February 2000

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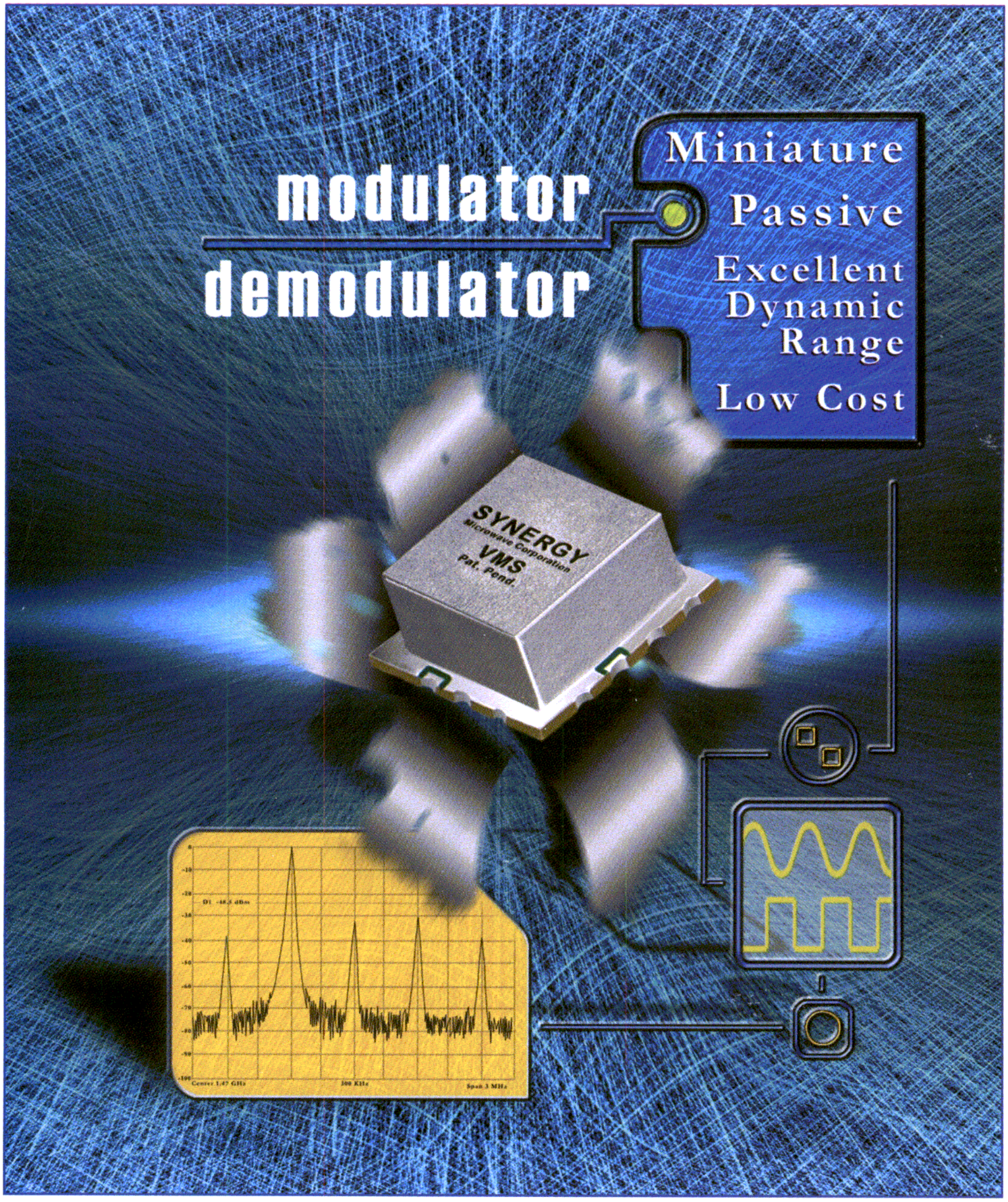
www.amwireless.com



Circuit Design

Phase Lock Loop Stability Analysis

Wireless Systems

Unfiltered FQPSK: Interpretation and Enhancements

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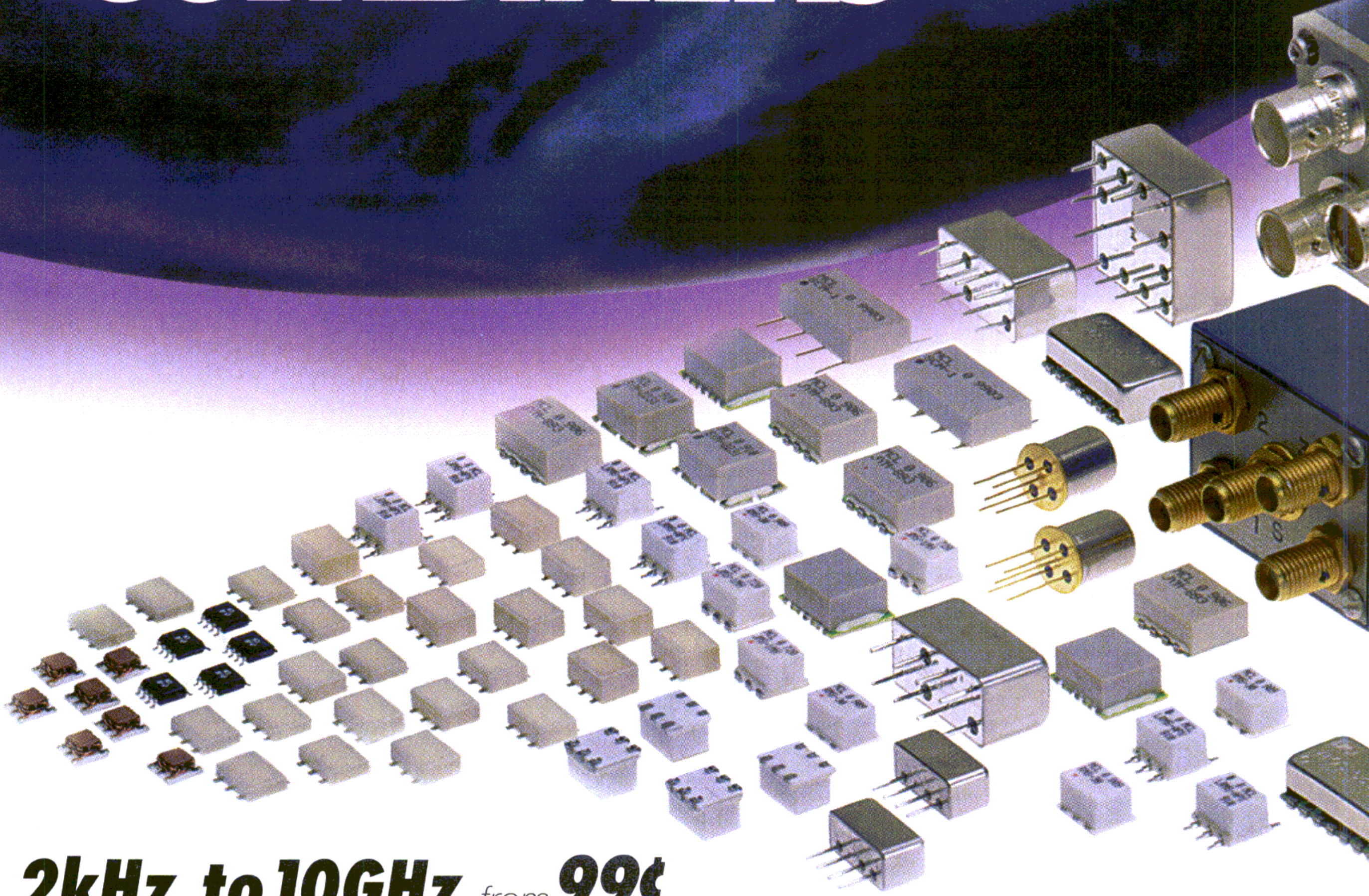

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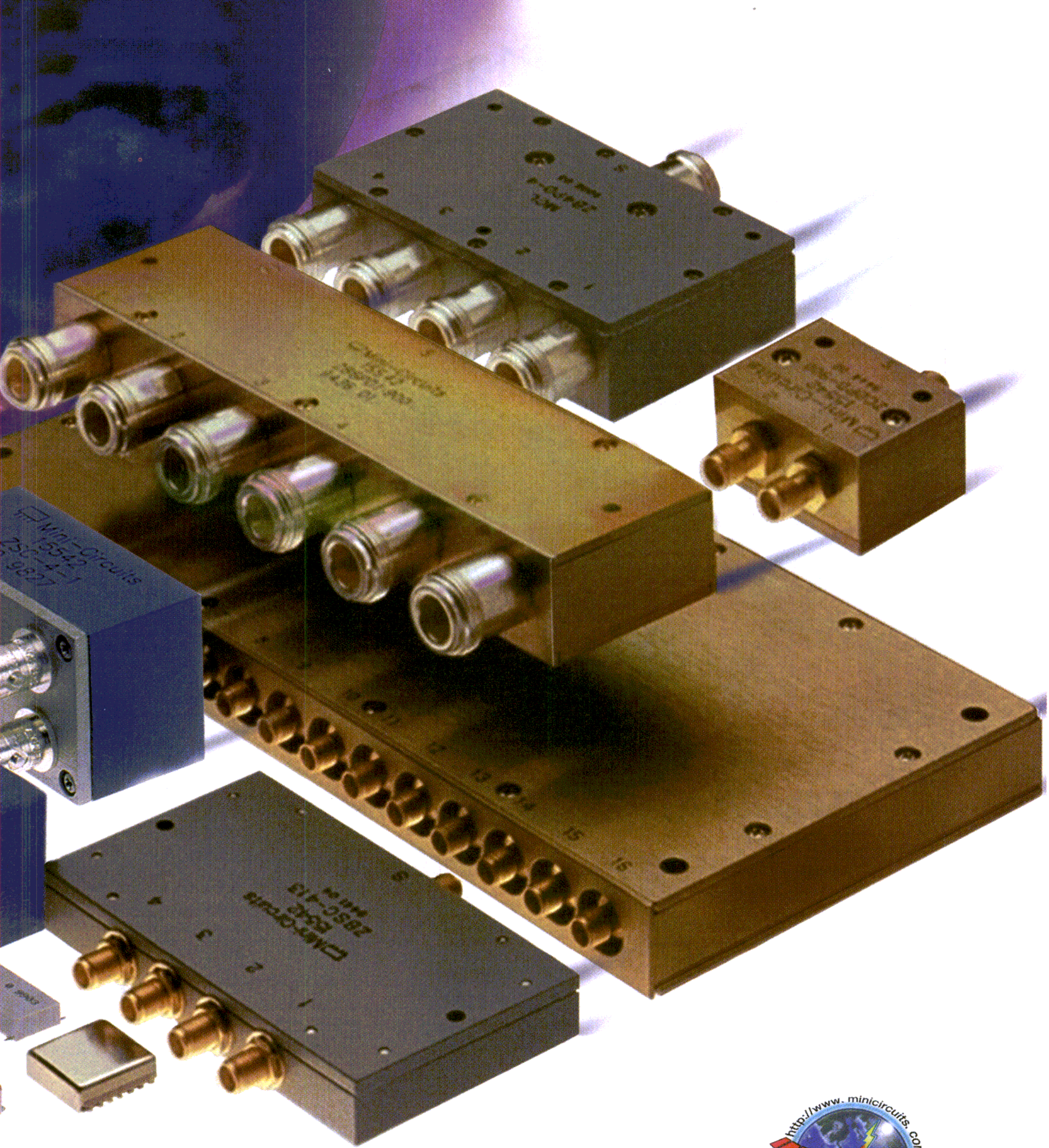

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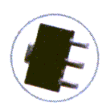
SPECIFICATION MATRIX	SGA-6286 SGA-6289	SGA-6386 SGA-6389	SGA-6486 SGA-6489
Frequency (GHz)	DC-3.5	DC -3.0	DC-1.8
Gain (dB)	13.8	15.4	19.7
TOIP (dBm)	34.0	36.0	34.0
P1dB (dBm)	20.0	20.0	20.0
N.F. (dB)	3.9	3.8	2.9
Supply Voltage (Vdc)	4.2	5.0	5.2
Supply Current (mA)	75	80	75

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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February 2000
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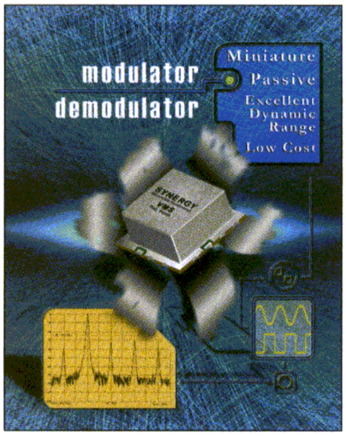
CONTENTS



76

On Our Cover

A High Dynamic Range Miniature Modulator/ Demodulator

Synergy Microwave

The company's new VMS series of I/Q modulators and demodulators offers designers both low noise and high isolation in a small, low cost package. The patent-pending technology also provides EMI/ RFI protection and moisture resistant capability.

Photo provided by Synergy Microwave.

TECHNICAL FEATURES

PRODUCTS & TECHNOLOGIES

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February 2000
Vol. 12 No. 2

Applied MICROWAVE & WIRELESS

PRODUCTS & TECHNOLOGIES

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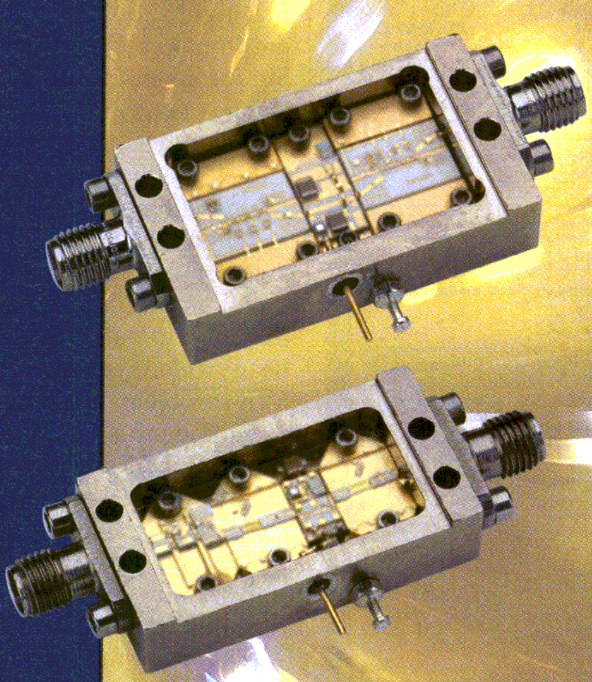
MODEL NUMBER	FREQ. MHZ	GAIN dB NOM	NOISE FIG. dB TYP	3RD ORDER ICP dB MIN
CELLULAR-BAND AMPLIFIERS				
JCA01-C01	800-960	24	0.9	32
JCA01-C02	800-960	24	1.1	37
JCA01-C03	800-960	24	1.2	40
JCA01-C04	800-960	24	1.3	42
JCA01-C05	800-960	42	0.9	32
JCA01-C06	800-960	42	1.1	37
JCA01-C07	800-960	42	1.2	40
JCA01-C08	800-960	42	1.3	42
PCS-BAND AMPLIFIERS				
JCA12-PC01	1710-1990	24	0.9	32
JCA12-PC02	1710-1990	24	1.1	37
JCA12-PC03	1710-1990	24	1.2	40
JCA12-PC04	1710-1990	24	1.3	42
JCA12-PC05	1710-1990	40	0.9	32
JCA12-PC06	1710-1990	40	1.1	37
JCA12-PC07	1710-1990	40	1.2	40
JCA12-PC08	1710-1990	40	1.3	42
WLL-BAND AMPLIFIERS				
JCA23-W01	2300-2500	24	1.0	32
JCA23-W02	2300-2500	24	1.2	37
JCA23-W03	2300-2500	24	1.3	40
JCA23-W04	2300-2500	24	1.5	42
JCA23-W05	2300-2500	41	1.0	32
JCA23-W06	2300-2500	41	1.2	37
JCA23-W07	2300-2500	41	1.3	40
JCA23-W08	2300-2500	41	1.5	42



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Editorial

The Y2K Prevention Effort was Almost Good Enough

By Gary A. Breed
Publisher

Because I work in a high-tech industry, my friends would often ask me if I thought the Y2K problem was real and how it might affect them. My standard answer was, "I'm sure all the really big problems will be found, but some small ones will inevitably occur." One of the examples I used was that some companies might find problems processing their customers' credit cards.

I didn't realize that I was predicting my own company's future!

This column is being written on January 13 and I don't know yet if it has been widely reported that a significant Y2K problem affected many companies, including ours. Our credit card processing software had a serious bug that was not identified until three business days of the new year had passed. In those three days, credit card transactions were not cleared from temporary files when each day's activities were closed out. As a result, our bookstore customers who made purchases on January 3 got the charges to their cards repeated on the 4th and 5th! Others got double charges.

The software company claims that they began notifying users in February 1999 with a notice on their web site, but we are a registered user and received no letter, e-mail or other warning. We had no reason to visit their web site because their product was working fine and we were using an earlier upgrade that had been represented as being Y2K compatible.

For us, the end of the story is not too bad. Only a small number of people were affected and the credit card companies are working with us and the software vendor to clear up the problems. But, some of the large companies using the faulty software may have had hundreds or even thousands of transactions go wrong.

So if you have been wondering whether all that money spent on Y2K testing and software updates was worth it, the answer is "yes." Although many companies and governments around the world did nothing and suffered few problems, some potentially dangerous or disruptive problems were uncovered and fixed. (And one moderate-sized problem was missed!)

Like medical and dental exams, bank audits and termite inspections, prevention seems costly — but never as costly as the problems it prevents. ■

Calendar

CONFERENCES

FEBRUARY

February 22-25, 2000
Wireless/Portable 2000
San Jose, CA
Information: Wireless/Portable
Internet: http://www.WirelessPortable.com

February 28-March 1, 2000
CTIA Wireless 2000
New Orleans, LA
Information: CTIA
Tel: 1-800-421-6338 or 415-979-2250
Internet: http://www.ctiashow.com

MARCH

March 6-9, 2000
CARTS 2000 — Capacitor and Resistor Technology Symposium
Long Beach, CA
Information: CTI-EIA/ECA
E-mail: pwalsh@eia.org
Internet: http://www.eia.org

March 28-30, 2000
RTS 2000
Paris, France
Information: BRIP
Tel: + 33 01 53 17 11 44
E-mail: crey@birp.fr
Internet: http://www.birp.com/rts

APRIL

April 8-13, 2000
NAB2000
Las Vegas, NV
Information: NAB
Internet: http://www.nab.org.conventions

April 10-11, 2000
2000 IEEE Emerging Technologies Symposium on Broadband Wireless Internet Access
Dallas, TX
Information: Dr. Jon Velhl
Tel: 972-952-4190
Internet: http://www.ieeedallas-els.org

April 17-19, 2000
48th International Relay Conference
Lake Buena Vista, FL
Information: NARM EIA/ECA
E-mail: pwalsh@eia.org

MAY

May 1-4, 2000
2000 GaAs MANTECH Conference
International Conference on Gallium-Arsenide Manufacturing Technology
Washington, DC
Information: Wes Mickanin
Tel: 503-615-9253; Fax: 503-615-8903
E-mail:wem@tqs.com
Internet: http://www.gaasmantech.org

May 17-19, 2000
Eurocomm 2000
Munich, Germany
Information: Jacques Keevers
Tel: +32 2 7706643; Fax: +32 2 7708505
E-mail: j.keevers@ieee.org
Internet: http://www.eurocomm.org/2000

May 20-26, 2000
IEEE International Conference on Phased Array Systems and Technology
Dana Point, CA
Information: Dr. Michael Thorburn
Tel: 310-336-2197; Fax: 310-336-6225
E-mail: m.a.thorburn@IEEE.org
Internet: http://www.ieee.org

May 21-24, 2000
50th Electronic Components and Technology Conference
Las Vegas, NV
Information: EIA/ECA-IEEE/CPMT
E-mail: pwalsh@eia.org
Internet: http://www.ectc.org

JUNE

June 7-9, 2000
2000 IEEE/EIA International Frequency Control Symposium and Exhibition
Kansas City, MO
Information: IEEE Ultrasonics, Ferroelectrics and Frequency Control Society
E-mail: pwalsh@eia.org
Internet: http://www.ieee.org/uffc/fc

June 11-13, 2000
2000 IEEE Radio Frequency Integrated Circuits Symposium
Boston, MA
Information: Jyoti Mondal
Tel: 847-259-9600, ext. 4130
E-mail: mondajy@mail.northgrum.com
Internet: http://www.ims2000.org/rfic.htm

SHORT COURSES

Comsearch Center for Technical Training

Basics of Wireless for Non-Engineers
Reston, VA . . . . .February 7-8, 2000
Basics of Microwave Design
Reston, VA . . . . .February 9-10, 2000
WLL and PMP Overview
Reston, VA . . . . .February 15-16, 2000
WLL Theory and Design
Reston, VA . . . . .February 17-18, 2000
Information: Comsearch, Tel: 1-800-318-1234; Fax: 703-476-2720; E-mail: training@comsearch.com; Internet: www.comsearch.com.

Franklin Research Group

Understanding Telecommunications
San Jose, CA . . . . .February 7-8, 2000
Portland, OR . . . . .February 10-11, 2000
New York, NY . . . . .February 14-15, 2000
Atlanta, GA . . . . .February 28-29, 2000
Information: Franklin Research Group, Tel: 201-847-7882; Fax: 201-847-1821; Internet: www.franklinresearch.com.

RTT Programmes Limited

Cellular Design
London, England . . . . . . .February 7-9, 2000
RF Amplifier Design
London, England . . . . . . .February 14-16, 2000
RF Oscillator Design
London, England . . . . . . .March 13-15, 2000
RF/IF Processing
London, England . . . . . . .March 20-22, 2000
RF System Engineering
London, England . . . . . . .March 27-29, 2000
Information: Lorraine Gannon, Tel: +44 181 844 1811; Fax: +44 181 751 2616; E-mail: seminars@rttsys.com; Internet: www.rttsys.com.

Applied Technology Institute

Fundamentals of Radar Technology
Washington, DC . . . . .February 8-10, 2000
Radar Systems Design and Engineering
Washington, DC . . . . .February 15-17, 2000
Mathematical and Physical Wavelets
Newport, RI . . . . .March 13-16, 2000
Satellite Communication Systems Engineering
Los Angeles, CA . . . . .March 21-23, 2000
Information: ATI Defense Systems Training, Tel: 1-888-501-2100 or 410-531-6034; Fax: 410-531-1013; E-mail: ATIcourses@aol.com; Internet: www.ATIcourses.com.

Besser Associates

Wireless Measurements: Theory and Practice
Mountain View, CA . . . . . .February 14-18, 2000
Signal Integrity of High-Speed Digital Design
Mountain View, CA . . . . . .February 28-29, 2000
Multitone Amplifier Design
Mountain View, CA . . . . . .March 2-3, 2000
Introduction to RF Transceivers and Systems Components
Mountain View, CA . . . . . .March 6-7, 2000
RF Test Equipment Operation (laboratory course)
Mountain View, CA . . . . . .March 8, 2000
RF Testing for the Wireless Age (laboratory course)
Mountain View, CA . . . . . .March 9-10, 2000
Behavioral Modeling
Mountain View, CA . . . . . .March 13-15, 2000
RFIC Techniques for Wireless Applications
Mountain View, CA . . . . . .March 20-22, 2000
EMC/EMI and Thermal Issues for Electronic Packages and Systems
Mountain View, CA . . . . . .March 23, 2000
Applied RF Techniques II
Phoenix, AZ . . . . .March 26-31, 2000
Introduction to Receivers, Signals and Noise
Mountain View, CA . . . . . .March 27-28, 2000
Information: Annie Wong, Tel: 415-949-3300; Fax: 415-949-4400; E-mail: info@bessercourse.com; Internet: www.bessercourse.com.

University of Missouri-Rolla

Grounding and Shielding Electronic Systems
Atlanta, GA ....February 16-17, 2000
Orlando, FL ....February 22-23, 2000

Circuit Board Layout to Reduce Noise Emission and Susceptibility
Atlanta, GA ....February 18, 2000
Orlando, FL ....February 24, 2200

Information: Buddy Poe, Tel: 573-341-6061; Fax: 573-341-4992; E-mail: buddyp@umr.edu; Internet: www.umr.edu/~conted/ee.html.

University of Wisconsin at Milwaukee

EMC Printed Circuit Board Design
Tampa, FL ....Feb. 28-March 1, 2000

Information: Mark Schmidt, Program Assistant, Tel: 1-888-545-4700; Fax: 1-888-545-4600; E-mail: dschmidt@uwm.edu; Internet: www.uwm.edu/dept/ccee.

University of California at Los Angeles Extension

Digital Avionics Systems
Los Angeles, CA ....Feb. 28-March 3, 2000

Kalman Filtering
Los Angeles, CA ....March 13-16, 2000

Synthetic Aperture Radar: Understanding the Imagery
Los Angeles, CA ....March 13-17, 2000

Radar Interferometry: Principles, Processing and Applications
Los Angeles, CA ....March 20-24, 2000

Information: UCLA Extension, Short Course Program Office, Tel: 310-825-3344; Fax: 310-206-2815.

Northeast Consortium for Engineering Education

Antennas: Principles, Design and Measurements
San Diego, CA ....March 13-16, 2000
Orlando, FL ....May 22-25, 2000

Information: Kelly Brown, Tel: 407-892-6146; Fax: 407-892-0406; E-mail: stcloudof1@aol.com; Internet: www.usit.com/antenna.

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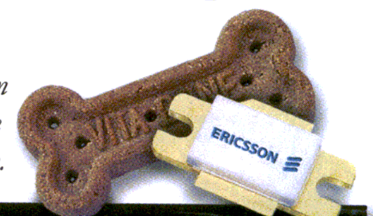

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PTF 10049	470–860MHz	85	12.0	32	–32	50	Input Matched
PTF 10159	470–860MHz	120	12.0	32	–35	58	Input Matched
PTF 10160	860–960MHz	85	16.0	26	–30	54	I/O Matched
PTF 10036	860–960MHz	85	11.0	28	–30	55	Input Matched
PTF 10020	860–960MHz	125	11.0	28	–30	55	Push Pull
PTF 10100	860–960MHz	165	12.0	28	–30	47	Input Matched
PTF 10149	925–960MHz	70	16.0	26	–30	50	Input Matched
PTF 10021	1.4–1.6 GHz	30	11.0	28	–30	48	I/O Matched
PTF 10125	1.4–1.6 GHz	135	11.5	28	–30	45	I/O Matched
PTF 10035	1.9–2.0 GHz	30	11.0	28	–30	35	I/O Matched
PTF 10112	1.8–2.0 GHz	60	11.0	28	–28	41	I/O Matched
PTF 10120	1.8–2.0 GHz	120	10.0	28	–30	40	I/O Matched
PTF 10048	2.1–2.2 GHz	30	10.0	28	–30	39	I/O Matched
PTF 10122	2.1–2.2 GHz	50	9.5	28	–30	39	I/O Matched
PTF 10134	2.1–2.2 GHz	100	10.0	28	–30	36	I/O Matched

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Deadline: March 17, 2000

EuMC 2000 — 30th European Microwave Conference

October 2-5, 2000 — Paris, France

Topics: Microwave materials, passive components and circuits, active circuits and systems, MMICs, CAD of microwave circuits, CAD for passive components, electromagnetic field theory, antenna design, modeling and measurement, millimeter and submillimeter wave components and circuits, microwave measurements, superconducting components and subsystems, EM compatibility, packaging and remote sensing, including active and passive radars.

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October 4-5, 2000 — Paris, France

Topics: Low cost, high speed silicon integrated circuits, propagation modeling for wireless multimedia, linearity at the active device and component levels, new modulation and coding schemes, advanced multi-user receivers, spatial channels and smart antennas, satellite networks, wireless LAN, UMTS, wireless access to information infrastructure and impact of technology on system performance.

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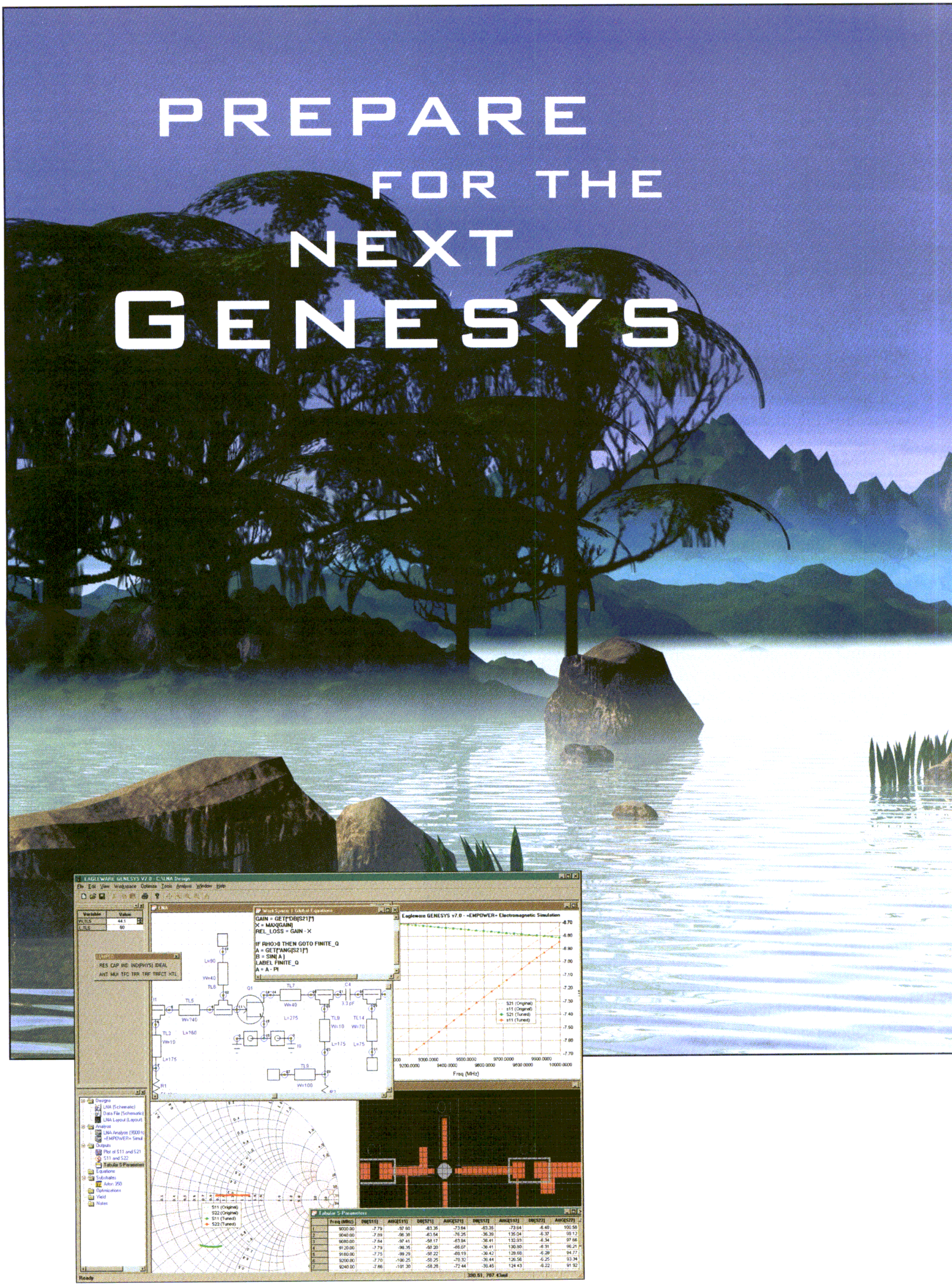
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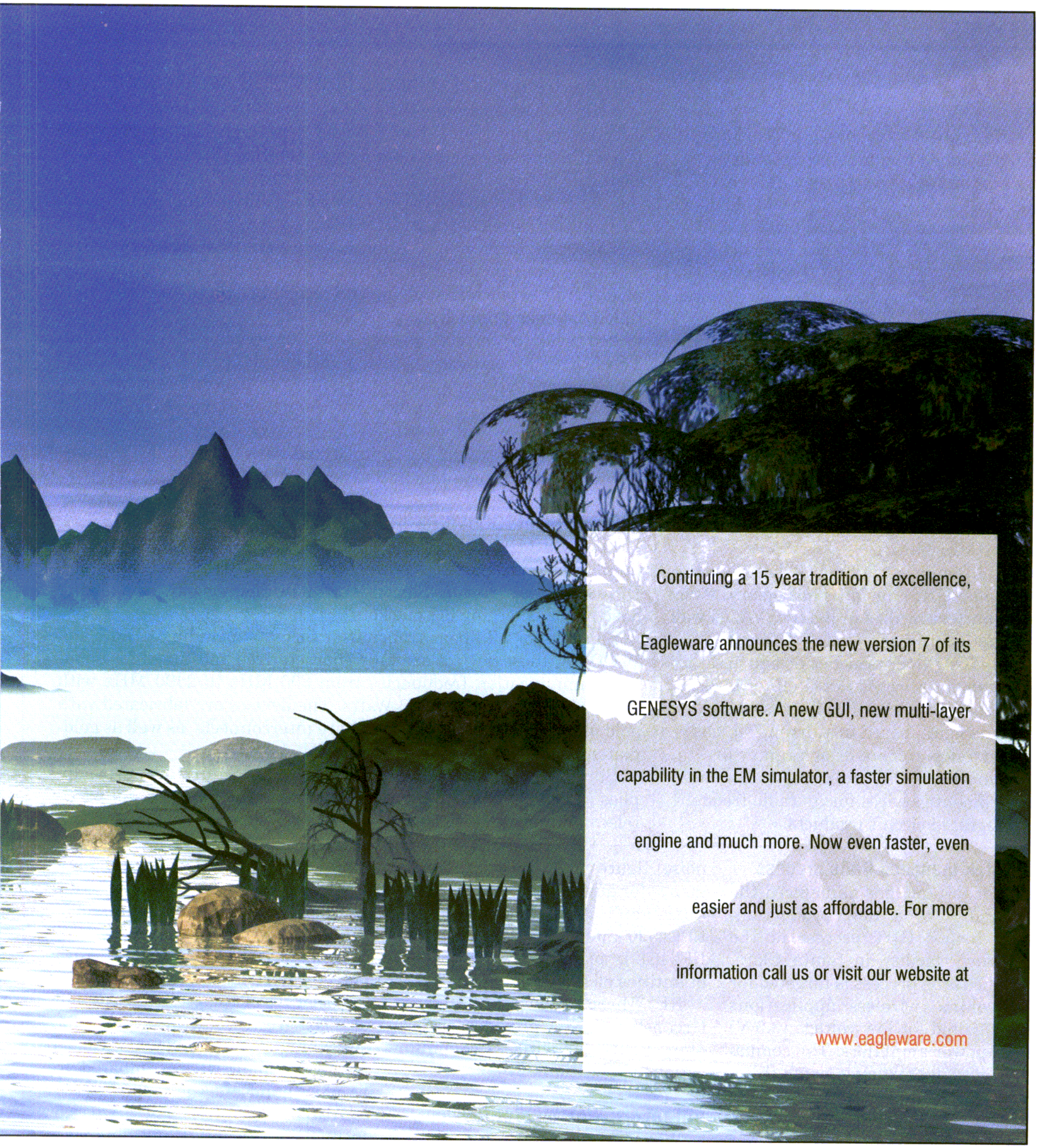
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• Stanford Microdevices, based in Sunnyvale, CA, has announced the opening of a new design center in Kanata, Ontario, Canada. The facility will house RF design and test capabilities.

• L-3 Communications' Global Network Solutions (GNS) business has opened a new headquarters facility in Reston, VA, as well as regional offices in Dallas, TX, and Atlanta, GA. GNS provides design and management services for satellite, wireless local loop and microwave radio systems.

• Anadigics, based in Warren, NJ, has opened a new remote design center in Newbury Park, CA, providing design and development services for its radio frequency integrated circuit products.

Andrew transverter receives first MMDS certification

Andrew Corporation, through its recently acquired Conifer Corporation subsidiary, has received type certification from the Federal Communications Commission for the Ultraflex® two-way customer premise transmit/receive transverter. This certification is the first to be given to an MMDS (Multichannel Multipoint Distribution Service) transverter.

The transverter, one of Andrew's wireless local loop subscriber products, can function as a multi-channel analog and digital video receive unit, while transmitting and receiving data to and from multiple customer premise modems and terminals. A GPS frequency-locked oscillator corrects for signal drift over time and temperature. The FCC type certification covers variable data rates and modulation schemes, allowing service providers to use different wireless cable modems and set-top boxes without certifying the individual brand designs.

Andrew Corporation, based in Orland Park, IL, supplies communications systems equipment and services worldwide.

Stanford Microdevices introduces LDMOS device line

Stanford Microdevices Inc. has announced the availability of a family of high-linearity power transistors based on Laterally Diffused Metal Oxide Silicon (LDMOS) technology, supporting designs for emerging 3G markets.

Suitable for all linear transmitter formats including CDMA and W-CDMA, these devices are fully characterized and tuned for popular communication frequencies from 800 MHz to 2400 MHz with power levels from 10 to 120 Watts. The devices are fabricated with gold metallization and gold bond-wire interconnects, as well as gold-plated packages for higher reliability.

Stanford Microdevices, headquartered in Sunnyvale, CA, supplies radio frequency integrated circuits for the wireless and wired telecommunications markets.

Celeritek launches GaAs process

Celeritek Inc. has announced the availability of a new production-ready InGaP HBT process at the company's facility in Santa Clara, CA. The GaAs process is targeted for high-volume wireless applications and is suited for low voltage (3.0 V) linear power amplifiers. The company is also introducing a family of InGaP HBT amplifier modules.

Celeritek develops GaAs radio frequency integrated circuits for wireless communications.

RF Micro Devices offers new line

RF Micro Devices has introduced the company's first component produced using SiGe HBT technology.

The RF2461 is a CDMA/FM low noise amplifier/mixer 900 MHz downconverter providing excellent noise figure and linearity performance for dual-mode CDMA/FM cellular applications. Offered in an LPCC-20 4 mm × 4 mm package, the RF2461 is priced at $1.87 each in quantities of 10,000.

RF Micro Devices, based in Greensboro, NC, provides proprietary RFICs for wireless communications applications.

ABI report: WAP market to boom

Wireless Access Protocol-based handsets will be the next "big thing" in wireless, according to a report from Allied Business Intelligence.

By 2005, one-third of all user handsets will use WAP, the report says, with more than 500 million WAP-based handsets expected to be shipped between 2000 and 2005.

Digital will remain important, the report says, with the digital/dual-mode handset marketplace to grow at 29 percent a year through 2005. Analog-only users, however, will drop from 34 percent in 1998 to 5 percent by 2005.

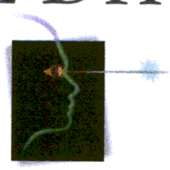
Other wireless access devices will also see growth, but the market for smartphones and handhelds will be relatively small when compared to wireless handsets, the report says.

ABI is an Oyster Bay, NY-based technology research think tank specializing in communications and emerging technology markets.

Companies, organizations and institutions may submit information for our News section to: Shannon O'Connor, Managing Editor, Applied Microwave & Wireless, 4772 Stone Drive, Tucker, GA, 30084; Fax: (770) 939-0157; E-mail: amw@amwireless.com.

IC

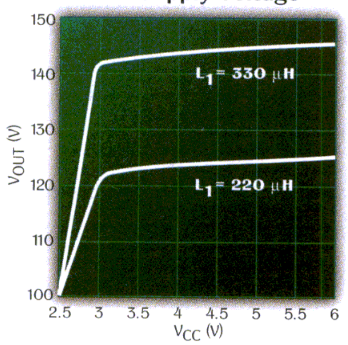
A Quiet Revolution In EL Drivers



Why the TK659xx?

Because it is a brighter and quieter EL driver.

Peak to Peak Lamp Voltage vs. Supply Voltage



Output is independent of variations of input voltage.

Available in 10 lamp frequencies

175 Hz	300 Hz
200 Hz	325 Hz
225 Hz	350 Hz
250 Hz	375 Hz
275 Hz	400 Hz

The new TK659xx series of EL drivers operates with a breakthrough voltage waveform with controlled rise and fall times. This improves efficiency while reducing audible and RF noise.

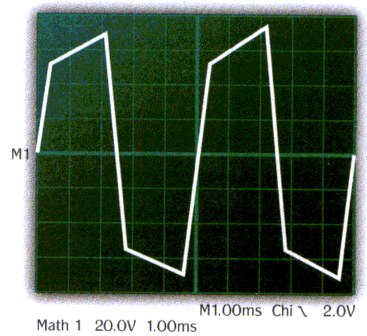
Further noise reduction is achieved with a fixed frequency PWM boost circuit operating above 20kHz. This boost circuit regulates the output independent of variations in battery voltage, to assure consistent brightness and color.

The peak current control of the TK659xx series allows use of Toko's miniature EL coils (D31FU, D32FU, and D52FU) without fear of saturation, while reducing overall circuit space.

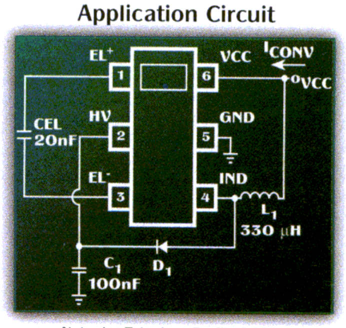
The TK659xx series is housed in a small, SOT-23L package, making it ideal for devices where noise and board space are critical. This includes wireless phones, caller IDs, PDAs, MPEG players, handheld computers, GPS, digital cameras, watches, clocks, radios, calculators, portable instruments, radar detectors — anything with an EL panel.

Call us or check our web site for details on the TK659xx.

TK659xx Voltage Waveform Controlled Rise and Fall Times



Application Circuit



Note: L_1=Toko Low Profile D52FU

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BUSINESS AND FINANCE

Signal Technology acquires Advanced Frequency Products

Signal Technology Corporation has announced that it has acquired privately-held Advanced Frequency Products (AFP), based in Andover, MA. Financial terms of the transaction were not disclosed.

Signal Technology, based in Danvers, MA, produces electronic components and systems for telecommunications. AFP provides high-frequency millimeter wave and microwave transceivers for the broadband wireless communications infrastructure market.

Motorola NSS receives contracts

Motorola Inc.'s Network Solutions Sector has received two new contracts to provide cellular network equipment for systems in Israel and Ghana.

Under a contract with Pele-Phone Communications, Motorola NSS will expand the company's existing CDMA digital cellular network in Israel with additional infrastructure equipment. The contract is valued at $122 million.

A separate agreement with Ghana Telecommunctions Company Ltd. calls for the installation of a new GSM digital cellular network in Ghana. The contract is worth $41.5 million.

Motorola, based in Schaumburg, IL, provides semiconductors, integrated communications solutions, embedded electronic systems and components.

Adaptive Broadband signs contract

Adaptive Broadband™ Corporation has signed a $225 million, five-year contract to provide its AB-Access fixed wireless broadband equipment to Telecom Wireless Corporation. Adaptive Broadband, based in Sunnyvale, CA, is a data networking solutions company developing technology for the deployment of broadband wireless communication over the Internet.

Giga-tronics awarded contract

The Instrument Division of Giga-tronics Inc., based in San Ramon, CA, has received a follow-on order for VXI synthesizers and down converters from ManTech Test Systems Inc., of Chantilly, VA. The contract is worth $3.3 million.

Giga-tronics' Instrument Division manufactures instruments and automatic test systems for wireless communications. ManTech Test Systems is a subsidiary of ManTech, which provides automatic test equipment to the defense and commercial markets.

BUSINESS AND FINANCE

Crown Castle, One 2 One sign contract for mobile network in Northern Ireland

Crown Castle International Corp.'s UK subsidiary, Crown Castle UK Limited, and wireless service provider One 2 One have signed a $160 million agreement for a new mobile network in Northern Ireland.

Crown Castle will provide a turnkey solution for One 2 One, including all cell planning, acquisition, design, build, operation and maintenance for the mobile network. The agreement will run for an initial term of eleven years. Crown Castle already operates approximately 100 tower sites in Northern Ireland and these sites will provide the backbone for the new network. One 2 One is the world's first all digital mobile network and is a wholly owned subsidiary of Deutsche Telekom.

Crown Castle International Corp., based in Houston, TX, provides communication sites and wireless network services as well as related infrastructure and network support services, in the United States and the United Kingdom. One 2 One, the world's first all digital mobile network, is a wholly owned subsidiary of Deutsche Telekom.

L-3 purchases training division from Raytheon

L-3 Communications has acquired Raytheon's Training Devices and Training Services (TDTS), headquartered in Arlington, TX, and Binghampton, NY. The transaction is valued at $160 million.

TDTS provides high fidelity, fully integrated simulator training products, such as flight simulators and pilot training systems for aircraft and helicopters, combat vehicle trainers, and training support services. Its products and services are used by U.S. and foreign military agencies and prime contractors.

L-3 Communications supplies secure communication systems and products, microwave components, avionics and ocean systems and telemetry, instrumentation, space and wireless products.

Proxim acquires Wavespan Corporation

Proxim Inc., based in Sunnyvale, CA, has acquired Wavespan Corporation of Mountain View, CA. The transaction is valued at approximately $14.4 million.

Proxim is a supplier of spread spectrum wireless LAN products to OEMs and wireless solutions providers. Wavespan is a developer of wireless broadband access solutions.

ECI Telecom receives $20 million order

ECI Telecom Ltd.'s wholly-owned subsidiary, InnoWave ECI Wireless Systems Ltd., has announced a $20 million order from Atlantic Telecom, a subsidiary of Atlantic Telecom Group PLC, to deliver fixed wireless access solutions in the UK.

Under the agreement, InnoWave's MultiGain Wireless (MGW) systems will be deployed in Atlantic Telecom's new license area in northwest England, as well as its existing operating systems in Scotland. The system uses frequency hopping CDMA technology.

InnoWave ECI Wireless Systems Ltd. provides fixed wireless access systems and wireless local loop (WLL) products and solutions. ECI Telecom, based in Petah Tikva, Israel, provides integrated network solutions for digital communications and data transmission systems.

TSMC, TASMS announce merger plans

Taiwan Semiconductor Manufacturing Company (TSMC) and TSMC-Acer Manufacturing Corporation (TASMC) have announced that they have signed an agreement to merge TASMC with and into TSMC. The financial value of the transaction was not disclosed.

TSMC, based in Hsin-Chu, Taiwan, is a semiconductor foundry offering processes including CMOS logic, mixed-mode and BiCMOS.

Wireless Facilities receives new contract

Wireless Facilities Inc., based in San Diego, CA, has signed a three-year contract with Advanced Telecom Corp. to provide network deployment services for the company's build-out of its Internet Protocol (IP) networks. The value of the contract was not disclosed.

Wireless Facilities designs, deploys and manages wireless networks worldwide. Advanced Radio Telecom Corp., headquartered in Bellevue, WA, provides local IP services to metropolitan areas nationwide.

Micronetics receives order for assemblies

Micronetics Wireless Inc., based in Hudson, NH, has announced that its Micronetics/Components Group has been awarded an order from Daimler Chrysler Aerospace to supply microwave integrated assemblies for the European Fighter Aircraft. The value of the agreement was not disclosed.

Micronetics Wireless manufactures broadband test equipment and components, as well as microwave components and subassemblies.

Com-Net acquires Ericsson's radio systems business

Com-Net Critical Communications, based in Pittsburgh, PA, has acquired Ericsson Inc.'s Private Radio Systems operations, located in Lynchburg, VA. The new company will be called Com-Net Ericsson Critical Radio Systems. The value of the transaction was not disclosed.

Com-Net provides radio network solutions for the public safety and service markets. Ericsson, based in Stockholm, Sweden, provides communications products and services worldwide.

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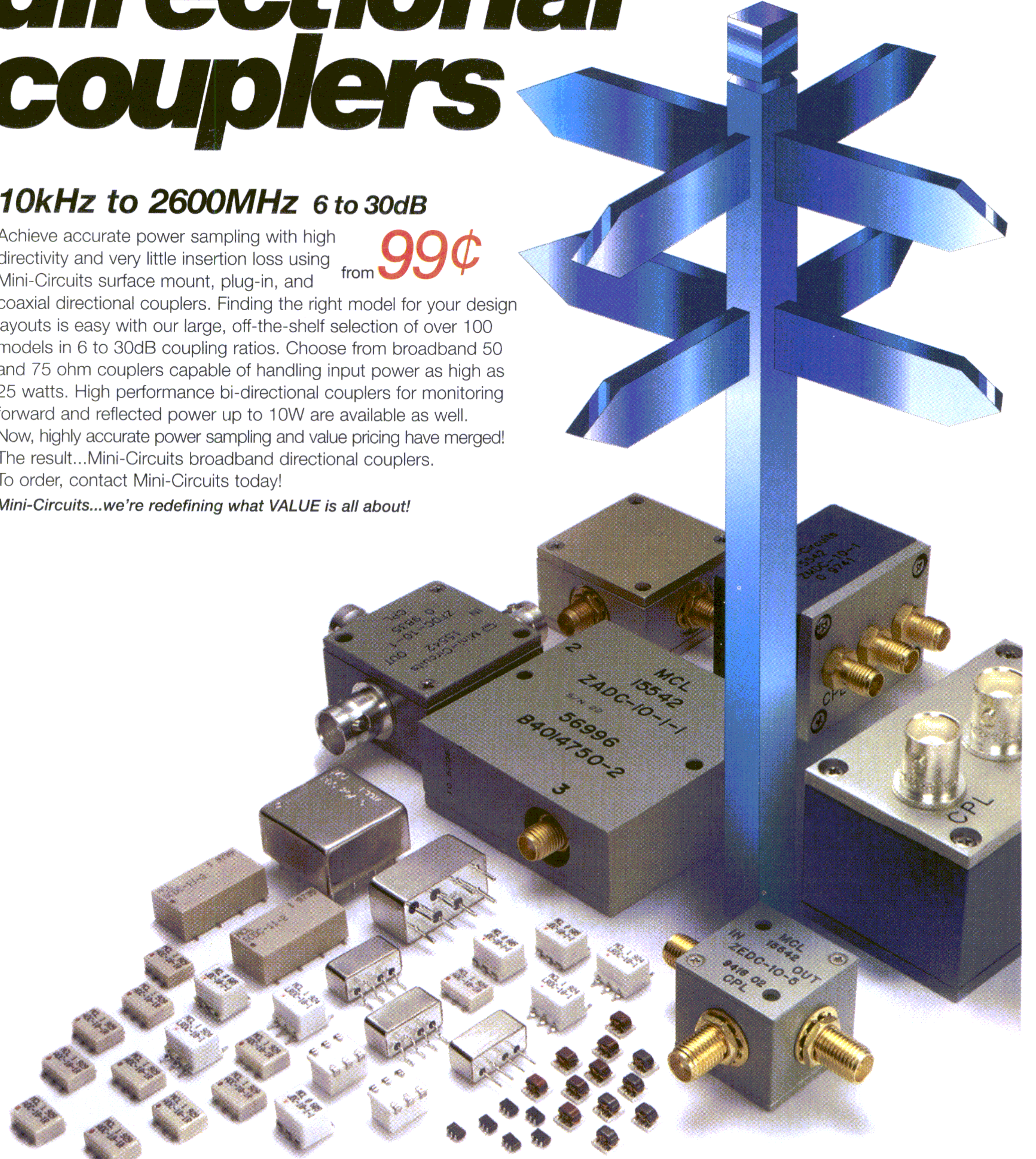

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COMPONENTS

Phase Lock Loop Stability Analysis

This article presents a review of PLL transfer functions with attention to the conditions required for steady-state stability

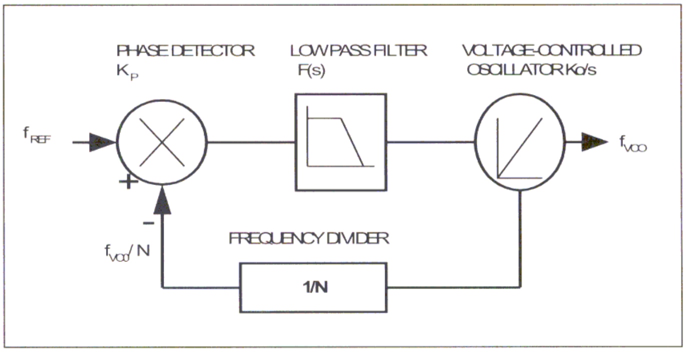
By Arun Mansukhani
Motorola, Inc.

Phase lock loops (PLLs) are key components of modern communication systems. Frequency synthesizers, FM demodulators and clock recovery circuits are some applications of PLLs. An important aspect of PLL design is the steady state stability of the loop. This article examines this aspect of PLL design, particularly the effect of loop filter on PLL stability.

PLLs are negative feedback control systems comprising of a phase-frequency detector (PD), a loop filter, a voltage-controlled oscillator (VCO) and a frequency divider. The function of the PD is to generate an output waveform based on the difference in phase (and frequency) between the input signal and a fixed reference. This is followed by a loop filter, normally a lowpass filter (LPF), whose function is to filter out any high frequency harmonics from the phase detector and to provide a DC signal output; followed by a VCO that generates a high frequency signal controlled by the DC input signal. A sample of the VCO output signal is then fed back to the input of the PD as the input waveform and compared in phase (and frequency) to a fixed reference. In a locked condition, the PLL output signal is locked in phase (and frequency) to the fixed input reference, i.e. the output phase has a fixed differential from the input phase.

According to control loop theory, the transfer function of the PLL is (assuming N = 1):

$$(\theta_1 - \theta_0)K_P F(s) K_0 / s = \theta_0$$
$$(1 + K_P F(s) K_0 / s)\theta_0 = \theta_1 K_P F(s) K_0 / s \qquad (1)$$



▲ **Figure 1. Block diagram of a PLL.**

$$T(s) = \frac{\theta_0}{\theta_1} = \frac{K_P F(s) K_0 / s}{(1 + K_P F(s) K_0 / s)} = \frac{K_P F(s) K_0}{s + K_P F(s) K_0} = \frac{G(s)}{1 + G(s)}$$

where

$T(s)$ is the closed loop PLL transfer function in the frequency domain (θ_I and θ_0 being the input and output signal respectively),

$G(s) = K_P F(s) K_0 / s$ is the open loop transfer function (complex) of the PLL,

K_P is the transfer function of the phase detector in Volts/ Hz,

K_0/s is the transfer function of the VCO in Hz/ volts, and

$F(s)$ is the transfer function of the loop filter.

Note that this is the transfer function of the PLL when the loop is closed. The block diagram of the PLL is shown in Figure 1.

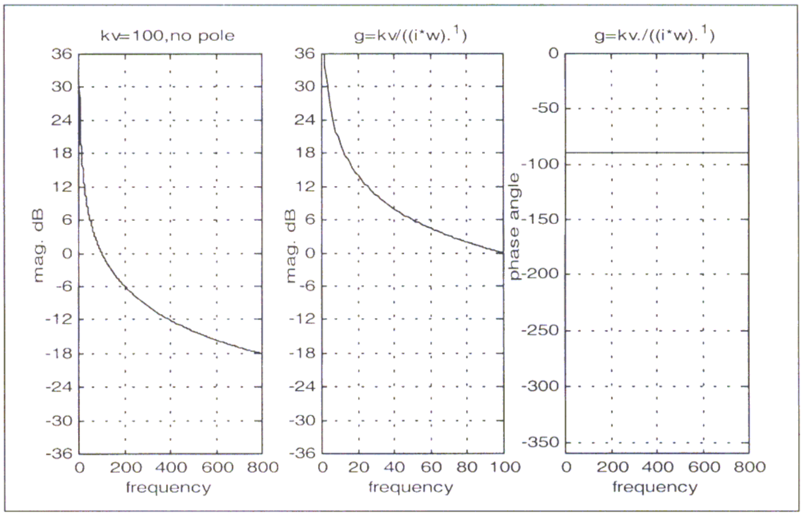
Before we examine the PLL closed loop transfer function in detail, it is important to examine the stability of the PLL. A PLL is unstable when the denominator of the closed loop transfer function is equal to zero. For this to occur,

$$1 + G(s) = 0$$

$$G(s) = -1, \text{ or} \qquad (2)$$

$$G(s) = 0 \text{ dB} @ \angle{-180^\circ} \text{ (magnitude/phase angle)}$$

Hence, the PLL is unstable at the frequency where the magnitude of the open loop transfer function is unity and the phase angle is –180 degrees. Because stability is an important aspect of any PLL design, the condition of unity open loop gain and a phase angle of 180 degrees must be avoided.

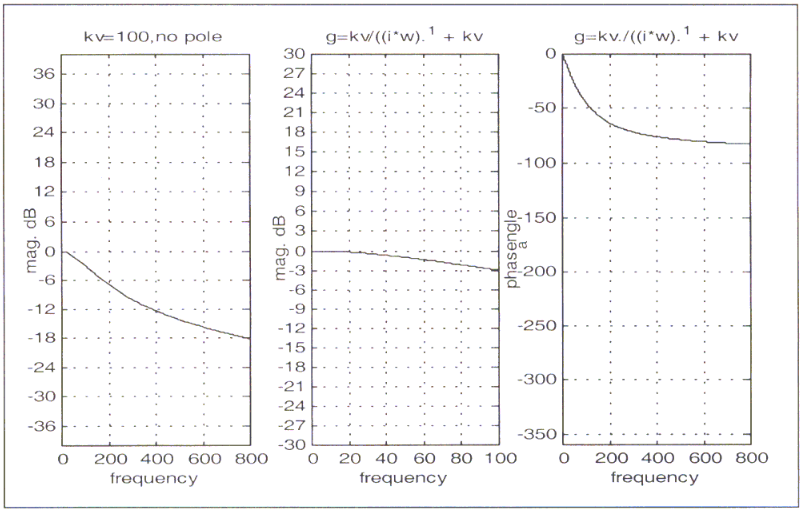


▲ **Figure 2. PLL open loop transfer function with no loop filter.**

PLL response with no loop filter

To further understand the PLL transfer function response and stability, let us examine the case when there is no loop filter. With $F(s) = 1$ (i.e. no loop filter), the PLL closed loop transfer function becomes (3),

$$\frac{\theta_0}{\theta_I} = \frac{K_P F(s) K_0 / s}{\left(1 + K_P F(s) K_0 / s\right)} = \frac{K_P K_0}{s + K_P K_0}$$

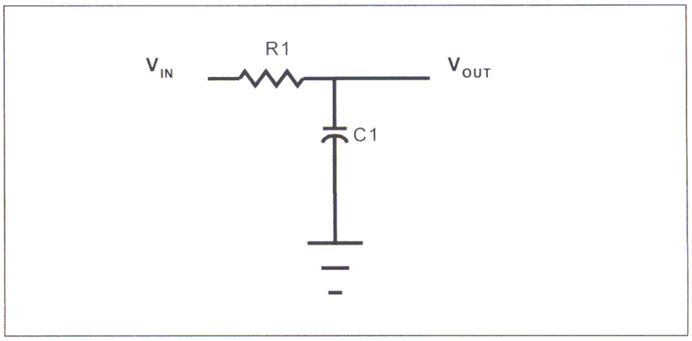


▲ **Figure 3. PLL closed loop transfer function with no loop filter.**

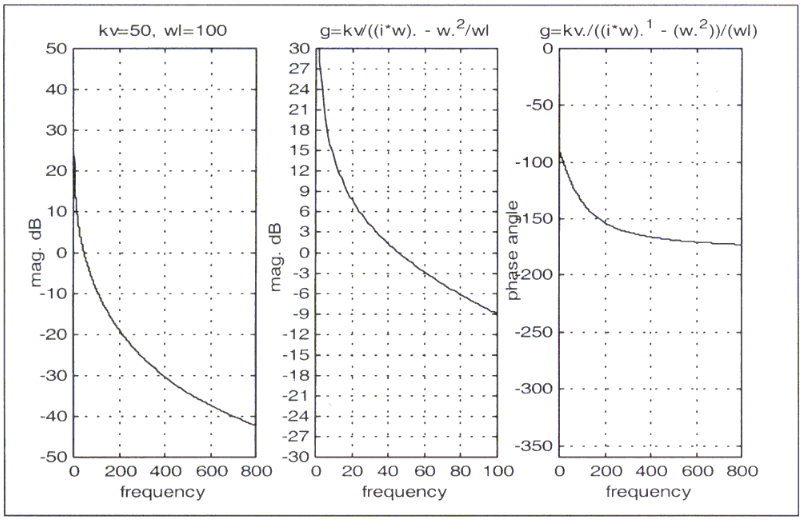
which is the transfer function of a LPF with DC gain of unity and a 3 dB cutoff frequency of $K_P \times K_0$. Therefore, an increase in the DC gain of the phase detector and/or the VCO results in a wider loop, which in turn results in higher phase noise in the PLL. Also, the open loop gain has a slope of 6 dB per octave or 20 dB per decade for all frequencies. The phase angle is always –90 degrees at all frequencies. Hence, with no low-pass filter in the loop, the PLL is always stable, according to the stability criteria. But the main drawback of a PLL design with no loop filter is that the designer has little or no control over the loop response. Figures 2 and 3 show a plot of the open and closed loop transfer function (gain and phase vs. frequency). The plot was done on MatLab using the absolute value function (called *abs*) to compute the magnitude of the transfer function and the angle function (called *angle*) to compute the phase angle. Also, the plot commands — plot (w, abs) and plot (w, angle) — were used to plot magnitude and phase vs. frequency.

PLL with a single loop filter

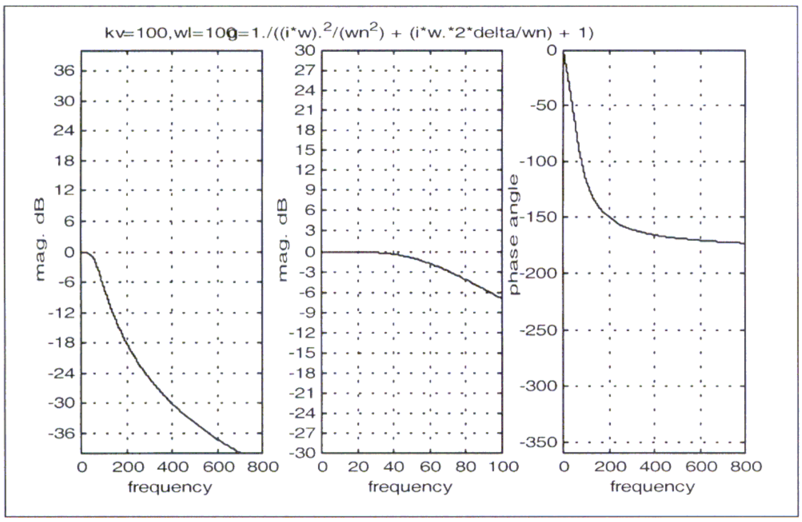
In most PLL designs, a low pass filter is used. The function of a LPF is to filter out any high-frequency harmonics in the loop that might cause the loop to go out of lock, and also to stabilize the loop. Adding a LPF also affects the loop response including parameters such as the loop time response τ_r, loop bandwidth ω_C and the damping factor ∂ of the loop. Figure 4 shows the low



▲ **Figure 4. Single pole loop filter.**



▲ **Figure 5. PLL open loop response with single pole LPF.**



▲ **Figure 6. PLL closed loop response with single pole LPF.**

pass filter that is commonly used in PLL designs. The filter is a one pole low-pass filter with a 3 dB cutoff frequency at ω_L.

Therefore, in this case,

$$F(s) = \frac{1}{s/\omega_L + 1} \tag{4}$$

Substituting equation 4 in equation 1 gives

$$T(s) = \frac{\theta_0}{\theta_I} = \frac{K_P F(s) K_0 / s}{1 + K_P F(s) K_0 / s} = \frac{K_V / s(s/\omega_L + 1)}{1 + \dfrac{K_V}{s(s/\omega_L + 1)}} = \frac{1}{(s^2/\omega_N{}^2) + s(2\partial/\omega_N) + 1} \tag{5}$$

where

$$K_V = K_P K_0$$
$$\omega_N = (K_V \omega_L)^{1/2}$$
$$\partial = \omega_N/(2K_V)$$

Using a low-pass filter with a cutoff frequency of ω_L, the PLL closed loop response is a 2nd order low-pass filter transfer function, centered at the VCO frequency. A characteristic of the second order low pass response is that the slope of the filter drops at a rate of 12 dB/octave. The term ω_N is defined as the natural frequency of the loop and the term ∂ is defined as the damping factor of the loop. Using equation 5 we can derive ω_C, which is

$$\omega_C = \omega_N\left(1 - 2\partial^2 + \left(2 - 4\partial^2 + 4\partial^4\right)^{1/2}\right)^{1/2}$$

for $\partial < 1$

ω_C is the 3 dB bandwidth of the PLL. Knowing ω_C, we can determine the time it takes for the PLL output to rise to 90 percent of its final value, which is approximately

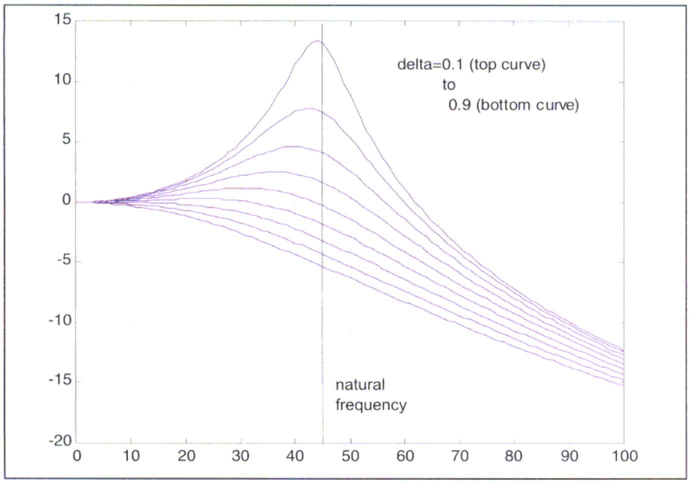
$$\tau_r = 2.2/\omega_C$$

Therefore, given the cutoff frequency ω_L of the low pass filter and the values of K_P and K_0, we can determine the natural frequency ω_N of the loop, and subsequently determine the PLL loop bandwidth ω_C. Knowing this, we can then calculate the time τ_r it takes for the PLL to settle to its final frequency value.

In order to determine the stability of the loop with a single pole low-pass filter in the loop, we must examine the open loop transfer function of the PLL. As stated before, the open loop transfer function is given by the function $G(s)$ and is equal to:

$$G(s) = K_P F(s) K_0 / s = K_P K_0 / s(s/\omega_L + 1) = \frac{K_V}{s(s/\omega_L + 1)} \tag{6}$$

The open loop transfer function has two poles — one at DC and the other at ω_L. Note that at every pole, the gain slope drops at a slope of 6 dB per octave. The gain at DC is infinite; as the frequency increases, the magnitude of the transfer function drops at a slope of 6 dB per octave (due to the presence of the pole at DC). When the frequency reaches ω_L, the gain drops at a slope of another 6 dB per octave (a total slope of 12 dB per octave after



▲ **Figure 7. Closed loop gain response vs. frequency for different damping factors (¶).**

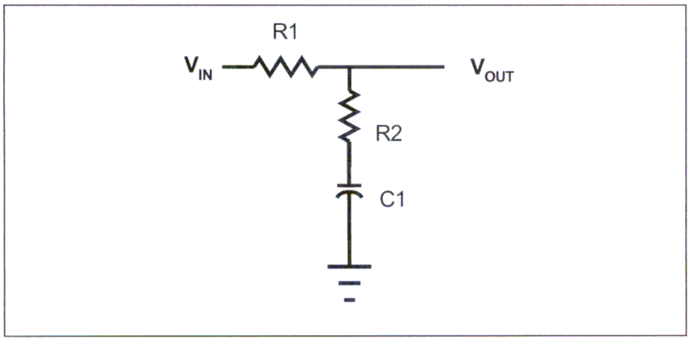
two poles). The PLL is unstable at the frequency where open loop gain crosses the unity gain line at a slope of 12 dB per octave and the phase is –180 degrees. This condition can be observed by plotting the open loop gain and phase response using MatLab. The condition of instability can be avoided by the selection of the pole frequency of the loop filter. As the pole frequency ω_L decreases, the open loop gain intersects the unity gain axis at 12 dB/ octave, the phase angle approaches –180 degrees and the PLL becomes unstable. The PLL approaches instability. This can be observed either by decreasing ω_L while observing the gain and phase response on Matlab or by examining the phase response of the open loop transfer function, which is

$$\varphi = -\tan^{-1} - (\omega_L/\omega)$$

Clearly, as ω_L decreases, the phase angle approaches –180 degrees.

One disadvantage of using a single pole filter is that both the closed loop bandwidth and the damping factor of the closed loop response of the PLL depend on the loop filter bandwidth. The designer cannot independently set the loop bandwidth without affecting the amount of transient overshoot. This deficiency can be easily overcome if a pole-zero loop filter is used. See Figures 5 and 6 for the open and closed loop response of the PLL with a single pole loop filter.

When designing a PLL, it is important to choose the damping factor such that the loop time response has very little overshoot. The percent of overshoot is defined as the time it takes for a PLL to settle at a given frequency. A high percent overshoot can cause the loop to go out of lock. Figure 7 is a plot of the closed loop gain response vs. frequency for different values of damping (from = 0.1 to 0.9 in increments of 0.1, 0.707 being the design goal).



▲ **Figure 8. Pole-zero filter.**

PLL response with a pole-zero loop filter

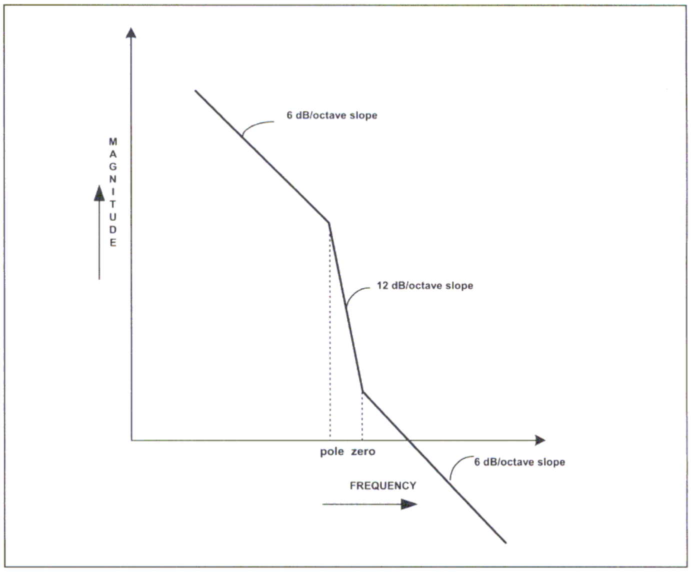
Another way to control the loop response is by using a pole-zero filter in the PLL (see Figure 8). A pole-zero filter is a low pass filter with a pole frequency ω_P and a zero frequency ω_Z. The addition of a pole in the transfer function causes the transfer function slope to drop at a rate of 6 dB per octave whereas the addition of a zero in the PLL transfer function has the opposite effect. For example, the addition of a zero frequency increases the slope by a 6 dB/ octave. This phenomenon is illustrated in Figure 9.

The pole-zero filter transfer response is given by

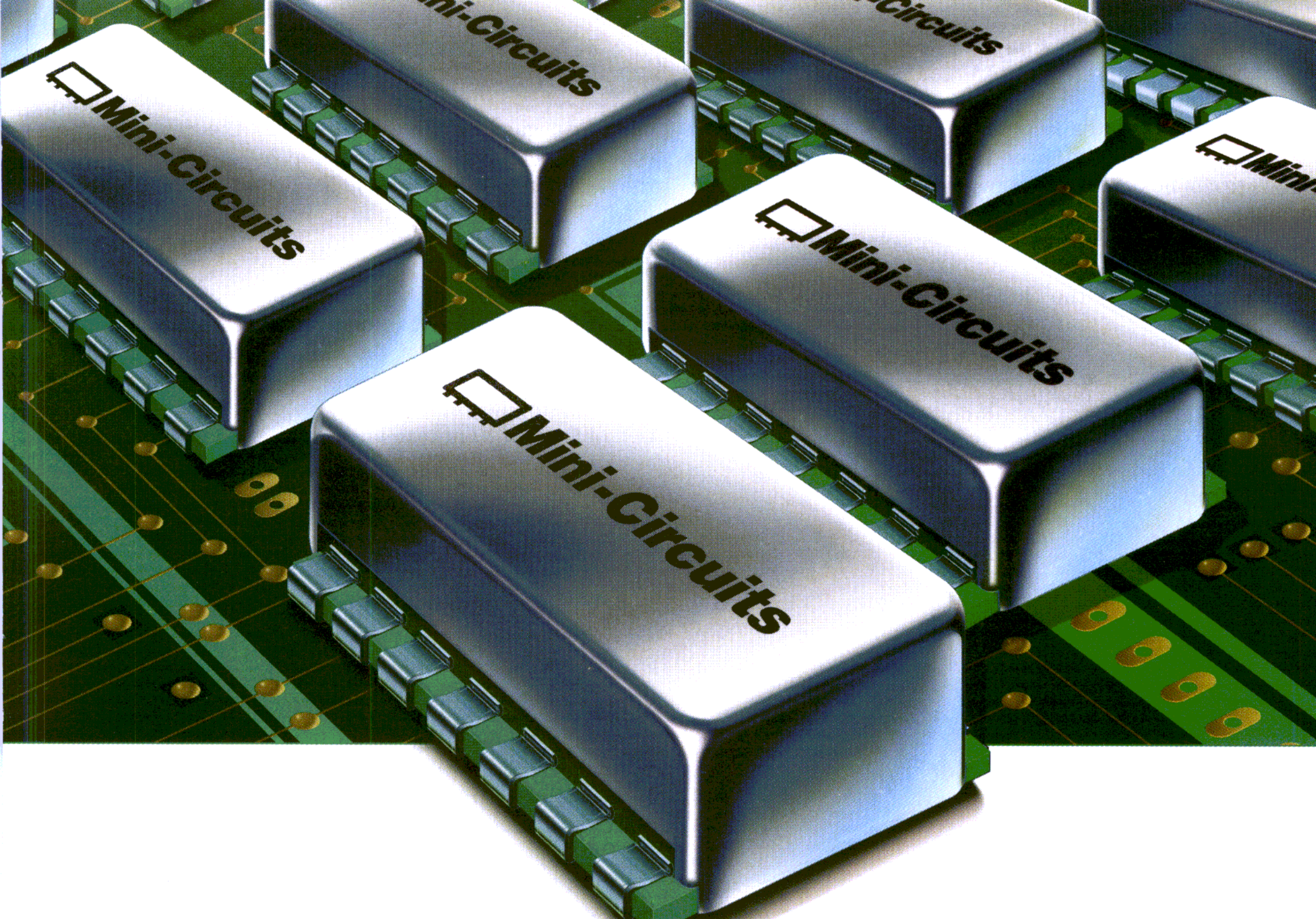
$$F(s) = \frac{s/\omega_z + 1}{s/\omega_p + 1} \tag{7}$$

where ω_Z is the zero frequency and ω_P is the pole frequency. The open loop transfer function is:

$$G(s) = \frac{K_V F(s)}{s} = \frac{K_V(s/\omega z + 1)}{s(s/\omega_P + 1)} \tag{8}$$



▲ **Figure 9. PLL open loop gain response with a pole-zero loop filter.**



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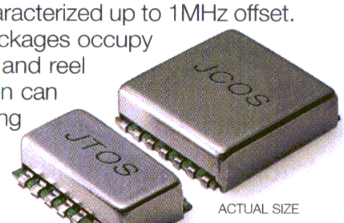

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JTOS-75	37.5-75	-110	-27	16V	20	13.95
JTOS-100	50-100	-108	-35	16V	18	13.95
JTOS-150	75-150	-106	-23	16V	20	13.95
JTOS-200	100-200	-105	-25	16V	20	13.95
JTOS-300	150-280	-102	-28	16V	20	15.95
JTOS-400	200-380	-102	-25	16V	20	15.95
JTOS-535	300-525	-97	-28	16V	20	15.95
JTOS-765	485-765	-98	-30	16V	20	16.95
JTOS-1000W	500-1000	-94	-26	18V	25	21.95
JTOS-1025	685-1025	-94	-28	16V	22	18.95
JTOS-1300	900-1300	-95	-28	20V	30	18.95
JTOS-1550	1150-1550	-101	-20	***	30	19.95
JTOS-1650	1200-1650	-95	-20	13V	30	19.95
JTOS-1750	1350-1750	-101	-16	***	30	19.95
JTOS-1910	1625-1910	-92	-13	12V	20	19.95
JTOS-1950	1550-1950	-103	-14	***	30	19.95
JTOS-2000	1370-2000	-95	-11	22V	30 (@8V)	19.95
JTOS-3000	2300-3000	-90	-22	***	25 (@5V)	20.95
JCOS-820WLN	780-860	-112	-13	***	25 (@9V)	49.95
JCOS-820BLN	807-832	-112	-24	14V	25 (@10V)	49.95
JCOS-1100LN	1079-1114	-110	-15	***	25 (@8V)	49.95

Notes: *Prices for JCOS models are for 1 to 9 quantity. **Required to cover frequency range. ***Tuning Voltage for JTOS-3000 is 0.5 to 12V, JTOS-1550, JTOS-1750, and JTOS-1950 is 0.5 to 20V, and JCOS-820WLN and JCOS-1100LN is 0 to 20V. For additional spec information and details about 5V tuning models available, consult RF/IF Designer's Guide, our Internet Site, or call Mini-Circuits.

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In this case, the location of the pole is always before the zero frequency. Given the pole frequency location, a zero can be placed after the pole (as shown in the Figure 9) so as to avoid the magnitude from crossing the unity gain axis at a slope of 12 dB per octave, and therefore avoiding instability.

To determine the closed loop response, simply plot $T(s)$,

$$T(s) = \frac{\theta_0}{\theta_I} = \frac{K_P F(s) K_0 / s}{\left(1 + K_P F(s) K_0 / s\right)} \qquad (9)$$

where

$$F(s) = \frac{s/\omega_Z + 1}{s/\omega_P + 1}$$

The transfer function of $T(s)$ is

$$T(s) = \frac{(s\omega_Z + 1)}{s^2/\omega_P K + (1 + K/\omega_Z)s + 1} \qquad (10)$$

where $\omega_N = (\omega_P K)^{1/2}$ and

$$\partial = (.5)(1/K + 1/\omega_Z)\omega_N$$

Therefore, selecting the pole frequency sets the natural frequency (and subsequently the loop bandwidth) and selecting the zero (based on the pole location in the open loop gain response) determines the desired percentage overshoot. Therefore, a pole-zero filter allows the designer to select the loop bandwidth and the damping factor independently and still achieve stability.

Summary

Steady-state stability is an important criterion in PLL design. Stability can be determined by examining the transfer function of the PLL in an open state. As seen, a condition of open loop gain of unity and a phase angle of –180 degrees must be avoided for stable operation of the PLL. This can be accomplished by the proper selection of the loop filter parameters. ■

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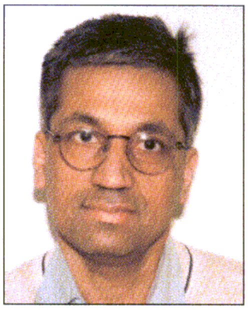
2. G. Nash, "Phase-Locked Loop Design Fundamentals," Motorola Application Notes (AN-535), Motorola Semiconductors, Phoenix, AZ.

Author information



Arun M. Mansukhani is an RF Systems Engineer with Motorola, Inc. in Piscataway, NJ. He has more than 16 years of experience in RF design. Presently, he is working on next-generation digital cellular system design. Please e-mail him for a copy of the PLL MatLab programs. He may be reached at 732-878-8435 or by e-mail at Arun_Mansukhani-W15392@email.mot.com.

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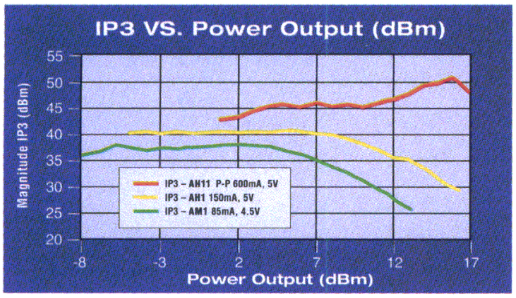

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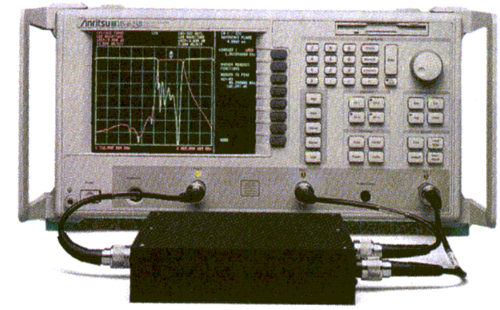
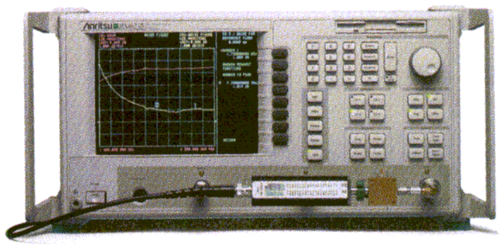
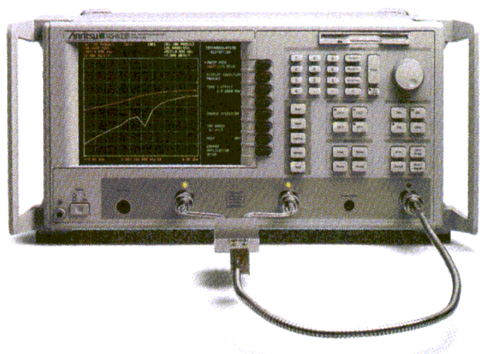
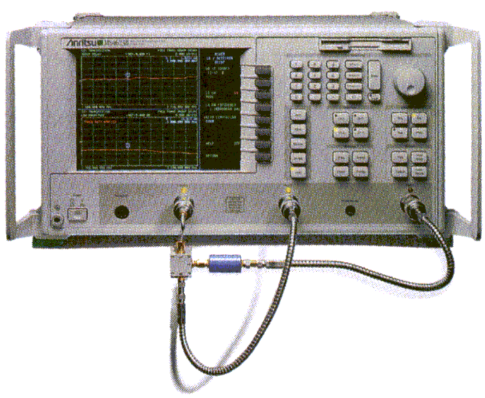








An ASIC Driver for GaAs FET Control Components

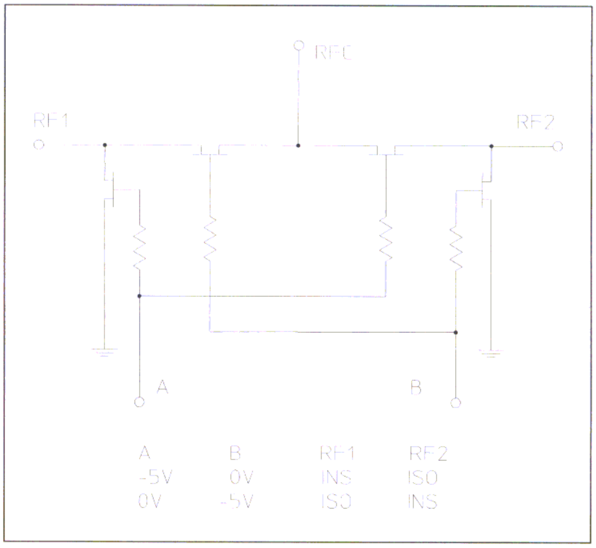
This application note covers driver selection and circuit design for FET switches and attenuators, including non-linear performance as V_{opt} is varied

By Christopher Weigand
M/A-COM, Inc.

The use of a plastic packaged ASIC driver enables the designer to convert from a TTL or CMOS input voltage to the complementary drive signals required for GaAs FET switches and digital attenuators. These ASICs are available in single (SWD-109) and quad (SWD-119 and DR65-0001) channel devices.

GaAs FET control components require complementary drive voltages, because most switching devices have both series and shunt FETs. An example is the case of a SPDT switch, with the series shunt configuration shown in Figure 1. In the "on" arm, the series FET would be on and the shunt FET would be off. When a sufficient negative voltage is applied to the gates of the FETs, they are turned off. This negative voltage gradually becomes lower as the pinch-off voltage of the FETs is lowered for low voltage applications. The negative operating voltage for "off" FETs was typically –7 volts when GaAs control components were first introduced. This standard is now typically –5 volts, with many GaAs control components operating at –3 volts in battery powered devices.

Because the FETs operate in depletion mode, they are "on" when no volts are applied to the gate, and "off" when a negative voltage less then the pinch-off voltage is applied to the gate. It should be noted that GaAs FETs in control components have the same doping in both the source and drain, and that the gate is centrally located between the source and drain. Therefore, the source and drain are interchangeable. Most data sheets show that the control voltage for an on FET is between –0.2 and 0 volts. Due to the increasing use of FETs with lower pinch-off voltages, it is recommended that CMOS logic be used to drive GaAs FET control components so that the "low" control voltage for on FETs is between –0.1 and 0.0 volt.



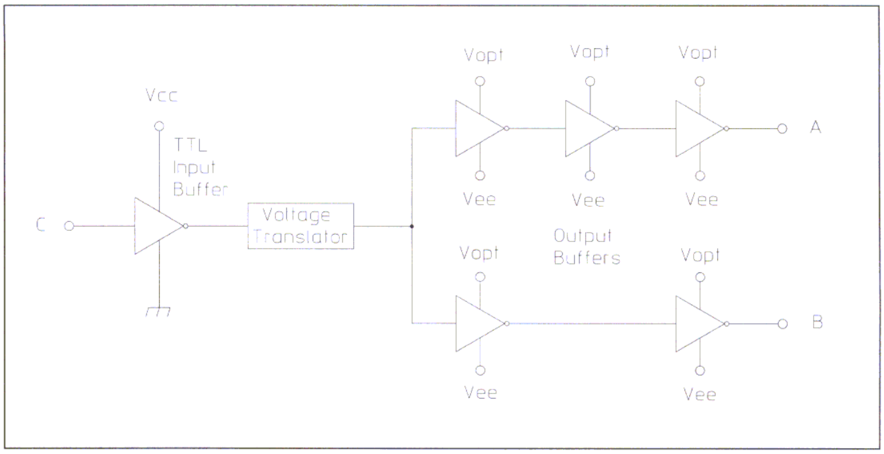
▲ **Figure 1. A typical SPDT FET switch circuit.**

Block diagram of ASIC GaAs FET driver

There are three basic elements in a GaAs FET driver circuit — the TTL input buffer section, a voltage translator, and complementary output buffer stages. The input section has TTL trip points, but it was designed to accept CMOS drive circuitry as well. There is an advantage to driving the ASIC with CMOS in that the positive quiescent current will be under 100 μA when a CMOS control is used. ΔI_{cc} is approximately 1.5 mA at V_{cc} of – 2.1 volts, which means that the positive bias current will increase when the input logic is a TTL high.

The TTL input stage provides an input to the voltage translator that is either 0 volts or 5

volts. The voltage translator takes the signal from the output of the TTL buffer and translates it down to between V_{ee} (the negative supply voltage), and V_{opt}. The output buffering serves two purposes. There is an additional inverting buffer on the path to the A_n output than there is going to the B_n output, which provides the complementary outputs of V_{opt} and V_{ee}. The successive buffer stages use larger transistors at each stage to be able to drive capacitive loads on the control lines.



▲ **Block diagram of an ASIC driver.**

Advantages of using an ASIC FET driver

There are several key advantages of using an ASIC silicon driver to drive GaAs circuitry. A silicon driver has quiescent currents that are in microamps, where on-chip GaAs drivers use several milliamps. Because the silicon driver provides complementary V_{opt} and V_{ee} voltages, the GaAs MMIC can operate down to DC when V_{opt} is set to zero volts. With an on-chip GaAs driver and a standard CMOS drive, the RF input would need to be centered at +5 volts potential, which would require pull-up resistors and DC blocking capacitors. Because the silicon ASIC driver provides negative control voltages, the ground bond pads on the GaAs MMICs can be bonded directly to ground, which improves the high frequency performance compared to floating grounds, which would be required if standard CMOS drivers were used with 0 volts and V_{cc} for the rails.

M/A-COM sells three ASIC drivers. DR65-0001 is a quad FET driver designed to be a low cost driver. To minimize cost and die size, the load capacitance requirement was dropped from 25 pF to 4 pF. SWD-119 is a quad FET driver that has fast switching speed, on the order of 20 ns typical, and is designed to have a pinout that is directly compatible with quad digital attenuators sold by M/A-COM. SWD-109 is a single channel FET driver with the same performance as SWD-119.

Interface with RF switches

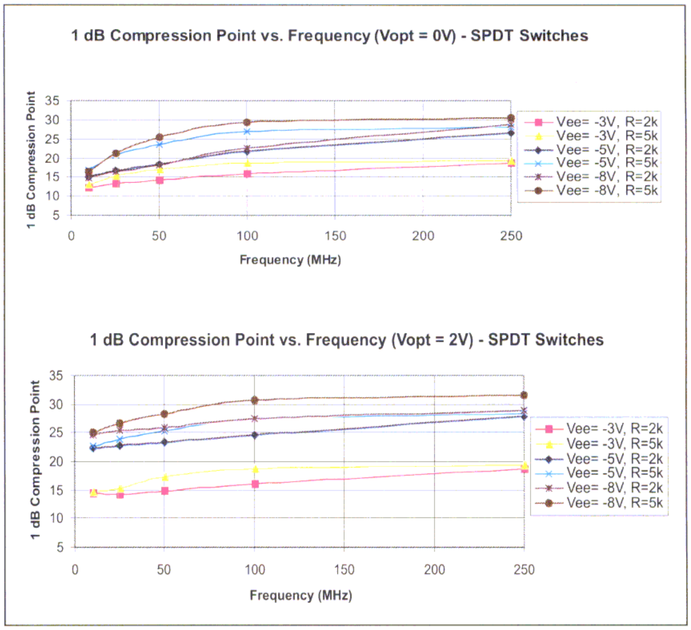
For RF switches, it is usually desirable to have a TTL "1" be the control for the "on" arm. For this to be true, the A output of the FET drivers is connected to the series FETs and the B output is connected to the shunt FETs. To maximize switching speed, the FET driver should be located physically close to the switch.

For fast RF switching the RF lines of the switch need to be at ground potential (see Figure 7). The control voltages on a switch are the gate voltages. V_{gs} should be 0 volts for an "on" FET, and –5 volts typically for an "off" FET. If the RF lines are blocked by capacitors, the source voltage is DC blocked, so Vgs is not well defined and will vary depending on the MMIC leakage currents. When a DC block is required, in most cases a 10 kohm shunt resistor or a shunt inductor on the RF common arm will usually suffice to provide adequate switching performance.

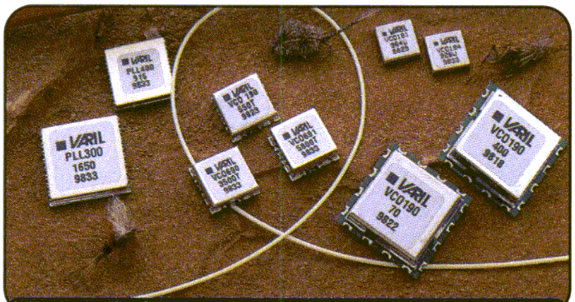
Interface with digital attenuators

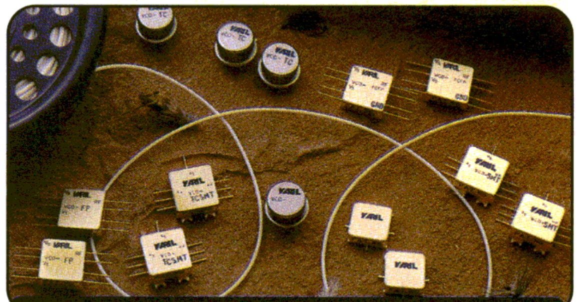
The SWD-119 is designed to interface directly with digital attenuators such as the AT-220 four bit attenuator, as shown in the diagram of Figure 8. With the controls to the SWD-119 coming in from the left, the drive lines to the FETs are a direct pin-pin connection to the GaAs FET attenuator. The logic is set up so that a control "0" gives the insertion loss state, and a control "1" sets a bit to the attenuation state.

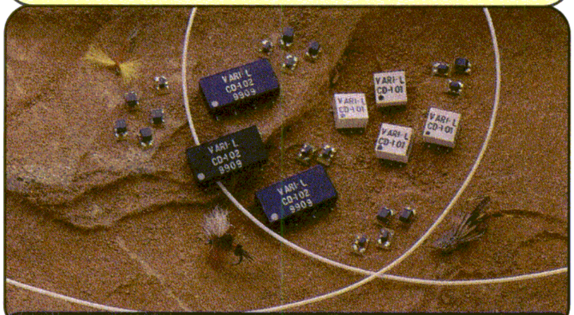
To drive a 5 bit attenuator a single and quad driver would be used. In most applications the driver outputs are connected directly to the attenuator inputs. In

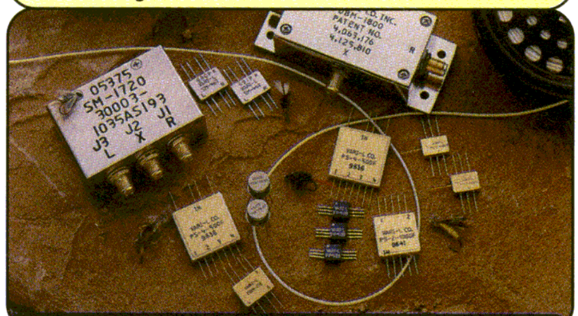


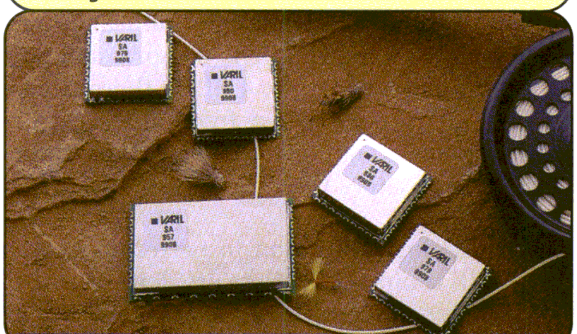
▲ **Figure 3. 1 dB compression point vs. V_{opt}, V_{ee}, gate resistors, and frequency (DC to 250 MHz).**

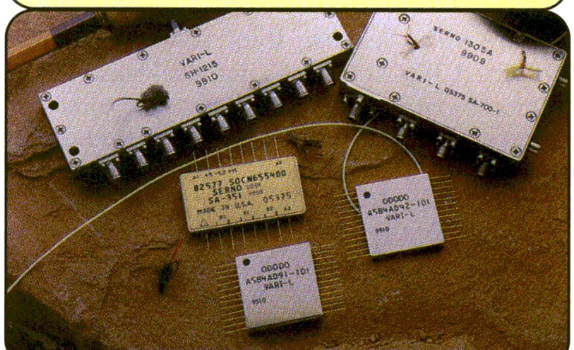


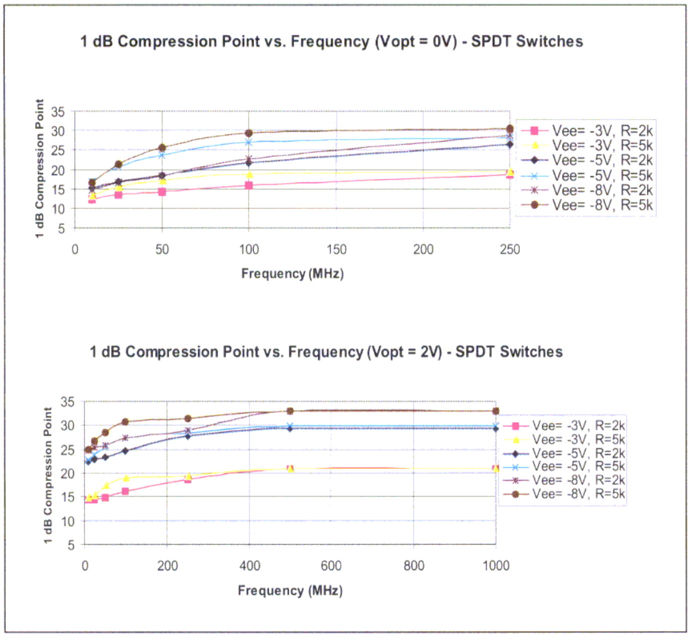




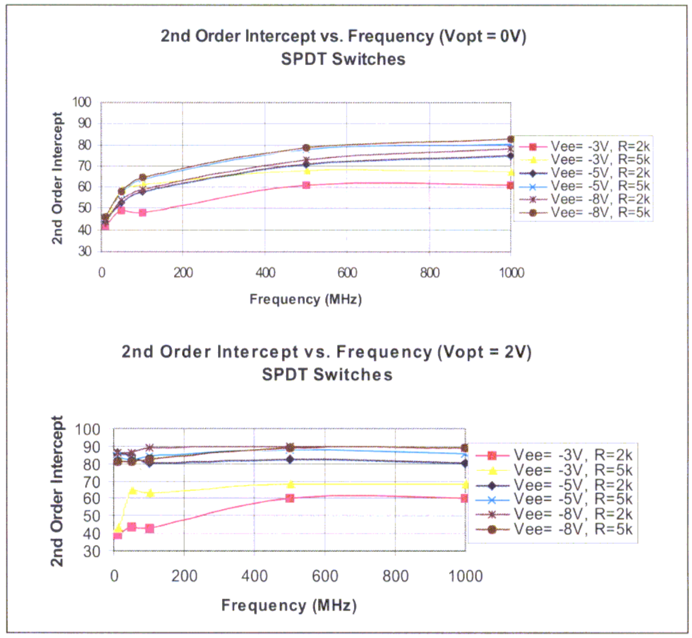








▲ **Figure 4. 1 dB compression point vs. V_{opt}, V_{ee}, gate resistors, and frequency (DC to 1000 MHz).**



▲ **Figure 5. Second order intercept vs. V_{opt}, V_{ee}, gate resistors and frequency.**

instances where video leakage from the control lines must be minimized, filtering can be installed between the driver and the attenuator. The filtering would consist of a single shunt capacitor in most cases, which would slow the switching speed.

Most of the newer M/A-COM attenuators have an internal 10 kohm shunt resistor, but in older attenuators it may be necessary to provide a DC return at either the RF IN or RF OUT port.

Without the DC return, all bits will be accurate relative to each other, except that the reference loss state will have degraded performance, and may have up to 0.5 dB more loss than it would be with a DC return. The other bits do not exhibit this problem because they contain a shunt element that provides a DC return.

Non-linear performance with V_{opt} = 0 V and 2 V

1 dB compression, second order intercept, and third order intercept were evaluated as a function of V_{opt}, V_{ee}, gate resistors, and frequency. This was done to reevaluate the non-linear performance after the pinch-off voltages of the FETs were lowered and the ion implanter was changed to increase the FET breakdown voltages. We measured the non-linear performance of two SPDT switches. One switch was the SW-313, which uses the MASW4030G die, which has 5 kohm gate resistors. In the other switch, we replaced the MASW4030G die with the MASW6010G die, which has 2 kohm gate resistors. See the graphs in Figures 3, 4, and 5 for plots of 1 dB compression, second order intercept, and third order intercept vs. V_{opt}, V_{ee}, gate resistors and frequency.

There are some key points to note relative to the plots. For the –3 volts data, there are significant unit-to-unit variations below 100 MHz. The compression points in a single wafer lot varied up to 4 dB so there needs to be a significant guard band at low frequencies. With –3 volts control, there was significantly better performance with 5 kohm gate resistors than there was with 2 kohm gate resistors.

There was little variation in performance above 500 MHz so 1 dB compression plots are done over two ranges to better demonstrate the performance at low frequencies. The performance at low frequencies will be comparable for IP2 and IP3.

Switching speed vs. gate resistance

There are two primary factors that limit the switching speed of GaAs FET switches, the transmission line inductance between the driver and the MMIC and the RC time constant on the MMIC. The RC time constant on the MMIC will increase by a factor of 2.5 as the gate resistance is increased from 2 kohm to 5 kohm. Unless very fast switching speed is required, a 5 kohm is sufficient for adequate switching performance.

Summary

The basic performance of the ASIC driver for GaAs control components was described. Special attention was paid to the performance as V_{opt}, R_{gate} and V_{ee} were varied as a function of frequency. For 3 volt performance there was significant unit-unit variation at low frequencies, and increasing gate resistance improved non-linear performance. At V_{ee} = –5 volts there was improvement in the non-linear performance, with only small changes

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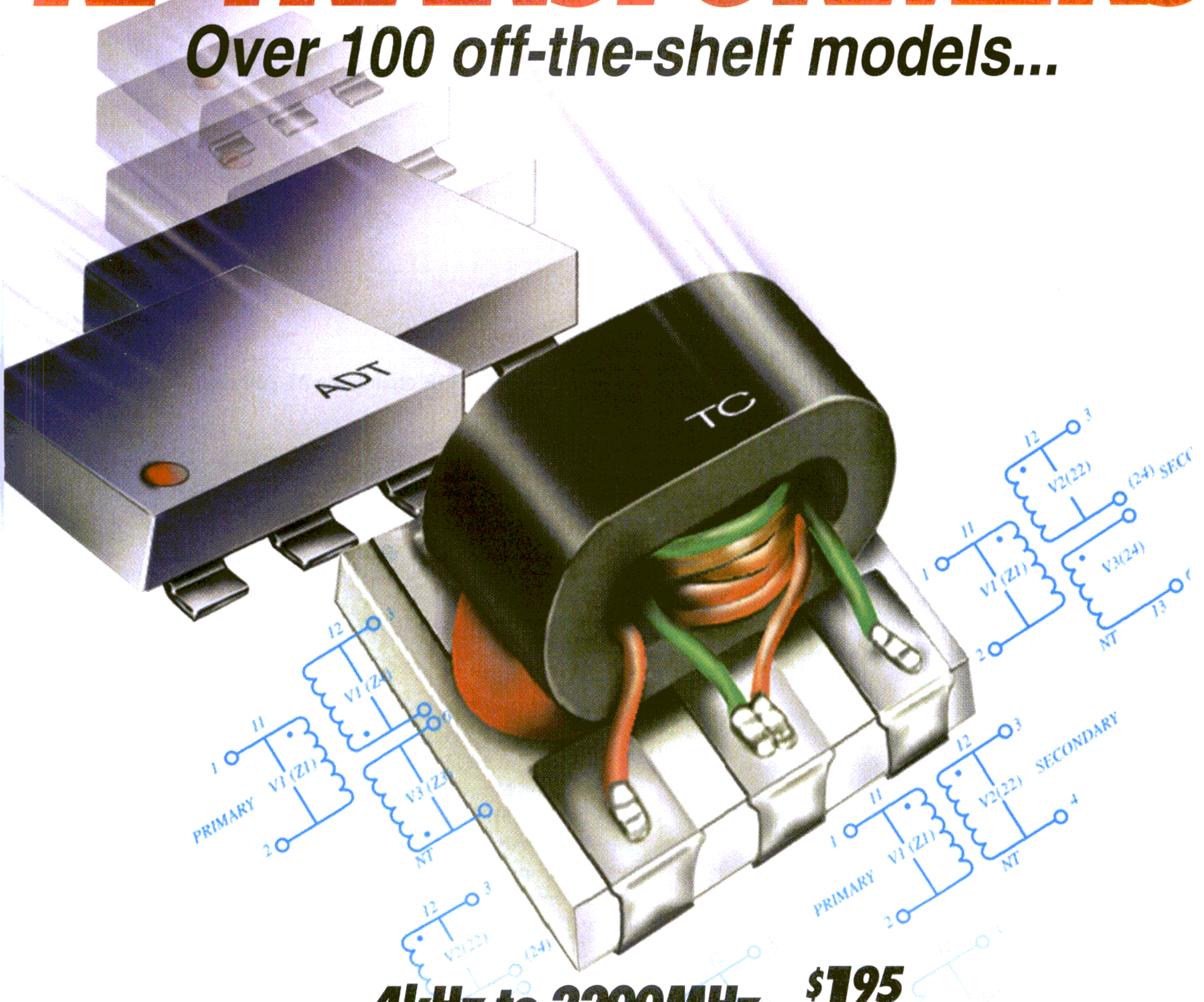

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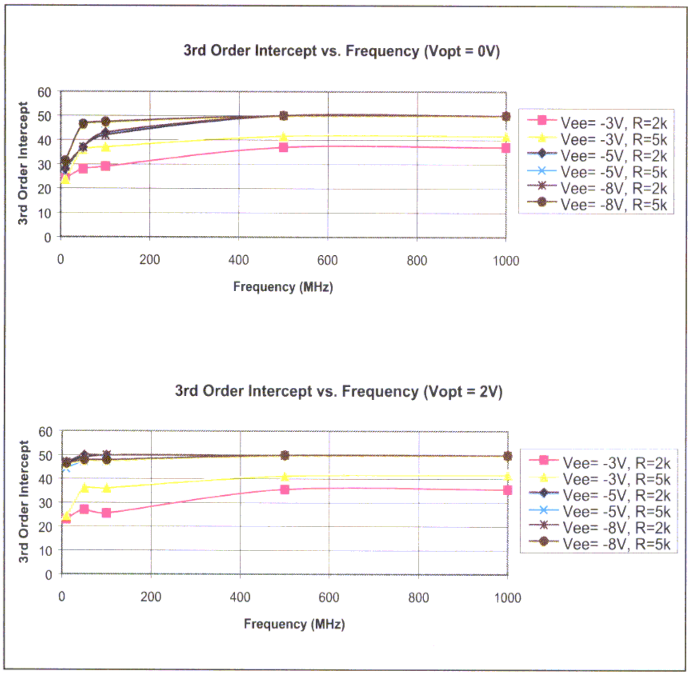
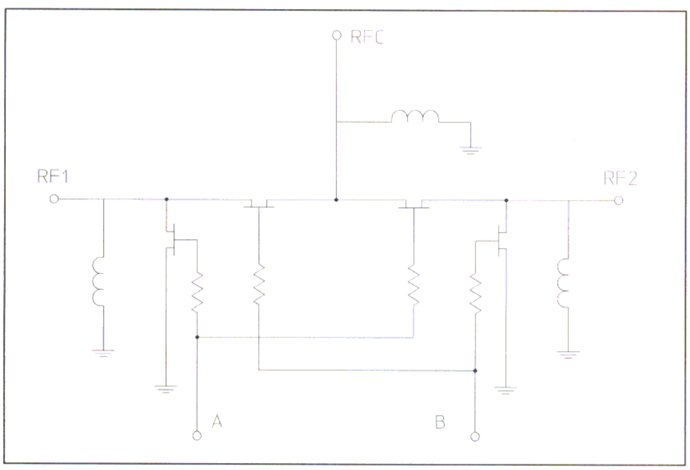
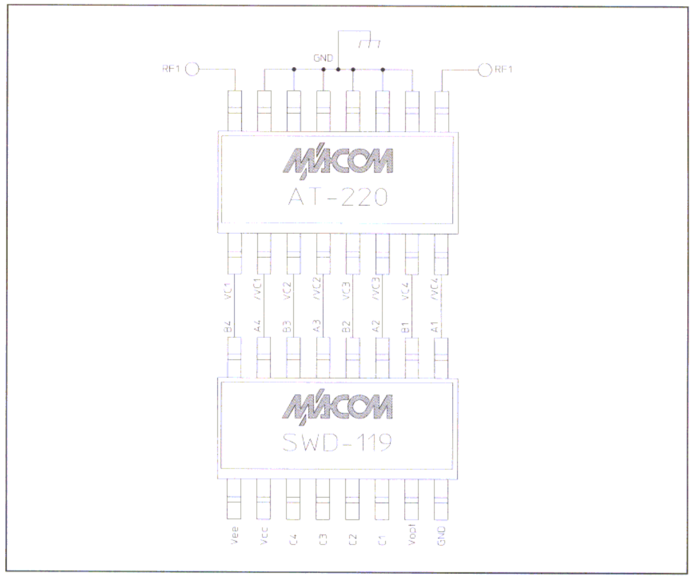


▲ **Figure 6. Third order intercept vs. V_{opt}, V_{ee}, gate resistors and frequency.**

as gate resistance was varied. There was still fairly significant roll-off in non-linear performance with V_{ee} = –5 volts. With V_{ee} = –8 volts, and V_{opt} = +2 volts, the non-



▲ **Figure 7. SPDT switch with GND returns on RF ports.**



▲ **Figure 8. SWD-119 driving the AT-220 digital attenuator.**

linear performance was the best. There was much less variation across frequency for V_{ee} = –8 volts.

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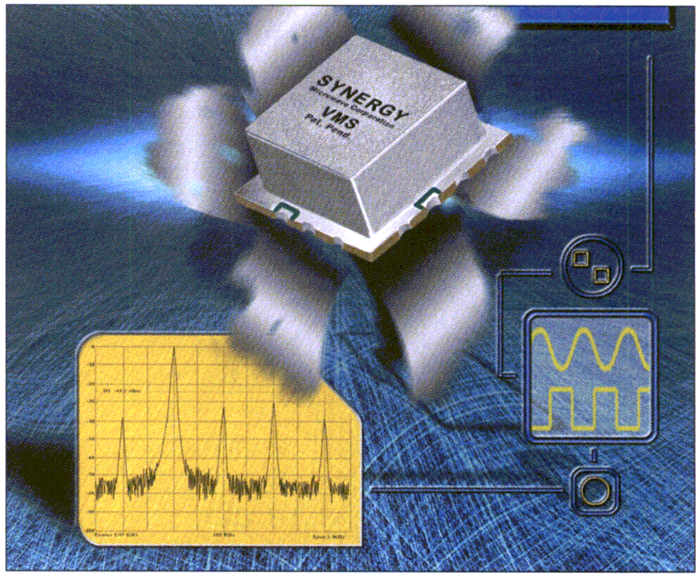
Author information

Christopher Weigand received a BSEE from Tufts University in 1977 and an MSEE from Tufts University in 1979. From 1979 until 1981, he worked at Hughes Aircraft Company on the design and development of satellite ground stations. From 1981 until 1989, he designed passive components, phase modulators and subassemblies for ANZAC. He has been with M/A-COM since 1989, and currently is a Principal Engineer in the RF & Microwave Components product line. He has designed GaAs MMIC switches, digital attenuators and digital phase shifters. He has been the project engineer for the PHS diversity switch product line. He may be reached via email at weigand@tycoelectronics.com.

A High Dynamic Range Miniature Modulator/Demodulator

This month's cover features a new patent-pending product line with low noise figure and high isolation

Most commercially available integrated modulators/demodulators operating at cellular band frequencies and higher require many external components and lack the high dynamic range and low noise figure demanded by modern communication systems. Synergy's unique patent pending VMS series of I/Q modulator and demodulator technology allows the designer to select a low cost design with a high dynamic range in a package as small as 0.5 in × 0.5 in × 0.22 in., capable of withstanding harsh reflow environment during automated assembly. The package is EMI/RFI protected with moisture resistant capability.

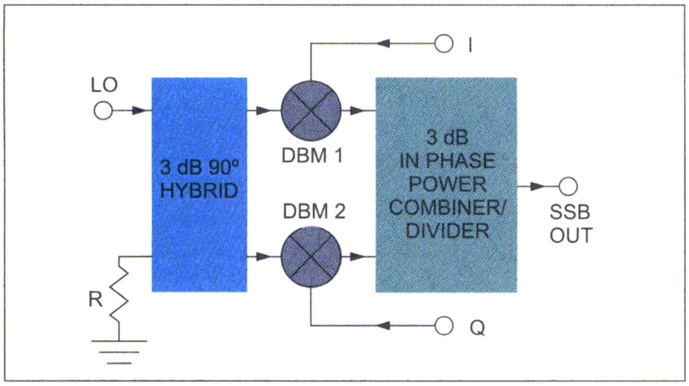


▲ **Synergy Microwave announces the VMS series of I/Q modulator/demodulator products.**

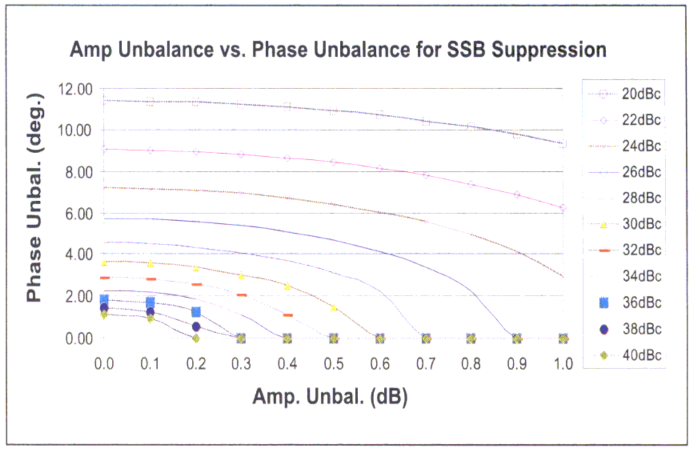
Description

Many manufacturers of active modulators claim high dynamic range when operated with low level (–20 dBm or lower) signals. Active modulators based on Gilbert Cell technology need plenty of external circuitry, including supply decoupling and tuning networks. When direct current (DC) is involved for biasing the internal devices, designers may run into unwanted spurious signals unless special care is taken in decoupling various parts of the modulator. The external circuitry also increases the required p.c. board real estate. Another suggested approach makes use of sub-harmonic modulators where the even harmonic of the local oscillator (LO) mixes with two incoming intermediate frequency (IF) signals of the same strength. The two IF signals are separated by 90 degrees in phase, resulting in either an upper sideband or a lower sideband RF signal in the case of a single sideband (SSB) modulator. In the case of a demodulator, this approach will result in two IF signals of similar strength differing in phase by 90 degrees, commonly known as a quadrature IF mixer. The carrier rejection is high due to the fact that the sub-harmonic mixers used in the modulator offer higher isolation at the even harmonics of the LO. However, the signal handling capability may be lower (depending on the design) than that of a fundamental type, resulting in inferior intermodulation products.

Figure 1 shows the block diagram of a typical modulator where two double balanced mixers (DBMs), whose IF ports are DC coupled, are interconnected with a 90 degree hybrid on one



▲ **Figure 1. Block diagram of a typical modulator where IF ports are DC coupled and are interconnected with a 90 degree hybrid on one side and an in-phase divider/combiner on the other side.**

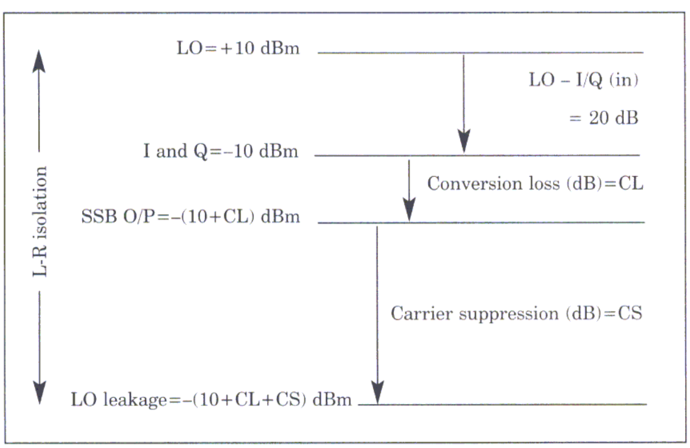


▲ **Figure 2. The plot of amplitude unbalance vs. phase unbalance for various values of SSB rejection.**

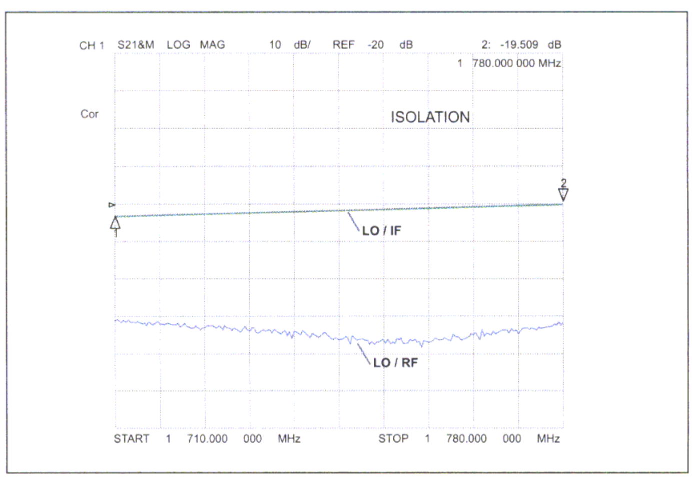
side and an in-phase divider/combiner on the other side. The carrier rejection of the modulator is directly related to the LO to RF isolation of the mixers, whereas, the single sideband (SSB) rejection is related to the combined amplitude and phase balance of the various components within the modulator. Figure 2 shows the plot of amplitude unbalance vs. phase unbalance for various values of SSB rejection. From the figure, it is evident that for 40 dB SSB rejection, one has to maintain the phase and amplitude balances as low as 1 degree and 0.1 dB respectively. The noise figure (NF) in decibels (dB) of the passive modulator is typically no more than 1 dB higher than the conversion loss of the modulator, whereas active modulators have poorer noise figure resulting in lower signal-to-noise ratio.

The modulator used in the VMS series (patent pending) reduces component count, including transformers, which are the main source of degraded phase and amplitude response at higher frequencies. The transformers used in this modulator provide higher frequency coverage with excellent phase and amplitude balance, crucial to higher sideband rejection. Also, the transformers provide improved isolation resulting in better carrier rejection. The 1 dB compression point of the modulator, coupled with carrier and sideband rejection, determines the dynamic range of the modulator. The 1 dB compression can be improved (different model) by using higher level Schottky diodes (four are used to form a ring/cross over quad), where the level is expressed by the forward voltage drop (V_f) of the diode for a given current. There is a practical limit to this. The typical values for V_f are 225 mV, 350 mV and 550 mV. One can take advantage of the fact that the effective value of V_f can be increased by connecting as many diodes as practical in series, keeping in mind the package size and the frequency response limitations. When too many diodes are housed in a single package (multiples of 4 or as required), the cost becomes prohibitive.

Assuming the package size and frequency response of the diode is the same irrespective of the V_f, the basic design of the modulator remains unchanged. The bandwidth limitation comes from that of the 90 degree hybrid used in the modulator. For the smallest unit, the bandwidth is typically 10 percent. The third order intermod-

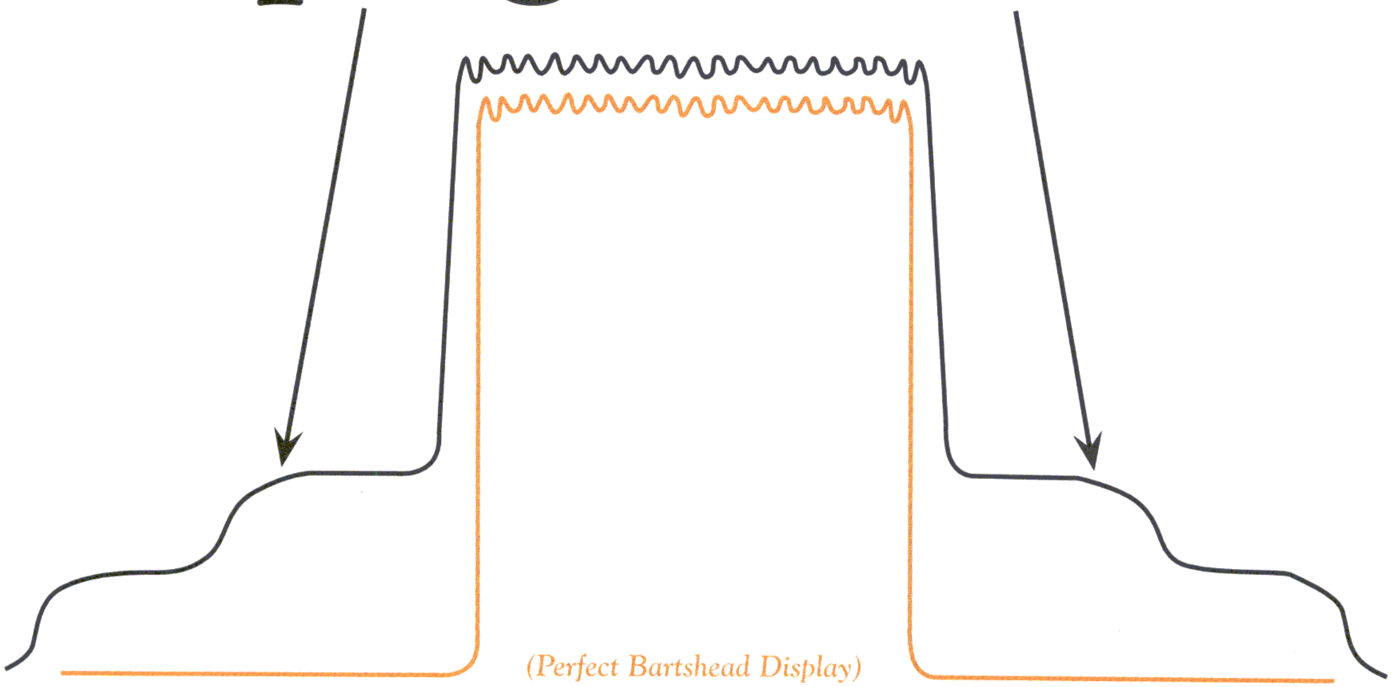


▲ **Diagram 1. Relationships to help calculate the LO/RF isolation for each of the DBMs used in the modulator.**



▲ **Figure 3. Typical isolation for a PCS band unit.**

Danger. Sloping Shoulders.



Spectral growth can compromise the most precise CDMA technologies. Often, the difficulty lies in the transmission equipment, a potent source of distortion or the "shoulders" you see on your display. Now there are amplifiers that help you eliminate shoulders, keep the space between carriers narrow and optimize bandwidth.

AR "S" Series amplifiers are uniquely linear. Thanks to their exceptional design, CDMA signals get amplified but distortion stays low. You quickly pinpoint problems when testing your driver amp (or other transmission equipment) because you know the one place it's not coming from (the AR test amp).

"S" Series amplifiers offer a broad band that accommodates the 0.8 to 0.9 GHz, 1.85 to 1.99 GHz and 3.5 GHz frequencies used in wireless transmission. Plus spurious emissions and noise figures that are unusually low.

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Circle 32

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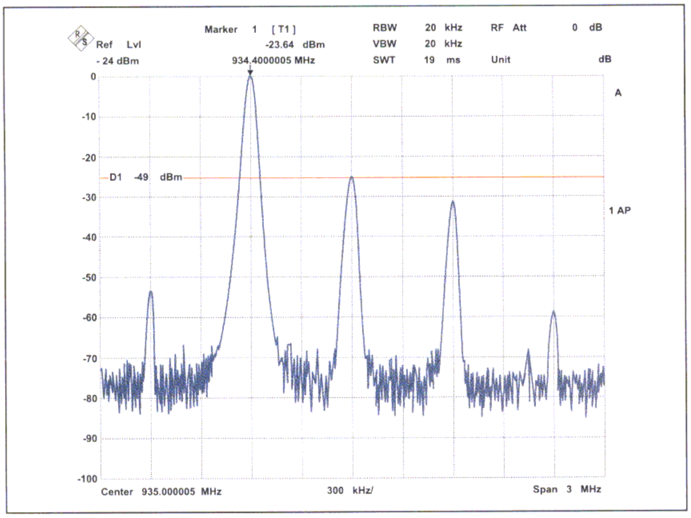
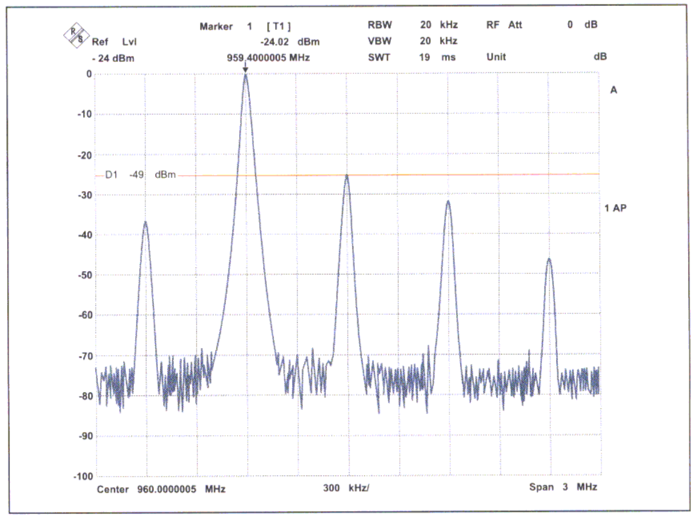


▲ **Figure 4. Carrier and sideband performance at low cellular band for the VMS-935 model.**



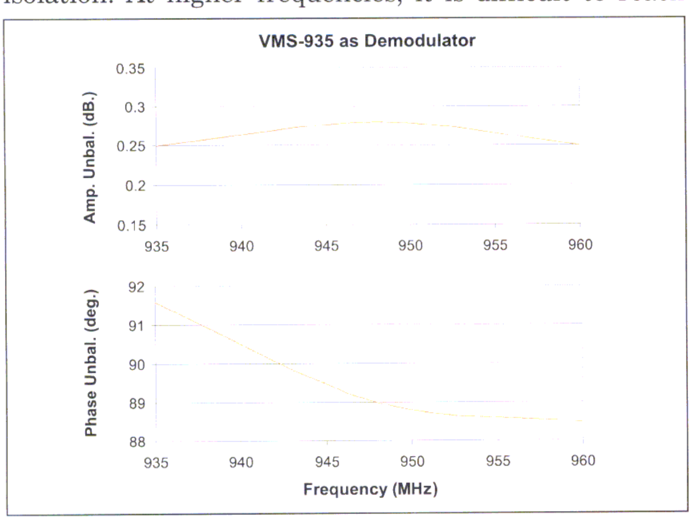
▲ **Figure 5. Carrier and sideband performance at high cellular band for the VMS-935 model**

ulation product (IP3) depends on the V_f of the Schottky diode, the L/R, L/I and R/I isolations in the mixers and the impedance match at all ports.

Importance of LO/RF isolation

As pointed out earlier, the LO to RF isolation determines the carrier rejection in the modulator. Diagram 1 helps the designers to calculate the LO/RF isolation for each of the DBMs used in the modulator. The isolation number for each of the DBMs will decrease by 3 dB, due to the combining effect in the in-phase combiner. In the diagram below, we will assume that the modulator is being driven by a LO drive of +10 dBm and the I and Q levels are –10 dBm each.

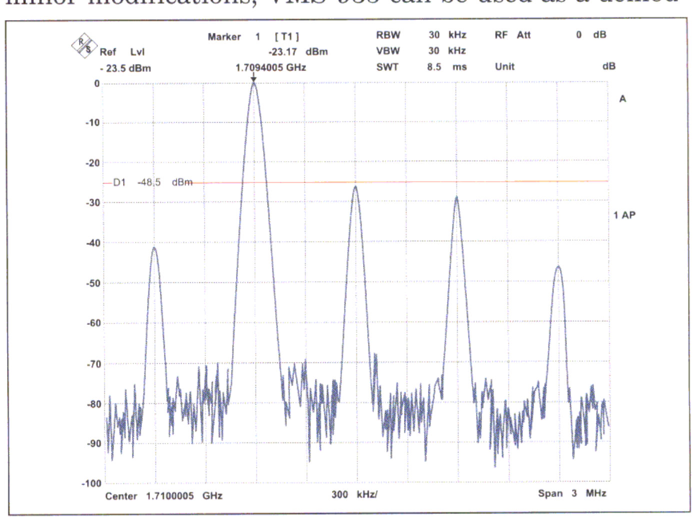
For a carrier rejection of 30 dB and a typical conversion loss of 5 dB, the LO/RF isolation of the modulator will be 55 dB and each of the DBMs will have 52 dB of isolation. At higher frequencies, it is difficult to reach this figure with a conventional approach, while the new approach can easily achieve this in smaller size. Figure 3 shows a plot of typical isolation for a PCS band unit. LO/RF isolation is better than 50 dB, while the LO/IF isolation is 20 dB.



▲ **Figure 6. The amplitude and phase unbalances across the full band of 935-960 MHz.**

Performance

Two models, VMS-935 and VMS-1710, operating in the popular 935 to 960 MHz cellular and 1710 to 1780 MHz PCS frequency bands, are selected to highlight the performance of the new technology. For each of these applications, the I and Q levels are set at –10 dBm each with a 90 degree phase difference between them at 600 kHz, while the LO level is fixed at +10 dBm. Figures 4 and 5 display the carrier and sideband performance in the cellular band for the VMS-935 model, showing that the SSB rejection is 30 dB (min.) across the band. With minor modifications, VMS-935 can be used as a demod-



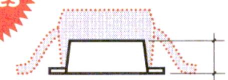
▲ **Figure 7. Carrier and sideband performance at low PCS band for the VMS-1710 model.**

New NEC Bipolar Transistors

Higher f_Ts, Lower 1/f

New, Smaller Packages

NEW Packages



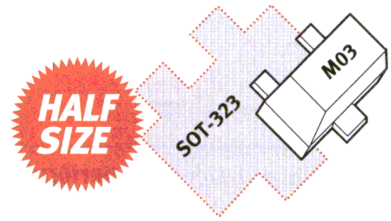
- **Flat Lead** design reduces parasitics and improves electrical performance
- **Low Profile** package is ideal for PCMCIA and other space-constrained designs

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.*



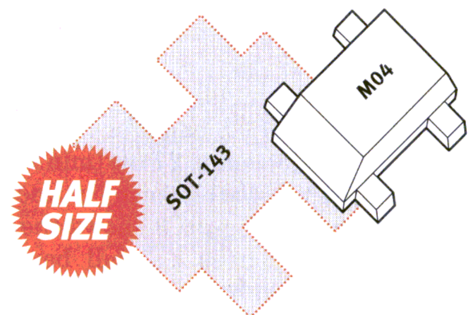
New M03

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**

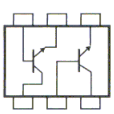


New M04

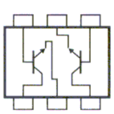
Half the footprint area of a SOT-143

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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Circle 29

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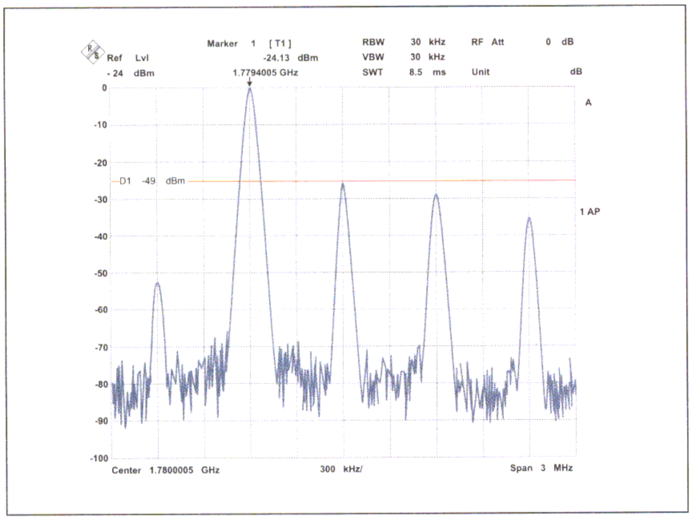
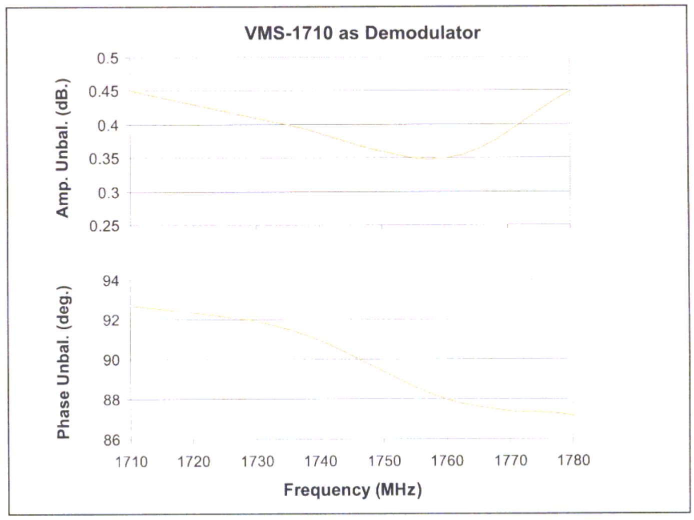


▲ **Figure 8. Carrier and sideband performance at high PCS band for the VMS-1710 model.**

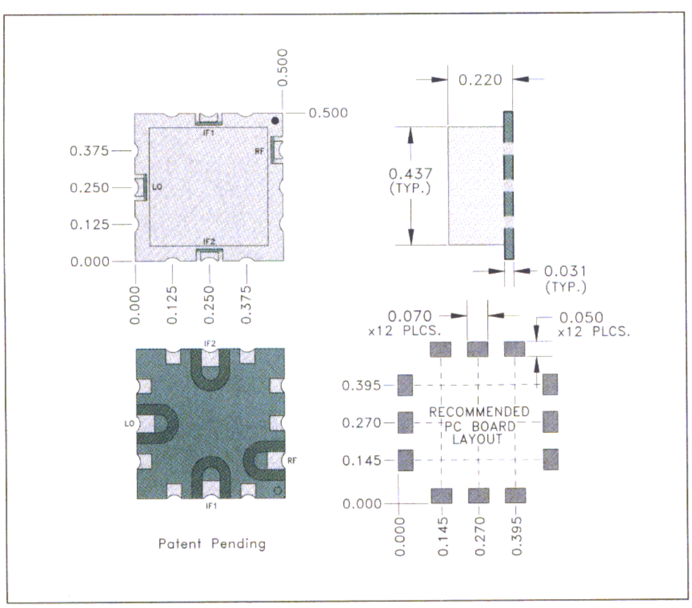


▲ **Figure 9. The amplitude and phase unbalances across the full band of 1710-1780 MHz.**

ulator, and Figure 6 shows the amplitude and phase unbalances across the full band of 935-960 MHz. The amplitude balance is within 0.3 dB and the phase balance is within ±1.7 degrees.

Figures 7 and 8 highlight the carrier and sideband performance in the PCS band for the VMS-1710 model, where the SSB rejection is 27 dB (min..) and the carrier rejection is 25 dB. VMS-1710 can be tuned to function as a demodulator, and Figure 9 shows the amplitude and phase unbalances across the full band of 1710-1780 MHz. The amplitude balance is within 0.45 dB and the phase balance is within ±3.0 degrees. The plots shown in Figures 4, 5 ,7 and 8 also display products that are related to the harmonics of the I and Q signals mixing with the LO. For higher dynamic range, it is highly recommended that the modulator be operated in the linear range. Units designed for higher compression points offer lower levels of the above products.

The outline drawing and the recommended layout of mounting pads for the VMS series modulators is shown in Figure 10.



▲ **Figure 10. The outline drawing and the recommended layout of mounting pads for the VMS series modulators.**

Conclusions

The exceptional RF performance of the new lower cost modulator is demonstrated through the above examples. This new design can be manufactured consistently in large volume and will be available in various packages as needed. The frequency of operation can be easily extended to 3 GHz and more using printed transmission lines for transformers, 90 degree hybrids and an in-phase combiner/divider. This unique approach combines a better noise figure and dynamic range with reduced cost in external circuitry, and requires much less real estate on printed circuit board. ■

For more information, contact:

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Paterson, NJ 07504
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Fax: 973-881-8361
E-mail: sales@synergymwave.com

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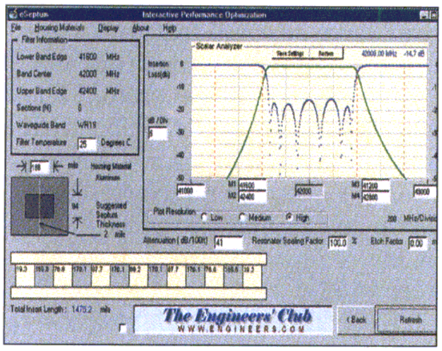
CAD/CAE SOFTWARE

Waveguide filter program

The Engineers' Club offers eSeptum©, an e-plane waveguide filter design and evaluation program. It allows the user to manipulate up to 14 variables, including etch factor, waveguide tolerances,



temperature, scaling factors, band-pass ripple, septum widths and resonator lengths. eSeptum can investigate manufacturing tolerances. Housing materials include pre-programmed temperature coefficients. Output data can be graphical or tabular, and the program's graphing utility allows exporting of files into plotting and graphing programs. The program runs on Windows® 95/98/NT and is priced at $895 plus shipping.

The Engineers' Club
Circle #157

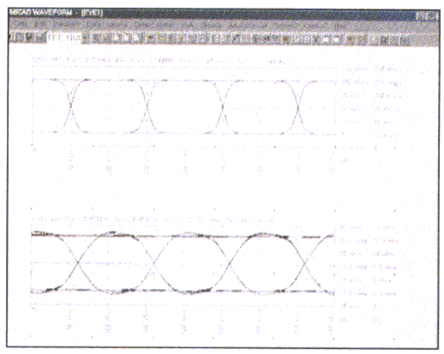
DesignGuides help set up initial design

Agilent Technologies has introduced a new option for its Advanced Design System (ADS) EDA automation software. DesignGuides provide key specifications and circuit display page templates to allow users to begin the design and simulation process immediately. The DesignGuides have been developed cooperatively by industry experts and Agilent's EEsof EDA group. Electronic seminars are provided with some of the guides on CD-ROM to provide further insight into the topic. The topics available now include power amplifier design, linearization techniques for power amplifiers, oscillator design, phase-locked loop design and microwave passive circuit design. Future guides are planned for RF board, RFIC and communication system design. Agilent DesignGuides are priced between $5000 and $10,000 each, depending on the complexity.

Agilent Technologies, Inc.
Circle #158

Time domain analysis

Optotek announces the availability of a time domain transient analysis simulator, MMICAD WAVEFORM™, which allows the prediction of waveforms in fast,



nonlinear circuits where *s*-parameter data can be imported and used in the simulation. With this capability, the user can address complex frequency-dependent effects in passive circuits, such as the variation of effective dielectric constant and loss of a microstrip with frequency in a broadband circuit. GaAs FET and HEMT models are included in addition to the SPICE diode model. The program is designed for stand-alone operation, and it can be used with other components of the MMICAD software suite. MMICAD WAVEFORM is priced from $7,500.

Optotek Ltd.
Circle #159

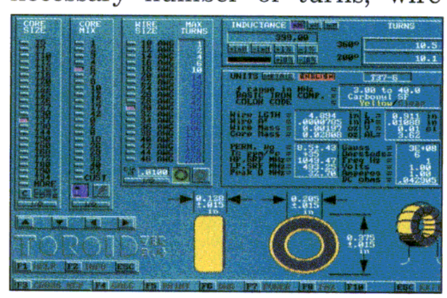
EM software adds optimization and parametrics

Ansoft Corporation has introduced Ansoft HFSS version 7.0, integrating a new module called Optimetrics™. New design features allow users to predict unwanted resonances in components, reduce problem size with more powerful absorbing boundary conditions and customize the interface to streamline the design process. Ansoft HFSS is a full-wave 3D electromagnetic simulator that enables the design of connectors, high frequency packages and antennas. With Optimetrics, the user can import parametric simulation results directly into the optimizer. The engineer can quickly identify the highest performance design that is least sensitive to manufacturing tolerances. A new matrix solver has been developed to obtain the speed necessary for optimization. On Windows machines, multiple processors are supported to additional speed improvement.

Ansoft Corporation
Circle #160

Toroid inductor design

The program TOROID version 3.4 is available from J. Bellora & Associates. TOROID calculates the necessary number of turns, wire

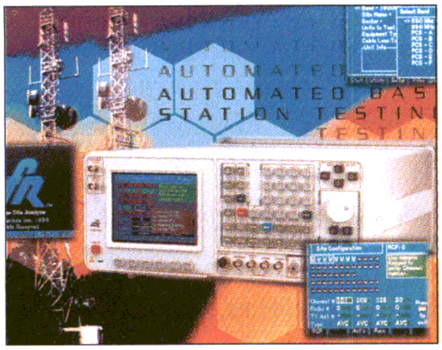


length, self-resonant frequency and other parameters required for accurate design using iron core toroids. Complete data on standard cores is included, showing color codes, iron composition, temperature coefficient and physical dimensions. Custom sizes and permeabilities may also be selected. The program can be used by small laboratories or hobbyists to quickly evaluate surplus toroids or unmarked parts. The TOROID PRO version is priced at $99.95 and $49.95 for additional copies.

J. Bellora & Associates
Circle #161

TEST EQUIPMENT

TIA/EIA-136 automated base site analyzer

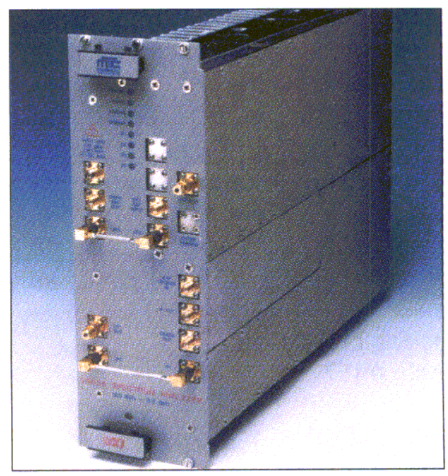


The 1900 Base Site Analyzer (BSA) from IFR, Inc. is a comprehensive tester that addresses field installation and troubleshooting needs for TIA/EIA-136 infrastructure equipment. Test times are reduced using the 1900 with IFR's AutoCell automated test software, which completely verifies modulation quality, adjacent channel power, analog and digital mode output power. It will also perform RSSI calibration test for dual-diversity and newer quad-diversity receivers. The 1900 also has the ability to fully test AMPS base stations, and it can serve as a full-feature 2 GHz service monitor with spectrum analyzer and tracking generator. The 1900 is priced at $43,995, with options available for testing Lucent, Ericsson and Nortel base station equipment.

IFR, Inc.
Circle #162

VXI spectrum analyzer

Morrow Technologies now offers the Model V9035, a fully programmable Spectrum Analyzer covering the frequency range of 100 kHz to 3500 MHz. It is designed for use in VXI cellular and PCS production test systems as well as other VXI-based test applications. The analyzer achieves 0.5 ppm frequency accuracy with digital tuning and it has an amplitude range of –120 to +20 dBm with an absolute level accuracy of ±0.5 dB. Overlay drawings establish test limits and thresholds for any application. Also featured is a fast measurement speed of <150 μs/step and 2 Hz resolution.

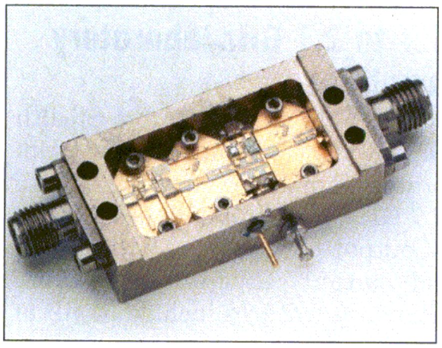


Morrow Technologies Corp.
Circle #163

AMPLIFIERS

Ultra-broadband amplifier

JCA Technology introduces the JCA222-B01 ultra-broadband amplifier for EW and ECM applications. The amplifier covers 2.0 to 22.0 GHz with a minimum gain of 30 dB

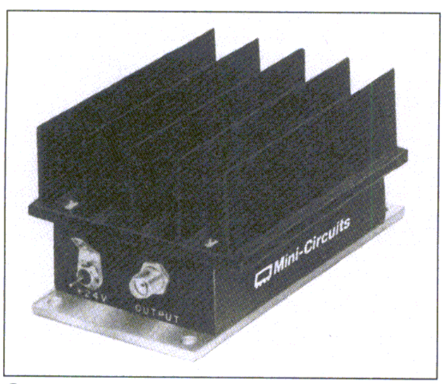


and a maximum gain response variation of ±2.0 dB. Typical noise figure is 3.75 dB and power output is +18 dBm at 1 dB compression. Options include drop-in style packages, alternate gains and temperature compensation.

JCA Technology
Circle #164

+29 dBm power for VHF and cellular applications

Mini-Circuits' new ZHL-3010 coaxial amplifier provides high IP3 of +46 dBm (typ.) over the frequency range of 50 to 1000 MHz.

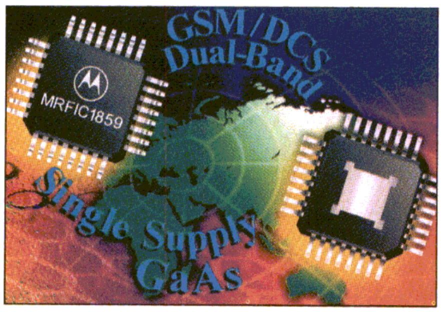


Operating from a 12 VDC power supply, the amplifier delivers a typical gain of 35 dB ±0.6 dB and +29 dBm power output at 1 dB compression. VSWR is typically 2.0:1 input and 1.6:1 output in the mid to upper frequency range. The ZHL-3010 includes a heat sink and SMA female connectors. The unit is priced at $179.95 in small quantities (1 to 9).

Mini-Circuits
Circle #165

GaAs dual-band integrated power amplifier

Motorola's MRFIC1859 offers a single-chip solution for power amplification in dual-band GSM/DCS phone designs. This integrated power amplifier includes an on-chip voltage generator that permits single-supply operation. A gate-drain



switching function protects the devices from over current situations by applying gate bias before the drain voltage is switched on. At 900 MHz, the amplifier provides 3 watts peak power for nominal 2 watt phone designs, with 2 watt peak power at 1800 MHz for 1 watt phones. The device is GPRS compatible and can also be used in PCS1900 cellular phones. The 3.6 volt MRFIC1859 is priced at $5.95 each in $10,000 quantity.

Motorola SPS
Circle #166

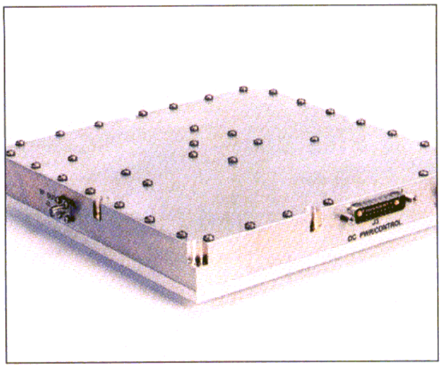
7 watt CDMA amplifier

Ophir RF offers Model GRF2030, providing 7 watts linear power output in the 1.805 to 1.880 GHz frequency band. High efficiency operation is obtained using unique microstrip networks and advanced GaAs FET devices. Reliability is enhanced with a voltage-regulated bias supply and EMI/RFI filtering. The amplifier can be used in many general purpose linear applications.

Ophir RF
Circle #167

15 watt CDMA amplifier

MPD Technologies offers a new 15 watt CDMA power amplifier for the PCS transmit band of 1930 to 1990 MHz. The modular amplifier can be used in base station or tower-top installations, operating with a base plate temperature range of –30 to +65° C. A circulator provides protection for transmission into any load mismatch. The DC power requirement is 24 to 26 volts and the amplifier case measures 8.0 × 8.0 × 1.25 inches. Monitor lines are provided for over-temperature, over-current, VSWR and device failure status.



MPD Technologies, Inc.
Circle #168

1.7 to 2.7 GHz laboratory amplifier

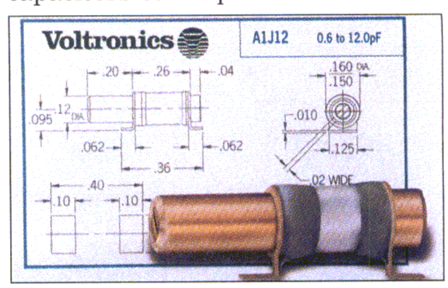
The SM1727-37HR from Stealth Microwave is a compact class A laboratory amplifier covering 1.7 to 2.7 GHz. The unit supplies 5 watts of output power (P_{1dB}) with a typical output IP3 of +51 dBm. Standard features include 33 dB of linear gain, output mismatch protection and internal cooling. Options include 20 dB of linear gain and a digital output power display. The unit includes a power supply for 110 VAC operation and measures 11 × 8 × 3.5 inches in size.

Stealth Microwave
Circle #169

PASSIVE COMPONENTS

New 12 pF trimmer

Voltronics extends the range of the A1 series miniature trimmer capacitors to 12 pF. With 13 turns

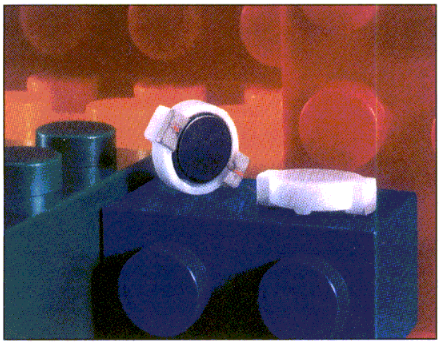


of adjustment and positive stops, the model A1_12 is well-suited for tuning applications in amplifiers, filters and oscillators. The A1_12 is offered in PCB and SMT mountings. High voltage and non-magnetic versions are also offered. Q is >1000 at 200 MHz and self-resonance at 12 pF is 1.2 GHz. The A1_12 is $4.00 in 1,000 quantities.

Voltronics Corporation
Circle #170

Low profile inductors

The new DO1606 PowerWafer™ inductors from Coilcraft combine high inductance values with low profile construction. 19 inductance values are offered, from 1 μH to 1

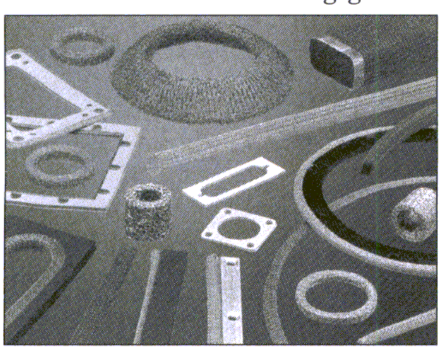


mH, with versions that are capable of handling up to 2.5 amps. The inductors are 2 mm high with a 5.3 × 6.5 mm footprint. A ceramic cover offers an optimum surface for pick and place handling. The company provides PSPICE models at its web site. A typical inductor costs $0.60 in quantities of 10,000.

Coilcraft
Circle #171

Wire mesh gaskets

Chomerics offers a broad line of metal-based EMI shielding gaskets



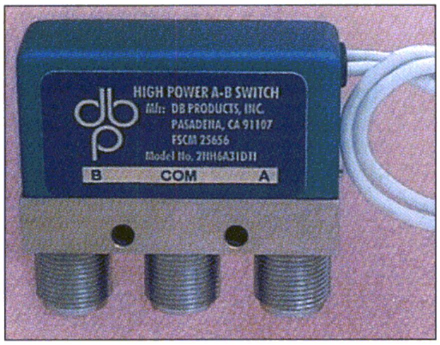
including knitted wire mesh, mesh over an elastomer core, oriented wires in elastomer and metal screen impregnated elastomers. These gaskets typically provide 60 to 100 dB of EMI shielding between 20 MHz and 10 GHz in applications where they are applied to seams and apertures in metal enclosures. Wire mesh prices start below $0.25 per foot.

Chomerics
Circle #172

SIGNAL PROCESSING

SPDT RF switch handles DC to 3 GHz

DB Products introduces a high power SPDT coaxial switch. This A-B selector switch is rated for DC to

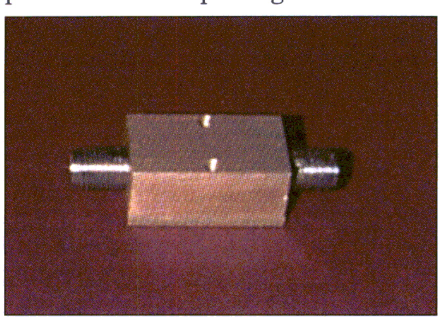


6 GHz with power handling of 1 kW CW at 1 GHz and 600 watts CW at 3 GHz. Specifications at 3 GHz include a maximum insertion loss of 0.15 dB, VSWR of 1.15:1 or better and isolation of at least 90 dB. RF connectors are 50 ohm type N. Coils may be 12, 15, 24 or 28 VDC. Special voltages can be provided, including AC.

DB Products
Circle #173

Small, low loss 2 GHz highpass filter

Sage Laboratories offers a highpass filter in a package size under



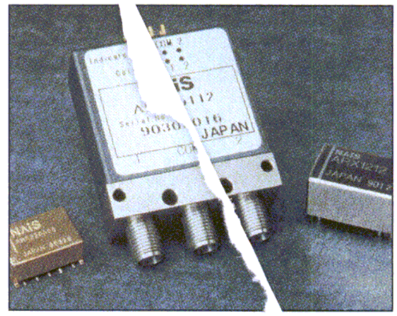
one inch. This selective 2.0 GHz filter has low loss of less than 3.0 dB from 2.0 to 16.0 GHz, VSWR of 2.0:1 maximum and 30 dB rejection at 1.9 GHz. Power handling capability is one watt. The package measures 0.89 × 0.50 × 0.39 inches, plus standard SMA connectors.

Sage Laboratories
Circle #174

Products

RF relays switch signals up to 26.5 GHz

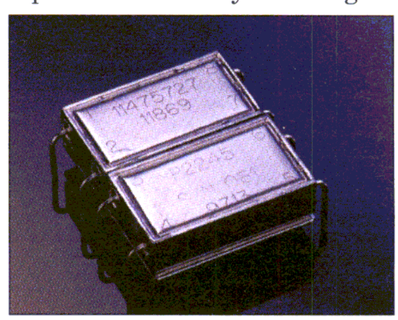
Aromat announces three new NAiS brand relays for switching high frequency 50 ohm signals. The RX small microwave relay covers 300 kHz to 2.5 GHz with 0.2 dB or less insertion loss and 60 or greater isolation. The relay also features 1.2:1 (max.) VSWR. Ultra-small RA relays are used in LAN systems, ADSL, cable modems and many other applications. They include a metal cover for grounding and shielding, and operate with signals up to 1 GHz. Insertion loss is 0.3 dB or less with a minimum isolation of 20 dB. The cost competitive RD coaxial SPDT relay provides stable RF performance and long life. It has a range up to 26.5 GHz with power consumption of 700 mW. Insertion loss is 0.5 dB and isolation is 60 dB at 18 GHz. The RD relays are available in latching, fail safe and TTL drive types, in 12 and 24 VDC.



Aromat
Circle #175

High stability filter sets

Model P-2245 from Signal Technology Corporation is a matched set of low loss linear phase diplexer filters for use in I/Q modulation/demodulation systems. They are designed for demanding applications including space systems. Available in the VHF frequency range, the filter sets feature amplitude and delay tracking of ±0.1 dB and ±100 ps respectively over a wide temperature range of –40 to +85° C. Insertion loss if 0.1 ±0.1 dB, phase linearity is ±2 degrees and unit isolation is 80 dB minimum. A welded hermetic housing is standard. Pricing is $480 per set in quantities of 100 to 249.



Signal Technology Corporation
Circle #176

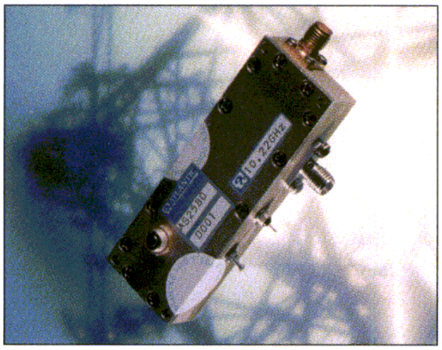
Two watt absorptive solid state switch

American Microwave's model SWN-218-2DT, with Options 0518 and HPR2W, is a SP2T non-reflective switch that will handle 2 watts power from 500 MHz to 18 GHz with 0.85 dB max. insertion loss and 80 dB isolation at 18 GHz, and 95 dB isolation at 500 MHz.

American Microwave
Circle #177

Microwave 6 to 16 GHz frequency converters

Atlantic Microwave's AFC series of frequency converters are designed for use in terrestrial microwave links. A GaAs FET dielectric resonator oscillator is the heart of each unit, feeding a



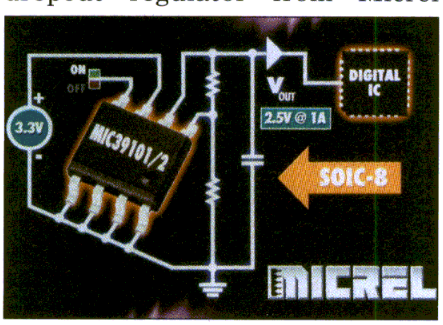
microstrip double balanced mixer that can act as either an up- or downconverter. Units can be mechanically tuned over ±30 MHz and after being set maintain a frequency stability of ±120 ppm over –10 to +60° C. A higher performance mixer diode is optional, providing a +14 dBm 1 dB compression point and +24 dBm input IM3. Another option is electrical tuning or frequency modulation up to ±3 MHz deviation.

Atlantic Microwave
Circle #178

SEMICONDUCTORS

Ultra low dropout regulator

Low power wireless devices can benefit from a new ultra low dropout regulator from Micrel

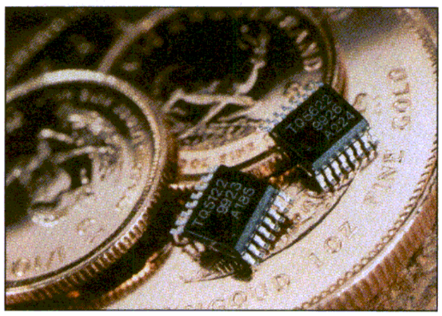


Semiconductor. The regulators convert 3.3 to 2.5 volts, or 2.5 to 1.8 volts, at up to one amp to power low voltage DSP and PLD ICs. The MIC39101 is a fixed output regulator, while the MIC39102 has an adjustable output voltage down to 1.25 volts. The dropout voltage at full load is 410 mV. These devices are provided in a thermally enhanced SO-8 package and are priced starting at $1.50 each in 1,000 piece quantities.

Micrel Semiconductor
Circle #179

TDMA receiver ICs

TriQuint Semiconductor now offers the TQ5122 cellular band receiver and TQ5622 PCS band receiver ICs. These complementary



devices are designed for use in IS-136, TDMA or equivalent wireless handsets. The ICs include power down modes to "sleep" at less than 100 μA current drain. Conversion gain is 18.5 dB and 17.5 dB for the cellular and PCS models respectively. IIP3 is typically greater than –9 dBm and the noise figure is better than 2.8 dB for each device. Unit price is $2.43 in quantities of 1,000.

TriQuint Semiconductor
Circle #180

GaAs low noise amplifier

RF Micro Devices has introduced the RF2365, a low noise amplifier with high dynamic range for front end applications in the PCS and 2.4 GHz frequency range. The device features 1.6 dB and 1.75 dB noise figures at 1850 and 2.45 GHz respectively, with 18 dB and 15.5 dB gain at those frequencies. Offered in an extremely small SOT23-5 plastic package, the ICs are priced at $0.66 each in quantities of 10,000.

RF Micro Devices
Circle #181

FREQUENCY CONTROL

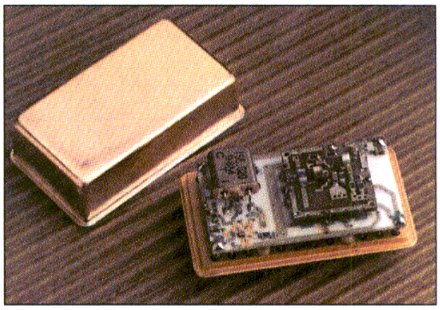
VCXO covers telecom frequencies

Connor-Winfield offers an extensive range of 5 × 7 mm HCMOS surface mount VCXOs specifically designed for telecom applications. The VCXO series is offered in both 3.3 and 5 volts designs, at frequencies from 4 to 45 MHz at 3.3 volts, or 1 to 52 MHz at 5 volts. Standard or extended temperature ranges are offered, with frequency tolerances to 25 ppm and voltage control deviations up to ±100 ppm. Standard telecom frequencies are available from stock.

Connor-Winfield Corporation
Circle #182

PLL clock modules

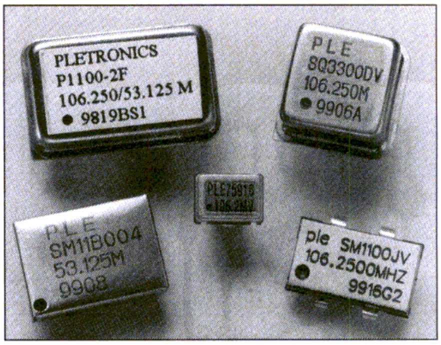
Champion Technologies offers custom designed PLL modules for high speed clock, tracking and synchronization applications at frequencies up to 2.488 GHz. A VCXO portion using high frequency inverted-mesa crystals is combined with a low noise microstrip resonator based VCO to obtain the final output frequency.



Champion Technologies
Circle #183

Oscillators for fibre channel

Pletronics offers oscillators in 106.25 MHz and 53.125 MHz frequencies for use in fibre channel telecom systems. Options include packaging, temperature range and frequency stability. Prices start at $1.50 each for 10,000 pieces.

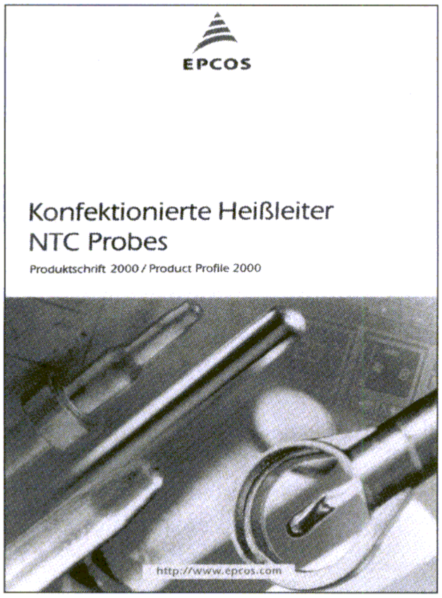


Pletronics, Inc.
Circle #184

LITERATURE

Two new catalogs

EPCOS, Inc. has published two new catalogs, NTC Probes and NTC Thermistor Sensors. Thermistor-controlled electronics are the foundation of many products designed for comfort and convenience. Producers of automobiles, appliances and heating installations rely on the high measuring accuracy of NTC thermistors for exact, energy-saving temperature regulation. The NTC Thermistor Probes catalog provides photographs, schematics and descriptions of 17 devices for home appliance, automotive and heating appliance applications. The NTC Thermistor Sensors catalog provides detailed schematics and technical information about four glass-encapsulated and two ceramic devices for home appliance, automotive, data system and power engineering applications.



EPCOS, Inc.
Circle #185

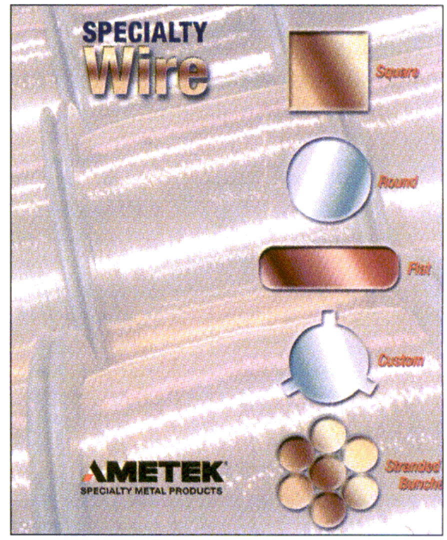
DSP book

Analog Devices has announced a new DSP book that explains the techniques of digital signal processing (DSP). *The Scientists and Engineers Guide to Digital Signal Processing*, by Steven W. Smith, Ph.D., is directed at scientists and engineers who need the power of DSP but do not have time to learn the rigorous theory and mathematics. The book covers a wide range of topics, starting with the elementary techniques of convolution, Fourier analysis and analog-to-digital conversion. Advanced chapters describe the methods used in audio and image processing, data compression and neural networks. An extensive section on digital filters describes how easy-to-use algorithms can solve difficult filtering problems. The author holds advanced degrees in Physics and Electrical Engineering from the University of Utah and has spent the last 15 years directing the development of instrumentation systems.

Analog Devices
Circle #186

Specialty wire brochure

A new illustrated brochure describes the full range of specialty wire products from AMETEK Specialty Metal Products, a manufacturer of custom-shaped wire. Specialty wire products are available in a variety of alloys, including aluminum, copper, nickel, stainless steel and custom-engineered alloys. AMETEK offers wire in a broad range of shapes: round, square, rectangular, half-round, hex, flat, bunched and custom shapes and in sizes from rod size to fine wire. Among the available packaging options are reels, spools, barrels, hex pails, coils, cut-to-length, bundles, and layer level wound reels. AMETEK is a manufacturer of metal powder, strip, wire, and bonded products, holding numerous patents in technically advanced metallurgical products. Its expertise in wrought powder technology enables it to manufacture nickel

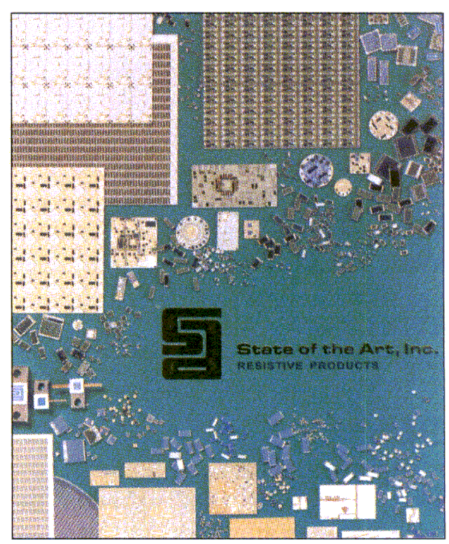


and custom alloy wire with purity, consistency and close compositional control. It uses a variety of forming technologies: drawing, rolling, and shaping; has in-house tooling capabilities; and provides experienced engineering support.

AMETEK
Circle #187

Resistive products catalog

State of the Art, Inc. has published a 28-page catalog covering its entire line of resistive products.



The catalog includes SOTA's high-frequency resistors and chip attenuators, which are designed for RF applications such as base stations and antennas, and in microwave applications for power dividers and power amps, where reduction in an electrical signal is required. Also included is information of SOTA's high-reliability thick and thin film chip resistors, available with TCRs from ±25 ppm to ±100 ppm; tolerances to 0.5 percent; and with wire or epoxy bondable, or solderable terminations. Also detailed are SOTA's MIL-PRF-55342 QPL thick and thin film chip resistors, which have established reliability levels of M, P, R, S, and T (space level). The film resistors are designed for use in mission-critical applications, such as manned space flight, life support systems, and human implantable devices. Special application resistors are described, such as high power, high voltage, power moisture, and high value resistors, zero ohm, multi-tap resistors and attenuators. A 5-page section covers both thick and thin film surface mount resistor networks, including SOTA's MIL-PRF-914 QPL.

State of the Art, Inc.
Circle #188

Frequency sources and SAW devices catalog

Micro Networks Corporation has just released a comprehensive 24-page shortform catalog for its wide range of microelectronic products, used in a variety of applications, including network servers, telecommunication systems, instrumentation, and defense electronics. Products include frequency sources (fixed-frequency oscillators, voltage-controlled oscillators, and phase-locked products and synthesizers); Surface Acoustic Wave (SAW) products (filters, correlators, receivers, and delay lines); data conversion products; and custom microelectronics. Engineers will find product descriptions and specifications, along with mechanical package drawings and ordering information.

Micro Networks
Circle #189

Wireless communications cables catalog

A new 16-page catalog featuring low loss wireless communications cables, including coaxial antenna feeders, jumper cables and low inductance power cables, is being offered by Montrose/CDT of Auburn, Massachusetts. The *Cables for Wireless Communications Systems* Catalog describes WCX Series coaxial cables which feature gas-injected foam polyethylene dielectrics for ultra low loss characteristics, and the RGX Series with irradiated dielectrics. These cables provide electrical or cost advantages over conventional 50 ohm types such as RG 8/, 58, 142, 213, 316, and 400, claims the firm. Providing an overview of their

manufacturing capabilities and quality procedures, the 16-page catalog also includes complete product descriptions, specifications and ordering information for both their WCX and RGX Series coaxial cables.

Montrose/CDT
Circle #190

Semiconductor products selection guide

Richardson Electronics, Ltd. announces the release of United Monolithic Semiconductors' new selection guide. Richardson is the worldwide distributor for United Monolithic Semiconductors (UMS). Included in the UMS selection guide are components for point-to-point and point-to-multipoint microwave radios to 60 GHz. Also included are 5.8 GHz ISM band products, amplifiers, switches, frequency generation products, mixers and attenuators for application from DC to 80 GHz. Best known for their work in millimeter wave, UMS designs and produces devices and integrated circuits based on advanced GaAs MMICs in high volume quantities with guaranteed quality. UMS provides key components for mobile communication, automotive radar and point-to-point and point-to-multipoint radio links.

Richardson Electronics, Ltd.
Circle #191

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Our editors select dozens of products to feature in each monthly issue — we choose them for their importance to RF/microwave and wireless system designers, to present new companies and new product lines, and to give as many companies as possible their share of attention.

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Include a color photograph, if available.

Telecom transformers

Just released by Bel Fuse Inc., a manufacturer of magnetic components for high speed data transmission and networking applications, is a 274-page Telecom Magnetics catalog representing more than 270 Bel components. This comprehen-

sive new literature is designed for easy reference, boasting tabbed sections for devices engineered for high frequency, ISDN, DSL, modem and power applications. To further assist engineers in specifying IC-compatible transformers and magnetic components, a "Product Locator" that indicates part numbers and corresponding data sheets, types of ports supported (single, dual and octal) and cross references to transceiver IC manufacturers, is provided. The Bel telecom transformers featured are available in surface mount and through-hole packages, and are engineered to interface with all major IC manufacturers' chip sets including Globespan, Conexant, Level One, Motorola, Lucent, AMD, Siemans and Analog Devices. Chokes, POTS filters, and inductors are also offered. Moreover, the catalog provides complete electrical specifications, schematics, circuit diagrams and application notes.

Bel Fuse Inc.
Circle #192

NEW PRODUCTS

NO.67

RF/IF MICROWAVE COMPONENTS

FROM $4.45

RF TRANSFORMERS HAVE 4:1 IMPEDANCE 1.5 TO 600MHz

These broad band TCM4-6T surface mount RF transformers from Mini-Circuits operate in the 1.5 to 600MHz band with 4:1 impedance ratio. Referenced to midband loss (0.6dB typ), insertion loss is 1dB from 3MHz to 350MHz, 2dB in the 2 to 400MHz range, and 3dB band wide when operated within -20°C to +85°C (max.). Open case design has plastic base with solder plated leads, and applications include CATV plus VHF/UHF transmitters and receivers. RF power is 250mW (max.).

FROM $8.95

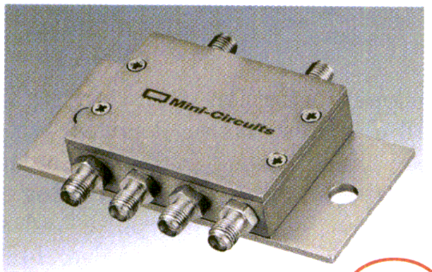
0.07" MIXERS PERFORM IN HIGHER FREQUENCY DESIGNS

Mini-Circuits patented family of Blue Cell™ mixers deliver a unique combination of low conversion loss, superb temperature stability, thin profile, and low cost to higher frequency designs. This level 7 (LO) MBA-671 model spans 2400MHz to 6700MHz with 36dB L-R, 26dB L-I isolation and low 6.5dB midband conversion loss (all typ). Operating temperature is -40°C to +85°C (max.) and applications include satellite, ISM, and PCMCIA. Available off-the shelf.

FROM $6.95

LEVEL 7 (LO) MIXERS IDEAL FOR VHF/UHF RECEIVERS

Mini-Circuits low profile 2500MHz to 3200MHz ADE-32 frequency mixers measure only 0.112" (max.) in height allowing engineers to develop smaller surface mount wireless products. Open case design also allows water wash to drain eliminating the possibility of residue entrapment. Electrically, these mixers typically display low 5.4dB conversion loss, +15dBm IP3 at center band, and good 32dB L-R, 30dB L-I isolation. Maximum RF power rating is 200mW. Patent pending.



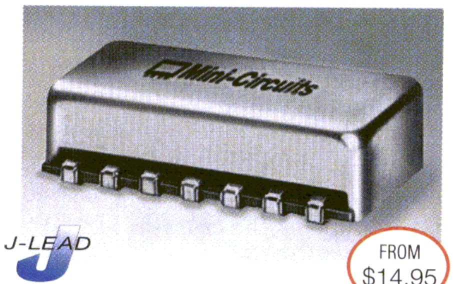
FROM $89.95

DC TO 5GHz SPDT SWITCH HAS VERY HIGH ISOLATION

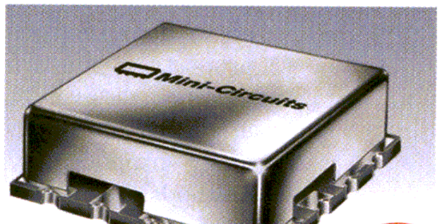
Mini-Circuits ZASW-2-50DR is a DC to 5GHz connectorized single pole double throw (SPDT) reflective switch incorporating a high speed TTL driver for fast 10nsec (typ) switching speed. Typically, isolation is 75dB at 2GHz, 1dB compression is 19dBm at center band, and maximum operating temperature range is -55°C to +100°C. The switch is ideal for transmitter/receiver isolation and automated switching networks. Available from stock.

2WAY 150 TO 350MHz SPLITTER FOR SURFACE MOUNT DESIGNS

Mini-Circuits has unveiled the JSPQ-350, a 2way-90° power splitter/combiner operating in the 150 to 350MHz band with 20dB isolation, 0.5dB amplitude and 1 degree phase unbalance, and very good 1.28:1 in/out VSWR (all typ.). Insertion loss is 0.5dB typical, which is an average of coupled outputs less 3dB. Package height is 0.250" (max.).



FROM $14.95



FROM $12.95

12V VCO PROVIDES 50 TO 100MHz OCTAVE BAND TUNING

Mini-Circuits has introduced a compact, value priced 50 ohm voltage controlled oscillator. Typically, this 12V, 20mA (max. current) ROS-100 model provides 50 to 100MHz octave band tuning, low -105dBc/Hz SSB phase noise at 10kHz offset, and excellent -30dBc harmonic suppression. The miniature 0.5"x0.5"x0.18" size conserves real estate, and applications include test instruments such as signal generators. Maximum operating temperature is -55°C to +85°C.

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Unfiltered FQPSK: Another Interpretation and Further Enhancements

Part 1: Transmitter implementation and optimum reception

By Marvin K. Simon and Tsun-Yee Yan
Jet Propulsion Laboratory

A new interpretation of unfiltered Feher-Patented Quadrature Phase-Shift-Keying (FQPSK) is presented, readily identifying a means for spectral enhancement of the transmitted waveform as well as an improved method of reception. The key to these successes is the replacement of the half-symbol by half-symbol mapping, originally used to describe FQPSK by a symbol-by-symbol mapping operation combined with memory. The advantages of such an interpretation are two-fold. In particular, the original FQPSK scheme can be modified so that the potential of a waveform slope discontinuity at the boundary between half symbols is avoided without sacrificing the "constant" envelope property of the transmitted waveform. Also, a memory receiver can be employed to improve error probability performance relative to previously proposed symbol-by-symbol detection methods. The analysis presented in this paper does not include other versions of FQPSK, such as FQPSK-B, which is currently being considered for military applications.

This is part one of a two-part article. Part two will be published in the March 2000 issue of *Applied Microwave & Wireless* magazine.

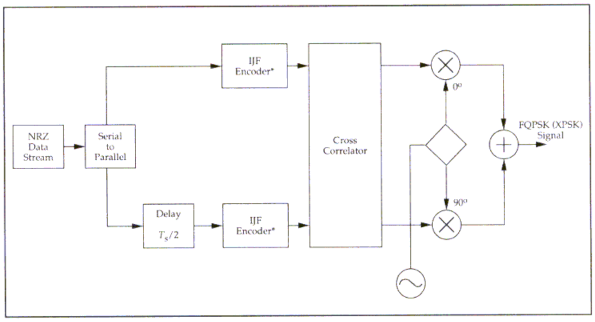


Figure 1. Conceptual block diagram of FQPSK (XPSK). Note that what is referred to as an "IJF Encoder" is in fact a mapping function without any error-correcting capability.

Introduction

In its generic form, Feher-Patented Quadrature Phase-Shift-Keying (FQPSK), as patented [1] and reported in the recent literature [2-3], is conceptually the same as the *cross-correlated phase-shift-keying* (XPSK) modulation technique introduced in 1983 by Kato and Feher [4].[1] This technique was in turn a modification of the previously introduced (by Feher et al [6]) *interference and jitter free QPSK* (IJF-QPSK), with the express purpose of reducing the 3 dB envelope fluctuation characteristic of IJF-QPSK to 0 dB, thus making it appear constant envelope[2], which is beneficial in nonlinear radio systems. (It is further noted that using a constant waveshape for the even pulse and a sinusoidal waveshape for the odd pulse, IJF-QPSK becomes identical to the staggered quadrature overlapped raised cosine (SQORC) scheme introduced by Austin and Chang [7]).

The means by which Kato and Feher achieved their 3 dB envelope reduction was the introduction of an intentional but controlled amount of cross correlation between the inphase (I) and quadrature (Q) channels. This cross correlation operation was applied to the IJF-QPSK (SQORC) baseband signal prior to its modulation onto the I and Q carriers (Figure 1). Specifically, this operation was described by mapping in each half symbol the 16 possible combinations of I and Q channel waveforms present in the SQORC signal into a new[3] set of 16 waveform combinations, chosen in such a way

that the cross correlator output is time continuous and has unit (normalized) envelope[4] at all I and Q uniform sampling instants.

Because the cross correlation mapping is based on a half symbol characterization of the SQORC signal, there is no guarantee that the slope of the cross correlator output waveform is continuous at the half symbol transition points. In fact, we will show that for a random data input sequence, such a discontinuity in slope occurs one quarter of the time.

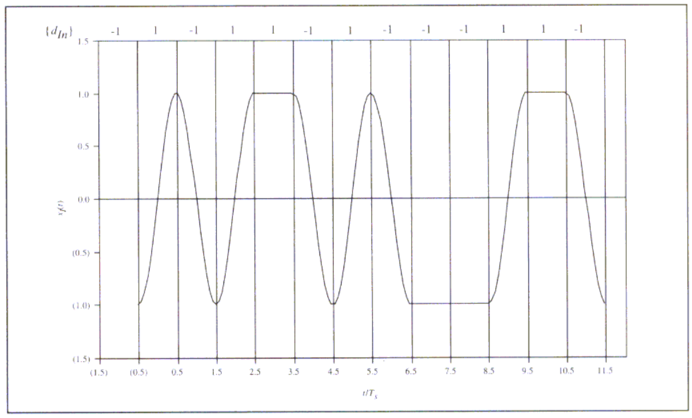
It is well known that the rate at which the side lobes of a modulation's power spectral density (PSD) roll off with frequency is related to the smoothness of the underlying waveforms that generate it. That is, the more derivatives of a waveform that are continuous, the faster its Fourier transform decays with frequency. Thus, since the first derivative of the FQPSK waveform is discontinuous (at half symbol transition instants) on the average one quarter of the time, one can anticipate that an improvement in PSD rolloff could be achieved if the FQPSK cross correlation mapping could be modified so the first derivative is always continuous.

Restructuring the cross correlation mapping into a symbol-by-symbol representation will place the slope discontinuity referred to above in evidence and will be particularly helpful in suggesting a means to eliminate it. This representation also has the advantage that it can be described directly in terms of the data transitions on the I and Q channels. Thus, the combination of IJF encoder and cross correlator can be replaced simply by a single modified cross correlator. The replacement of the conventional FQPSK cross correlator by this modified cross correlator, which eliminates the slope discontinuity, leads to what we shall refer to as *enhanced FQPSK*. We will show that not only does enhanced FQPSK have a better PSD (in the sense of reduced out-of-band energy) than conventional FQPSK, but from a modulation symmetry standpoint, it is a more logical choice.

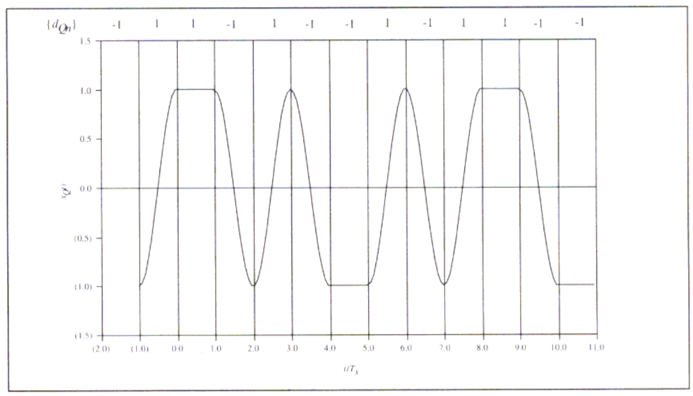
A further and more important advantage of the reformulation as a symbol-by-symbol mapping is the ability to design a receiver of FQPSK or enhanced FQPSK that specifically exploits the correlation introduced into the modulation scheme to significantly improve power efficiency or equivalently, error probability performance. Such a receiver, which takes a form analogous to those used for trellis-coded modulations, will be shown to yield significant performance improvement over receivers that employ symbol-by-symbol detection, thus ignoring the inherent memory of the modulation.

Review of IJF-QPSK and SQORC

The IJF-QPSK scheme (alternately called FQPSK-1) is based on defining waveforms $s_o(t)$ and $s_e(t)$, which are respectively odd and even functions of time over the symbol interval $-T_s/2 \le t \le T_s/2$, and then using these and their negatives $-s_o(t)$, $-s_e(t)$ as a 4-ary signal set for



▲ **Figure 2a. Inphase IJF encoder output.**



▲ **Figure 2b. Quadrature phase IJF encoder output.**

transmission in accordance with the values of successive pairs of data symbols in each of the I and Q arms. Specifically, if d_{In} denotes the I channel data symbols in the interval, $(n-½)T_s \le t \le (n+½)T_s$, then the transmitted waveform $x_I(t)$ in this same interval would be determined as follows:

$$\begin{aligned} x_I(t) &= s_e(t-nT_s) \triangleq s_0(t-nT_s) \text{ if } d_{I,n-1}=1,\ d_{I,n}=1 \\ x_I(t) &= -s_e(t-nT_s) \triangleq s_1(t-nT_s) \text{ if } d_{I,n-1}=-1,\ d_{I,n}=-1 \\ x_I(t) &= s_o(t-nT_s) \triangleq s_2(t-nT_s) \text{ if } d_{I,n-1}=-1,\ d_{I,n}=1 \\ x_I(t) &= -s_o(t-nT_s) \triangleq s_3(t-nT_s) \text{ if } d_{I,n-1}=1,\ d_{I,n}=-1 \end{aligned} \quad (1)$$

The Q channel waveform $x_Q(t)$ would be generated by the same mapping as in (1) using instead the Q channel data symbols $\{d_{Qn}\}$ and then delaying the resulting waveform by one half a symbol. If the odd and even waveforms $s_o(t)$ and $s_e(t)$ are defined by

$$\begin{aligned} s_e(t) &= 1, \quad -T_s/2 \le t \le T_s/2 \\ s_o(t) &= \sin\frac{\pi t}{T_s}, \quad -T_s/2 \le t \le T_s/2 \end{aligned} \quad (2)$$

then typical waveforms for the I and Q IJF encoder outputs are illustrated in Figure 2.

An identical modulation to $x_I(t)$ (and likewise for $x_Q(t)$) generated from the combination of (1) and (2) can be obtained directly from the binary data sequence $\{d_{In}\}$ itself, without the need for defining a 4-ary mapping based on the transition properties of the sequence. In particular, if we define the two-symbol wide raised cosine pulse shape

$$p(t)=\sin^2\left(\frac{\pi(t+T_s/2)}{2T_s}\right),\quad -T_s/2\le t\le 3T_s/2 \tag{3}$$

then the I modulation

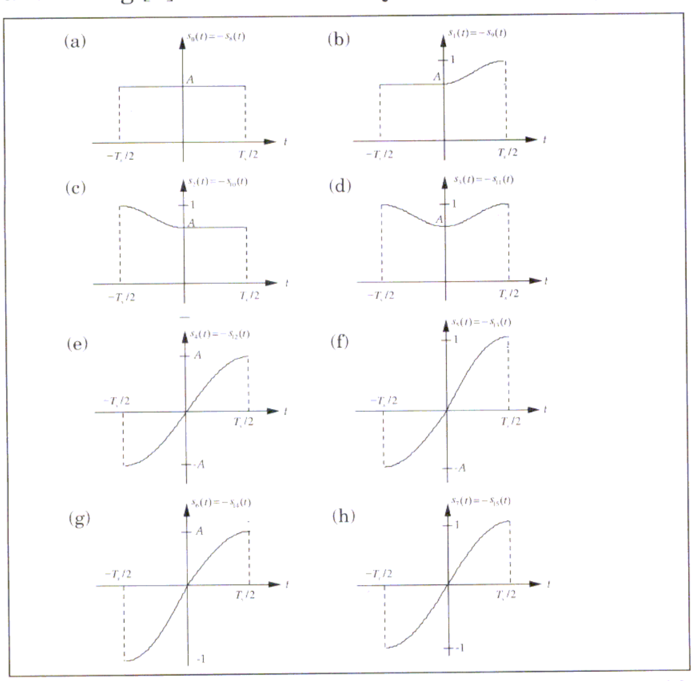
$$x_I(t)=\sum_{n=-\infty}^{\infty} d_{In}\,p(t-nT_s) \tag{4}$$

will be identical to that generated by the above IJF scheme. Similarly,

$$x_Q(t)=\sum_{n=-\infty}^{\infty} d_{Qn}\,p\left(t-\left(n+\frac{1}{2}\right)T_s\right) \tag{5}$$

would also be identical to that generated by the above IJF scheme. A quadrature modulation scheme formed from $x_I(t)$ of (4) and $x_Q(t)$ of (5) is precisely what Austin and Chang [7] referred to as SQORC modulation, namely, independent I and Q modulations with overlapping raised cosine pulses on each channel. The resulting carrier modulated waveform is described by

$$x(t)=x_I(t)\cos\omega_c t+x_Q(t)\sin\omega_c t \tag{6}$$



▲ **Figure 3. FQPSK full-symbol waveforms: (a) $s_0(t) = -s_8(t)$ vs. t; (b) $s_1(t) = -s_9(t)$ vs t; (c) $s_2(t) = -s_{10}(t)$ vs. t; (d) $s_3(t) = -s_{11}(t)$ vs. t; (e) $s_4(t) = -s_{12}(t)$ vs. t; (f) $s_5(t) = -s_{13}(t)$ vs. t; (g) $s_6(t) = -s_{14}(t)$; and (h) $s_7(t) = -s_{15}(t)$ vs. t.**

Symbol-by-symbol cross correlator mapping for FQPSK

Before revealing the modification of FQPSK, which results in a transmitted signal having a continuous first derivative, we first recast the original characterization of FQPSK in terms of a cross correlation operation performed on the pair of IJF encoder outputs every half symbol interval into a mapping performed directly on the input I and Q data sequences every full symbol interval. To do this, we define sixteen waveforms $s_i(t); i=0,1,2,\ldots,15$ over the interval $-T_s/2 \le t \le T_s/2$, which collectively form a transmitted signaling set for the I and Q channels. The particular I and Q waveforms chosen for any particular T_s-sec signaling interval on each channel depends on the most recent data transition on that channel as well as the two most recent successive transitions on the other channel (Figure 3).

$$\begin{aligned}
&s_0(t)=A,\ -T_s/2\le t\le T_s/2,\ s_8(t)=-s_0(t)\\
&s_1(t)=\begin{cases}A, & -T_s/2\le t\le 0\\ 1-(1-A)\cos^2\dfrac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases},\\
&\qquad s_9(t)=-s_1(t)\\
&s_2(t)=\begin{cases}1-(1-A)\cos^2\dfrac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ A. & 0\le t\le T_s/2\end{cases},\\
&\qquad s_{10}(t)=-s_2(t)\\
&s_3(t)=1-(1-A)\cos^2\frac{\pi t}{T_s},\ -T_s/2\le t\le T_s/2,\\
&\qquad s_{11}(t)=-s_3(t)
\end{aligned} \tag{7a}$$

$$\begin{aligned}
&s_4(t)=A\sin\frac{\pi t}{T_s},\ -T_s/2\le t\le T_s/2,\\
&\qquad s_{12}(t)=-s_4(t)\\
&s_5(t)=\begin{cases}A\sin\dfrac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ \sin\dfrac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases},\\
&\qquad s_{13}(t)=-s_5(t)\\
&s_6(t)=\begin{cases}\sin\dfrac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ A\sin\dfrac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases},\\
&\qquad s_{14}(t)=-s_6(t)\\
&s_7(t)=\sin\frac{\pi t}{T_s},\ -T_s/2\le t\le T_s/2,\\
&\qquad s_{15}(t)=-s_7(t)
\end{aligned} \tag{7b}$$

Note that for any value of A other then unity, $s_6(t)$ and $s_7(t)$ as well as their negatives $s_{13}(t)$ and $s_{14}(t)$, will have a discontinuous slope at their midpoints (i.e., at $t=0$), whereas the remaining twelve waveforms all have a continuous slope throughout their defining interval. Also, all 16 waveforms have zero slope at their endpoints, and thus concatenation of any pair of these will not result in a slope discontinuity.

Next, we define the mapping function for the baseband I-channel transmitted waveform $y_1(t)=s_I(t)$ in the nth signaling interval $(n-\frac{1}{2})T_s \le t \le (n+\frac{1}{2})T_s$, in terms of the transition properties of the I and Q data symbol sequences $\{d_{In}\}$ and $\{d_{Qn}\}$, respectively.

8a

I. If $d_{I,n-1}=1$, $d_{I,n}=1$ (no transition on the I sequence, both data bits positive), then

A. $y_I(t)=s_0(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$, results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

B. $y_I(t)=s_1(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

C. $y_I(t)=s_2(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

D. $y_I(t)=s_3(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

8b

II. If $d_{I,n-1}=-1$, $d_{I,n}=1$ (a positive going transition on the I sequence), then

A. $y_I(t)=s_4(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

B. $y_I(t)=s_5(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

C . $y_I(t)=s_6(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

D. $y_I(t)=s_7(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

8c

III. If $d_{I,n-1}=-1$, $d_{I,n}=-1$(no transition on the I sequence, both data bits negative), then

A. $y_I(t)=s_8(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

B. $y_I(t)=s_9(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

C. $y_I(t)=s_{10}(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

D. $y_I(t)=s_{11}(t-nT_s)$ if $d_{Q,n-2}$,

$d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

8d

IV. If $d_{I,n-1}=1$, $d_{I,n}=-1$ (a negative going transition on the I sequence), then

A. $y_I(t)=s_{12}(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

B. $y_I(t)=s_{13}(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in no transition and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

C. $y_I(t)=s_{14}(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in no transition.

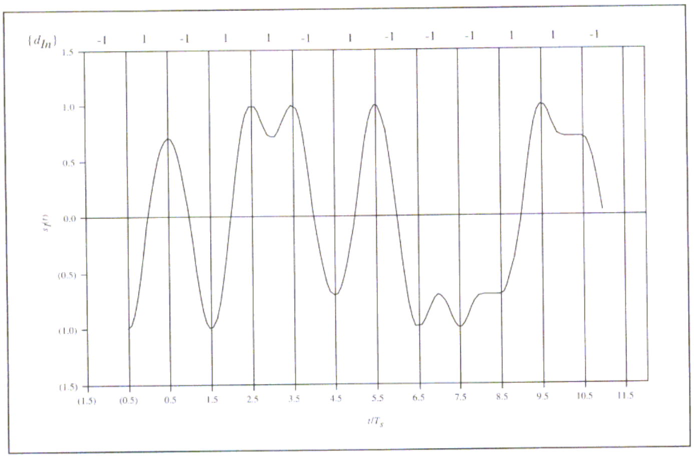
D. $y_I(t)=s_{15}(t-nT_s)$ if $d_{Q,n-2}$, $d_{Q,n-1}$ results in a transition (positive or negative) and $d_{Q,n-1}$, $d_{Q,n}$ results in a transition (positive or negative).

Making use of the signal properties in (7a) and (7b), the mapping conditions in (8a-8d) for the I channel baseband output can be summarized in a concise form (Table 1). A similar construction for the baseband Q-channel transmitted waveform $y_Q(t) = s_Q(t-T_s/2)$ in the nth signaling interval $nT_s \le t \le (n+1)T_s$ in terms of the transition properties of the I and Q data symbol sequences $\{d_{in}\}$ and $\{d_{Qn}\}$, respectively, can be obtained analogous to (8a-d). The results are summarized in Table 2. *Note that the subscript i of the transmitted signal $s_i(t-nT_s)$ or $s_i(t-(n+½)T_s)$ as appropriate is the binary coded decimal (BCD) equivalent of the three transitions.*

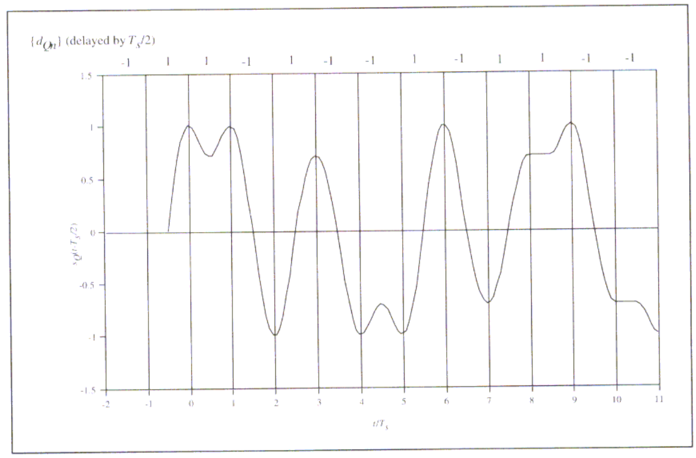
Applying the mappings in Tables 1 and 2 to the I and Q data sequences of Figure 2 produces the identical I and Q baseband transmitted signals to those that would be produced by passing the I and QIJF encoder outputs of this figure through the cross-correlator (half symbol mapping) of the QPSK (XPSK) scheme as described in [4] (Figure 4). Thus, we conclude that *for arbitrary I and Q input sequences, FQPSK can alternately be generated by the symbol-by-symbol mappings of Tables 1 and 2 as applied to these sequences.*

A new and improved FQPSK

As discussed above, the symbol-by-symbol mapping representation of FQPSK identifies the fact that four out of the sixteen possible transmitted waveforms, namely $s_5(t)$, $s_6(t)$, $s_{13}(t)$ and $s_{14}(t)$ have a slope discontinuity at their midpoint. Thus, for random I and Q data symbol sequences, on the average the transmitted FQPSK waveform will likewise have a slope discontinuity at one quarter of the uniform sampling time instants. To prevent this from occurring, we now redefine



▲ **Figure 4a. Inphase FQPSK (XPSK) output.**



▲ **Figure 4b. Quadrature phase FQPSK (XPSK) output.**

these four transmitted signals in a manner analogous to $s_1(t)$, $s_2(t)$, $s_9(t)$ and $s_{10}(t)$, namely,

$$s_5(t)=\begin{cases}\sin\frac{\pi t}{T_s}+(1-A)\sin^2\frac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ \sin\frac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases},$$

$$s_{13}(t)=-s_5(t) \qquad (9)$$

$$s_6(t)\begin{cases}\sin\frac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ \sin\frac{\pi t}{T_s}-(1-A)\sin^2\frac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases},$$

$$s_{14}(t)=-s_6(t)$$

Note that the signals $s_5(t)$, $s_6(t)$, $s_{13}(t)$ and $s_{14}(t)$ as defined in (9) do not have a slope discontinuity at their midpoint nor, for that matter, anywhere else in the defining interval. Also, the zero slope at their endpoints has been preserved. Thus, using (9) in place of the corresponding signals of (7b) will result in a modified FQPSK signal that has no slope discontinuity anywhere in time *regardless of the value of A*. Figure 5 illustrates a comparison of the signal $s_6(t)$ of (9) with that of (7b) for a value of $A=1\sqrt{2}$. Figure 6 illustrates the power spectral density of conventional FQPSK and its enhancement obtained by using the waveforms of (9) as replacements for those in (7b). The significant improvement in spectral rolloff rate is clear from a comparison of the two.

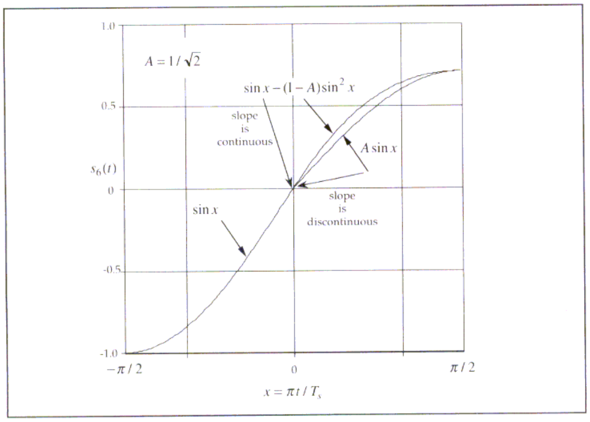
As it currently stands, the signal set chosen for enhanced FQPSK has a symmetry property for $s_0(t)$, $s_1(t)$, $s_2(t)$ and $s_3(t)$ that is not present for $s_4(t)$, $s_5(t)$, $s_6(t)$ and $s_7(t)$. In particular, $s_1(t)$ and $s_2(t)$ are each composed of one half of $s_0(t)$ and one half of $s_3(t)$, i.e., the portion of $s_1(t)$ from $t=-T_s/2$ to $t=0$ is the same as that of $s_0(t)$, whereas the portion of $s_1(t)$ from $t=0$ to $t=T_s/2$ is the same as that of $s_3(t)$, and vice versa for $s_2(t)$. To achieve the same symmetry property for $s_4(t)-s_7(t)$, one would have to reassign $s_4(t)$ as

$\left\lvert\frac{d_{In}-d_{I,n-1}}{2}\right\rvert$	$\left\lvert\frac{d_{Q,n-1}-d_{Q,n-2}}{2}\right\rvert$	$\left\lvert\frac{d_{Qn}-d_{Q,n-1}}{2}\right\rvert$	$s_I(t)$
0	0	0	$d_{In}s_0(t-nT_s)$
0	0	1	$d_{In}s_1(t-nT_s)$
0	1	0	$d_{In}s_2(t-nT_s)$
0	1	1	$d_{In}s_3(t-nT_s)$
1	0	0	$d_{In}s_4(t-nT_s)$
1	0	1	$d_{In}s_5(t-nT_s)$
1	1	0	$d_{In}s_6(t-nT_s)$
1	1	1	$d_{In}s_7(t-nT_s)$

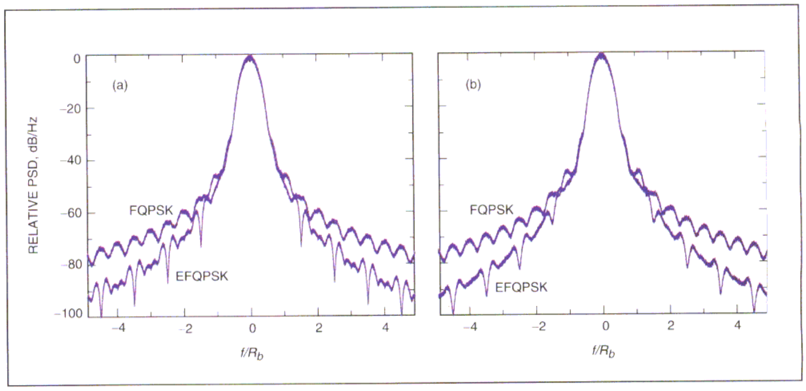
▲ **Table 1. Mapping for inphase (I)-channel baseband signal $y_I(t)$ in in the interval $(n-½)T_s \le t \le (n+½)T_s$.**

$\left\lvert\frac{d_{Qn}-d_{Q,n-1}}{2}\right\rvert$	$\left\lvert\frac{d_{I,n}-d_{I,n-1}}{2}\right\rvert$	$\left\lvert\frac{d_{I,n+1}-d_{I,n}}{2}\right\rvert$	$s_I(t)$
0	0	0	$d_{Qn}s_0(t-nT_s)$
0	0	1	$d_{Qn}s_1(t-nT_s)$
0	1	0	$d_{Qn}s_2(t-nT_s)$
0	1	1	$d_{Qn}s_3(t-nT_s)$
1	0	0	$d_{Qn}s_4(t-nT_s)$
1	0	1	$d_{Qn}s_5(t-nT_s)$
1	1	0	$d_{Qn}s_6(t-nT_s)$
1	1	1	$d_{Qn}s_7(t-nT_s)$

▲ **Table 2. Mapping for quadrature (Q)-channel baseband signal $y_Q(t)$ in the interval $nT_s \le t \le (n+1)T_s$.**



▲ **Figure 5. Original and new FQPSK shapes.**



▲ **Figure 6. Power spectra of conventional and enhanced FQPSK: (a) without SSPA; and (b) with SSPA.**

$$s_4(t)=\begin{cases}\sin\frac{\pi t}{T_s}+(1-A)\sin^2\frac{\pi t}{T_s}, & -T_s/2\le t\le 0\\ \sin\frac{\pi t}{T_s}-(1-A)\sin^2\frac{\pi t}{T_s}, & 0\le t\le T_s/2\end{cases}, \qquad (10)$$

$$s_{12}(t)=-s_4(t)$$

This minor change, which produces a complete symmetry in the waveform set, has an advantage from the standpoint of hardware implementation and produces a negligible change in spectral properties of the transmitted waveform. Nevertheless, for the remainder of the discussion, we shall ignore this minor change and assume the version of enhanced FQPSK first introduced in this section.

Interpretation of FQPSK as a trellis-coded modulation

The I and Q mappings given in Tables 1 and 2 are alternately described in terms of the (0,1) representation of the I and Q data symbols and their transitions. Specifically, define

$$D_{In}\triangleq(1-d_{In})/2,\ D_{Qn}\triangleq(1-d_{Qn})/2 \qquad (11)$$

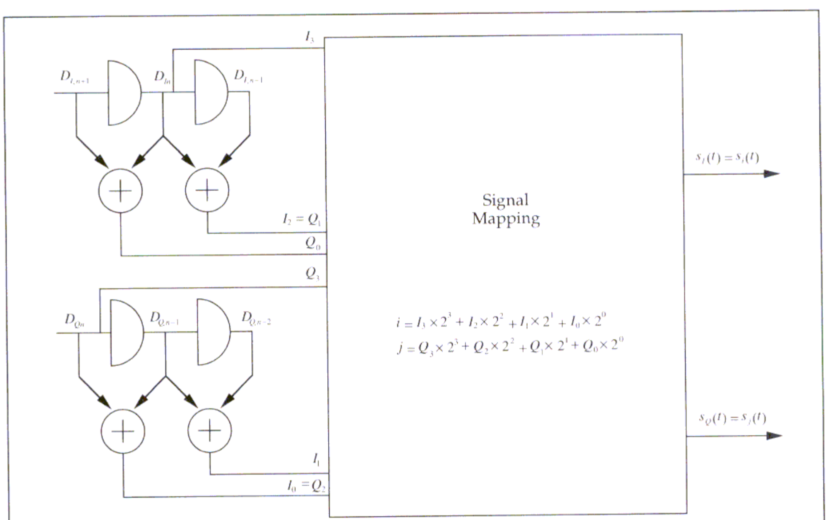
which both range on the set (0,1). Then, defining the BCD representation of the indices i and j by

$$i=I_3\times2^3+I_2\times2^2+I_1\times2^1+I_0\times2^0$$
$$j=Q_3\times2^3+Q_2\times2^2+Q_1\times2^1+Q_0\times2^0 \qquad (12)$$

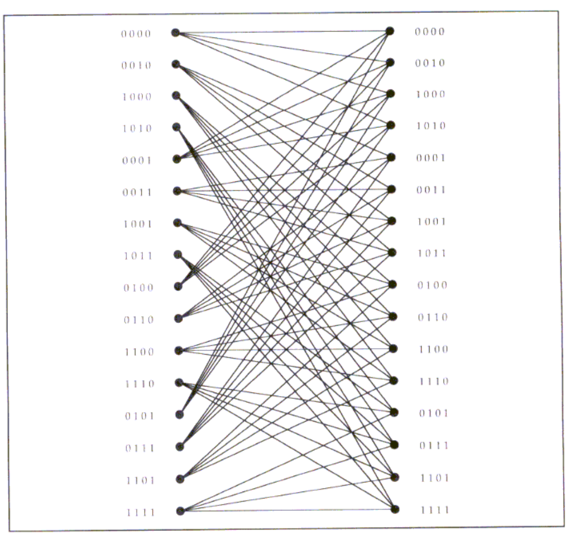
with

$$I_0=D_{Qn}\oplus D_{Q,n-1},\ Q_0=D_{I,n+1}\oplus D_{In}$$
$$I_1=D_{Q,n-1}\oplus D_{Q,n-2},\ Q_1=D_{In}\oplus D_{I,n-1}=I_2 \qquad (13)$$
$$I_2=D_{In}\oplus D_{I,n-1},\ Q_2=D_{Qn}\oplus D_{Q,n-1}=I_0$$
$$I_3=D_{In},\ Q_3=D_{Qn}$$

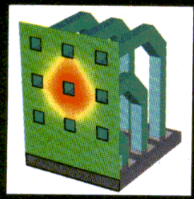
we have $y_I(t)=s_i(t-nT_s)$ and $y_Q(t)=s_j(t-(n+½)T_s)$. That is, in each symbol interval $((n-½)T_s\le t\le(n+½)T_s$ for $y_I(t)$ and $nT_s\le t\le(n+1)T_s$ for $y_Q(t)$), the I and Q channel baseband signals are each chosen from a set of 16

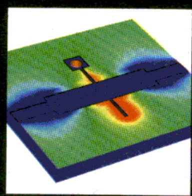


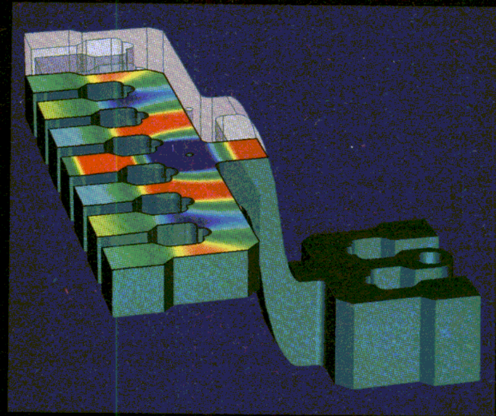
▲ **Figure 7. Alternate implementation of FQPSK baseband signals.**

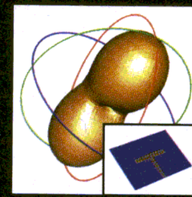


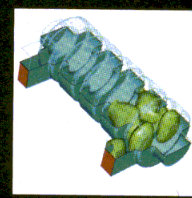
▲ **Figure 8. 16-state trellis diagram for FQPSK.**











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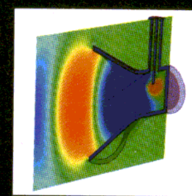

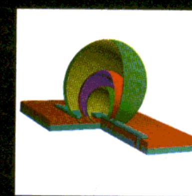

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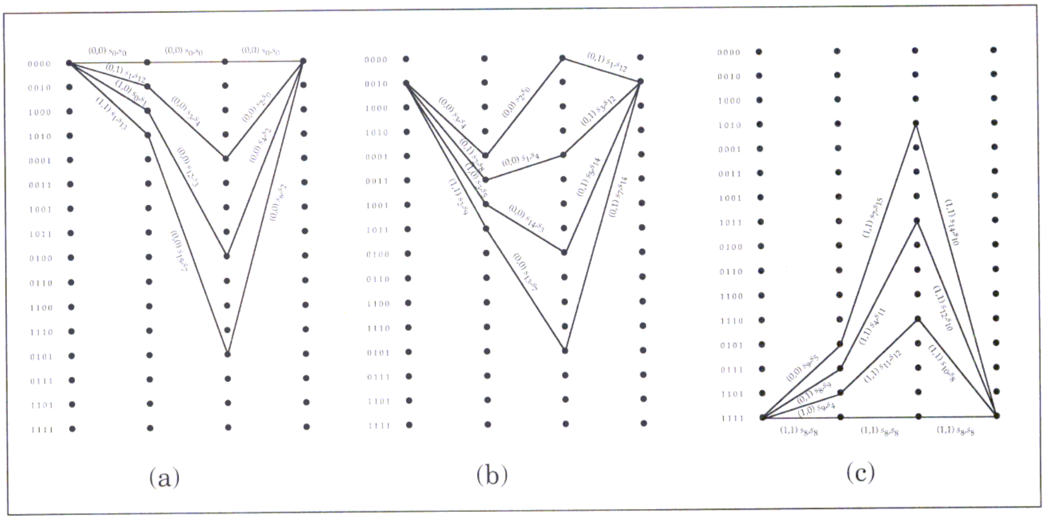
signals, $s_i(t)$, $i=0,1,\ldots,15$, in accordance with the 4-bit BCD representations of their indices defined by (12) together with (13). A graphical illustration of the implementation of this mapping is given in Figure 7.

Another interpretation of the mapping in Figure 7 is as a 16-state trellis code with two binary (0,1) inputs $D_{I,n+1}$, D_{Qn}, and two waveform outputs $s_i(t)$, $s_j(t)$, where the state is defined by the 4-bit sequence D_{In}, $D_{I,n-1}$, $D_{Q,n-1}$, $D_{Q,n-2}$. The trellis is illustrated in Figure 8, and the transition mapping is given in Appendix 1. In this table, the entries in the column labeled "input" correspond to the values of the two input bits that result in the transition, while the entries in the column "output" correspond to the subscripts i and j of the pair of symbol waveforms $s_i(t)$, $s_j(t)$ that are outputted.



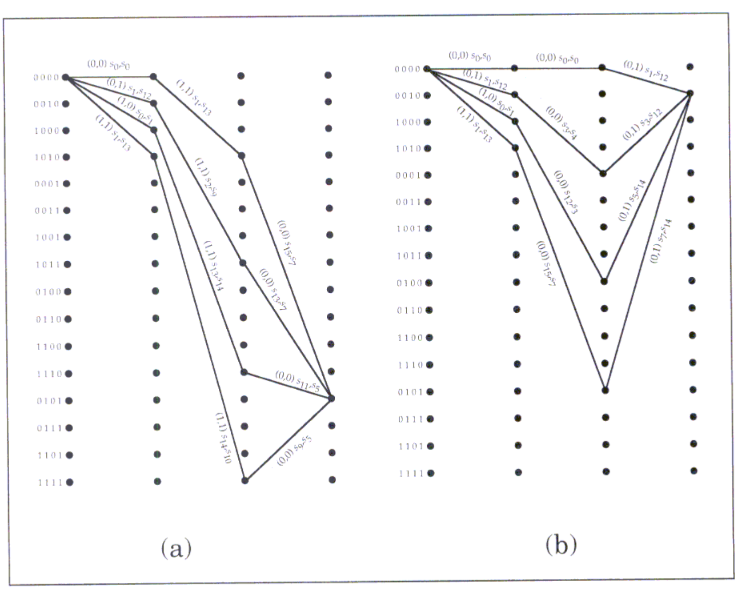
▲ **Figure 9. (a) Paths of length 3 branches starting and ending in State 1; (b) paths of length 3 branches starting and ending in State 2; and (c) paths of length 3 branches starting and ending in State 16.**

To compute the performance of this trellis coded modulation (TCM), we need to determine the minimum Euclidean distance between pairs of error event paths that leave a given state and first return to that or another state a number of branches later. The smallest length error event for which there are at least two paths that start in one state and remerge in the same or another state is three branches. For each of the 16 starting states, there are exactly four such error event paths that remerge in each of the 16 end states. Figures 9a and 9b are examples of these error event paths corresponding to the first two states, respectively, for the case where the start and end states are the same. The remaining length three error event paths for states 9-16 are the mirror images of the ones for states 1-8 (for example, Figure 9c, which should be compared with Figure 9a). Also, the paths for states 9-16 will have identical Euclidean distance properties to those for states 1-8, since the output symbols along their branches will be the negatives of those along their mirror images. Figures 10a and 10b are examples of the groups of four error event paths that start in a given state and remerge in another state. A similar mirror image symmetry exists for these groups of paths and thus once again it is sufficient to consider only the first eight starting states.

It is important to note that the trellis code defined by the mapping in Appendix 1 is not uniform, e.g., it is not sufficient to consider only the all zeros path as the transmitted path in computing the minimum Euclidean distance. Rather, one must consider all possible pairs of error event paths starting from each of the 16 states (eight states is sufficient in view of the above-mentioned distance symmetry properties) and ending in each of the 16 states. This determines the pair having the minimum Euclidean distance.

The following example illustrates the procedure for the groups of paths that start and end in the same state.

Upon examination of the squared Euclidean distance between all pairs of paths in the above-mentioned figures, it can be shown that the minimum of this distance, for example, $d^2_{\min}$, occurs between the first and third paths of Figure 9b. Thus, based on the output symbols



▲ **Figure 10. (a) Paths of length 3 branches starting in state 1 and ending in State 13; (b) paths of length 3 branches starting in State 1 and ending in State 2.**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that occur along this pair of paths, we have

$$
\begin{aligned}
d_{\min}^2 = \int_{-T_s/2}^{T_s/2} [& (s_3(t)-s_3(t))^2 + (s_4(t)-s_5(t))^2 \\
& +(s_2(t)-s_{14}(t))^2 + (s_0(t)-s_3(t))^2 \\
& +(s_1(t)-s_5(t))^2 + (s_{12}(t)-s_{14}(t))^2]dt
\end{aligned} \quad (14)
$$

Evaluation of the squared Euclidean distances between the pairs of waveforms required in (14) using (7a) and (7b) for their definition results after much algebra in

$$
d_{\min}^2 = \left[\frac{7}{4} - \frac{8}{3\pi} - A\left(\frac{3}{2} + \frac{4}{3\pi}\right) + A^2\left(\frac{11}{4} + \frac{4}{\pi}\right) \right] T_s = 1.552 T_s \quad (15)
$$

The average signal (I+Q) energy is obtained from

$$
\begin{aligned}
E_{av} &= \frac{1}{256}\sum_{i=0}^{15}\sum_{j=0}^{15}\int_{-T_s/2}^{T_s/2}[s_i^2(t)+s_j^2(t)]dt = \\
& 2\left[\frac{1}{16}\sum_{i=0}^{15}\int_{-T_s/2}^{T_s/2} s_i^2(t)dt\right] = \frac{1}{4}\sum_{i=0}^{7}\int_{-T_s/2}^{T_s/2} s_i^2(t)dt
\end{aligned} \quad (16)
$$

which again using (7a) and (7b) evaluates to

$$
E_{av} = \left(\frac{7+2A+15A^2}{16}\right) T_s = 0.9946 T_s \quad (17)
$$

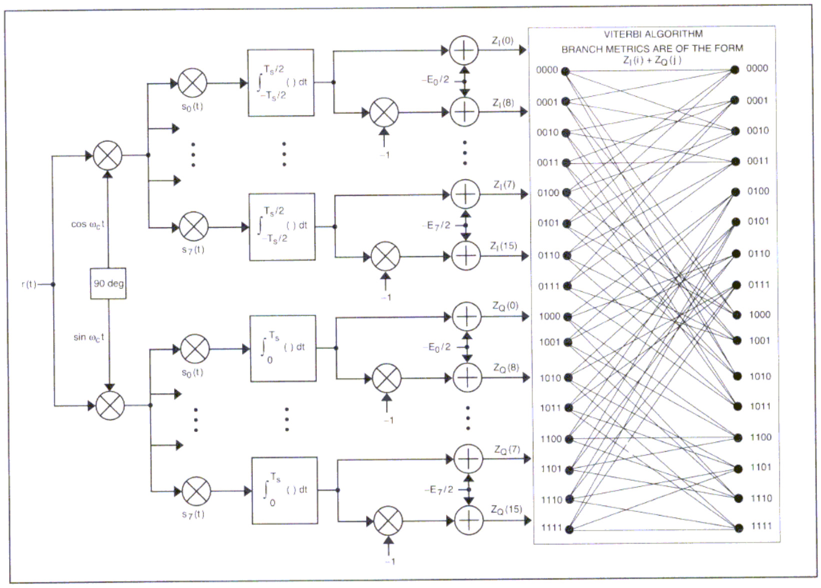
Since the average signal (symbol) energy is twice the average energy per bit, then the normalized minimum squared Euclidean distance for the paths corresponding to starting and ending in the same state is

$$
\frac{d_{\min}^2}{2E_b} = \frac{16\left[\frac{7}{4} - \frac{8}{3\pi} - A\left(\frac{3}{2}+\frac{4}{3\pi}\right) + A^2\left(\frac{11}{4}+\frac{4}{\pi}\right)\right]}{(7+2A+15A^2)} = 1.56 \quad (18)
$$

Upon examination of all length 3 error event paths that begin in one state and end in another, e.g., Figures 10a and 10b, no pair of paths with smaller normalized minimum squared Euclidean distance was found.

Furthermore, by exhaustive search, it can be shown that the minimum squared Euclidean distance of (18) is the smallest over all pairs of paths that start in any state and end in any state regardless of the length of the path. Thus, *the normalized minimum squared Euclidean distance for the FQPSK scheme is given by (18).*

For the spectrally enhanced FQPSK using the waveforms of (9) as replacements for their equivalents in (7b), the minimum squared Euclidean distance over all length 3 trellis paths occurs, for example, between the first and second paths, starting and ending in state "0000" and is given by (see Figure 9a)



▲ **Figure 11. The optimum trellis-coded receiver for FQPSK.**

$$
\begin{aligned}
d_{\min}^2 = \int_{-T_s/2}^{T_s/2} [& (s_0(t)-s_1(t))^2 + (s_0(t)-s_{12}(t))^2 \\
& +(s_0(t)-s_3(t))^2 + (s_0(t)-s_4(t))^2 \\
& +(s_0(t)-s_2(t))^2 + (s_0(t)-s_0(t))^2]dt
\end{aligned} \quad (19)
$$

Once again, evaluation of the squared Euclidean distances between the pairs of waveforms required in (19) using (7a) and (7b) together (now with (9) for their definition) results after much algebra in

$$
d_{\min}^2 = \left[\frac{3-6A+15A^2}{4}\right] T_s = 1.564 T_s \quad (20)
$$

Likewise, the average signal energy is now

$$
E_{av} = \left(\frac{\frac{21}{8} - \frac{8}{3\pi} - A\left(\frac{1}{4} - \frac{8}{3\pi}\right) + \frac{29}{8}A^2}{4} \right) T_s = 1.003 T_s \quad (21)
$$

Thus, the normalized minimum squared Euclidean distance is

$$
\frac{d_{\min}^2}{2E_b} = \frac{(3-6A+15A^2)}{\frac{21}{8} - \frac{8}{3\pi} - A\left(\frac{1}{4} - \frac{8}{3\pi}\right) + \frac{29}{8}A^2} = 1.56 \quad (22)
$$

which coincidentally is identical to that for FQPSK.

Again there is no other pair of paths starting in any state and ending in any other that produces a smaller normalized minimum squared Euclidean distance. Thus, we shall conclude that *the enhancement of FQPSK provided by using the waveforms of (9) as replacements for their equivalents in (7b) is significantly beneficial from a spectral standpoint with no penalty in receiver performance.*

In accordance with the foregoing representation of FQPSK as a trellis coded modulation with 16 states, the optimum receiver (employing a Viterbi algorithm) for FQPSK is illustrated in Figure 11.

In later sections of this article, we shall illustrate average bit error probability (BEP) results obtained from a simulation of this receiver. For the moment, we shall just compare its asymptotic (limit of infinite energy-to-noise ratio) performance with that of the optimum receiver for conventional uncoded offset QPSK (OQPSK).

Since for the latter, $d^2_{\min}/2\bar{E}_b = 2$, which is the same as that for BPSK [8], we see that as a trade against the significantly improved power spectrum afforded by FQPSK and its enhanced version relative to that of OQPSK, *an asymptotic loss of only* 10 log(2/1.56)=1.07 *dB is experienced.*[6]

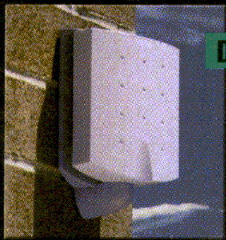
Symbol-by-symbol FQPSK detection

In the second part of this article, which will be published in the March issue of *Applied Microwave & Wireless*, we will examine the performance of FQPSK when the detector makes decisions on a symbol-by-symbol basis — that is, the inherent memory introduced by the trellis coding is ignored at the receiver. Signal representation, suboptimum receivers and average bit error probability performance will also be covered in the conclusion to this article. ■

End notes

[1] Other versions of FQPSK, referred to as FQPSK-B [5], include proprietary designed filtering for additional spectrum containment. Such filtering is not germane to our discussions in this paper and will not be considered.

[2] The reduction of the envelope from 3 dB to 0 dB occurs only at the uniform sampling instants on the inphase (I) and quadrature (Q) channels. It is for this reason that XPSK is referred to as being "pseudo" or



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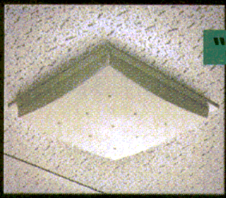

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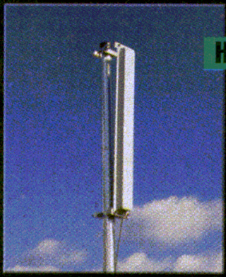

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"quasi" constant envelope, i.e., its envelope has a small amount of fluctuation between the uniform sampling instants.

[3] Of the 16 possible cross correlator output combinations, only 12 of them are in fact new, i.e., for four of the input I and Q combinations, the cross correlator outputs the identical combination.

[4] Actually, in its generic form, Kato and Feher allow (through the introduction of a transition parameter $k = 1-A$) for a controlled amount of envelope fluctuation. For quasi-constant envelope, one should choose $A = 1/\sqrt{2}$.

[5] This also occurs between several other pairs of paths starting and ending in the same state.

[6] Needless to say, at smaller (finite) SNRs, the loss between uncoded OQPSK and trellis decoded FQPSK will be even less.

Acknowledgment

The authors wish to gratefully acknowledge Meera Srinivasan for simulating the optimum trellis-coded receiver of FQPSK (Figure 11). Additional thanks is also due Meera for having spent considerable time independently verifying much of what is presented here, which is immeasurably important in establishing the credence of our results.

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Appendix 1. Trellis state transitions.

current state	input	output	next state	current state	input	output	next state
0 0 0 0	0 0	0 0	0 0 0 0	0 1 0 0	0 0	4 2	0 0 0 0
0 0 0 0	0 1	1 12	0 0 1 0	0 1 0 0	0 1	5 14	0 0 1 0
0 0 0 0	1 0	0 1	1 0 0 0	0 1 0 0	1 0	4 3	1 0 0 0
0 0 0 0	1 1	1 13	1 0 1 0	0 1 0 0	1 1	5 15	1 0 1 0
0 0 1 0	0 0	3 4	0 0 0 1	0 1 1 0	0 0	7 6	0 0 0 1
0 0 1 0	0 1	2 8	0 0 1 1	0 1 1 0	0 1	6 10	0 0 1 1
0 0 1 0	1 0	3 5	1 0 0 1	0 1 1 0	1 0	7 7	1 0 0 1
0 0 1 0	1 1	2 9	1 0 1 1	0 1 1 0	1 1	6 11	1 0 1 1
1 0 0 0	0 0	12 3	0 1 0 0	1 1 0 0	0 0	8 1	0 1 0 0
1 0 0 0	0 1	13 15	0 1 1 0	1 1 0 0	0 1	9 13	0 1 1 0
1 0 0 0	1 0	12 2	1 1 0 0	1 1 0 0	1 0	8 0	1 1 0 0
1 0 0 0	1 1	13 14	1 1 1 0	1 1 0 0	1 1	9 12	1 1 1 0
1 0 1 0	0 0	15 7	0 1 0 1	1 1 1 0	0 0	11 5	0 1 0 1
1 0 1 0	0 1	14 11	0 1 1 1	1 1 1 0	0 1	10 9	0 1 1 1
1 0 1 0	1 0	15 6	1 1 0 1	1 1 1 0	1 0	11 4	1 1 0 1
1 0 1 0	1 1	14 10	1 1 1 1	1 1 1 0	1 1	10 8	1 1 1 1
0 0 0 1	0 0	2 0	0 0 0 0	0 1 0 1	0 0	6 2	0 0 0 0
0 0 0 1	0 1	3 12	0 0 1 0	0 1 0 1	0 1	7 14	0 0 1 0
0 0 0 1	1 0	2 1	1 0 0 0	0 1 0 1	1 0	6 3	1 0 0 0
0 0 0 1	1 1	3 13	1 0 1 0	0 1 0 1	1 1	7 15	1 0 1 0
0 0 1 1	0 0	1 4	0 0 0 1	0 1 1 1	0 0	5 6	0 0 0 1
0 0 1 1	0 1	0 8	0 0 1 1	0 1 1 1	0 1	4 10	0 0 1 1
0 0 1 1	1 0	1 5	1 0 0 1	0 1 1 1	1 0	5 7	1 0 0 1
0 0 1 1	1 1	0 9	1 0 1 1	0 1 1 1	1 1	4 11	1 0 1 1
1 0 0 1	0 0	14 3	0 1 0 0	1 1 0 1	0 0	10 1	0 1 0 0
1 0 0 1	0 1	15 15	0 1 1 0	1 1 0 1	0 1	11 13	0 1 1 0
1 0 0 1	1 0	14 2	1 1 0 0	1 1 0 1	1 0	10 0	1 1 0 0
1 0 0 1	1 1	15 14	1 1 1 0	1 1 0 1	1 1	11 12	1 1 1 0
1 0 1 1	0 0	13 7	0 1 0 1	1 1 1 1	0 0	9 5	0 1 0 1
1 0 1 1	0 1	12 11	0 1 1 1	1 1 1 1	0 1	8 9	0 1 1 1
1 0 1 1	1 0	13 6	1 1 0 1	1 1 1 1	1 0	9 4	1 1 0 1
1 0 1 1	1 1	12 10	1 1 1 1	1 1 1 1	1 1	8 8	1 1 1 1

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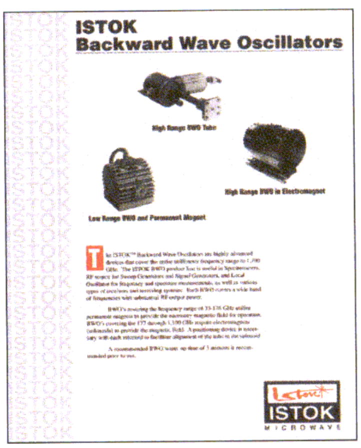

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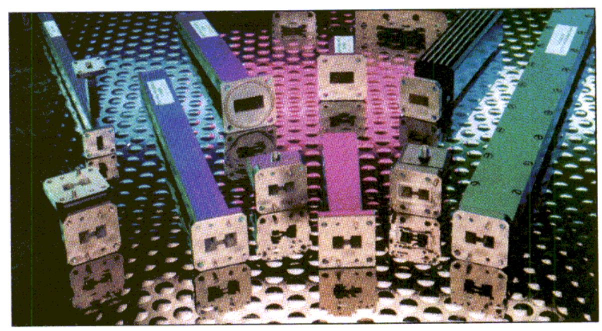

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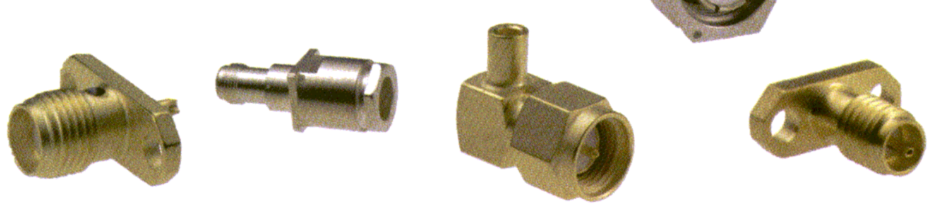

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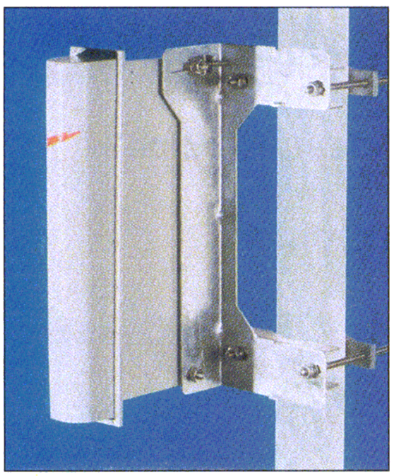
Sector antennas for broadband services

Andrew Corporation now offers three new models in its Broadband Communications Antenna (BCA) series. The High Gain Sector Antenna covers the 24.25 to 26.5 GHz band, delivering 20.5 dBi gain with 90 degree azimuth coverage. Coverage of 45 or 90 degrees is available in either vertical or horizontal polarization, and the vertical pattern is contoured to avoid nulls in the coverage area. Two models of the BCA Series Mark II give enhanced pattern performance in the 27.35 to 28.5 GHz and 37.0 to 40.0 GHz bands, with 21 dBi and 22 dBi gain respectively at 90 degree sector coverage. Azimuth patterns and cross-polarization rejection are improved over earlier models for better rejection outside the desired sector.



Andrew Corporation
Circle #196

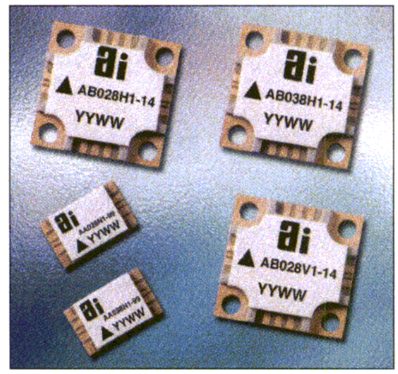
3.5 mm coaxial calibration kits

Maury Microwave has announced the 8060A series of coaxial TRL/LRL (Through-Reflect-Line, Line-Reflect-Line) 3.5 mm calibration kits. The kit is designed to work with Hewlett Packard (Agilent) vector network analyzers at frequencies up to 34 GHz. They are provided with either 3.5 mm in-series phase matched adapters or 2.4 to 3.5 mm phase matched adapters. A full 2-port calibration can be made over the entire bandwidth. In addition, 1-port or 2-port SOLT (Short-Open-Load-through) fixed load calibration can be performed, and a 1-port Short-Open (airline plus load) calibration can be performed for gated measurements. Recommended options include the A034B and A034E push-on and thread-on connector gages and the 8799A1 8 in/lbs torque wrench.

Maury Microwave Inc.
Circle #197

Packaged broadband mm-wave amplifiers

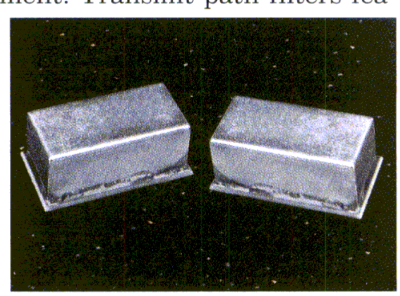
Alpha Industries has added five packaged amplifiers to its product line, supporting mm-wave sensor and communications applications. The AB028H1-14 is a high gain amplifier for 23 to 30 GHz, with 36 dB gain, +16 dBm power output and 3 dB noise figure. The AB038H1-14 offers 30 dB gain, +16 dBm power output and 4 dB noise figure. Both of these amplifiers operate from a single supply. The AB028V1-14 is a variable gain amplifier covering 23 to 30 GHz. It features 30 dB available gain, 30 dB attenuation range, +16 dBm power output and 3 dB noise figure. Two new low noise amplifiers, the AA028N1-99 and AA038N1-99, cover 23 to 30 GHz and 26 to 42 GHz respectively. The lower frequency model has 3 dB noise figure, 18 dB gain and +9 dBm output power, while the high frequency version has 3.8 dB noise figure, 18 dB gain and +5 dBm power output. The AA028N1-99 is priced at $30.98 in quantities of 10,000.



Alpha Industries
Circle #198

IF filters are designed for LMDS

Piezo Technology, Inc. (PTI) offers the 8600 series of LC filters for IF and tuner applications in LMDS/MMDS wireless network equipment. Transmit path filters feature center frequencies in the 450 to 650 MHz range, with receive path filters operating in the 1200 to 1500 MHz range. At 1 dB, passband responses are generally 75 MHz and 375 MHz for the transmit and receive paths respectively. The filters are provided in hermetically sealed packages on tape and reel for automated manufacturing.



Piezo Technology, Inc.
Circle #199

Test Systems Enhance RF Component and Power Amplifier IC Testing

Agilent Technologies, Inc. has announced new test instrumentation for RF component and RFIC production. The Agilent 87050E multiport test set can greatly improve throughput and accuracy in high volume production of 50 ohm components, while the Agilent 84000 Series Model A20e addresses high volume RFIC testing by providing a complete solution for power amplifier ICs.

The 87050E multiport test set works with the Agilent 8712E series network analyzers to characterize devices operating up to 3 GHz with as many as 12 ports. All transmission paths and port reflection characteristics can be measured with a single set of connections.

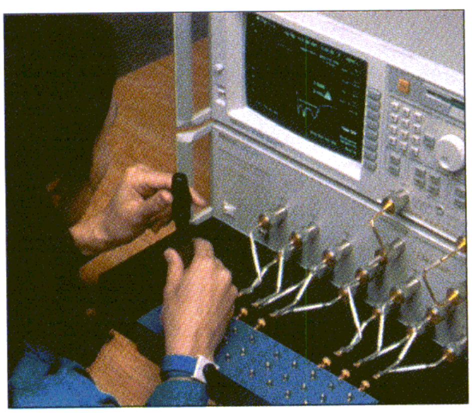
Calibration time is reduced with two types of calibration. Periodic full calibration using short, open and load standards is simplified by requiring only one connection at each port while the unit calibrates all measurement paths. In addition, a self calibration routine uses the unit's internal standards to correct for system drift. A full calibration may not be needed for as long as a month. The unit is priced from $8,250 to $14,250, depending on the configuration.

The Model A20e expands the Agilent 84000 RFIC series with the capability for complete testing of power amplifier RFICs. The system performs 2-port tests, with vector measurements from 750 MHz to 3 GHz and scalar measurements up to 8 GHz. The product is compliant with CDMA, NADC, PDC, PHS, GSM, DECT and TETRA. Measurements include input/output power, gain, ACPR, harmonics, *s*-parameters, current and voltage. The base price is under $400,000. ■

For more information, contact:

Agilent Technologies, Inc.
(A subsidiary of Hewlett-Packard Company)
5301 Steven Creek Blvd. – MS 54LAK
Santa Clara, CA 95052
Tel: 800-452-4844 ext. 6849 (87050E)
ext. 6866 (Model A20e)
Internet: http://www.agilent.com

Or circle Reader Service #201 (Agilent 87050E) or #202 (Agilent 84000 Series Model A20e)

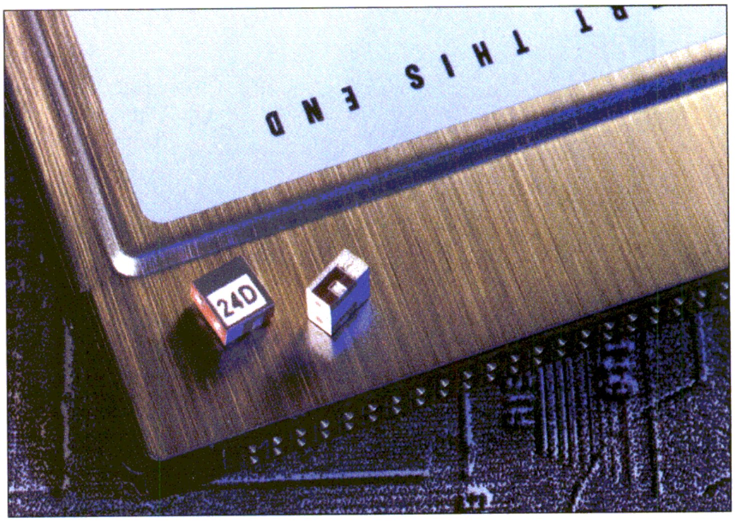


▲ **Agilent introduces new production test equipment, the Agilent 87050E multiport test set (above) and the 84000 Series Model A20e RFIC power amplifier test system (right).**

Miniature 2.45 GHz Dielectric Filter Targets PCMCIA Card Applications

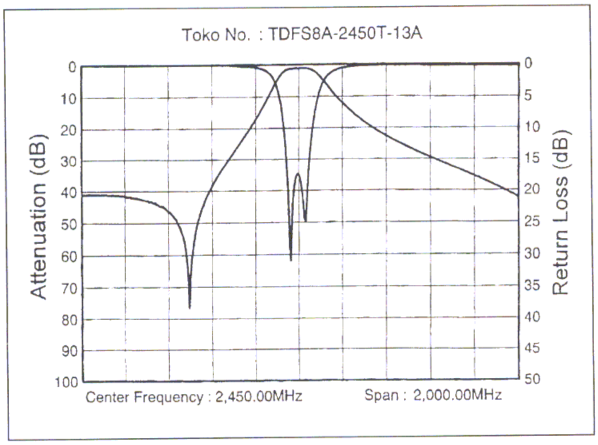
Toko announces a new dielectric bandpass filter for 2.450 GHz wireless local area networking designs using PCMCIA card for factors. The TDFS8A-2450T-13A has a footprint of 3.2 × 3.0 mm and a low profile of only 1.8 mm.

The TDFS8A-2450T-13A is a bandpass filter with an insertion loss of <2.0 dB and an average minimum attenuation of 40 dB at 1950 MHz. The filter's center frequency is 2450.0 MHz and it has a passband of $f_0 \pm 50.0$ MHz. Passband amplitude ripple is 0.7 dB maximum and VSWR in the passband is 2.0:1 maximum. The filter is designed for 50 ohm input and output impedances. Figure 1 shows a plot of the filter's frequency response and return loss.



▲ **Toko announces the TDFS8A-2450T-13A, a miniature bandpass filter for 2.45 GHz PCMCIA card applications.**

The filter is manufactured using Toko's proprietary dielectric material and process. It is provided on 2000-piece tape and reel in for automated circuit assembly. Pricing in 10,000 unit quantities is less than $1.00 each. Other Toko products include miniature inductors, integrated circuits, DC-DC converters, imaging products and electronic component assemblies. ■



▲ **Figure 1. Frequency response and return loss of the miniature 2.45 GHz filter.**

For more information, contact:

Toko America, Inc.
125 Feehanville Drive
Mount Prospect, IL 60056
Tel: 847-297-0070
Fax: 847-699-7864
E-mail: info@tokoam.com
Internet: http://www.tokoam.com

Or circle Reader Service #203

Quadrature Modulator IC Operates up to 2.5 GHz

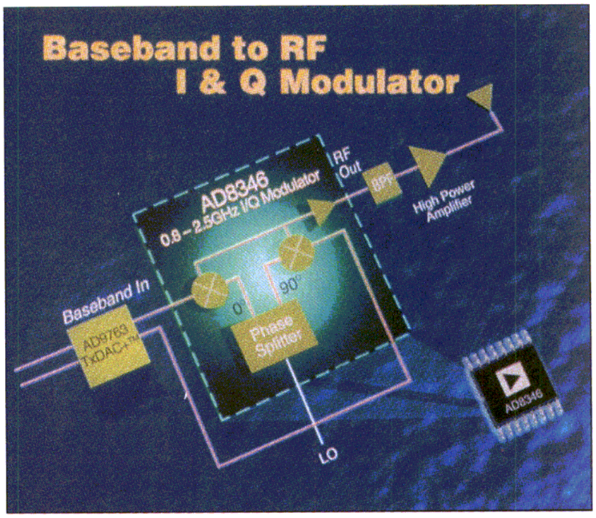
A new RF integrated circuit for wireless communications has been introduced by Analog Devices. The AD8346 is a direct quadrature (I/Q) modulator with high phase accuracy and amplitude balance over its rated frequency range of 800 to 2500 MHz.

Quadrature modulation is used in many modern communications systems, including GSM, CDMA and W-CDMA. The AD8346 provides quadrature modulation directly at the transmitter output frequency, eliminating IF stages and upconversion processes. A low power transmitter could be built using the AD8346 with an AD97xx series TxDAC and an appropriate LO signal source (e.g. frequency synthesizer). An external amplifier can be used for higher power. Key specifications of the device are summarized in Table 1.

Packaging for the AD8346 is a 16-lead TSSOP and it is specified for operation over a temperature range of –40 to +85° C. Its 2.7 to 5.5 volt supply range is suitable for today's typical design requirements. The IC is fabricated using Analog Devices' 2.5 GHz bipolar silicon process. The device is priced at $4.96 each in 1000-piece quantities.

Parameter	Specification
Output frequency	0.8 to 2.5 GHz
Quadrature phase error	1 degree
I/Q amplitude balance	0.2 dB
Output power (typ)	–10 dBm
Output VSWR (typ)	1.25:1
Output P_{1dB} (typ)	–3 dBm
Carrier feedthrough (typ)	–42 dBm
Sideband suppression (typ)	–36 dBc
IM3 suppression (typ)	–60 dBc
Output noise floor (typ)	–147 dBm/Hz
LO input drive level (min/max)	–12 dBm/–6 dBm
LO input VSWR (typ)	1.9:1
Supply voltage (min/max)	2.7 V/5.5 V
Supply current (min/max)	35 mA/55 mA

▲ Table 1. AD8346 key specifications (for operation at 1900 MHz).



▲ Analog Devices' new AD8346 offers direct quadrature modulation up to 2.5 GHz.

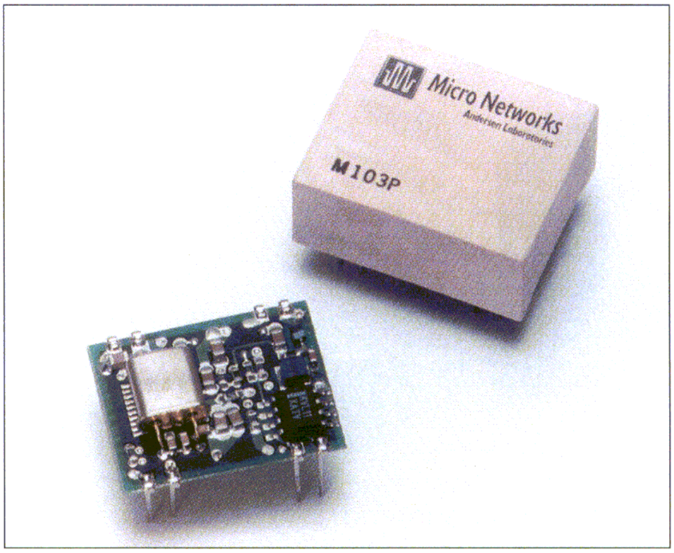
For more information, contact:

Analog Devices, Inc.
Ray Stata Technology Center
804 Woburn Street
Wilmington, MA 01887
Tel: 800-ANALOGD (800-262-5643)
Fax: 781-937-1021
Internet: http://www.analog.com

Or circle Reader Service #204

High Speed VCXOs and Master Clock Oscillators Operate at 3.3 Volts

Micro Networks/Andersen Laboratories has introduced new low voltage (3.3 V) oscillator products for high speed, high performance computing and telecommunications equipment. Applications include network servers, instrumentation, base stations, datacom/ telecommunication transmission, defense and ATE systems.



▲ **Micro Networks has developed a series of low voltage Master Clock Oscillators for high performance computer and telcommunications systems.**

Master Clock Oscillators

The M103 series of Master Clock Oscillators features 2 ps phase jitter for timing applications at frequencies up to 800 MHz. The oscillators provide a precise, stable clock source in either a SMT or DIP package for state-of-the-art or first generation products.

PECL logic is used in the M103, operating at 3.3 volts. Optimum jitter is obtained in a 50 ohm environment. temperature stability is maintained within <30 ppm variation from 0 to 70° C at frequencies up to 800 MHz. Any application requiring high frequency and precise timing can benefit from the M103 performance. OEM delivery is from stock to 18 weeks.

High performance, high speed VCXOs

The expanded line of Micro Networks' products includes the M303 series of 3.3 volt VCXOs. Originally designed for SONET applications, the M303 series has been expanded to include frequencies from 100 to 800 MHz. With 3 ps jitter in a 50 ohm environment, these VCXOs are suited for any high speed clock application.

The M303 series is available in either SMT or DIP packaging, operating from 3.3 volts with a temperature stability specification of <30 ppm over the range of 0 to 70° C. OEM delivery is rom stock to 12 weeks.

Micro Networks is ISO-9001 registered and certified to MIL-PRF-38534. ■

For more information, contact:

Micro Networks/Andersen Laboratories
324 Clark Street
Worcester, MA 01606
Tel: 508-852-5400 ext. 3110
Fax: 508-853-8296
E-mail: sales@mnc.com

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Product Spotlight

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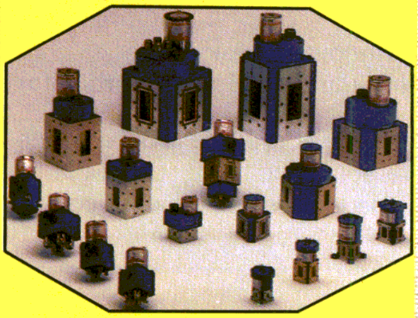

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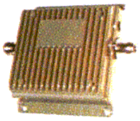

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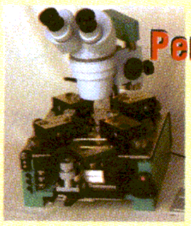
Circle 102

Personal Probe Station



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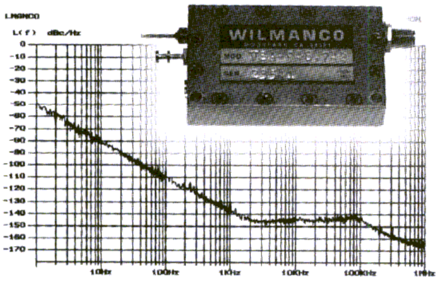

Typical measured phase noise of (2) model VSA-100 MHz Freq. Source.

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FREQ. STABILITY: *±5ppm, 0°c to +50°c*
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Circle 106

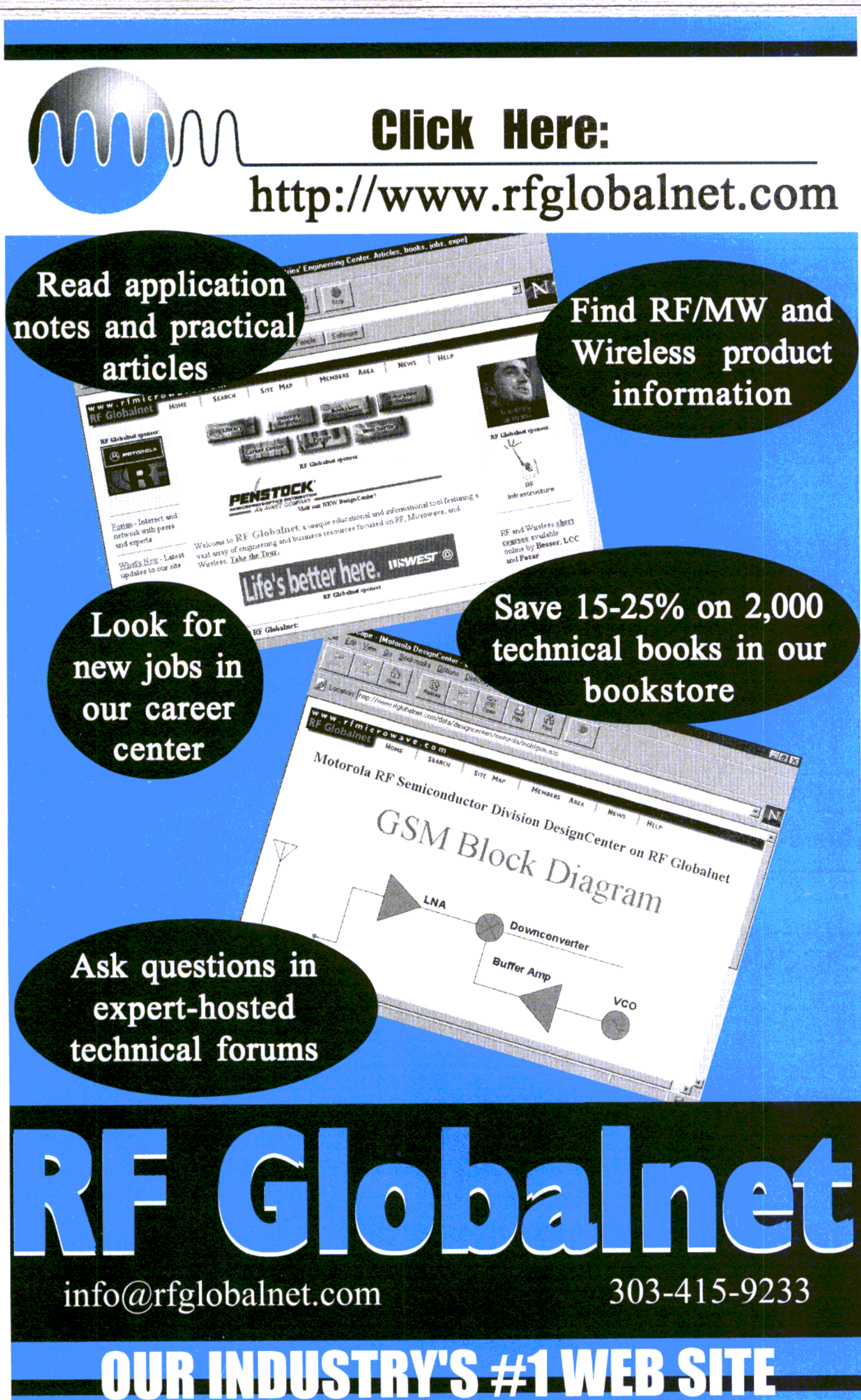
How to place Product Spotlight and Classified ads

Our Product Spotlight section allows you to show off your company's products or services in an economical 1/9 page color ad. Ads may be sent camera ready, or we'll make up an ad for you! All you'll need to provide is a product photo, company logo and a description of the product or service you want to advertise.

Our Classifieds section is the place to advertise career opportunities, consulting services, used test equipment and just about anything else. Special Classifieds rates are available and are agency-commissionable if material is supplied. Or, if you prefer, our staff can make up the ad for you.

Classified display ads are available in color or black and white in standard ad sizes, or in black and white by column inch. (Column-inch ads are 2-1/4 inches wide.)

For more information on advertising in these two sections, please contact Applied Microwave & Wireless, 4772 Stone Drive, Tucker, GA 30084; Tel: (770) 908-2320; Fax: (770) 939-0157; E-mail: amw@amwireless.com



Circle 65

Applied Microwave & Wireless

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We will notify you if we want to publish your work and will ask you to submit the final article. We'll need to receive your article at least eight weeks before the date of publication. You may consult our Editorial Calendar and indicate a topic on your proposal, although we cannot guarantee publication in a specific issue.

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Applied Microwave & Wireless (USPS 011-596) (ISSN 1075-0207), printed in the U.S.A., is published monthly by Noble Publishing Corporation, 4772 Stone Drive, Tucker, GA 30084. February 2000. Twelve issues are mailed in the United States for $30, outside the U.S. for $60, or provided free, with a completed and signed subscription form, to qualified professionals engaged in electronics engineering at 1 MHz to lightwave frequencies. Single issues, when available, are $7 in the U.S. and $12 outside the U.S. The material contained in this magazine is believed to be true and correct; however, the responsibility for the contents of articles and advertisements rests with the respective authors and advertisers. Periodical Rate postage paid at Tucker, GA 30084 and additional mailing offices.

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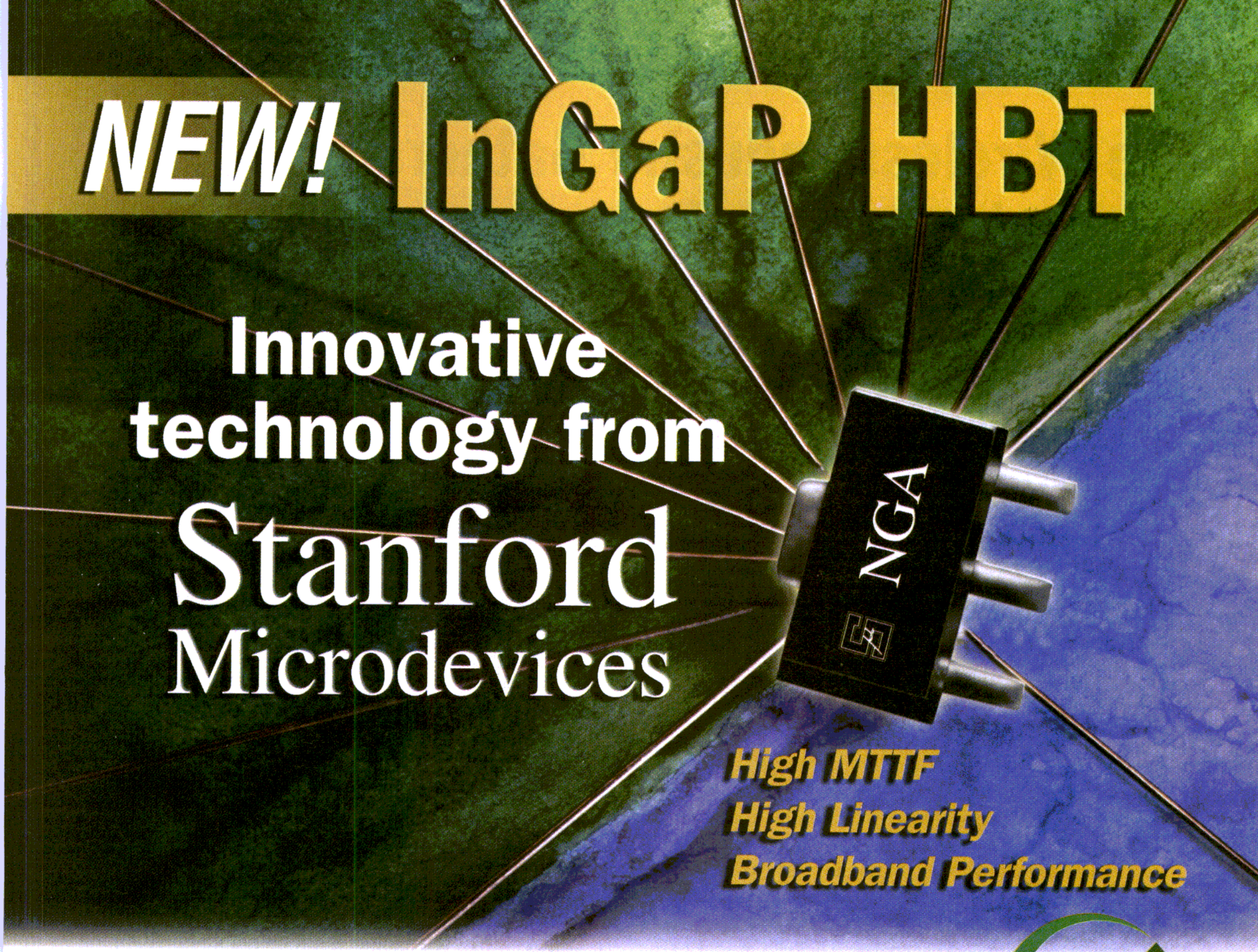

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- Cascadable 50Ω: 1.5:1 VSWR
- Low positive voltage supply
- Low thermal resistance package
- High linearity

SPECIFICATION MATRIX

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

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

NGA-589 DC-6 GHz

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

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- High gain
- High P1dB

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Circle 44

See us at Wireless Symposium, booth #921

The Wireless Industry Gathers for Business and Technology

Two major trade shows are scheduled for the end of this month. First, the Wireless Symposium will be held February 22-24 in San Jose, CA, then CTIA's Wireless 2000 event will take place in New Orleans, LA, February 28 to March 1. The first show covers components, technical forums and design matters. The CTIA show emphasizes the business of wireless and the overall technologies that drive the industry.

Wireless Symposium

The Eighth Wireless Symposium includes an exhibition with 400 companies displaying their products and services for wireless design engineers and system developers. In addition, some 75 technical papers describe new products and engineering techniques useful for the design, testing and manufacture of wireless products.

A significant part of this event is a series of workshops, providing concentrated technical instruction. These courses include:

Wireless Made Simple — Instructor: Al Scott, Besser Associates, Part I, Tuesday February 22; Part II, Wednesday, February 23.

RF Fundamentals — Instructor: Rick Fornes, Besser Associates, Part I, Tuesday February 22; Part II, Wednesday, February 23.

Topical Issues in RF Power Amplifier Design — Instructor: Steve Cripps, Besser Associates, Part I, Tuesday February 22; Part II, Wednesday, February 23.

Measuring Wireless — Instructor: Morris Engleson, JMS, Tuesday, February 22.

Fundamentals in CDMA Technology for PCS — Instructor: Darryl Schick, Linear Lightwave, Wednesday, February 23.

DSP Made Simple — Instructor: Rick Lyons, Besser Associates, Part I, Thursday, February 24; Part II, Friday, February 25.

Digital Modulation — Instructor: Joe Boccuzzi, Besser Associates, Part I, Thursday, February 24; Part II, Friday, February 25.

Oscillator Design — Instructor: Randy Rhea, Eagleware, Thursday, February 24.

3G Wideband CDMA Technology — Instructor: Darryl Schick, Linear Lightwave, Thursday, February 24.

Antennas and Propagation for Wireless Communications — Instructor: Steven Best, Cushcraft Corp., Friday, February 25.

Phase Locked Loops — Instructor: Eric Drucker, PLL Consultants, Friday, February 25.

RF Wireless System Fundamentals — Instructor: Rick Fornes, Besser Associates, Friday, February 25.

For more information on the Wireless Symposium, or to register, call 1-888-WireReg or go online to www.WirelessPortable.com.

CTIA Wireless 2000

New Orleans is a very good venue for a business-to-business trade show, with an excellent convention center and plenty to do after show hours. Following the theme "Wireless Today & Tomorrow," Wireless 2000 includes both a large exhibition and informational seminars. Instructional meetings covering business and technology topics include:

Wireless Data University, presented by Andrew Seybold and Alan Reiter, Sunday, February 27. Register by calling 831-338-7701.

CTIA's Latin American Conference, held on Wednesday, March 1.

Panel sessions, held Monday, February 28, and Tuesday, February 29 — *Management Issues in Business*, *Marketing Issues in Wireless*, *Technology Issues in Wireless Data*, *Wireless Data Consumer Products and Marketing*, *Distribution*, *Borderless Wireless* and *Public Policy*.

A highlight of the event will be the three keynotes featuring industry and government leaders:

International Roundtable — Monday, February 28, with Chris Gent, Chief Executive of Vodaphone AirTouch plc, Keiji Tachikawa, President of NTT Mobile Communications Network, and Anthony S.K. Wong, Director-General of Telecommunications Authority, OFTA.

Internet Anywhere — Tuesday, February 28.

A Conversation with Nelson Mandela — Wednesday, March 1, the former President of the Republic of South Africa will discuss issues of importance in a world where technology and social change continue to play major roles in our lives.

For information and registration for CTIA's Wireless 2000, call 1-800-421-6338 or 1-415-979-2250. On the Internet, show information can be found at the event's comprehensive Web site: www.ctiashow.com.

Applied Microwave & Wireless will be attending both events to learn about issues that affect our readers and advertisers, and to stay abreast of technical and business developments. ■