths exceeding 45 to 50 dB are possible. **Signal Technology Corp., Systems Operation, 37 Sutton Rd., Webster, MA 01570; (508) 943-7440, FAX: (508) 949-1804, e-mail: tcoppola@sigtech.com, Internet: http://www.sigtech.com.**

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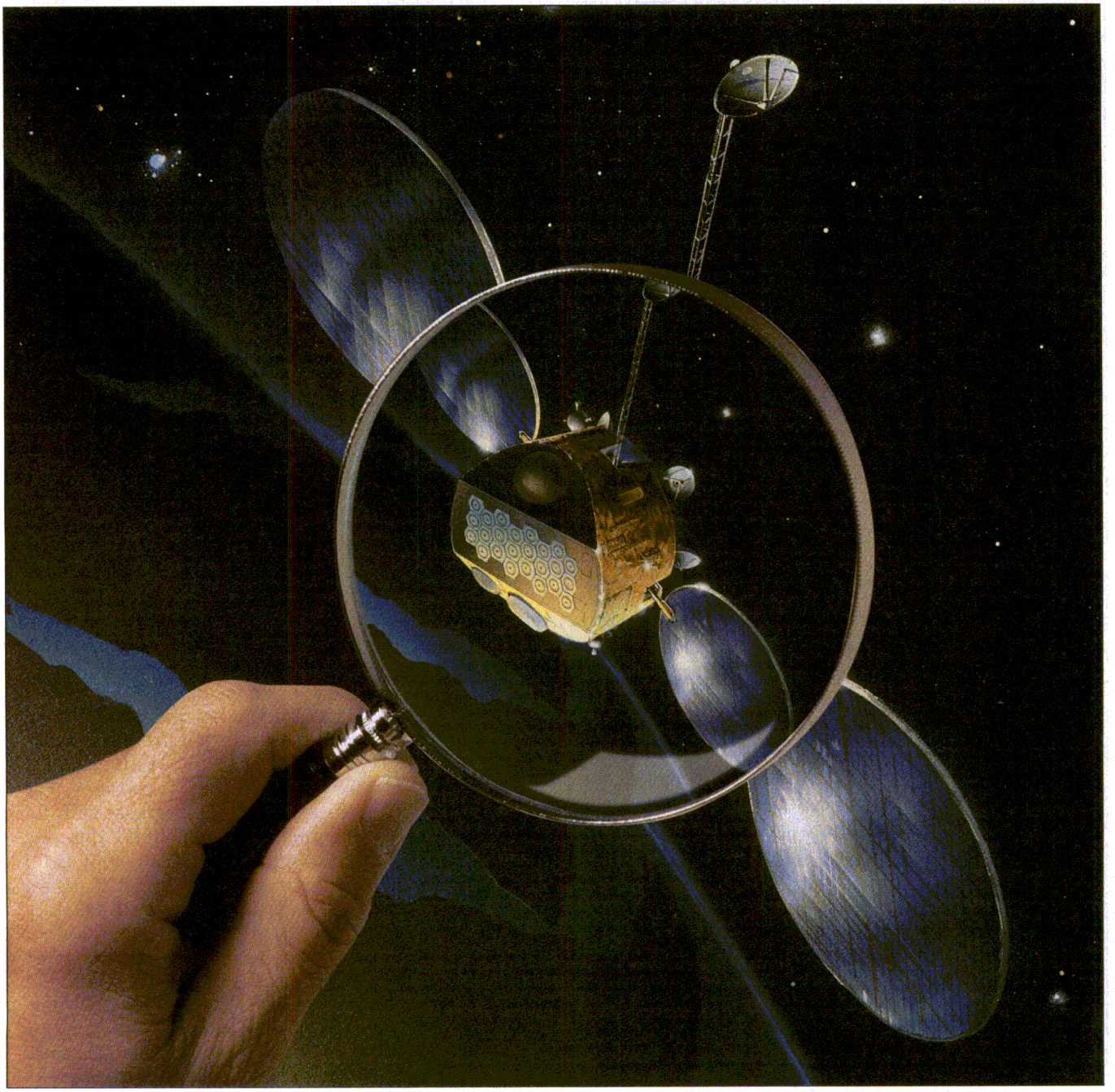
Reference

1. Theodore A. Coppola, "RF Notch Filter Having Multiple Notch and Variable Notch Frequency Characteristics," United States Patent No. 6,020,783, February 1, 2000.



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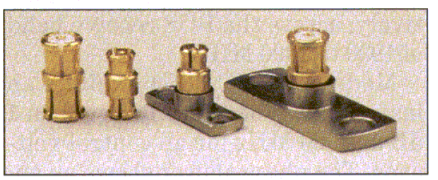

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CIRCLE NO. 213

SMT Phase Shifters Adjust PCS Designs

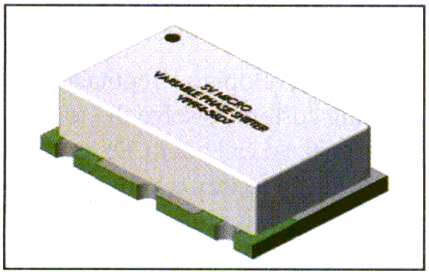
These lines of analog-voltage-variable phase shifters provide linear phase shifts as a function of voltage with low insertion loss and negligible group-delay variations.

JACK BROWNE
Publisher/Editor

PHASE shifters are the tuning elements in electronically steered phase-array antennas. They also make the noise cancellation in feedforward amplifiers possible. For designers in search of true surface-mount-technology (SMT) phase shifters with linear responses, several lines of analog-voltage-variable phase shifters from the Commercial Products Group of SV Microwave cover many of the wireless applications from 10 MHz to 3 GHz. In particular, several models have been developed for analog and digital cellular and personal-communications-services (PCS) applications with phase-shift ranges of 120 and 180 deg. to 2.5 GHz.

The compact phase shifters (see figure) are supplied in a six-lead SMT package measuring 0.60 × 0.38 × 0.20 in. (1.52 × 0.97 × 0.51 cm) They are compact enough for the smallest feedforward and multicarrier power amplifiers (MCPAs), and they operate with direct-current (DC) bias voltage. Phase shifting is accomplished through application of standard 0- to +12-VDC control voltages.

Specific models are available for code-division-multiple-access (CDMA) IS-95 or Narrowband Advanced Mobile Phone Service (NAMPS) transmit (824-to-860-MHz) and receive (869-to-894-MHz) bands, Global System for Mobile Communications (GSM) transmit (880-to-915-MHz) and receive (925-to-960-MHz) bands, and PCS transmit (1850-to-1910-MHz) and receive (1930-to-1990-MHz) bands as well as Institute of Electrical and Electronics Engineers (IEEE) 802.11 wireless local-area-network (WLAN) applications at 2.4 GHz. As an example, models VP192C-36C1 and VP192A-36C1 are voltage-variable phase shifters with minimum phase-shift ranges of 120 and 180 deg., respectively, across the PCS receive band of 1930 to 1990 MHz.



These true surface-mount voltage-variable phase shifters are suitable for amplifier and antenna applications in cellular and PCS bands.

Model VP192C-36C1 exhibits an insertion loss of typically less than 1 dB at 1930 MHz for all control voltages. The phase shifter shows an approximate straight-line relationship between control voltage and phase shift through phase shifts of about 120 deg. This phase shifter achieves a return loss of better than 18 dB and typically 20 dB for all phase-control voltages.

Model VP192A-36C1, with a minimum phase-shift range of 180 deg., can actually achieve phase shifts of 200 deg. or more, with degraded tuning linearity. This phase shifter achieves an insertion loss of better than 1.5 dB and typically less than 1.2 dB, with a return loss of better than 23 dB and typically 25 dB or better for all phase-control voltages. The loss variation versus phase shift for the model VP192C-36C1 120-deg. phase shifter is typically only ± 0.1 dB, and reaches only ± 0.25 dB for the 180-deg. model VP192A-36C1.

For these and the company's other SMT phase shifters, group-delay variations as a function of phase shift are almost negligible. With 0-deg. phase shift as the reference point (0-ps group-delay variation), total group-delay variations are approximately 500 ps for the entire phase-shift range.

Additional phase shifters are available at frequencies from 10 MHz to 3 GHz in bandwidths as wide as 20 percent. The firm also offers a line of low-loss voltage-variable attenuators (VVAs) in the same packages and frequency ranges as the phase shifters, with IP3 performance approaching +50 dBm. **SV Microwave, Commercial Products Group, 7247 Bryan Dairy Rd., Largo, FL 33777; (727) 541-5800, FAX: (727) 541-5869, e-mail: salescpg@svmicro.com, Internet: http://www.svmicrowave.com.**

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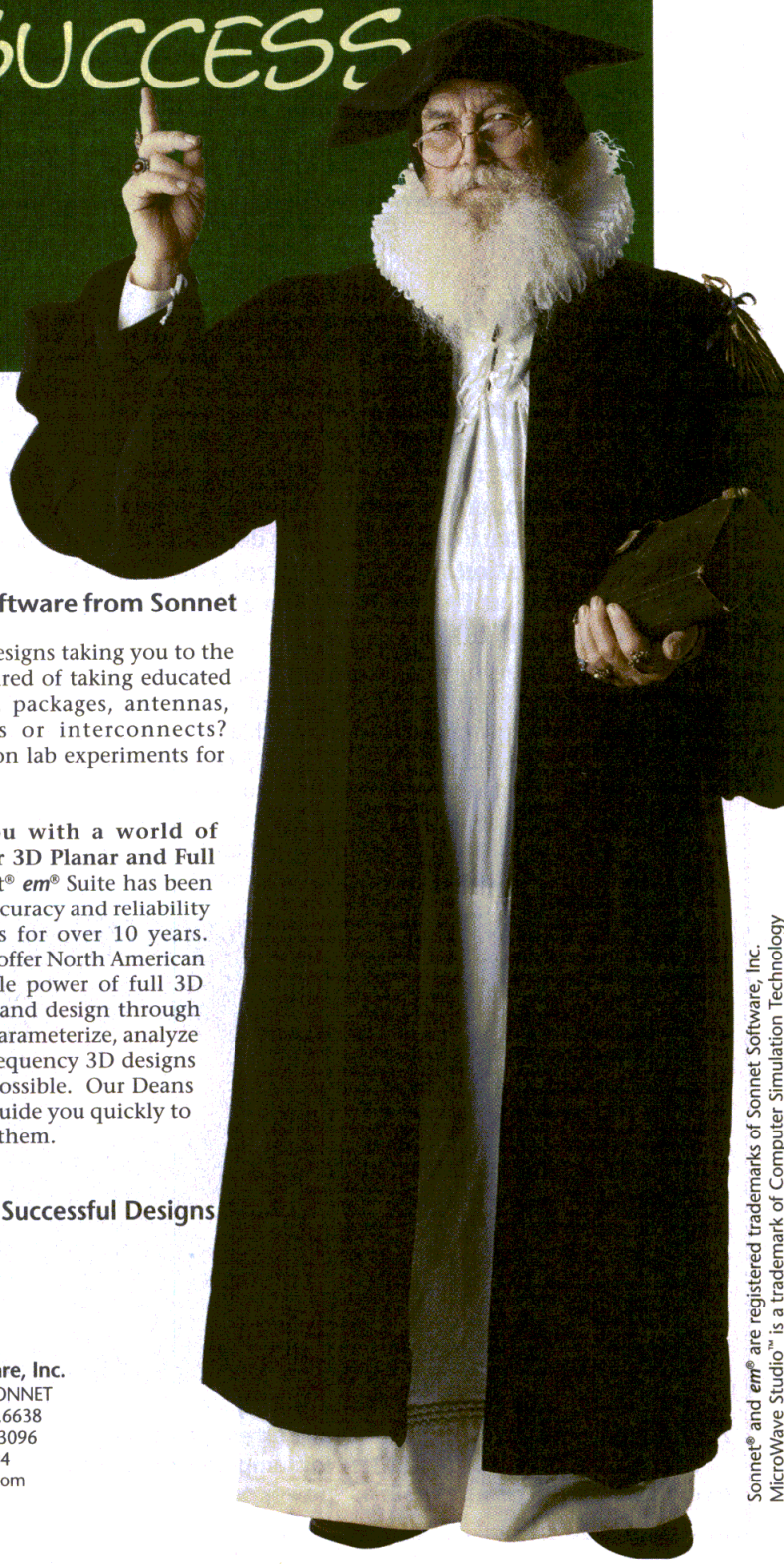
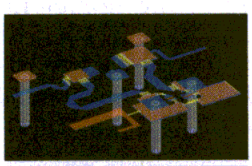
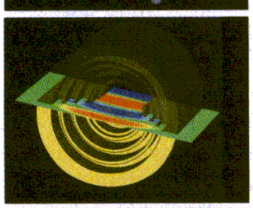
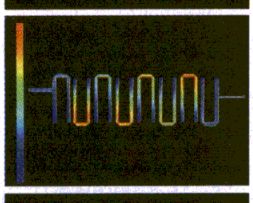
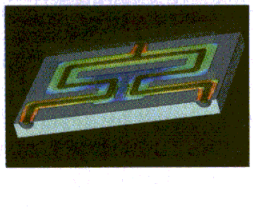


Smart Solutions.









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Detector/Controller IC

Stable IC Controls Amplifier Power

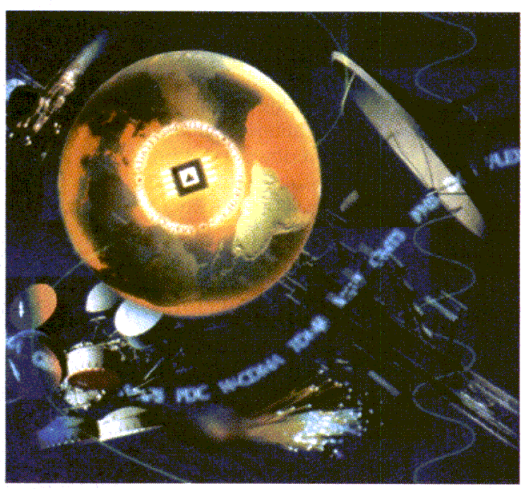
Supporting all current and emerging cellular and PCS standards, this device provides amplifier power control with ±0.5-dB accuracy.

JACK BROWNE
Publisher/Editor

TRANSMITTER designers must deal with swings in output power due to temperature and power-supply variations. Traditionally, power monitoring and control circuitry for transmitter amplifiers consists of diode-based detectors. But with the introduction of the AD8314 RF detector/controller integrated circuit (IC) from Analog Devices (Wilmington, MA), transmitter and amplifier designers can now reach for a single packaged solution for power-control problems. The AD8314 is designed to operate at frequencies from 100 to 2500 MHz with a typical dynamic range of 45 dB.

The AD8314 (see figure) builds upon the company's earlier AD8313 and AD8361 true root-mean-square (RMS) detector products (see *Microwaves & RF*, September 1999, p. 140). It is essentially a logarithmic amplifier, using multiple amplifier/limiter cells and passive detector stages to achieve the wide dynamic range. The amplifier/limiters each contribute approximately 10-dB gain and 3.5-GHz bandwidth. A full-wave rectifier is positioned at the output of each amplifier stage, serving as a square-law detector that converts the RF signal voltages to a fluctuating current with average value that increases with increasing signal level. In addition to four amplifier stages, five passive detector stages, separated by 10 dB each, span a dynamic range of approximately 50 dB. The output of the detector cells is in the form of differential currents, which can be readily summed.

The AD8314 remains stable over operating temperatures from –30 to +85°C. It responds to input signal voltages of 1.25 to 224 mV, which is equivalent to power levels of –45 to 0 dBm at 50 Ω. The AD8314 provides two voltage outputs. The first, which rises from close to ground to approximately +1.2 VDC for input signals from 1.25 to 224 mV, is intended for measurement mode. The second voltage output starts at +2.25 VDC for minimum-level input signals and drops close to ground at maximum input levels. The second voltage output is intended for analog control-loop applications.



The AD8314 RF detector/controller supports all current and emerging cellular standards for temperature-stable amplifier output-power control.

The transfer function of the AD8314 is characterized in terms of slope and intercept point. The log slope refers to the change in received-signal-strength-indication (RSSI) output voltage for a 1-dB change at the output. Nominally a slope of 21.5 mV/dB, a change at the input of 10 dB will result in a change at the output of approximately 215 mV. The 10-to-90-percent response time for a 10-dB input step is typically 70 ns. The full-scale (within 95-percent) settling time for an input step from –40 to 0 dBm is typically 150 ns.

The RF detector/controller is supplied in an eight-pin small-outline IC (SOIC) package. It is a single-package, temperature-stable alternative to discrete diode detectors that are designed to operate with supply voltages of +2.7 to +5.5 VDC. It draws typical quiescent current of 4.5 mA and only 40-μA current when disabled. P&A: $3.95 (1000 qty.); stock. **Analog Devices, Inc., Ray Stata Technology Center, 804 Woburn St., Wilmington, MA 01887; (800) ANALOGD, (781) 937-1428, FAX: (781) 937-1021, Internet: http://www.analog.com.**

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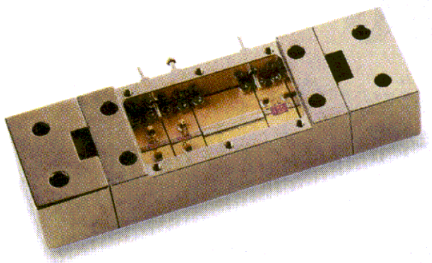

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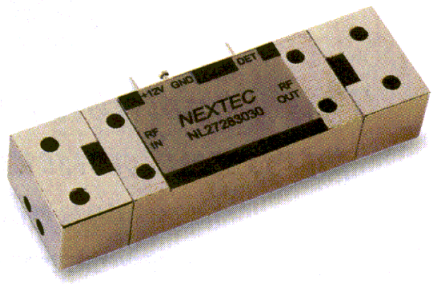

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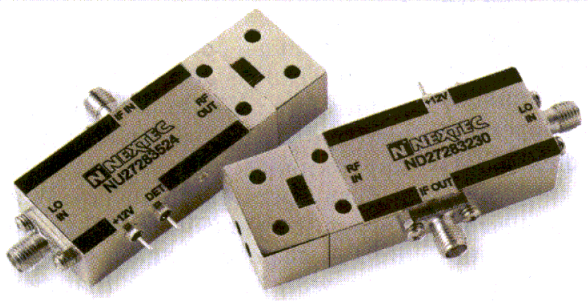

From 24 GHz to 30 GHz for LMDS and B-WLL

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- Small size/low cost

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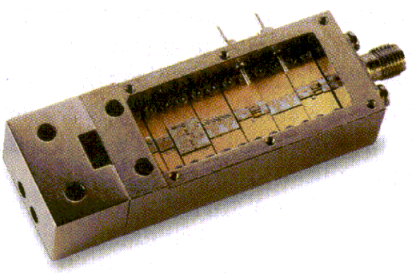

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CIRCLE NO. 233

VHF/UHF Transceiver

Transceiver Chip Targets PWC

This monolithic transceiver requires few external parts and conserves power for personal-wireless-connectivity applications.

DON KELLER
Senior Editor

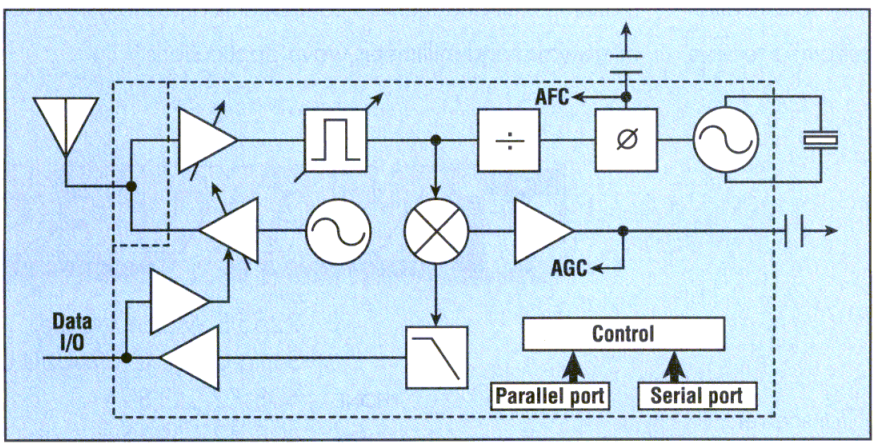
DATA transceivers are a necessity for personal-wireless-connectivity (PWC) devices such as two-way radios and pagers, and wireless personal-computer (PC) peripheral devices such as mobile computers and personal digital assistants (PDAs). The demand for these devices is on the rise, and market pressures are forcing original equipment manufacturers (OEMs) to shrink device size and extend battery life. To meet these requirements, Philsar Semiconductor, Inc. (Ontario, Canada) has introduced a single-chip, multipurpose data transceiver that operates in the 220-to-870- and 902-to-928-MHz industrial-scientific-medical (ISM) frequency bands.

The model PT800 MuRF data-transceiver chip boasts a low external-parts count and ultra-low power consumption. The transceiver operates in the half-duplex mode and uses amplitude-shift-key (ASK) or on-off-key (OOK) data-modulation/demodulation schemes. It is designed for use in low-power, short-range, wireless data-communications devices with data rates ranging from 1 to 10 kb/s.

The PT800 MuRF chip incorporates numerous functional components (see figure) which reduce the external-parts count in the final end-user device. The receiver section includes a two-gain-level, low-noise amplifier (LNA), a variable-gain, an active RF filter, a downconverter, and an active baseband filter. The transmitter section includes a voltage-controlled oscillator (VCO) and a gain-controlled power amplifier (PA). Both transmitter and receiver sections use a common data interface, baseband buffer, and phase-locked loop (PLL) synthesizer. The chip's serial and parallel control ports allow programmable power level, tuning, and control.

The nominal operating frequency of the PT800 is determined by the value of an off-chip inductor. The chip's programmable tuning range is approximately 10 percent of the nominal operating frequency. When transmitting at 870 MHz, the chip's RF PA can put out as much as 0-dBm RF power. At 315 MHz, that maximum figure jumps to +5 dBm. The amplifier can be programmed to operate at one of four power-attenuation levels: 0, −12, −24, and −36 dB. The receiver section has a typical noise figure of 6 dB and an average sensitivity of −107 dBm at 1-kb/s data rate.

The PT800 requires a single-voltage power supply operating at +2.7 to +3.3 VDC. In the receive mode, the chip typically draws 2.8 to 3.8 mA. In the 0-dBm transmit mode, the chip typically draws 12.5-mA current. For the −36-dBm mode, the typical current drain is 3.75 mA. When the chip is in power-down mode, current drain is typically 0.8 μA. **Philsar Semiconductor, Inc., 146 Colonade Road S., Nepean, Ontario K2E 7Y1, Canada; (613) 274-0922, FAX: (613) 274-0915, Internet: http://www.philsar.com.**



This block diagram shows the internal functional components of the MuRF.

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Instant Programming Customizes Oscillators

A unique combination of technologies enables an oscillator to be programmed for a specific frequency in seconds instead of weeks.

GENE HEFTMAN
Senior Editor

GETTING a prototype design or system up and running can be frustrating if it takes the usual few weeks to obtain the crystal oscillator that establishes the basic system timing. That scenario could be a thing of the past based on a concept called field instantly-programmable oscillators (FIPO) by its inventor, Cardinal Components, Inc. (Wayne, NJ). A FIPO is a specially fabricated oscillator that can be programmed within seconds using either of the company's two new field oscillator programmers—the PG-2000P (stand alone) or PG-2000 (see figure).

The FIPO system can be a boon to designers faced with unexpected design problems, timing modifications, a sudden ramp up in production, or other unforeseen situations. The company provides oscillator blanks, which are field-programmable devices available in a variety of package styles, supply voltages, stability ratings, and operating temperatures. These blanks—called the series CPP—can be programmed by the programmer instruments to any frequency from 1.0 to 133 MHz in 1.0-MHz increments. Oscillator blanks come in the full-size 14-pin dual-in-line package (DIP), the 8-pin DIP, ceramic 5 × 7-mm surface-mount device (SMD), and the plastic SMD. Only four pins are used in each package. Stability ratings range from 35 to 75 PPM, and the operating temperature ranges cover commercial (0 to 70°C) through industrial (−40 to 85°C). After programming the desired frequency, the device is a sealed, custom oscillator.

Programming with the PG-2000P is extremely simple since it is a self-contained, portable unit that operates without an external personal computer (PC) or frequency counter. To program a four-pin FIPO requires no special programming skills to set the desired frequency, and the oscillator is programmed within seconds. A key feature of the system is that once an oscillator is programmed, it can be programmed to a different frequency a second time (i.e., each oscillator can be programmed twice). This allows a designer to correct errors or reuse the same device for separate applications that have different frequency and/or output characteristics.

The PG-2000P incorporates a four-line liquid-crystal display (LCD) that shows the programmed frequency, control parameters, usage instructions, and error messages. The instrument programs over a frequency range of 1.0 to 133 MHz in 1.0-MHz increments and permits oscillators to be programmed for +3.0-, +3.3-, or +5.0-VDC supply voltages.

The PG-2000 instrument also programs four-pin CPP oscillators but it requires a PC with an Institute of Electrical and Electronics Engineers (IEEE) 488-interface and an external frequency counter as a reference. It is compatible with Windows 95/98/NT and Windows™ 2000.

The PG-2000P is available now and is priced at $1799. The PG-2000 will be available in July and will be priced at $499. Oscillator blanks are priced at approximately $1.80 (1000 qty.) depending on temperature range and stability specifications. **Cardinal Components, Inc., 155 Route 46 West, Wayne, NJ 07470; (973) 785-1333, FAX: (973) 785-0053, e-mail: cardinal@cardinalxtal.com, Internet: http://www.cardinalxtal.com.**

These field-oscillator-programmer instruments can program special oscillator devices over a wide frequency range in only seconds.

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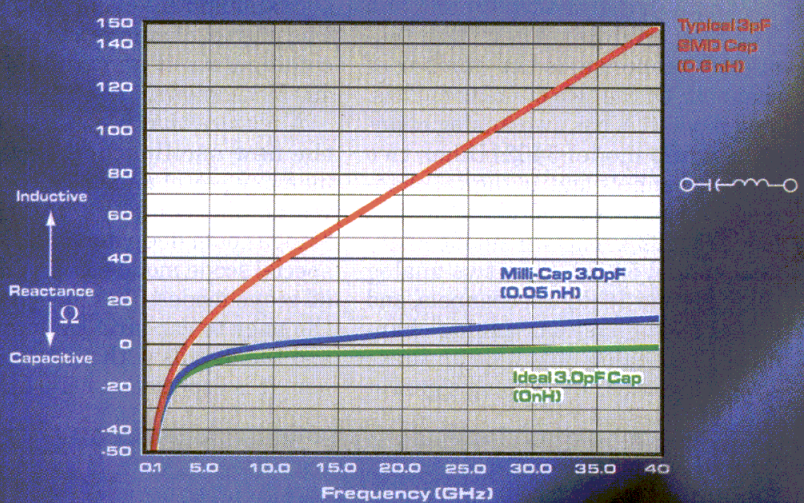

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CIRCLE NO. 235

Scopes Test Mixed Analog And Digital Designs

A series of oscilloscopes with upgraded channels, memory, display, and triggering helps engineers debug mixed analog and digital designs.

DON KELLER
Senior Editor

MIXED analog and digital design and devices, especially digital processors embedded in analog environments, have experienced sharp growth in recent years. This growth, in turn, has increased the demand for instruments that facilitate the testing of signals in both domains simultaneously. In response to this demand, Agilent Technologies, Inc. (Santa Clara, CA) has upgraded the HP 54600 series of oscilloscopes to include a number of new features that enhance their performance in such applications. The five new scopes include two 60-MHz models and three 100-MHz models. They build on the strengths of their six predecessors, but they are sufficiently different to be considered a new class of scopes (see table).

The main difference among the five new models, other than frequency response, is the number of channels. Of the two 60-MHz models, model 54621A has two analog channels, while model 54621D has two analog channels and 16 digital channels. Of the three 100-MHz models, model 54622A has two analog channels, model 54622D has two analog channels and 16 digital channels, and model 54624 has four analog channels. The analog channels on these scopes digitize the incoming signal at a maximum rate of 200 MSamples/s. The digital channels sample the incoming data at a maximum rate of 400 MSamples/s.

A feature that is common to all of the new oscilloscopes is 2 MB of memory per channel. This memory, which the company refers to as "MegaZoom deep memory," is not a special scope mode—it is available in all scope modes. This memory allows the user to capture the big picture and the details within with one trigger. It is especially useful for uncovering buried details in complex signals, discovering anomalies in the absence of good triggering events, and correlating high-speed, digital-control signals with slower analog signals.

The display on the new 54600 series of oscilloscopes has also been upgraded. The screen is the same size (7 in. diagonally) and it displays the same number of vertical points (256) as its predecessors, but the number of displayable horizontal points has been doubled from 500 to 1000, and the number of shades of gray has been increased from two to 32. The result is much clearer, sharper, higher-definition images of complex waveforms, similar to those seen on purely analog scopes.

The new scopes offer flexible triggering capabilities, including edge, pulse width, pattern, sequence, TV, inter-integrated-circuit bus (I^2C), and duration. These triggering modes can help designers isolate and analyze complex signals and conditions that are common to mixed analog and digital designs.

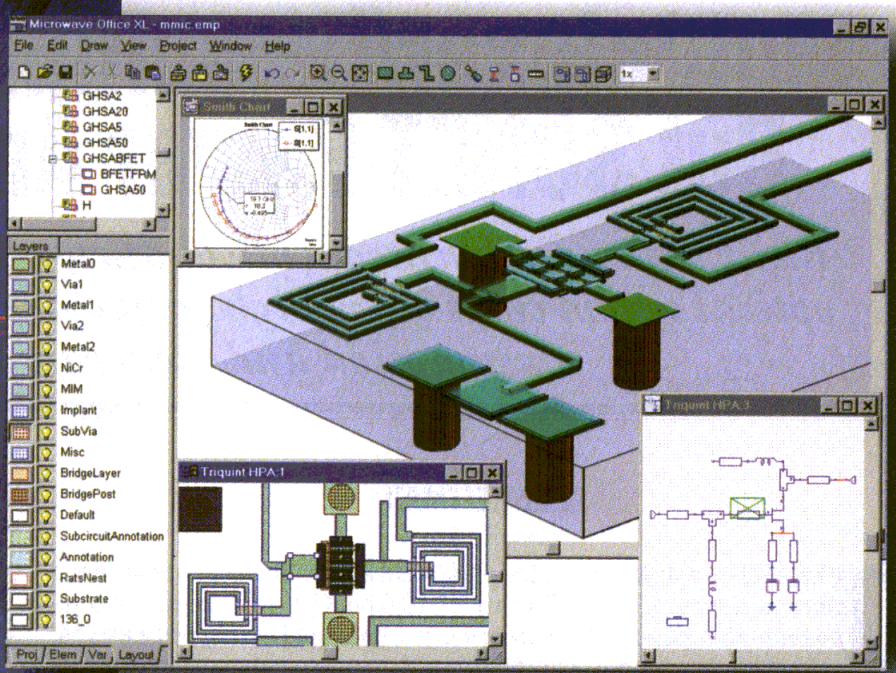
The new scopes also include a built-in floppy-disk drive, RS-232 as well as parallel interface ports, math and Fast Fourier Transform (FFT) functions, along with Quick Help in nine languages. **Agilent Technologies, Inc., Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844, Internet: http://www.agilent.com.**

CIRCLE NO. 56 or visit www.mwrf.com

The upgraded 54600 series scopes include five models

Model	Bandwidth (MHz)	Channels	Sample rate (MSamples/s)	Memory (MB)
54621A	60	2	200	2
54621D	60	2 + 16	200	2
54622A	100	2	200	2
54622D	100	2 + 16	200	2
54624A	100	4	200	2

Microwave Office™ 2000



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Multiplexer/Demultiplexer ICs Cut SONET Power And Cost

Built with an advanced semiconductor process, a pair of optical system chips offers high performance at low power consumption and moderate cost.

GENE HEFTMAN

Senior Editor

OPERATING a data-transmission system at the highest Synchronous Optical Network (SONET) rate of 10 Gb/s (OC-192) requires components that can meet the necessary speed and jitter specifications, yet be stingy on power consumption. A new two-chip multiplexer/demultiplexer combination from Vitesse Semiconductor Corp. (Camarillo, CA) is designed for this environment and is the first device manufactured using the company's high-integration digital gallium-arsenide (H-GaAs-V) process on its 6-in. fabrication line.

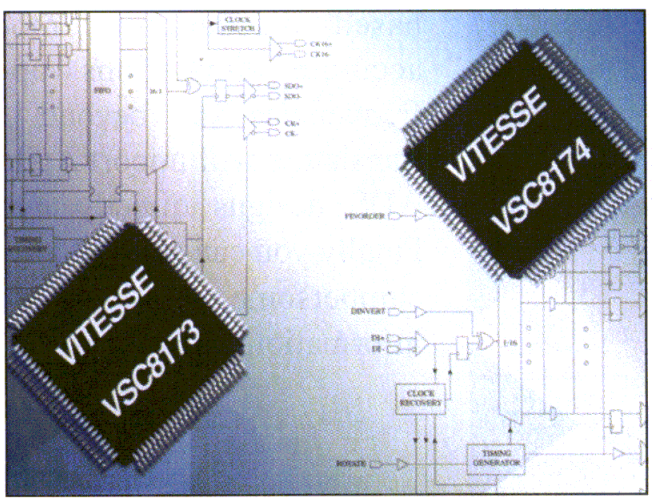
The VSC8173 is a 16:1 multiplexer with an on-chip clock multiplication unit (CMU) and a high-speed clock output to re-clock data at the laser driver. This clock output can be used to verify jitter performance during operation, ensuring SONET compliance. A complete voltage-controlled-oscillator/phase-locked-loop (VCO/PLL) circuit is also included on-chip to generate the required timing for multiplexing operation. The companion integrated circuit (IC) is the VSC8174 1:16 demultiplexer that incorporates the clock- and data-recovery functions. This device is specified for low 50-mV input data sensitivity that can decrease system cost by increasing the regenerator response due to the improvement in low-input signal response. Similar to the 8173, it has a VCO/PLL to generate a clock that remains phase-locked to the incoming data.

The 8173 is housed in an industry-standard 14 × 14-mm 80-pin plastic quad flat pack (PQFP) [see figure] or a 96-pin ball-grid array (BGA). The 8174 is available in a 100-pin PQFP or a 96-pin BGA. Typical power dissipation for each is 1.5 W when operating from a single +3.3-VDC supply. Operating at lower power reduces heat-sink, airflow, and supply requirements. The company claims the chip set runs at 30-percent less power than competing devices.

Both devices run at the standard SONET line rates of 9.953 and 10.66 Gb/s, supporting the use of forward-error-correction (FEC)/parity-check schemes in data transmission without changing devices or boards. Both also incorporate bit-order swap that allows a designer to interface to a device with either an increasing or decreasing bit order without crossing signal lines. This also enables quick recovery from circuit-design errors. For additional flexibility in system design, the chip set provides data-polarity inversion. Jitter performance exceeds all relevant specifications of Bellcore and the International Telecommunications Union (ITU). Both devices are fully compliant with SONET/SDH specifications.

The VSC8173/8174 are aimed at a variety of applications including telecommunications, dense wavelength-division multiplexing (DWDM), undersea transmission, and test equipment.

Samples of both ICs will be available next month with full production scheduled for August. P&A: $500 for the two-chip set (1000 qty.). **Vitesse Semiconductor Corp., 741 Calle Plano, Camarillo, CA 93012; (805) 388-3700, FAX: (805) 987-5896, Internet: www.vitesse.com.**



A two-chip multiplexer/demultiplexer set, the VSC8173/8174 offers high performance and integration for SONET/SDH systems.

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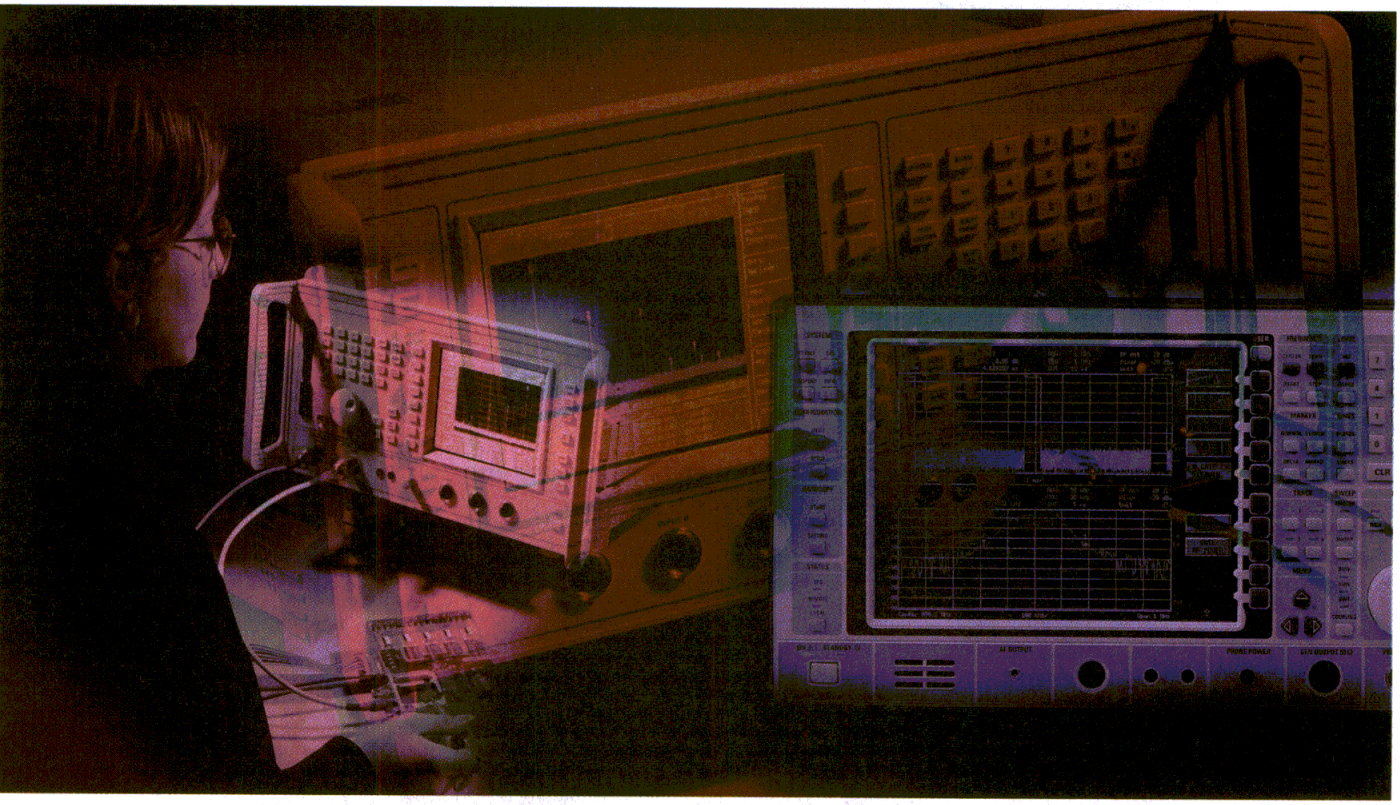

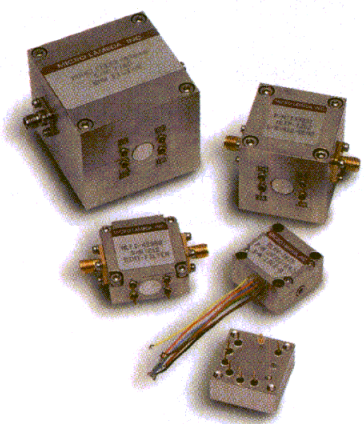

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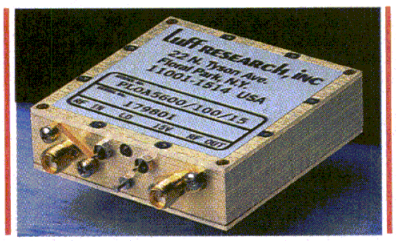

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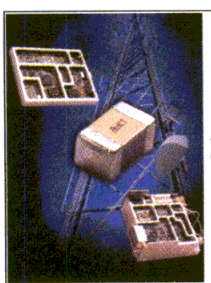

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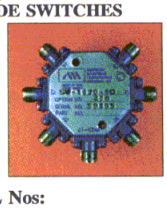

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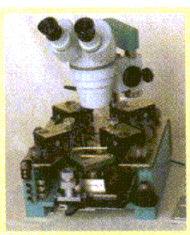

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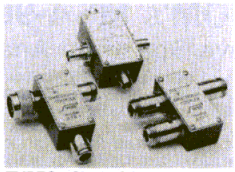

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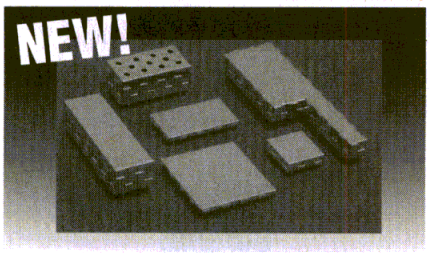

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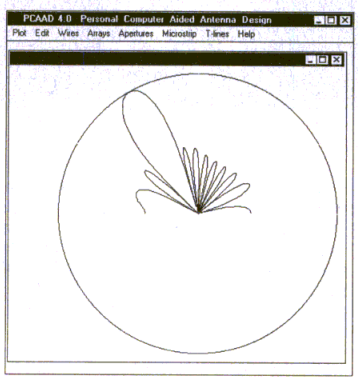

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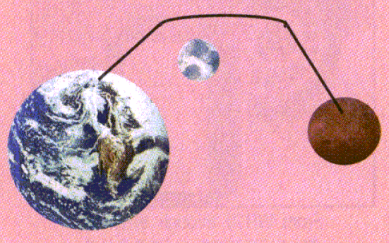

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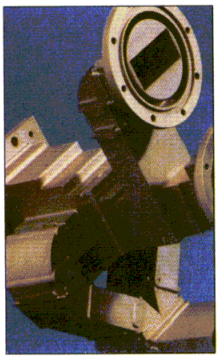

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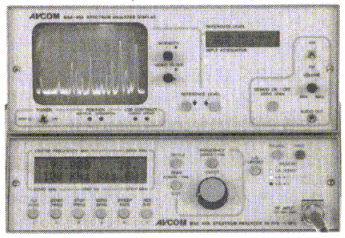

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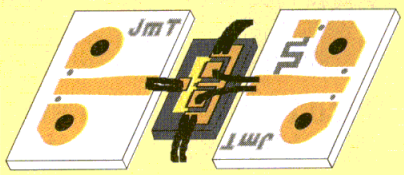

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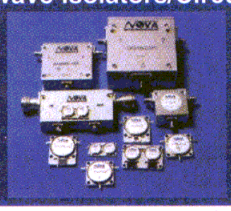

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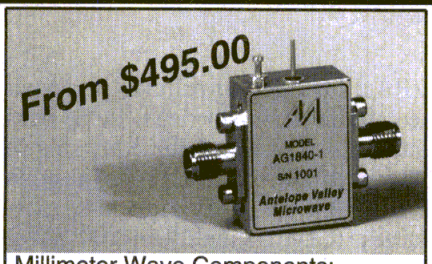

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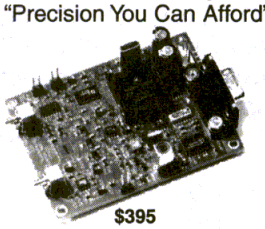

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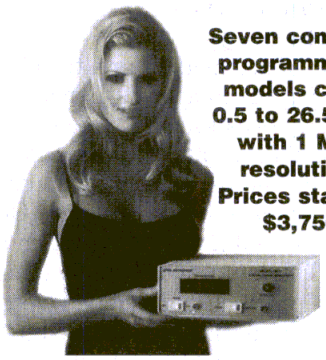

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A new family of optical transceivers operates at 2.125 Gbaud/s, double the rate of conventional 1.0625-Gbaud/s transceivers, for fibre channel applications. The transceivers are available in four package styles. The first is a 1 × 9-channel transceiver housed in a wave-solderable, 9.8-mm-high package compatible with mezzanine-card systems. The second is a hot-pluggable gigabit interface converter (GBIC) for SC fiber-optic connector. The third is a small-form-factor (SFF) transceiver for an LC fiber-optic connector. The fourth is a media-interface-adapter (MIA) extender. **Methode Electronics, Inc., 7444 W. Wilson Ave., Chicago, IL 60706; (708) 867-9600, FAX: (708) 867-9130, Internet: http://www.methode.com/2gfc.**

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Tiny VCXO saves space

The model VS-9000 voltage-controlled crystal oscillator (VCXO) is housed in a miniature ceramic-metal, surface-mount-technology (SMT) package measuring 5 × 7 × 2 mm. The oscillator is available at operating frequencies from 1 to 70 MHz. It is designed for use in wireless handsets and other high-density, size-sensitive applications. Modulation bandwidth is 5 kHz or better, and linearity ranges from ± 5 to ± 20 percent, and output jitter is 50 ps peak-to-peak. Its input impedance is at least 10 kΩ and it can drive a 15-pF load. The oscillator draws less than 50 mA from a +3.3-VDC power source, and less than 35 mA from a +5.0-VDC source. **Raltron Electronics Corp., 10651 NW 19th St., Miami, FL 33172; (305) 593-6033, FAX: (305) 594-3973, Internet: http://www.raltron.com.**

CIRCLE NO. 89 or visit www.mwrf.com

Base-station amp spans 1.7 to 2.5 GHz

The model STX-289 gallium-arsenide heterojunction-bipolar-transistor (GaAs HBT) amplifier operates from 1.7 to 2.5 GHz and is ideal for use as a driver stage for power amplifiers (PAs) installed in cellular personal-communication-services (PCS) base stations. The amplifier conforms to time-division-multiple-access (TDMA), code-division-multiple-access (CDMA), and PCS-1900 modulation schemes. It offers a typical linearity of +40 dBm at 1.9 GHz in 50-Ω systems. The device operates from a single +5-VDC power supply and is housed in a plastic SOT-89 package with backside ground. **Stanford Microdevices, Inc., 522 Alanor Ave., Sunnyvale, CA 94086; (800) 746-6642, FAX: (408) 739-0970, Internet: http://www.stanfordmicro.com.**

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Directional-antenna combo serves digital wireless

A directional-antenna package includes a model PMANT-19 parabolic grid-reflector antenna and a model PMPF-1 50-Ω, passive-feed, dipole antenna. The package suits 2.4-GHz digital wireless communication applications such as point-to-point links between a company's campus location and its off-site facilities. The antenna combo operates from 2.3 to 2.5 GHz and affords 19-dBi gain and –12-dB return loss. It operates at temperatures from –45 to +70°C. The complete antenna and feed assembly weighs less than 4 lbs. (1.8 kg) and measures 16.7 × 9.4 × 23.6 in. (42.42 × 23.88 × 59.95 mm). **Pacific Wireless, 2844 Mar Vista Dr., Suite 101, Aptos, CA 95003; (831) 419-5119, FAX: (831) 684-2494, Internet: http://www.pacwireless.com.**

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Attenuators use T^2 technology

A line of attenuators and terminations using the company's T^2 bonding technology boasts increased temperature capabilities and greater tensile strength. The proprietary manufacturing process bonds copper (Cu) thick-film conductors to thick film at 900°C. This process is said to eliminate the possibility of lead movement during installation, reduce material leaching, and increase connection reliability. The direct-bond process also is said to provide the attenuators and terminations with greater tensile strength, greater connection reliability, and more resistance to lead peeling than conventional soldered connections. **Bird Component Products, 10950 72nd St. North, Suite 107, Largo, FL 33777-1527; (727) 547-8826, FAX: (727) 547-0806, Internet: http://www.bird-fla.com.**

CIRCLE NO. 92 or visit www.mwrf.com

Software manages PC IR appliances

The xConnect 1.0 suite of software utilities manages infrared (IR) data connections between IR-enabled personal computers (PCs) funning Windows 98 and other IR-equipped computers and appliances such as personal digital assistants (PDAs), digital cameras, video projectors, mobile phones, and printers. When this appliance is brought within range of the PC's IR port, the software suite automatically detects and connects the appliance and launches the appropriate application. The software's Windows display indicates which applications are associated with which device, and an icon in the Windows "tray" displays the status of the connection. **Calibre, Inc., 1762 Technology Dr., Suite 226, San Jose, CA 95110-1308; (408) 573-3890, FAX: (408) 573-3899, Internet: http://www.calibre-inc.com.**

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Detector/controller IC replaces discrete diode

The model AD8314 RF detector/controller integrated circuit (IC) operates to 2500 MHz and can be used to replace discrete diode detectors to control RF power amplifiers (PAs). The IC boasts a precise, temperature-stable PA control with ±0.5-dB accuracy. It also has a dynamic range of 45 dB, which, the manufacturer says, allows it to support all current and emerging cellular and personal-communications-services (PCS) standards. The chip is housed in an 8-pin, microSOIC package. **Analog Devices, Inc., Ray Stata Technology Center, 804 Woburn St., Wilmington, MA 01887; (800) 262-5643, FAX: (781) 397-1021, Internet: http://www.analog.com.**

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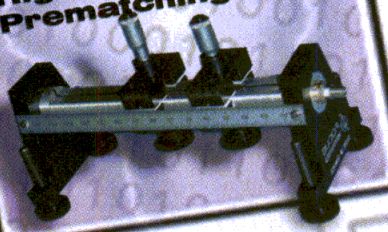

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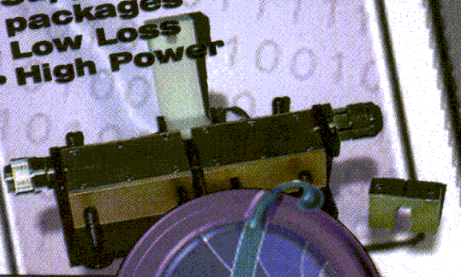

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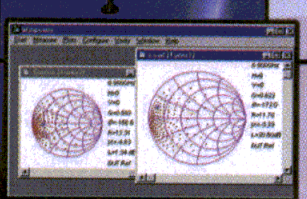

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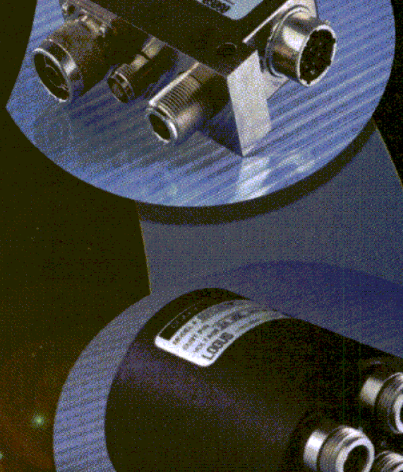

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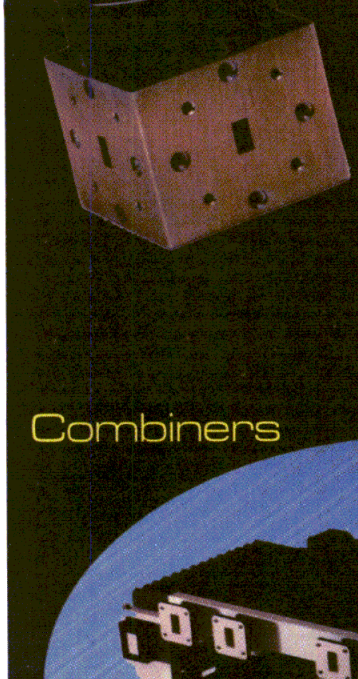

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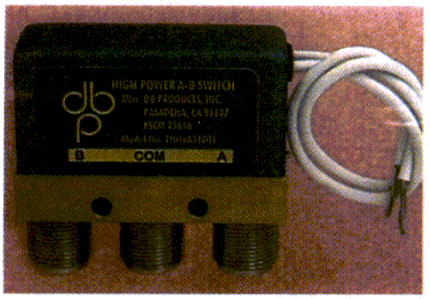

handles continuous-wave (CW) power levels to 1 kW at 1 GHz and 600 W at 3 GHz. Other specifications at 3 GHz include insertion loss of 0.15 dB or less, VSWR of 1.15:1, and minimum isolation of 90 dB. The switch is available at operational voltages of +12, +15, +24, or +28 VDC or other optional DC or AC voltages. **DB Products, Inc., 253 N. Vinedo Ave., Pasadena, CA 91107; (626) 449-3790, FAX: (626) 449-7169, Internet: http://www.dbprod.com.**

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PLLs boast low phase noise

The PE9600 family of phase-locked loops (PLLs) operates to 2.2 GHz and boasts a phase-noise floor of –159 dBc/Hz. The PLLs feature a 10/11 dual-modulus prescaler, counters, a phase comparator, and an on-chip charge pump. The counters are programmable through either a serial or parallel interface, and can also be directly hard wired. The PLLs can be combined with an external prescaler to generate the local-oscillator (LO) signals that are required by satellite-communication transceivers. The PLLs also boast a low power consumption of 60 mW and operate at temperatures ranging from –40 to +85°C. The devices are housed in a 44-pin, ceramic quad-flat-pack, J-lead (CQFJ) package. **Peregrine Semiconductor Corp., 6175 Nancy Ridge Dr., San Diego, CA 92121; (858) 455-0660, FAX: (858) 455-0770, Internet: http://www.peregrine-semi.com.**

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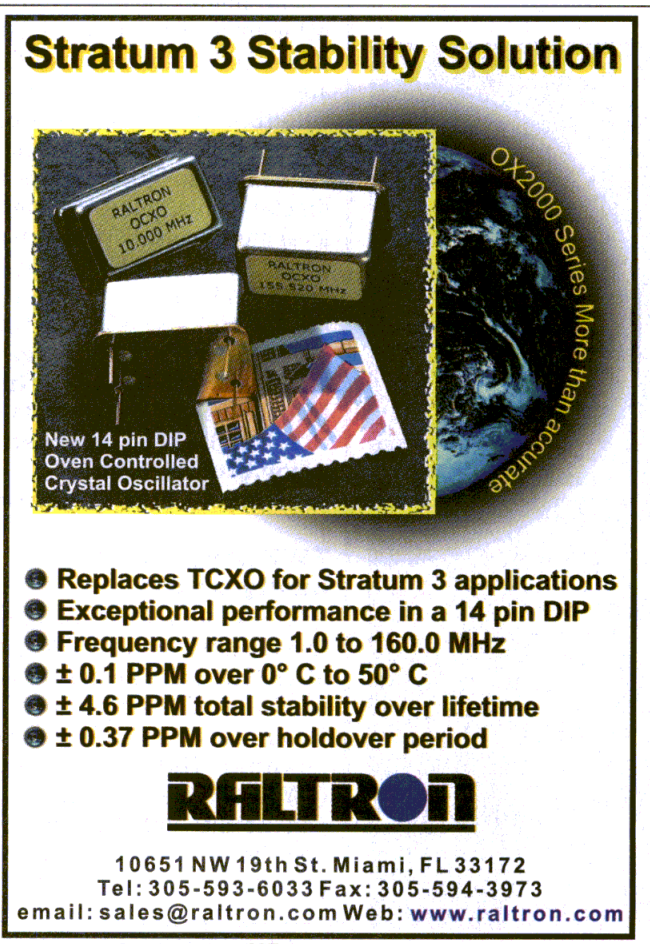

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CIRCLE NO. 451

Microwave relay spans 300 kHz to 2.5 GHz

The model RX microwave relay operates from 300 kHz to 2.5 GHz and offers less than 0.2-dB insertion loss and at least 60-dB isolation. This single-pole, double-throw (SPDT), 1 form C, printed-circuit-board (PCB) mount relay features a maximum VSWR of 1.2:1 and at least 20.8-dB return loss. The RX has a nominal operating power of 200 mW and measures 20.5 × 12.4 × 9.4 mm. Typical applications include hybrid-fiber-coaxial (HFC) systems, personal communications services (PCS), test equipment, local-area-network (LAN) equipment, VME-extension-for-instrumentation (VXI) bus test equipment, and cable modems. **Aromat Corp., 629 Central Ave., New Providence, NJ 07974; (908) 464-3550, FAX: (908) 464-8513, Internet: http://www.aromat.com.**

CIRCLE NO. 97 or visit www.mwrf.com

Distributed amp delivers +12-dBm gain

The model HMC261CB1, monolithic-microwave-integrated-circuit (MMIC) distributed amplifier operates from 20 to 32 GHz and delivers +12-dBm gain at its 1-dB compression point. The amplifier is ideally suited for use in microwave and millimeter-wave, point-to-point radios, local multipoint-distribution systems (LMDS), very-small-aperture terminal (VSAT), and other satellite-communications (SATCOM) applications. **Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, FAX: (978) 250-3373, Internet: http://www.hittite.com.**

CIRCLE NO. 98 or visit www.mwrf.com

TWT uses coupled-cavity circuit

The model ITW-28GC-150WA traveling-wave tube (TWT) operates at 28 GHz and handles power levels to 150 W. The single-beam TWT is ideally suited for use in local multichannel-distribution systems (LMDS), wireless-communications, and satellite-uplink amplifier service applications. The tube features a coupled-cavity circuit, which is said to provide greater reliability than the traditional helix design and is touted as the best technique for high-power, millimeter-wave tubes. EIA WR 28 waveguide and UG 599/U flange transitions provide RF input and output access. **Svetlana Electron Devices, Inc., 8200 S. Memorial Pkwy., Huntsville, AL 35802; (256) 882-1344, FAX: (256) 880-8077, Internet: http://www.istok.com.**

CIRCLE NO. 99 or visit www.mwrf.com

RF IC amp targets handheld systems

The model RF2162 RF integrated-circuit (RFIC) amplifier is designed for use as the final RF amp in dual-mode, +3-VDC, code-division multiple access (CDMA), time-division multiple access (TDMA), and Advanced Mobile Phone Service (AMPS) handheld cellular equipment, and other applications in the 800-to-960-MHz band. The amp offers +29-dBm linear output power, 29-dB gain, and 35-percent efficiency. **RF Micro Devices, 7625 Thorndike Rd., Greensboro, NC 27409-9421; (336) 664-1233, Internet: http://www.rfmd.com.**

CIRCLE NO. 100 or visit www.mwrf.com

Small EMC chamber offers rear access

The GTEM Lite series of self-contained, electromagnetic-compatibility (EMC) test chambers offers an alternative to open-site and anechoic chambers, performing radiated immunity tests to 5 GHz and emissions tests to 2 GHz. The cells are available in three septum-height sizes—350, 550, and 950 mm. The two smaller cells are ideal for testing small, battery-powered devices such as mobile phones and pagers. The chambers offer rear access to the equipment under test, which is said to facilitate test set up and adjustments between tests. **Schaffner-Chase EMC Ltd., Broadwood Test Centre, Rusper Rd. Capel, Dorking, Surrey RH5 5HF, United Kingdom; +44 1306 710205, FAX: +44 1306 713303, Internet: http://www.schaffner.com.**

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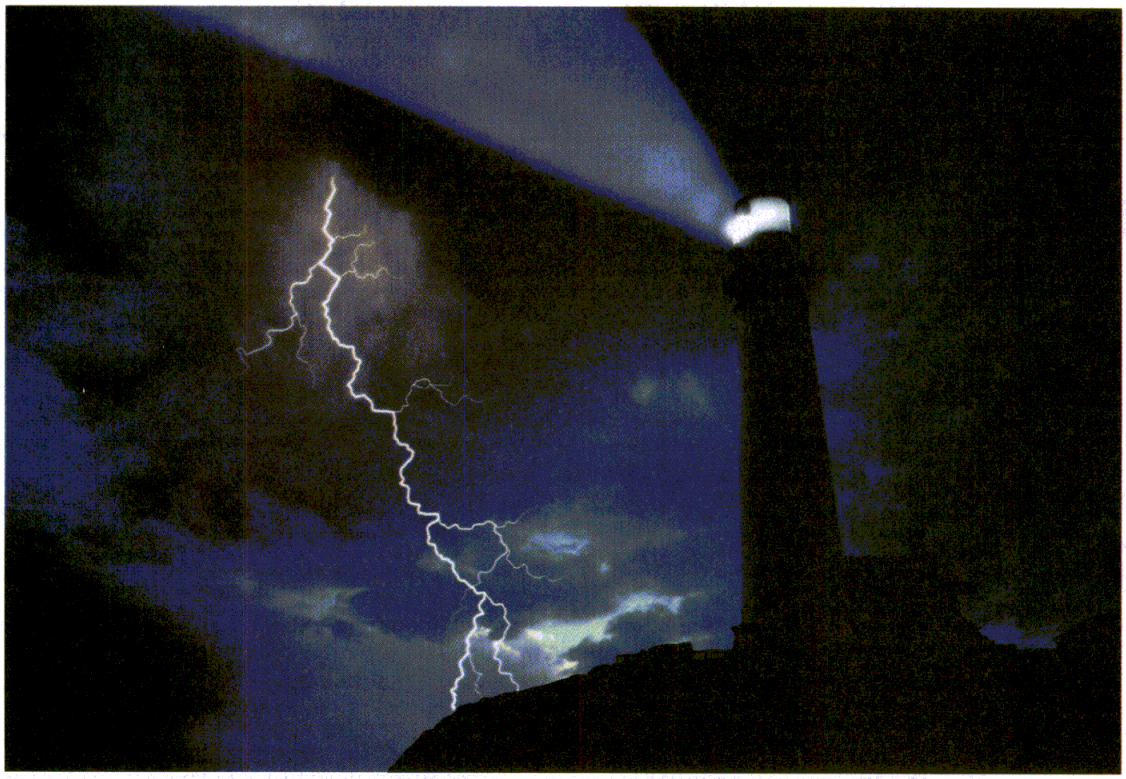

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Stanford Microdevices, Inc. (SMI) is a leading supplier of RF integrated circuits for the wireless and wired telecommunications markets and a supplier of choice of OEMs worldwide. Stanford Microdevices continues to be on the industry's leading edge because of our superior quality, outstanding value and innovative technological advances. SMI provides the tools to create wireless communications equipment that is smaller, lighter, more powerful at market leading prices.

Our SGA 6000 series of silicon germanium MMIC amplifiers offers the high intercept point, high efficiency and high integration level at high output power desired, while providing the low noise figure and low power consumption needed for all wireless applications.

SGA-6386 has 1dB compressed output power of +20dBm, output third-order intercept point of +36dBm and 15.5dB of gain at 900MHz. Pricing on the SGA-6386 is $1.21 in quantities of 10,000 pieces with availability from stock to eight weeks.

SPECIFICATION MATRIX

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

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

SiGe HBT MMIC features include:

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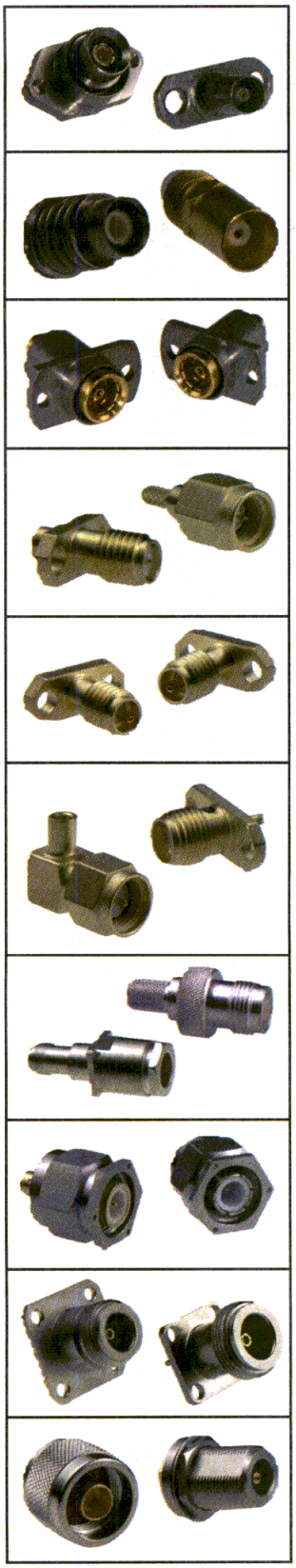

Frequency	Impedance	Cable Type	Coupling	Body & Contact Material	Finish
BMZ Subminiature Blindmate RF Coaxial Connectors					
DC to 18 Ghz	50 Ohms	Semi-Rigid Flexible	Blindmate/ Slide-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: Panel mount receptacles, PC board receptacles, hermetic seal jacks, in and between series adapters.					
BZ Miniature Blindmate RF Coaxial Connectors					
DC to 12.4 Ghz	50 Ohms	Semi-Rigid Flexible	Blindmate/ Slide-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: Panel mount receptacles, PC board receptacles, in and between series adapters.					
BMA Miniature Blindmate RF Coaxial Connectors					
DC to 18 Ghz	50 Ohms	Semi-Rigid Flexible	Blindmate/ Slide-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: In series jack to plug adapters					
SSMA Screw-on Subminiature RF Coaxial Connectors					
DC to 26 Ghz	50 Ohms	Semi-Rigid Flexible	Blindmate/ Slide-on	Stainless Steel Beryllium	Pass/Gold Gold
Also available: Panel mount jacks and plugs (hermetic), bulkhead mount jacks (hermetic), surface launch plugs and jacks, in and betw series adapters.					
HFSSMA Screw-on Subminiature RF Coaxial Connectors					
DC to 40 Ghz	50 Ohms	Semi-Rigid	Screw-on	Stainless Steel Beryllium	Pass/Gold Gold
Also available: Panel mount plugs and jacks, in and between series adapters.					
SMA Screw-on Miniature RF Coaxial Connectors					
DC to 18 Ghz	50 Ohms	Semi-Rigid Flexible	Screw-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: Panel mount jacks and plugs (hermetic and field replaceable), M.I.C. launcher jacks and plugs, bulkhead jacks (herme surface launcher jacks and plugs, end launcher jacks, in and between series adapters, dust caps, flange mount cable receptacles, g seals, 27 GHz versions.					
TNC Small-size Screw-on RF Coaxial Connectors					
DC to 11 Ghz	50 Ohms	Flexible	Screw-on	Brass Beryllium Copper	Silver or Nick Gold
Also available: Panel mount jacks and plugs (hermetically sealed), bulkhead jacks and plugs (hermetically sealed), surface launcher ja and plugs, end launcher jacks and plugs, in and between series adapters, dust caps.					
PTNC Precision Small-size Screw-on RF Coaxial Connectors					
DC to 15 Ghz	50 Ohms	Flexible Semi-Rigid	Screw-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: Panel mount jacks, bulkhead mount jacks, in and between series adapters, dust caps, hermetics.					
TYPE N Large-size RF Coaxial Connectors					
DC to 11 Ghz	50 Ohms	Flexible	Screw-on	Brass Beryllium Copper	Silver or Nick Gold
Also available: Panel mount jacks, surface end launch jacks and plugs, dust caps.					
PN Precision Large-size RF Coaxial Connectors					
DC to 12.4 Ghz	50 Ohms	Flexible Semi-Rigid	Screw-on	Stainless Steel Beryllium Copper	Pass/Gold Gold
Also available: In and between series adapters and 18 Ghz versions.					

Also Available: Cable Assemblies, Attenuators, Terminations and special products.

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SP2T switch boasts high isolation

The model SWN-218-2A-AKG option 011, 120, MIT, single-pole,

double-throw (SPDT) switch offers 100-dB isolation from 0.1 to 20 GHz. Insertion loss is less than 4 dB and VSWR is 2.0:1. It draws 75 mA at +5 VDC and 55 mA at –12 VDC. Turn-on time is less than 65 ns and turn-off time is less than 40 ns. The switch weighs 1.2 oz. and measures 1.2 × 1.0 × 0.5 in. (3.048 × 2.54 × 1.27 cm). **American Microwave Corp., 7311 G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, Internet: http://www.amwave.com.**

CIRCLE NO. 102 or visit www.mwrf.com

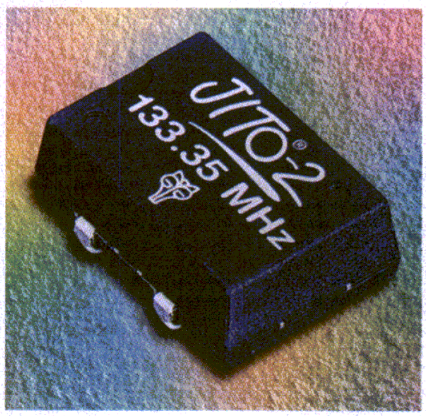
Surface-mount splitter targets CATV-HFC

The model PSCE2-35C2 surface-mount, two-way power splitter operates from 20 to 1000 MHz and is ideally suited for community-access television (CATV) and hybrid fiber/coaxial (HFC) applications. The splitter exhibits a nominal coupling of 3 dB, typical insertion loss of 0.4 dB, and frequency-response flatness of better than ±0.15 dB. Typical amplitude balance is 0.05 dB and typical phase balance is 0.5 deg. across the entire operating bandwidth. Return loss is typically better than 23 dB on all three ports, and typical isolation is 30 dB or better. The splitter is housed in a standard style-5M, J-leaded surface-mount package measuring 0.80 × 0.40 × 0.18 in. (2.032 × 1.016 × 0.4572 cm). **SV Microwave, 7247 Bryan Dairy Rd., Largo, FL 33777; (727) 541-5800, FAX: (727) 541-5869, Internet: http://www.microwave.com.**

CIRCLE NO. 103 or visit www.mwrf.com

Programmable oscillators span 0.34 to 250 MHz

The JITO-2P is a family of plastic-encased oscillator that can be programmed to generate a signal at any frequency from 0.34 to 250 MHz. The oscillators are available with a high-speed-complementary-metal-oxide-semiconductor (HCMOS) output. They are also available with a frequency stability of either ±100 or ±50 PPM, and an operating voltage of either +5.0 or +3.3 VDC. The +5.0-



VDC models draw 15 to 55 mA of input current, and the +3.3-VDC models draw 8 to 30 mA. Operating temperature range is –10 to +70°C. **Fox Electronics, 5570 Enterprise Pkwy., Fort Meyers, FL 33905; (888) 438-2369, FAX: (941) 693-1554, Internet: http://www.foxonline.com.**

CIRCLE NO. 104 or visit www.mwrf.com

Base-station monitor surveys VSWR

The "Bat" base-station monitor verifies antenna and VSWR status of code-division-multiple-access (CDMA), time-division-multiple-access (TDMA), and analog base-station antennas. It can measure as many as three multisector antenna arrays simultaneously. In the event that the VSWR exceeds a prescribed threshold, the monitor dials out via telephone modem alerting the system operator. Up to 16 Bat monitors can be interconnected via an internal network connection. **Berkeley Varitronics Systems, Inc., Liberty Corporate Park, 255 Liberty St., Metuchen, NJ 08840; (732) 548-3737, FAX: (732) 548-3404, Internet: http://www.bvsystems.com.**

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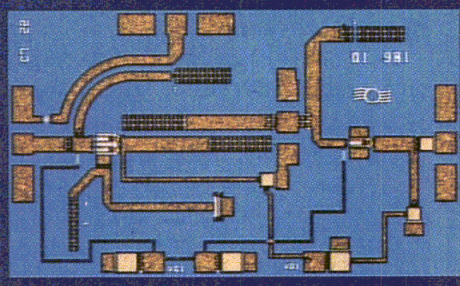
WITH A LITTLE TLC ...

TLC "O" MMIC Chip

The Worlds Most Versatile MMIC Chip



- *2 to 79 GHz Operation*
- *MULTIPLIER, VCO, ILO, DRO, MIXER, AMP*
- *> 16 dBm @ 38 GHz*
- *8 dBm @ 77 GHz*
- *-100 dBc/Hz @ 100 KHz @ 77 GHz*

The TLC "O" chip is a general purpose MMIC oscillator with an output buffer amplifier. The center frequency is voltage and mechanically tunable by plucking airbridges. Versatile option allows for higher Q restoration, injection locking, harmonic multiplication, sub-harmonic mix up and down.

The versatility of the TLCO01981 "O" chip oscillator makes it an excellent candidate for use in radar or communication systems from 2 to 79 GHz.

Patents Pending

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CIRCLE NO. 226

Transformers serve cable-TV equipment

Two surface-mount, wideband RF transformers—models C2020 and C2022—are designed to be used in cable modems, set-top boxes, Internet appliance products, and other high-speed, interactive applications. The model C2020 has an impedance ratio of 1:1, while the 2022's is 1:4. Both transformers operate at upstream frequencies from 5 to 80 MHz with less than 1-dB insertion loss. The six-pin, hermetically sealed transformers operate at temperatures from –40 to +85°C and measure 0.30 × 0.39 in. (7.62 × 9.9 mm). **Pulse Engineering, 12220 World Trade Dr., San Diego, CA 92128; (858) 674-8100, FAX: (858) 674-8262, Internet: http://www.pulseeng.com.**

CIRCLE NO. 106 or visit www.mwrf.com

SPST switch spans 0.5 to 18 GHz

A non-reflective/absorptive, single-pole, single-throw (SPST) switch operates from 0.5 to 18 GHz and handles power levels to 1 W. The switch's rise time can be adjusted in four discrete steps of 25±5, 50±5, 75±5, and 100±10 ns. Insertion loss is less than 1.75 dB at 0.5 GHz and less than 3.5 dB at 18 GHz. Isolation is greater than 80 dB at 0.5 GHz and greater than 85 dB at 18 GHz. Its typical VSWR is 2:1. Switch-on speed ranges from 130 to 210 ns, and switch-off speed ranges from 65 to 100 ns. The switch operates on +5 and –15 VDC and draws less than 45 mA from the +5-VDC supply and less than 40 mA from the –15-VDC supply. **American Microwave Corp., 7311 G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, Internet: http://www.amwave.com.**

CIRCLE NO. 107 or visit www.mwrf.com

Carbon-loaded foam absorbs microwaves

The ECCOSORB AN HR series of lightweight, carbon-loaded foam materials absorbs microwaves for applications such as lining high-performance, shrouded antennas to reduce back radiation, lining antenna caps for testing aircraft surface-mounted antennas, lining antenna housings to improve antenna performance, lining shielded antenna test boxes, reducing RF interference (RFI) around antenna feeds, as well as improving performance of antennas mounted onto ground planes. **Emerson & Cuming Microwave Products, 28 York Ave., Randolph, MA 02368; (800) 650-5740, FAX: (781) 961-2845, Internet: http://www.emersoncumingmp.com.**

CIRCLE NO. 108 or visit www.mwrf.com

OCXOs offer high stability

The OX-2000 series oven-controlled crystal oscillators (OCXOs) are available in frequencies from 1 to 160 MHz and boast a stability of ±0.1 PPM at temperatures from 0 to +50°C and claim to outperform similarly priced temperature-controlled crystal oscillators (TCXOs). The OCXOs meet ANSI Stratum-3 requirements, including ±4.5-PPM total lifetime stability, and ±0.37 PPM over a holdover period that includes variations in temperature, supply voltage, load variation, and 24-h aging. **Raltron Electronics Corp., 10651 NW 19th St., Miami, FL 33172; (305) 593-6033, FAX: (305) 594-3973, Internet: http://www.raltron.com.**

CIRCLE NO. 109 or visit www.mwrf.com

IR emitters boast fast switching

Model TSFF5200 and TSFF5400 gallium-aluminum-arsenide (GaAlAs) infrared (IR) emitters boast high switching speed and low power consumption for IrDA-compliant data applications in cell phones, palmtop computers, laptop computers, multimedia, as well as alarm and safety systems. Typically, the emitters are capable of radiating 40 mW of optical power. Forward voltage is +1.5 VDC, rise and fall times are 10 ns, and maximum modulation bandwidth is 35 MHz. **Vishay Intertechnology, Inc., 63 Lincoln Hwy., Malvern, PA 19355-2120; (610) 644-1300, FAX: (610) 296-0657, Internet: http://www.vishay.de.**

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Introducing the TrueTriangle™

Celeritek's NEW HBT amplifier modules.

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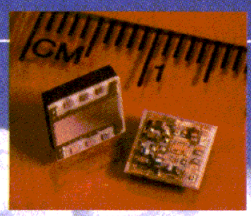

Celeritek's TrueTriangle

	CHP0230 Digital/Analog Cellular	CHP1232 Digital PCS
Frequency	824–849 MHz	1.85–1.91 GHz
Output Power	28.5 dBm	28.0 dBm
Linear Efficiency		
Analog	60%	—
Digital	40%	35%
Quiescent Current	50 mA	50 mA
Gain	30 dB	30 dB

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Isolator withstands severe environments

The model KX323 dual-junction waveguide isolator operates in the X-band and can handle power levels to 200 W average and 10 kW peak, forward and reverse. The isolator is designed to withstand severe military environments. It exhibits a maximum VSWR of 1.15:1 and 55-dB isolation. **Channel Microwave Corp., 480 Constitution Ave., Camarillo, CA 93012; (805) 482-7280, FAX: (805) 987-8794, Internet: http://www.channelmicrowave.com.**

CIRCLE NO. 111 or visit www.mwrf.com

Optical-fiber modules serve short-reach applications

The 623 family of parallel-fiber, optical-interconnect modules serves the emerging requirements for short-reach optical interconnect in large-scale computing and communications systems. The family includes 12-channel, eight-channel, and four-channel transmitters and receivers. At a channel capacity of 2.5 Gb/s, the modules offer 30-, 20-, and 10-Gb/s transmission over fiber lengths to 300 m. **Mitel Semiconductor Corp., 350 Legget Dr., P.O. Box 13089, Kanata, Ontario K2K 2W7, Canada; (613) 592-2122, FAX: (613) 592-4784, Internet: http://www.mitel.com.**

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BERT tests high-speed components

The 86130A BitAlyzer high-speed, bit-error-rate tester (BERT) characterizes, verifies, and tests optoelectronic and high-speed digital components and multiplexer/demultiplexer integrated circuits (ICs). This 3.6-Gb/s BERT can perform multiplexer and demultiplexer testing for telecommunications and system-area-network (SAN) ICs, and multiple transmitter and receiver testing in manufacturing. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844, Internet: ttp://www.agilent.com.**

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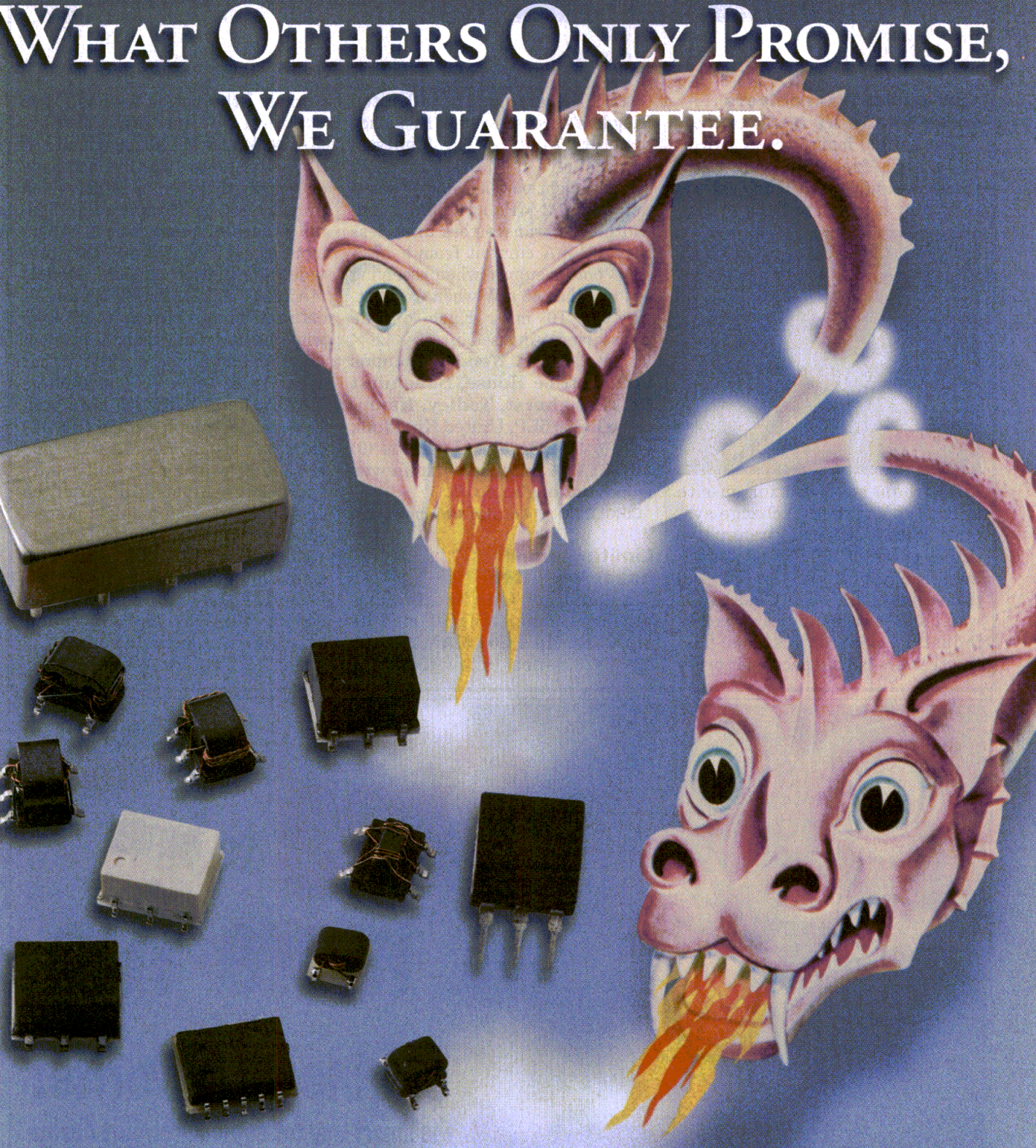

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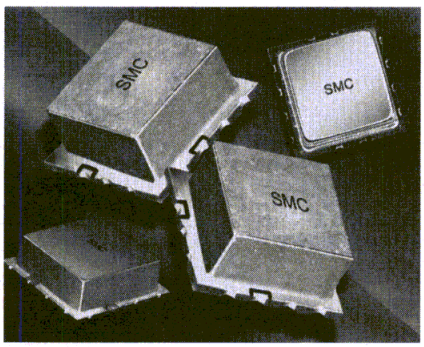
CIRCLE NO. 217

Low-power transceivers serve telemetry and control

The SX 100 family of low-power, very-high-frequency (VHF) transceivers takes advantage of the relatively uncluttered 125-to-180-



MHz frequency band for on-site telemetry and data-exchange systems. With a maximum transmitter output power of 10 mW and a maximum range of 2 km, the SX 100 transceivers are ideally suited for water and sewage utilities as well as industrial control systems. The transceivers can be specified in order to take advantage of the United Kingdom's license-exempt frequency band at 173 MHz. The family of transceivers supports either 112 serial or eight parallel channel selection over a 2-MHz switching bandwidth. The transceivers draw less than 40- mA current from a +7.2-VDC power supply when receiving and less than 65 mW when transmitting at 10 mW. Each of the transceivers is housed in a 60 × 38 × 17-mm package. **Wood & Douglas Ltd., Lattice House, Baughurst Rd., Baughurst, Tadley, Hampshire RG26 5LP, United Kingdom; +44 (0) 118 981 1444, FAX: +44 (0) 118 981 1567, Internet: http://www.woodanddouglas.co.uk.**

CIRCLE NO. 114 or visit www.mwrf.com

Limiting amps boast low jitter

A family of four limiting amplifiers—consisting of models MAX3264, MAX3265, MAX3268, and MAX3269—boasts low jitter for Gigabit Ethernet as well as Fibre Channel applications. Models MAX3264 and MAX3268 operate at data rates up to 1.25 Gb/s and have deterministic jitter of 14 ps, which provides 156-ps margin to the Gigabit Ethernet specifications. At an input signal level of 5 mV, they provide fully limited outputs. Models MAX3265 and MAX3269 operate at data rates that reach up to 2.5 Gb/s while having deterministic jitter of 11 ps and an input sensitivity that consists of 10 mV. The models MAX3264 and MAX3265 have positive-emitter-coupled-logic (PECL) outputs, while models MAX3268 and MAX3269 provide current-mode-logic (CML) outputs. All four of the devices operate on a single power-supply voltage that ranges between +3.3 and +5.0 VDC. **Maxim Integrated Products, 120 San Gabriel Dr., Sunnyvale, CA 94086; (408) 737-7600, Internet: http://www.maxim-ic.com.**

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CIRCLE NO. 215

Enclosure solutions

An 18-page brochure highlights a range of standard, modified, and custom enclosure solutions and packaging-systems capabilities for telecommunications, data-communications, test-and-measurement, computer, broadband, military/aerospace, and other applications. Cabinets and cases, subracks, plug-in units, high-volume stamping/chassis, and outdoor aluminum (Al) enclosures are described. The brochure also includes information on a process that guides customers through everything from preliminary design and engineering to production, testing, and assembly. **Pentair Electronic Packaging;** *(888) 550-9543, e-mail: info@pentair-ep.com, Internet: http://www.pentair-ep.com.*

CIRCLE NO. 63 or visit www.mwrf.com

Frequency-control products

A 284-page data book provides specifications on standard and custom frequency-control devices. Temperature-compensated crystal oscillators (TCXOs), voltage-controlled crystal oscillators (VCXOs), oven-controlled crystal oscillators (OCXOs), and rubidium (Rb) oscillators are covered. Customers can specify and order components off the page using the specifications and controlled issue numbers provided. **C-MAC Frequency Products;** *(919) 941-0430, FAX: (919) 941-0530, e-mail: fp@america.cfpwww.com, Internet: http://cfpwww. com.*

CIRCLE NO. 64 or visit www.mwrf.com

Crystal oscillators

A 70-page catalog focuses on quartz crystals and crystal oscillators. Electrical characteristics include supply voltage, rise and fall times, aging, duty cycle, and output impedance. Mechanical specifications, features, and descriptions are provided. **Reeves-Hoffman;** *(717) 243-5929, FAX: (717) 243-0079.*

CIRCLE NO. 65 or visit www.mwrf.com

Broadband products

A designer's handbook features more than 150 products from the power-amplifier (PA) digital cellular, silicon (Si) systems, and broadband product lines. Technical information, performance test data, specifications, schematics, and application notes are provided. **RF Micro Devices, Inc.;** *Internet: http://www. rfmd.com.*

CIRCLE NO. 66 or visit www.mwrf.com

VCOs and synthesizers

A catalog features low-noise voltage-controlled oscillators (VCOs) [to −112 dBc/Hz at 10-kHz offset] and low-noise synthesizers for wireless communication applications. The catalog lists the release of +5- and +3-VDC VCOs and a range of single-, dual-, and fixed-frequency synthesizers. Specifications are included. **Princeton Electronic Systems, Inc.;** *(609) 275-6500, FAX: (609) 799-7743, e-mail: pes@pesinc.com, Internet: http://www.pesinc.com.*

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Cascadable amplifiers

A 32-page catalog details RF and microwave components, high-reliability products and services, value-added testing and material services, as well as precision cable assemblies. Broadband amplifiers, limiting amplifiers, cascadable amplifiers, detectors, and cable assemblies are offered. High-reliability screening services, lot-acceptance testing, and element evaluation of semiconductor die are also covered. **Avnet Microwave Technical Solutions;** *(408) 360-4000, FAX: (408) 281-8802, Internet: http://www.avnetmarshall.com.*

CIRCLE NO. 68 or visit www.mwrf.com

Electronics enclosures

A four-page brochure describes a series of aluminum electronics enclosures combining contemporary design with solid construction. Flush-mounted front panels, durable black or gray finish, Electronic Industries Association (EIA) widths and heights, and equipment access are featured. Additional options are also discussed. **Buckeye ShapeForm;** *(800) 728-0776, (614) 445-8433, FAX: (614) 445-8224, Internet: http://buckeyeshape-form.com.*

CIRCLE NO. 69 or visit www.mwrf.com

Logic analyzers

A 22-page application note helps potential users comprehend the basics of logic-analysis measurements. Choosing the correct sampling method, transitional sampling, glitch capture, clocks, and state analyzer and cross-domain triggering are detailed. **Agilent Technologies, Inc.;** *(800) 452-4844 ext. 6980, Internet: http://www.tm.agilent.com/tmo/Notes/English/5968-8291E.html.*

CIRCLE NO. 70 or visit www.mwrf.com

DC-to-DC converters

An eight-page reference guide details standard usage of DC-to-DC converters. Key safety, thermal, and mechanical considerations are discussed. Appropriate connection and component practices, problems such as poor regulation, excessive noise at the load, crosstalk between circuits, system oscillation, and thermal difficulties are also covered. **Powercube;** *(800) 866-3590, FAX: (800) 866-3589, Internet: http://www.powercube.com.*

CIRCLE NO. 71 or visit www.mwrf.com

Power supplies

A 30-page shortform catalog features a selection of electronic test-and-measurement and power products. DC and AC power supplies, electronic loads, battery testers, amplitude modulation/frequency modulation (AM/FM) and RDS/RDBS signal generators are offered. Electronic safety testers, withstanding voltage testers, and oscilloscopes are also specified. **IFR Americas, Inc.;** *(800) 835-2352, (316) 522-4981, FAX: (316) 522-1360, e-mail: info@ifrsys.com, Internet: http://www.ifrinternational.com.*

CIRCLE NO. 72 or visit www.mwrf.com

Analog processing

A 72-page technical magazine covers topics such as products, applications, technology, and design techniques for analog, digital, and mixed-signal processing. A new direct-conversion radio chip set that eliminates intermediate-frequency (IF) stages and how digital signal processors (DSPs) enhance flexible third-generation (3G) base-station designs are overviewed. Analog-to-digital converters (ADCs) and choices for system design are examined. **Analog Devices, Inc.;** *(800) ANALOGD, FAX: (781) 937-1021, Internet: http://www.analog.com.*

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6002	5.25 – 5.35	20	20	30	4.75	600
6003	5.725 – 5.825	22	28.5	43	7.00	600
6004	5.725 – 5.825	22	27	41	7.00	600
6006	5.725 – 5.825	22	23	37	7.00	600
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	5.725 – 5.825	21	25	38		

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(continued from page 190)

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A Low Power 900-MHz Transmitter IC with Audio Baseband for ISM Applications Using 0.25-mm BiCMOS

MON4A-5
A Baseband to RF BiCMOS Transmitter RF IC for Dual-band CDMA/AMPS Wireless Handsets

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BROADBAND AMPLIFIERS AND VCOs
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(Student Paper) Ultra Low Power GaAs MMIC Low Noise Amplifier for Smart Antenna Combining at 5.2 GHz

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A Mixed Si and GaAs Chip Set for Millimeter-wave Automotive Radar Front Ends

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Refer to IMS Session TU2A.

TUE3A
WIRELESS CIRCUIT TECHNIQUES
JOINT IMS-RFIC SESSION
Refer to IMS Session TU3A.

TUE4A
INTEGRATED TRANSCEIVERS
JOINT IMS-RFIC SESSION
Refer to IMS Session TU4A.

2000 ARFTG Conference

The Automatic RF Techniques Group (ARFTG) will hold its 55th conference at the Hynes Convention Center on Friday, June 16. This year's conference will explore new methodologies in large-signal techniques with the theme "Going Beyond S-Parameters." Despite the fact that S-parameters have revolutionized the design, manufacturing and testing of linear high-frequency circuits, they fail to describe large-signal behavior.

MICROWAVE APPLICATION & PRODUCT SEMINARS (mAPS) TECHNICAL PROGRAM

Tuesday, June 13, 2000

EM Simulation

10:10 AM-11:50 AM

A Tutorial on Using the Free Sonnet Lite 3D Planar Electromagnetic Analysis Software, IE3D Planar and 3D Electromagnetic Simulation and Optimization Package, and others

1:10 PM-5:00 PM

Coplanar Circuit Design with COPLAN for ADS; MDSPICE Mixed Frequency and Time-domain SPICE Simulator; Topas: Advanced Nonlinear Modeling Software for FETs and BJTs; and others.

Wednesday, June 14, 2000

MILLIMETER WAVES AND FERRITES

9:30 AM-11:50 AM

A Family of User-friendly Self-biased Millimetre-wave LNAs; Passive Components: Microstrip Isolators and Circulators,; The "Alpha Two" Millimeter-wave Package,; Surface-mount Connectors for Applications up to Millimeter Waves, 4 mm^2 isolator for the 2-GHz band, Coaxial-Cable Assembly With a Discontinuous Outer Jacket, and others.

1:00 PM-5:00 PM

High Volume Filter and Diplexer Production by Investment Casting, Advanced Technology for Passive Microwave Components; New Multiplanar Ceramic Band Pass Filter for 2.4-GHz Applications, Maximizing RF Performance by Integrating Multiple Dielectric Constants and Embedding Passive Components, and others.

Thursday, June 15, 2000

FAB, AMPLIFIERS, MIXERS, AND OSCILLATORS

9:30 AM-11:50 AM

Micro-contamination Control in VLSI Processing; Direct Wafer Bonding of Silicon Wafers; Laser Structuring of Fine Lines; An Octave Bandwidth, 60-dB Dynamic Range, Microwave Successive Detection Logarithmic Video Amplifier, and others.

1:00 PM-2:20 PM

Wideband Photonic Links; A High Resolution Radar for Short Range Automotive Applications; A Telemetry Transmitter Chip Set for Ballistic Applications, and others.••

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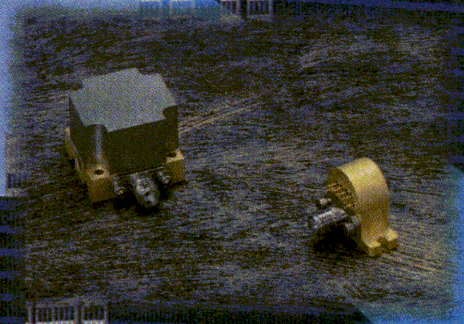
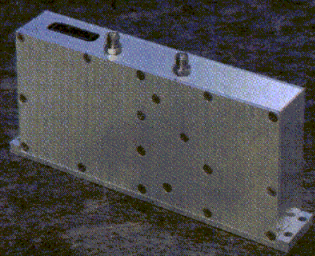
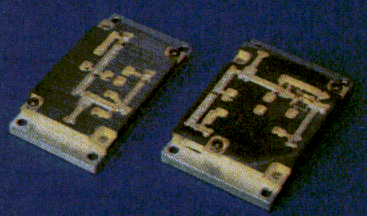
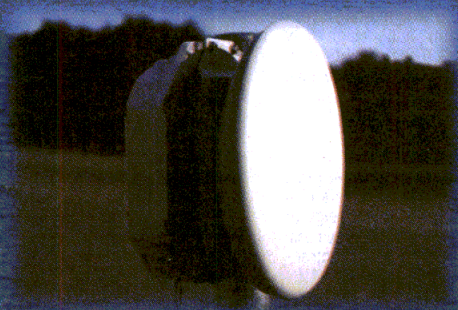


YIG Synthesizers



Millimeter Wave Transceivers/Receivers





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CIRCLE NO. 260

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Embedded System Engineer

This position will develop control and calibration firmware for broadband digital microwave radio subsystems. Objectives will be to provide customers with the external control interfaces they desire, to optimize product material costs with tradeoffs between software and hardware engineering, and to minimize labor in production tests. Experience with both hardware and software is desirable. Skills must include structured methods for software system engineering and developing at lower level languages for low-cost micro controllers. Requires an Engineering degree or equivalent experience, and 3-5 years of firmware development experience for control and calibration of hardware products. Experience in radio or RF module control and calibration is a plus. **Job #1**

Staff Microwave Design Engineer

This position, reporting to the vice president of engineering, includes the development of YIG tuned oscillators and other microwave circuits for use in low-cost, high-volume signal source and receiver systems. It affords the opportunity to work with the inventor of VertiCom's patented technology to bring products to market that enable much higher bandwidth digital radios. Experience should include low noise oscillator design and a track record of designs that perform in production with high yields. Frequencies range from 2 to 20 GHz. Experience in system design and analysis, manufacturing process development, designing amplifiers, couplers, mixers, and filters would be a definite advantage. Responsibilities include all phases of product life cycle from concept through early production. Applicant should have a minimum BSEE and 3-5 years of experience. Familiarity with HP/EESOF simulation tools or equivalent, design and layout of thin film hybrid circuitry and use of microwave test equipment are desired skills. **Job #2**

RF Design Engineer

This position will develop microwave circuit modules in the 20-40 GHz frequency range for use in frequency synthesizers and transceivers in the broadband microwave radio markets. Key success factors include optimized cost, manufacturing yield, and long-term reliability as well as adherence to development schedules. The successful candidate will possess strong analytical skills and use simulation tools to optimize designs, as well as strong measurement and device characterization skills to verify designs and models. Experience in use of standard and custom MMIC devices is a plus. Requires BSEE or MSEE with RF/Microwave emphasis, and 5 years' or more experience in millimeter wave transmit/receive module design. **Job #3**

Staff Mechanical Engineer

You will design packaging for RF/Microwave synthesizers and transceivers, and handle piece part and assembly tooling design for YIG oscillators. Requires a BSME and 5+ years' experience designing RF/Microwave modules (commercial radio products preferred); familiarity with #D CAD design, FEA, vibration and thermal simulation methods; expertise in vibration, stress and reliability testing; and knowledge of DFM with a record of designing products easily assembled with high yields. Background in high-volume manufacturing engineering preferred. **Job #4**

Marketing Manager

Successful candidate will be involved in extensive customer communications to develop and maintain long-term relationships. Will establish marketing goals to ensure share of market and profitability of products/services. Develop and execute marketing plans/programs to ensure the profit and expansion of company products/services. Research, analyze and monitor financial, technological, and demographic factors to capitalize on market opportunities and minimize competitive activity. Plan/oversee VertiCom's advertising and promotion activities including print, electronic, and direct mail outlets. Integral member of product development team to ensure timely, successful new product initiatives. Requires a degree in Business, Marketing, or Engineering. 5+ years' experience in Marketing or Sales. Must be an articulate self-starter and goal-orientated individual. **Job #5**

R/F Microwave Process Engineer

Position will build RF/Microwave hybrid micro assembly processes capable of supporting a 10x-volume ramp this year. Will problem-solve existing process problems and design new processes in support of production. Requires related B.S. (M.S. preferred), 5 years' experience minimum in a similar position, and a strong background in RF/Microwave production processes. Knowledge of hybrid microelectronics is required, including oscillator assembly techniques, bonding, die attach cleaning, welding and use of adhesives. **Job #6**

Product/Test Engineer

Solve problems and build production process capability to support a 10x-volume ramp this year. You will make design revisions to products, build test systems, train technicians, and establish process control. Requires a related BSEE (MSEE preferred); 5 years' experience in design or product engineering; and a strong background in RF/microwave frequencies, microwave test equipment and computerized automated test techniques. Knowledge of hybrid microelectronics is a plus. **Job #7**

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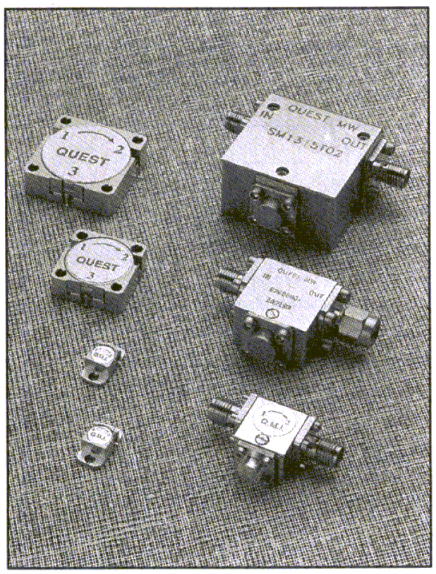
CIRCULATORS & ISOLATORS



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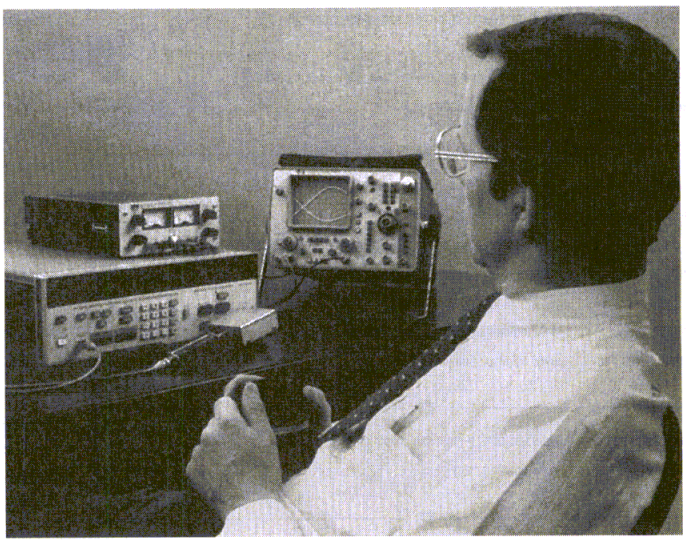


More than 19 years ago, a technical article by Mike Cuevas, a Product Manager at Hewlett-Packard Co. (Palo Alto, CA) addressed noise-figure measurements with the firm's HP 8970A receiver. The company, now known as Agilent Technologies, has introduced a replacement for the instrument on this month's front cover.

Microwaves & RF June Editorial Preview

Issue Theme: Military Electronics

News

Many high-frequency manufacturers have been lured away from military business by the tremendous growth of commercial communications markets. Yet, military business remains solid for those firms who have chosen to support it. What is the state of RF/microwave military business at present? For more on current RF/microwave military markets, don't miss the June issue of *Microwaves & RF*.

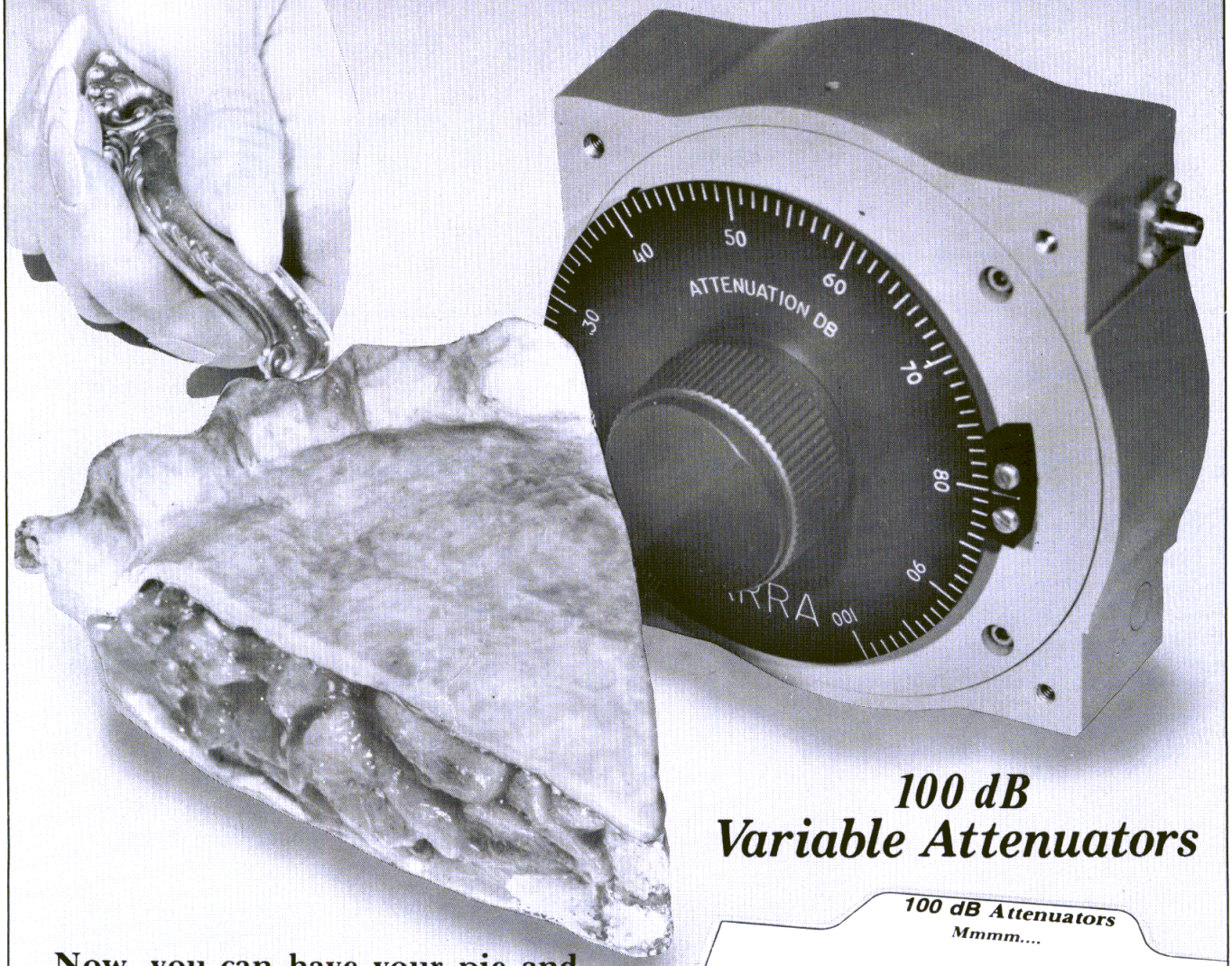
Design Features

The June issue of *Microwaves & RF* will offer numerous design and test techniques for engineers involved with commercial and military applications. For example, an author from Rockwell Collins will detail design techniques for miniature coupled-line directional couplers that can be used in avionics and wireless systems. Additional articles will examine a method of creating phase shifters for beam-forming networks using loaded lines, how to generate and use complex test signals for communications testing, and how to accurately compute phase-locked-loop (PLL) active-filter parameters.

Product Technology

June's Product Technology section will highlight commercial products including a three-part (transmitter, receiver, and software) analysis system for evaluating the performance of radio propagation environments, and a line of high-power transistors for avionics applications.

Get a piece of the ARRA pie...



100 dB Variable Attenuators

100 dB Attenuators
Mmmm....

Freq Range	Atten vs Freq dB	Model #
500 - 1000 MHz	1.5	2952 - 100
900 - 1300 MHz	0.75	2-3952 - 100
1000 - 2000 MHz	1.5	3952 - 100X
2000 - 4000 MHz	1.5	4952 - 100 X
4000 - 8000 MHz	1.5	5952 - 100X

Insertion loss - 6 dB
VSWR - 1.5
Power - 15 cw
Temperature -30 to +120 C

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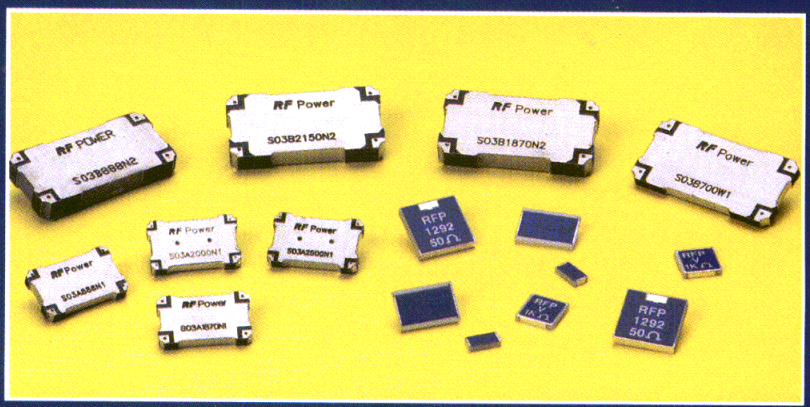
Surface Mount 90° Hybrid Couplers

- Burr-free manufacturing
- Full edge wrap
- Via holes for optimum connectivity

Terminations

- Non-Nichrome resistor for low IMD



Surface Mount

- Terminations
- 90° Hybrid Couplers
- Attenuators
- Combiners/Dividers
- Directional Couplers

Resistors

10-800 Watts, DC - 6 Ghz, SMD, flanged, coaxial

Attenuators

8-150 Watts, DC - 4 Ghz, SMD, flanged, coaxial

90° Hybrid Couplers

100-2000 Watts, 50 - 4200 Mhz, SMD, caseless, coaxial

Directional Couplers

100-2000 Watts, 4 - 6000 Mhz, SMD, caseless, coaxial

Combiners/Dividers

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

Custom Devices

Custom devices and assemblies

90° HYBRID COUPLERS

Model Number	Freq. Range (Mhz)	Power Watts. (CW)	Amp. Bal Max	Phase Bal. Deg Max	Isolation Min	VSWR	Insertion Loss Max.
S03B700W1	400-1000Mhz	200W	+/-0.65dB	+/-1.5	20dB	1.20:1	0.25dB
S03A888N1	815-960Mhz	100W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.25dB
S03B888N2	815-960Mhz	200W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.20dB
S03A1870N1	1750-1990Mhz	100W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.25dB
S03B1870N2	1750-1990Mhz	200W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.20dB
S03A1960N1	1930-1990Mhz	100W	+/-0.20dB	+/-1.5	20dB	1.25:1	0.25dB
S03B1960N2	1930-1990Mhz	200W	+/-0.10dB	+/-1.5	20dB	1.25:1	0.20dB
S03A2000N1	1500-2500Mhz	100W	+/-0.30dB	+/-2	20dB	1.20:1	0.25dB
S03B2150N2	2000-2300Mhz	200W	+/-0.20dB	+/-2	20dB	1.25:1	0.20dB
S03A2250N1	2000-2500Mhz	100W	+/-0.30dB	+/-2	20dB	1.20:1	0.25dB
S03A2500N1	2000-3000Mhz	100W	+/-0.35dB	+/-2	20dB	1.20:1	0.30dB
S03D3500NR5	3000-4000Mhz	50W	+/-0.30dB	+/-2	18dB	1.30:1	0.30dB

TERMINATIONS (CASE STYLE Z)

Reference	Watts	VSWR	Frequency
RFP-100200-4Z50-2	10	1.25:1	3 GHz
RFP-250250-4Z50-2	16	1.25:1	2 GHz
RFP-250250-6Z50-2	16	1.25:1	3 GHz
RFP-250375-4Z50-2	25	1.20:1	2 GHz
RFP-375375-6Z50-2	30	1.25:1	3 GHz

Call toll free 877-RFPC-INC (877-737-2462) for nearest representative.

Catalog request

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