A PENTON PUBLICATION FEBRUARY 2000

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FOR DESIGNERS AT HIGHER FREQUENCIES

Microwaves & RF

Semiconductor Issue

NEWS
Semiconductor technology heads for new territory

DESIGN FEATURE
Apply a monolithic directional detector

PRODUCT TECHNOLOGY
Noise generator tests high-speed satcom systems

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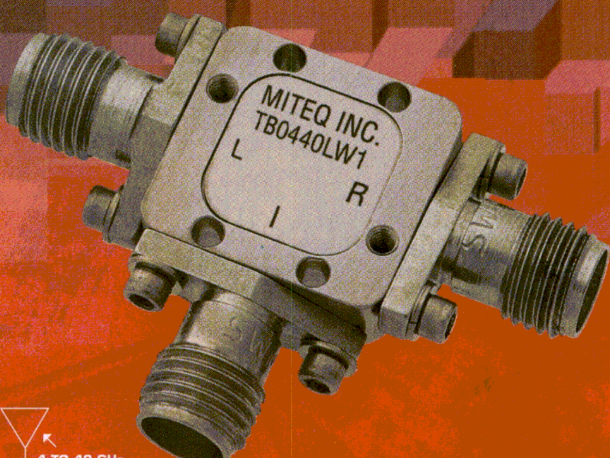
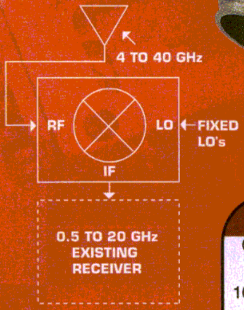
4 - 40 GHz BLOCK DOWNCONVERTER

MITEQ's Model TB0440LW1 allows the use of existing wide bandwidth receivers over millimeter frequency bands!



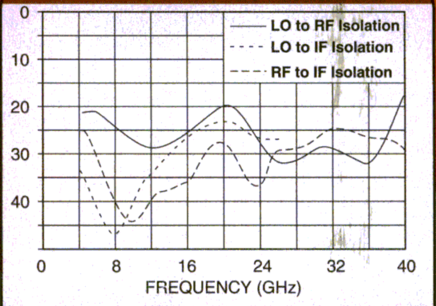
FEATURES:

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

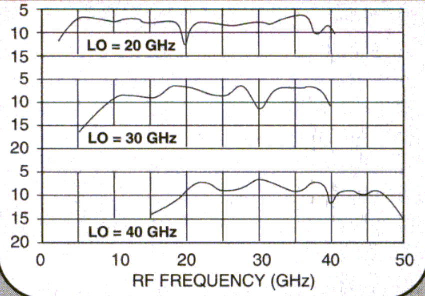


INPUT PARAMETERS	MIN.	TYP.	MAX.
RF frequency range (GHz)	4		40
RF VSWR (RF = -10 dBm, LO = +13 dBm)		2.5:1	
LO frequency range (GHz)	4		42
LO power range (dBm)	+10	+13	+15
LO VSWR (RF = -10 dBm, LO = +13 dBm)		2.0:1	
TRANSFER CHARACTERISTICS	**MIN.**	**TYP.**	**MAX.**
Conversion loss (dB)		10	12
Single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

ISOLATION LO/RF, LO/IF AND RF/IF



CONVERSION LOSS (LO=+13 dBm)



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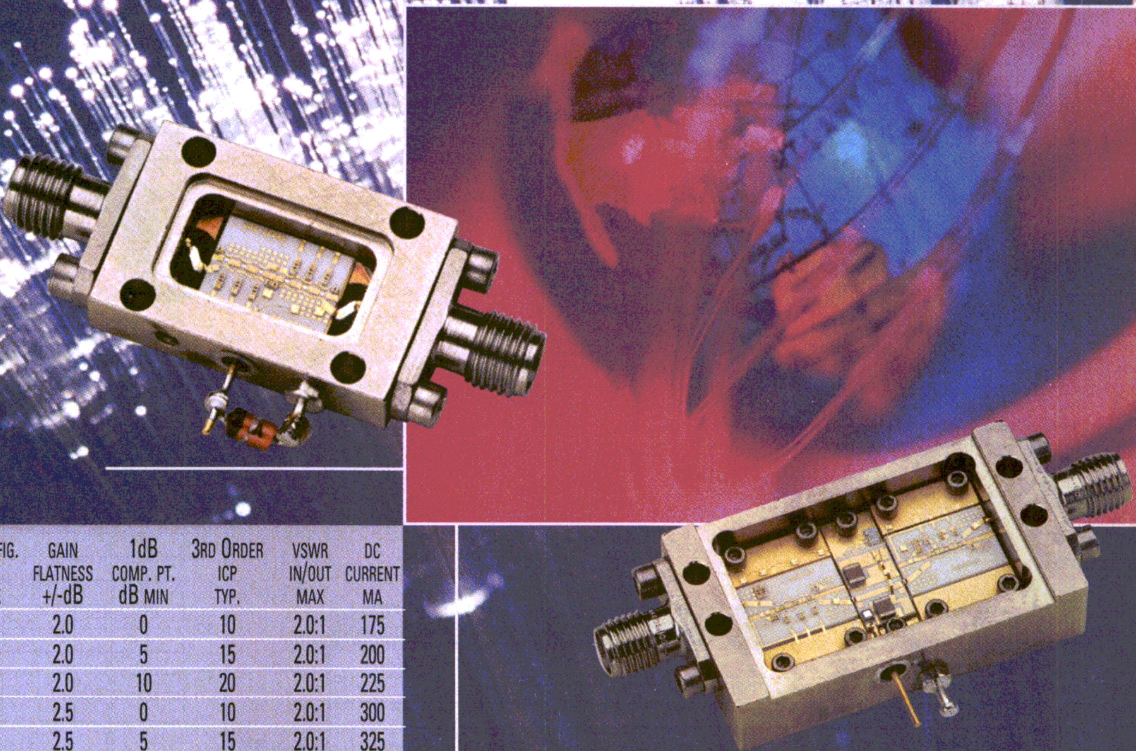

MODEL NUMBER	FREQ. RANGE GHZ	GAIN dB MIN	NOISE FIG. dB MAX	GAIN FLATNESS +/-dB	1dB COMP. PT. dB MIN	3RD ORDER ICP TYP.	VSWR IN/OUT MAX	DC CURRENT MA
JCA008-201	.01-8.0	25	*5	2.0	0	10	2.0:1	175
JCA008-202	.01-8.0	24	*5	2.0	5	15	2.0:1	200
JCA008-203	.01-8.0	22	*5	2.0	10	20	2.0:1	225
JCA008-301	.01-8.0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

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

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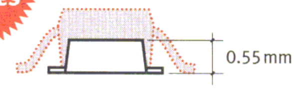

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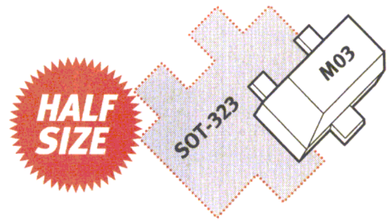
Part Number	Corner Freq*	V_{CE}	I_C	Package
NE856M03	3 KHz	3 V	30 mA	**M03**
NE685M03	5 KHz	3 V	5 mA	**M03**

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



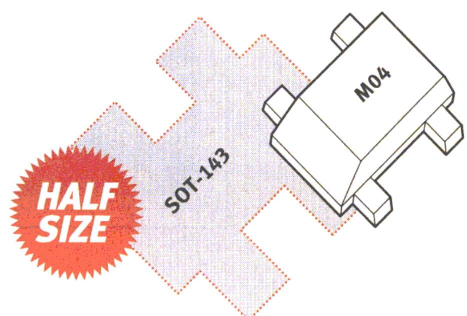
New M03

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**

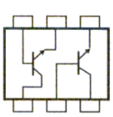


New M04

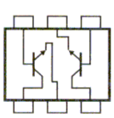
Half the footprint area of a SOT-143

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 NO. 288

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FEBRUARY 2000 VOL. 39, NO. 2

Microwaves & RF

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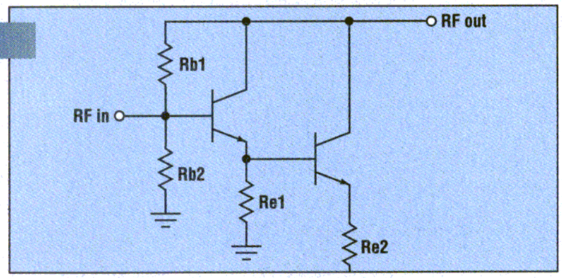
COVER FEATURE



DEPARTMENTS

NEWS

SPECIAL REPORT

DESIGN FEATURES

PRODUCT TECHNOLOGY

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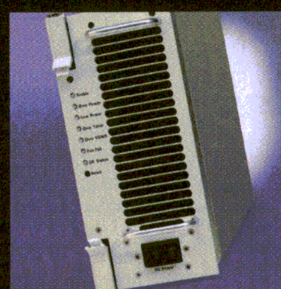

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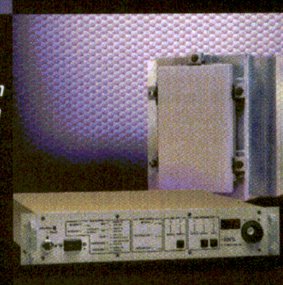

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Kit C124-2 (2% tolerance) $100

0805CS Series
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Kit C103-2 (2% tolerance) $85

NEW! 0805HQ *High Q* Series
2.5 nH - 51 nH, 5% tolerance
12 values (10 each) Kit C125 $60

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10 nH - 4.7 µH, 5% tolerance
36 values (10 each) Kit C100 $100
Kit C100-2 (2% tolerance) $120

NEW! 1008HQ *High Q* Series
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NEW! DO3316HC High Current Series
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DO1608/3316 Series
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DT1608/3316 Shielded Series
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DS5022 Shielded Series
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DO5022 Series
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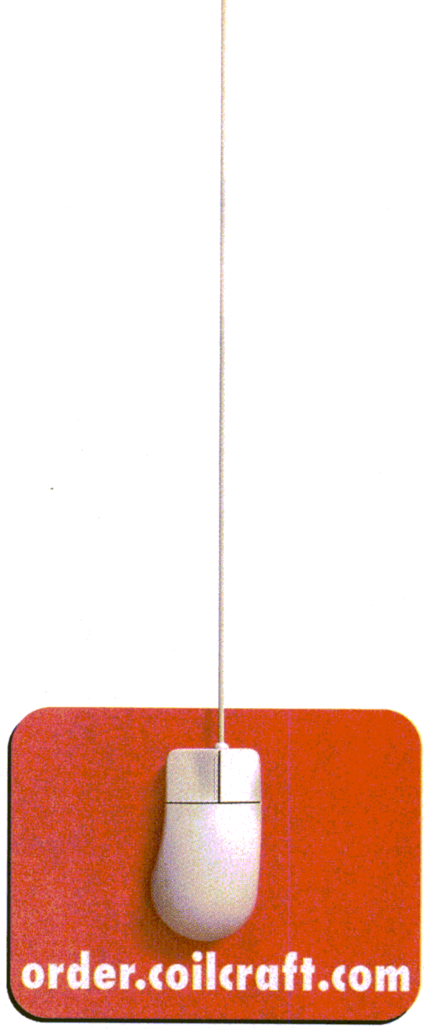

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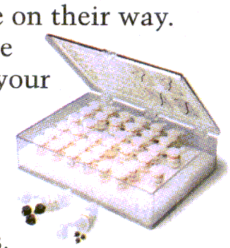

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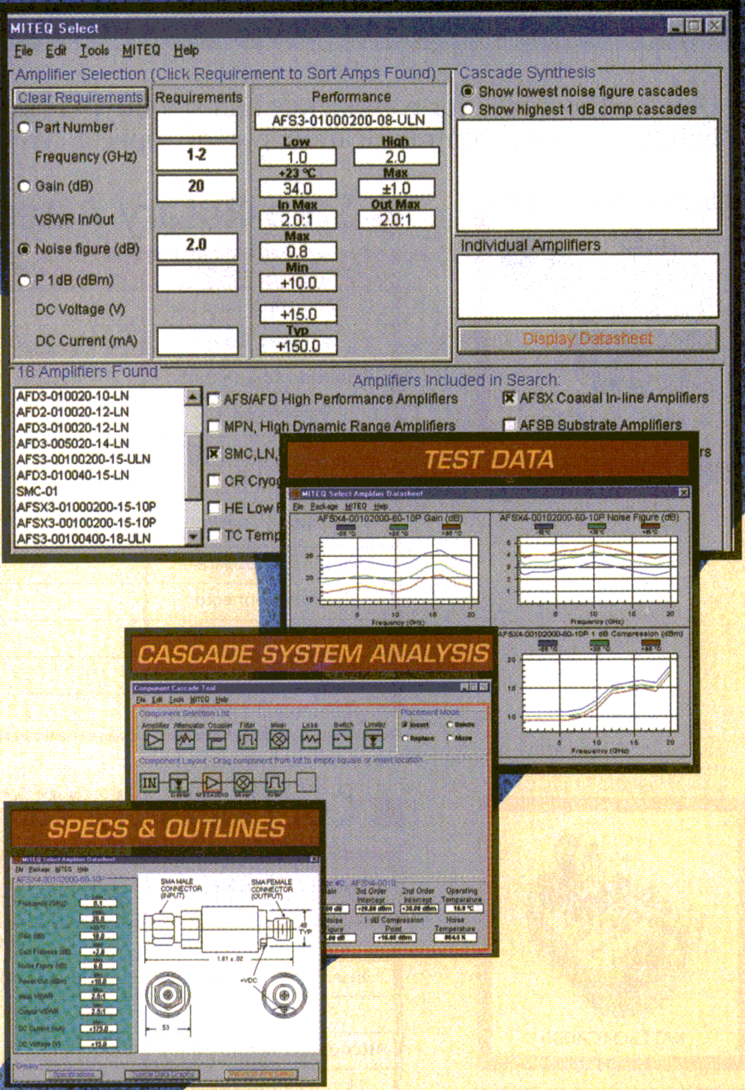

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CIRCLE NO. 232

REALISTIC EFFICIENCY

To the editor:

The November 1999 issue (on page 43) of *Microwaves & RF* reports that Matsushita has a 200-W RF amplifier that only uses 150-W DC (+15 VDC at 10 A) Now that is smokin'.

Doug McGarrett
Senior Engineer
Ademco
Syosset, NY

Editor's Note:

Mr. McGarrett is correct in noting that a 200-W amplifier driven by +15 VDC and 10 A (150 W) would be impressive. Even at 100-percent efficiency, the amplifier would require +20 VDC and 10 A. Unfortunately, these were the numbers provided in preview abstracts from the recent International Electron Devices Meeting (IEDM). More likely, the amplifier draws 10 A from a +48-VDC supply (or the equivalent of 20 A from a +24-VDC supply), which would yield a more realistic efficiency of slightly more than 41 percent.

DUPLICATED ERROR

To the editor:

I realize that mistakes do happen but this same mistake appeared twice. Despite the information in the December 1999 and January 2000 issues (pages 46 and 45, respectively) that claimed Noise Com "announced the execution of a definitive agreement to merge Boonton Electronics Corp. into a wholly owned subsidiary of Noise Com," this merger was announced to be terminated on October 26, 1999. Wireless Telecom Group, Inc. advised Boonton Electronics Corp. that it elected to terminate the merger that was originally announced on September 7, 1999, pursuant to the terms of the agreement.

Joanne Calandra
Noise Com, Inc.
Paramus, NJ

Editor's Note:

Our sincerest apologies to the folks at Noise Com as well as to the people of Boonton Electronics Corp. (Parsippany, NJ) since both companies were falsely represented in both the December and January issues of Microwaves & RF. *How could this happen twice? Unfortunately, in stepping up our magazine production schedules for 2000 (readers should be receiving each monthly issue about two weeks sooner this year), the production schedule for January was tightly compressed and directly followed the completion of the December 1999 issue. The same mistake did slip past our usually "eagle-eye" editors not once but twice and, again, to Noise Com, Boonton, and our readers, we apologize.*

Address letters to:

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600S

ULTRA-LOW ESR

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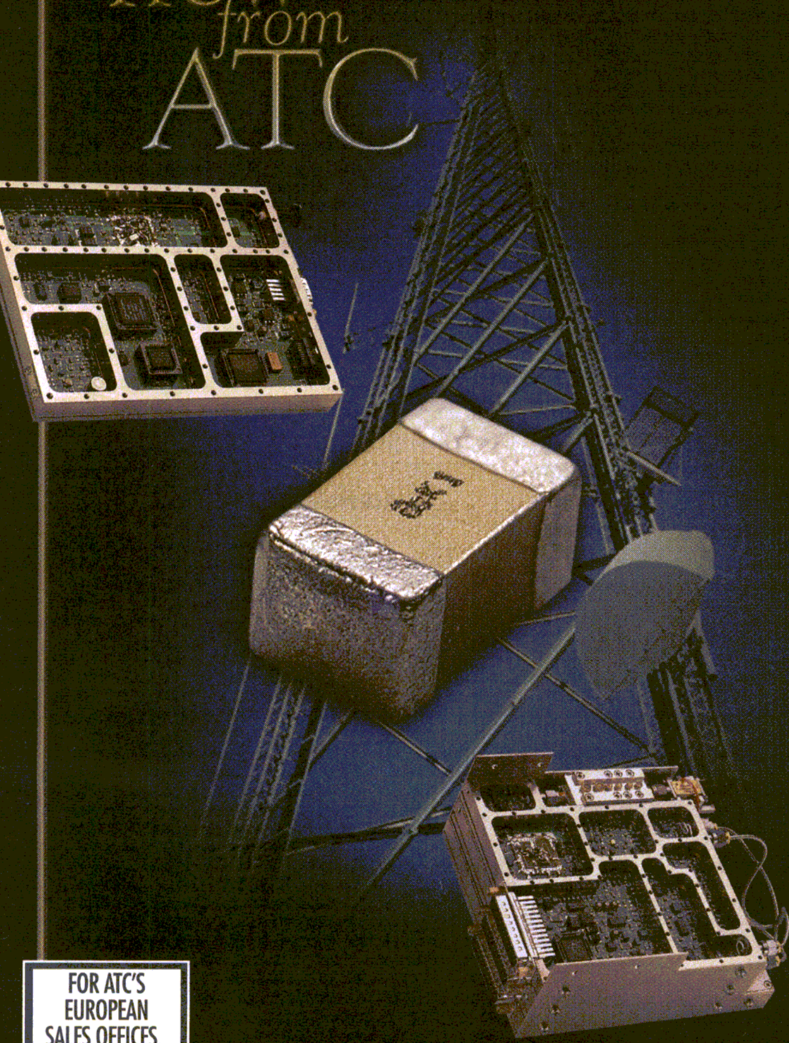
new *from* ATC



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SPECIFICATIONS

ATC SERIES	600 S
CASE SIZE:	0603
CAP. RANGE:	UP TO 100 PF
WVDC:	250 V
ESR @ 1 GHZ:	80 mΩ TYP.
DIELECTRIC:	NPO
MARKING:	EIA CODE
PACKAGING:	TAPE & REEL PLUS OPTIONS
OPTIONS:	SPECIAL TESTING & PACKAGING

American Technical Ceramics

CIRCUIT DESIGNER'S NOTEBOOK

Capacitor ESR Measurement Technique

Equivalent Series Resistance (ESR) is the summation of all losses resulting from dielectric (Rsd) and metal elements (Rsm) of the capacitor. Dielectric loss tangent of ceramic capacitors is dependant upon specific characteristics of the dielectric formulation, level of impurities, as well as microstructural factors such as grain size, morphology, and porosity (density). Metal losses are dependent on resistive characteristics of the electrode and termination materials, as well as the associated frequency dependent losses in electrodes due to skin effect. ESR is a key parameter to consider when utilizing capacitors in RF designs. A reliable and repeatable test method must be implemented in order to establish valid capacitor ESR characterizations.

Measurement Methodology:

Measuring the ESR of high Q ceramic chip capacitors requires a test system with an inherent Q higher than the device under test (DUT). A high Q coaxial resonant line is most commonly utilized for these measurements. The coaxial line resonator is typically constructed from copper tubing and a solid copper rod for its center conductor. The DUT is placed in series between the center conductor and ground.

Before performing ESR measurements the unloaded characteristics of the resonant line must be established. This is accomplished by providing RF excitation to the shorted coaxial line and ascertaining the 1/4 and 3/4 lambda bandwidth. The line is then open circuited after which the 1/2 and 1 lambda bandwidth measurements are established. This data is used to characterize the unloaded Q of the resonant line, fixture resistance and resonant frequency. The unloaded Q of the line is typically in the order of 1300 to 5000 (130MHz to 3GHz) with a fixture resistance rfo in the range of 5 to 7 milliohms.

The capacitor sample is placed in series with a shorting plunger located at the low impedance end of the line. The generator is tuned for a peak resonant voltage, and then re-tuned to 6dB down from the peak voltage on both skirts of resonance. A loosely coupled RF millivoltmeter probe located at the high impedance end of the line (approximately at 1/4 wavelength from the shorted end) will measure RF voltage at the 6dB points.

The DUT perturbs the Q of the line changing the resonant frequency and bandwidth as compared to the unloaded line. The corresponding 6dB down frequencies referred to as fa and fb are used in the calculation of the capacitor's ESR. This process is referred to as the Q perturbation method. See Figure 1.

Note: Since the capacitive reactance of the test sample is in series with the line, it will shorten its electrical length depending on capacitor value. Values above 10pF will yield reasonable measurement accuracies however, as we approach 1pF the measured ESR may develop substantial errors. The small capacitance values exhibiting high Xc will cause the electrical length of the line to drastically change. The reactance of the line is equal and opposite to that of the DUT, at resonance.

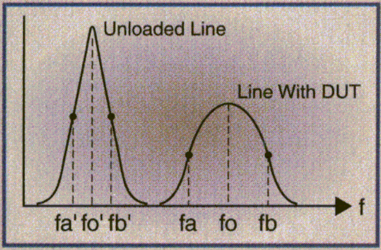


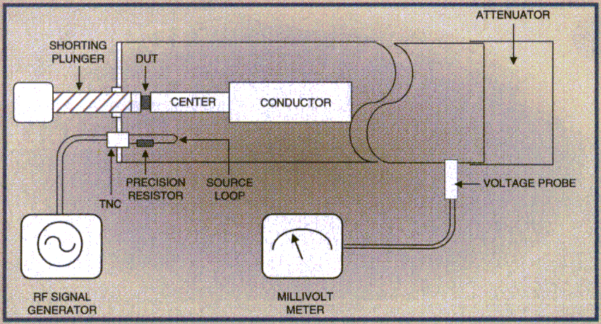
Figure 1: (Two Bandwidth Curves)

ESR Test System:

The test system most commonly used consists of a coaxial line (Boonton Model 34A) nominally (57.7cm) in length, with a resonant frequency of 130MHz and a characteristic impedance of 75 ohms. This impedance is chosen because it yields the highest line Q. Different line lengths may also be used for other frequency ranges.

A signal generator is connected to the low impedance end of the line and terminates in a non-inductive precision resistor. The resistor is mounted on a TNC connector and inserted into the DUT end of the line. It has an exposed loop that serves to loosely couple RF energy into the line. An RF excitation of 1 mw (0dBm) drives the shorted line through the source loop. The generator is swept until a peak resonant voltage is displayed on the RF millivoltmeter. The source loop is physically rotated until a 3 millivolt reference voltage is achieved at the high impedance end of the line. This procedure insures that the RF excitation does not load the line. See Figure 2.

An RF probe located at the high impedance end of the line is connected to a millivoltmeter to measure RF voltage at resonance. From these measurements the bandwidth and Q can be established. ESR is calculated by equating the change in bandwidth (BW) and Q, as compared to the initial unloaded shorted line condition. The BW data is put into an equation along with the initial line characterizations to calculate the ESR of the test sample. ESR measurements described here are performed in the series mode and can be achieved up to about 3 GHz.



See Figure 2: (Coaxial Resonator with DUT)

Factors Affecting ESR Measurement :

- Frequency measurement data for establishing BW require a minimum of four decimal places however, five places is desirable.
- Source and measurement probes must be loosely coupled to the line.
- The high impedance end of line should be shielded to reduce loss due to radiation to preserve Q. The shield is a cut-off attenuator offering16DB attenuation per radius.
- Placement of the DUT in the line fixture should be consistent.
- Keeping fixture contact surfaces clean is essential for good repeatability.

Richard Fiore
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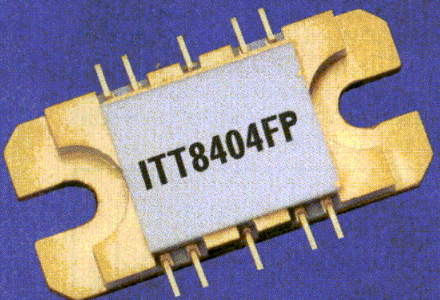

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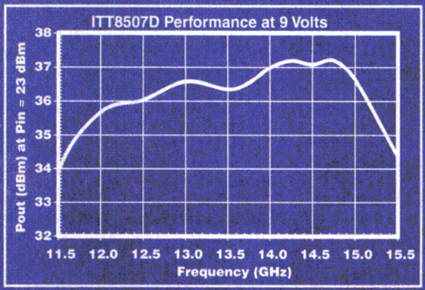

Part #	Description	Freq. (GHz)	Gain @ P1dB (dB)	P1dB (dBm)	IP3 (dBm
C Band					
ITT6401FN	2W, 2 Stage	4.5 - 6.9	20	33	42
ITT8403FP	4W, 2 Stage	5.25 - 7.0	18	34	46
ITT8404FP	6W, 2 Stage	6.0 - 7.2	16	37	43
Ku Band					
ITT8502FP	2W, 3 Stage	13 - 15	17	33	39.5
ITT8507FP	4W, 3 Stage	12.5 - 14.5	15.5	36	43
ITT8602FN	1W, 5 Stage	12 - 16	34.5	29.5	40

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MOURNING THE LOSS OF A QUIET LEADER

Voltronics Corp. is a small company, often regarded as a well-kept secret in the high-frequency industry. To its customers, it is a leading supplier of precision microwave trimmer capacitors. To outsiders, it is a quiet facility just off the main road through Denville, NJ. In many ways, the firm has always mirrored the personality of its founder, Richard J. Newman, a quiet leader who taught by example. For many, inside and outside Voltronics, Dick Newman served as an inspiration. Sadly, he passed away on the last day of 1999. He was 79 years old.

Dick Newman was more than just a business acquaintance. Oh, that may have been how our relationship started some 18 years ago, but it quickly evolved into much more. In the beginning, Dick served as engineering teacher to an eager young student, ever anxious to hear his tales of the microwave industry, of his times at other companies, and his opinions on engineering, business, and morality. Later, the relationship became one of true friendship, of lending an ear during painful times, and sharing a laugh during good times. And although this friendship included the continuous offer of a helping hand, Dick was not the type to ask for help. But he was certainly quick to give it.

Dick never lived to see the year 2000, but he was as forward-thinking an engineer or company executive that I have ever met. In themselves, trimmer capacitors are not dramatic. But Dick made them seem exciting because of his enthusiasm. He was fond of explaining a new product development or a new application for his company's trimmers. And he seemed to find never-ending ways to improve the many types of capacitors offered by Voltronics.

Although he founded Voltronics (in 1963), Dick was able to share the reins with those who worked for him, including his son, Scott. Dick gave his son tremendous responsibility, even during Scott's earliest days with the company. He displayed unwavering trust in Scott's own leadership and management choices during good periods as well as bad ones, as Scott earned the title of President and eventually took over the day-to-day operations.

During his later years, Dick would call on an almost monthly basis to meet for lunch. It was a time we both enjoyed, to catch up with each other and on the industry. And, although I knew that Dick and Voltronics often faced severe competition from other trimmer-capacitor manufacturers, he never uttered a derogatory word about a competitor, choosing instead to credit them for their efforts. Despite the business wars, Dick enjoyed sharing stories of his wife, his sons John and Scott, and his grandchildren. His priorities were well-placed. The company was important, but family was everything.

Dick Newman taught me the human side of engineering, that engineers have families and concerns that go far beyond formulas and calculations. He had many long-time friends probably because he understood some of the most important elements of friendship—compassion, loyalty, and honesty. He leaves behind a strong family, a solid company, and many friends who will miss him.

Jack Browne

Publisher/Editor

Microwaves & RF

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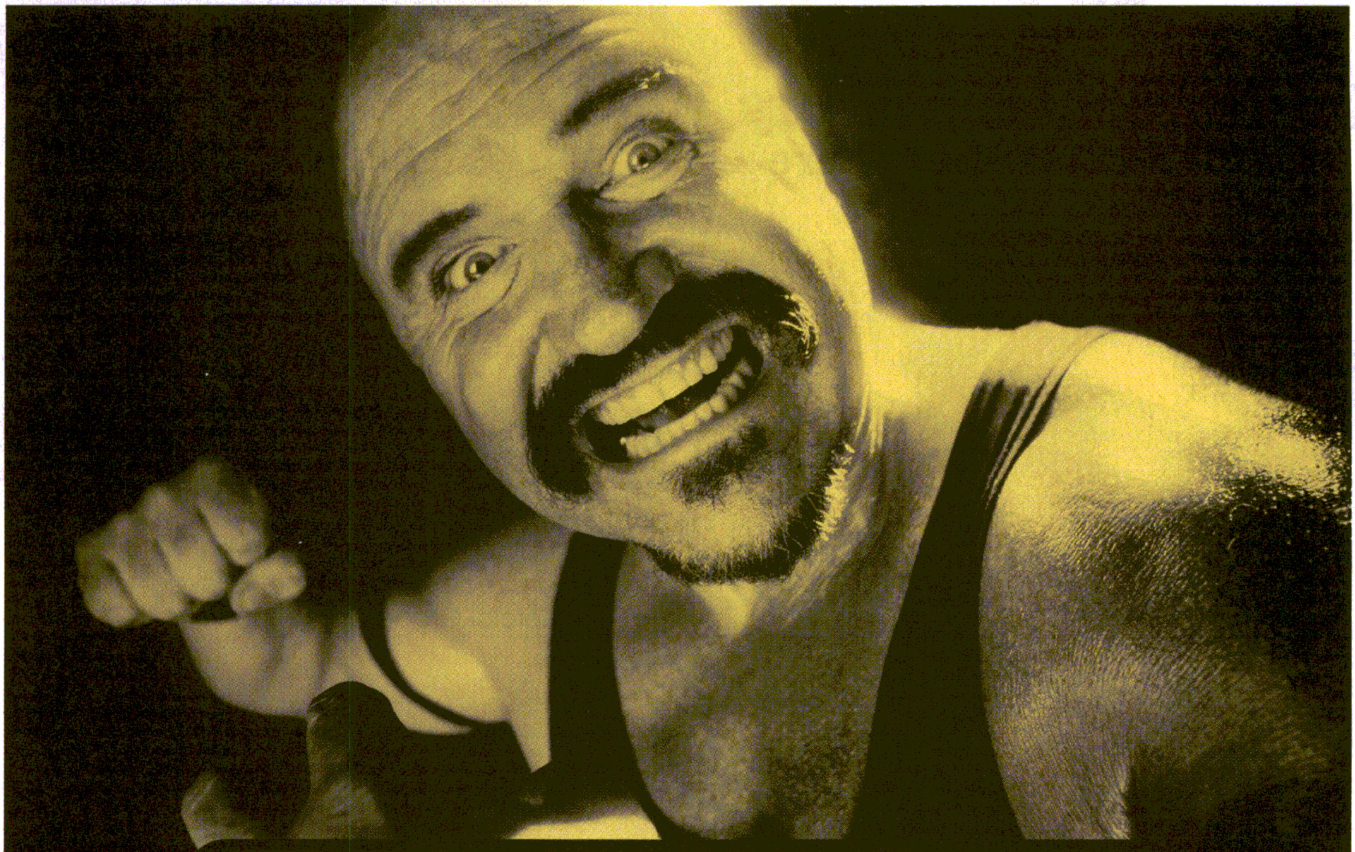

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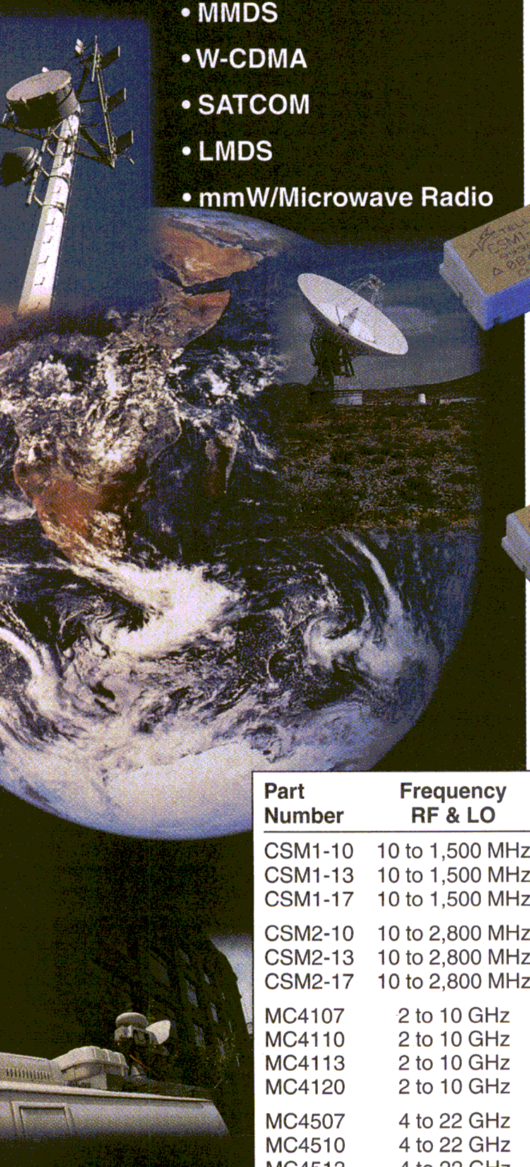

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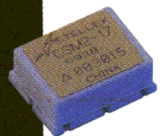

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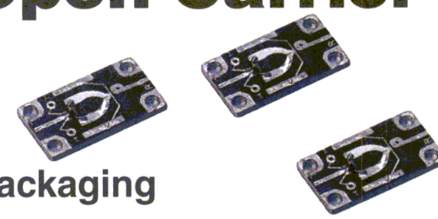

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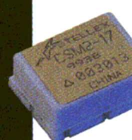

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CSM1-17	10 to 1,500 MHz	1 to 500 MHz	+17 dBm	40 dB	27 dBm	7.5 dB	Surface Mount
CSM2-10	10 to 2,800 MHz	10 to 2,000 MHz	+10 dBm	30 dB	20 dBm	7.5 dB	Surface Mount
CSM2-13	10 to 2,800 MHz	10 to 2,000 MHz	+13 dBm	30 dB	22 dBm	7.5 dB	Surface Mount
CSM2-17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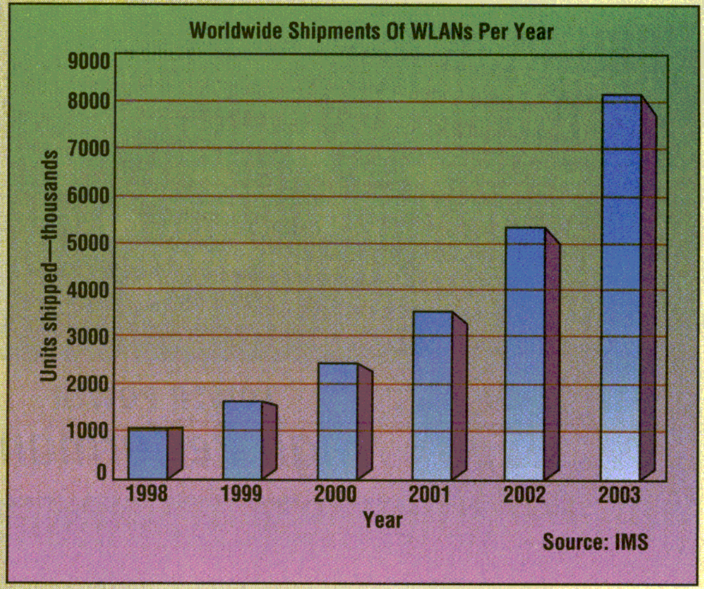
SoHo Adoption Set To Explode WLAN Market

WELLINGBOROUGH, NORTHANTS, ENGLAND—The worldwide market for wireless local-area networks (WLANs) is expected to explode to 7.6 times its present size according to Intex Management Services Ltd. (IMS), a market-research company. In an IMS report, "The Worldwide Market for Wireless LANs," the market is projected to be driven by the adoption of WLAN units due to the increase in transmission speed, the reduced cost of products, and consumer education.

As shown in the figure, the worldwide market for WLANs is projected to increase from 1.07 million units (i.e., NICs, access points, and building-to-building bridges) in 1998 to more than 8.14 million units by 2003. With recent developments from leading WLAN companies, data-transmission speed has increased from 2 to 10 Mb/s with the next generation of WLAN adapters. The increase in speed is attractive to many businesses, since the WLAN is now considered a direct alternative to a 10-Mb/s wired LAN, such as 10base-T Ethernet.



Additionally, the cost of WLANs and components is eroding to consumer cost points, making WLANs more cost-effective for small, medium, and large enterprises. Short-term growth is projected to be attributed to SME/SoHo implementation of the WLAN products. Medium-to-long-term growth is expected to be caused by penetration into the home/consumer market.

Joint Initiative Fosters Mobile Internet Applications

STOCKHOLM, SWEDEN—Ericsson and industry leaders IBM, Lotus, Oracle, Palm Computing, and Symbian are joining forces in the GPRS Applications Alliance (GAA), a cross-industry initiative that is designed to serve as a catalyst in the advancement of applications based on the mobile packet-switching technology, general-packet radio services (GPRS).

GPRS introduces packet data to mobile networks and is a first, vital step for Global System for Mobile Communications (GSM) and time-division-multiple-access (TDMA) operators in the evolution to third-generation (3G) mobile networks, enabling a range of new and enhanced services in a mobile environment.

The alliance is open to any organization that is interested in mobile communications, such as software developers, systems integrators, network operators, and other infrastructure vendors as well as device manufacturers.

Ericsson is initially establishing two application centers—one in Silicon Valley, CA and one in 'Mobile Valley,' Kista near Stockholm, Sweden. The new centers provide a unique environment for end-to-end testing of GPRS applications.

"The commitment of these major players in the industry will certainly add value and help the rapid take-off of the mobile Internet applications market," says Per Nordlöf, general manager of packet-switching systems at Ericsson Network Operators. "GAA participants are able to test and verify end-to-end their mobile Internet solutions and will be in a strong position to guide the further development of open standards for applications and services. Ericsson is involved from the infrastructure side and with its WAP-capable devices, Symbian's EPOC-based devices, and other WAP phones and wireless information devices."

The GAA home page can be found at http://www.gprsworld.com.

World's Largest Broadband Fiber In The Loop Deployment Is Planned

PITTSBURGH, PA—Marconi recently announced a major commitment by BellSouth to deploy voice, video, and asynchronous-transfer-mode (ATM)-based high-speed data technology throughout the BellSouth network using Marconi's newly introduced ATM-based DISC*S MX access platform. Marconi projects that BellSouth's planned deployment will generate more than $1 billion in sales revenue for the company over the next three years. This announcement follows the June 1999 commitment from BellSouth to Marconi for Integrated Fiber in the Loop (IFITL) and represents a strong next step in BellSouth's broadband deployment strategy. It builds on BellSouth's commitment to fiber-optic distribution deep into its network, based on Marconi's Deep Fiber technology.

Early this year, BellSouth began deploying Marconi's next-generation DISC*S MX products. DISC*S MX delivers enhanced, ATM-based broadband services in the framework of the "field-proven" DISC*S equipment.

BellSouth has been deploying Marconi's first-generation "fiber-to-the-curb" products since 1995 and serves approximately 500,000 homes with fiber-optic distribution capacity. During 1999, BellSouth announced plans to deploy 280,000 lines of fiber-optic distribution in the region. The deployment is concentrated in Atlanta and south Florida, and is part of BellSouth's commitment to deliver broadband voice, video, and data services to consumers. The success of this initial deployment program has led to the decision to accelerate fiber-optic deployment into other BellSouth-served geographies.

Breakthrough In New Transistor Material Is Announced

TEMPE, AZ—Motorola Labs recently announced that it has built the world's thinnest functional transistor using a new class of semiconductor materials that will enable future transistors to be exponentially smaller and faster while consuming less power. Motorola Labs has successfully built a working device that uses a class of perovskite materials never used in a transistor before. This represents the first fundamental change in the materials used to build transistors over the past 30 years.

The new technology enables the development of a transistor with an effective thickness that is initially three to four times thinner than those built with today's conventional semiconductor materials. While this technology is still in the early stages of development, it has already produced working devices that are electrically much thinner than those made with existing technology. Perovskites are a class of crystalline-oxide materials with unique material properties.

"As devices continue to shrink in size, the gate oxide of the transistor also needs to become thinner. However, we are quickly reaching the limit where we can no longer thin the silicon dioxide which has been used as a gate insulator for the last 30 years," says Jim Prendergast, vice president and general manager of Motorola's physical Sciences Research Lab (PSRL). "The solution is to use a new family of materials that appear electrically to be much smaller than their actual physical thickness."

Companies Cooperate In Development Of Bluetooth ICs

GENEVA, SWITZERLAND—STMicroelectronics recently announced that it has entered into an agreement with Ericsson Mobile Communications where the two companies will develop a range of devices based on the Ericsson Bluetooth core architecture. In addition, within the framework of this architecture, the companies are working together to approach the open market with optimized Bluetooth solutions. The agreement builds on both companies' positions in wireless technology and will significantly increase adoption of the proven Bluetooth core in products offering wireless communications capabilities. This agreement demonstrates STMicroelectronics' commitment to Ericsson's Bluetooth baseband architecture for the development of leading-edge solutions.

The first device will be implemented in ST's advanced 0.18-μm complementary-metal-oxide-semiconductor (CMOS) technology and will combine RF functions with logic and digital processing required for establishing and controlling communication links between Bluetooth-enabled devices. Specifically developed and thoroughly characterized in the RF domain, ST's RFCMOS8 technology permits the integration of complex digital circuits with RF transceivers operating in the industrial-scientific-medical (ISM) gigahertz frequency range.

Single-Chip Cable Tuner Is Developed

SAN DIEGO, CA—Silicon Wave, Inc. recently introduced its SiW100™ cable-tuner integrated circuit (IC), the first tuner to completely integrate all of the critical RF elements onto a single, low-power device for cable modems and cable set-top boxes. Designed to meet the price, power, and size requirements for consumer broadband-access devices, the SiW100 chip includes all of the front-end receiver and channel-selection functions of a high-performance, dual-conversion cable-television (CATV) tuner in a single 48-pin leadless package. Uniquely, the SiW100 chip includes the front-end low-noise amplifier (LNA), automatic gain control (AGC), voltage-controlled oscillator (VCO), and synthesizer without external resonant elements. Silicon Wave expects the price for the SiW100 chip to be less than $10 in 10,000 quantities. Limited quantities will be available in the second quarter of 2000.

"Silicon Wave is the first Bluetooth vendor to enter the single-chip TV tuner fray. Their approach using silicon-on-insulator gives them the advantage of integrating the low-noise amplifier onto the same die as the tuning circuit," says Gerry Kaufhold, principal analyst with Cahners In-Stat Group, a market-research firm in Scottsdale, AZ. "Next-generation set-top boxes will include as many as four individual tuners, so having the LNA inside the tuner gives them a size and cost advantage over other solutions."

Using the experience gained from developing 2.4-GHz transceiver ICs for the Bluetooth market, Silicon Wave is achieving system-level integration while still meeting the most-demanding performance specifications of distortion and low phase noise. The integrated VCO and synthesizer provide significant improvements by eliminating the need for externally 'tuning' the chip during production as well as emitting very-low phase noise. The SiW100 chip performs clear and concise on-chip tuning over the entire cable-system 800-MHz spectrum.

First Trial Of WAP Services In Russia Occurs

MOSCOW, RUSSIA—Mobile Telesystems CJSC (MTS), Russia's leading Global System for Mobile Communications (GSM) wireless phone-service operator, together with Motorola, Inc. have launched the first trials to deliver the Internet to mobile-phone users in Russia.

Using wireless-application-protocol (WAP)-enabled Motorola Timeport™ P7389 phones, mobile users have been accessing Internet information through major Internet portals such as Yahoo!, SkyGo, and AirFlash. The trial relies on the Motorola's open Mobile Internet Exchange™, or MIX™, platform and uses WAP, which allows Internet content to be viewed on a mobile-phone display. The new service will be available commercially to MTS subscribers early this year, once the tests are complete and local content is added.

"We are proud to offer our strategic customer in Russia a leading-edge solution that will allow MTS to offer the Internet to their subscribers yet mitigate seamlessly to GPRS and eventually to a third-generation solution," says Gene O'Rourke, vice president and general manager EMEA for Motorola, Inc.'s Network Solutions Sector. "Not only is the trial the first in this developing mobile-phone and Internet market, but it also makes Russia one of the top 10 countries in the world where such trials have been conducted," says O'Rourke.

Kudos

International Crystal Manufacturing (ICM) has been approved by the North Texas Women's Business Council (NTWBC) as a certified Women Business Enterprise, according to Beth Freeland, ICM president. ICM, a producer of crystal, oscillator, and filter products, offers clients expediated delivery services on standard and customized product orders...Gabriel Electronics, Inc. has announced that its trademark for the StealthWave™ antenna series has been formally registered with the US Patent and Trademark Office. StealthWave antennas use a patented technology developed by Gabriel to achieve optimum electrical performance in broadband microwave antennas...CTS Corp. has been named "Best-Managed Passives Component Company for 1999" by *Electronic Buyers' News*...GTE announced a contribution of $100,000 to Mayor Daley's High School Book Club at a recent press conference at Chicago's Orr High School. Mayor Daley's High School Book Club motivates teens in 75 Chicago-area high schools to become lifelong readers, helping them obtain good jobs.

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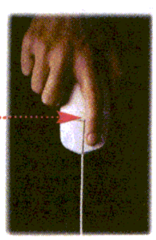

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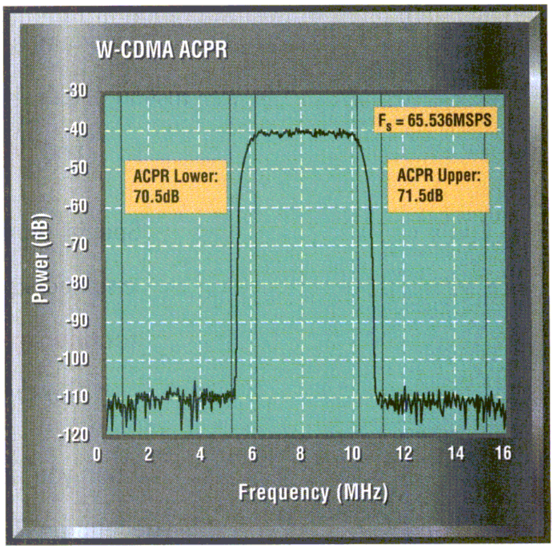

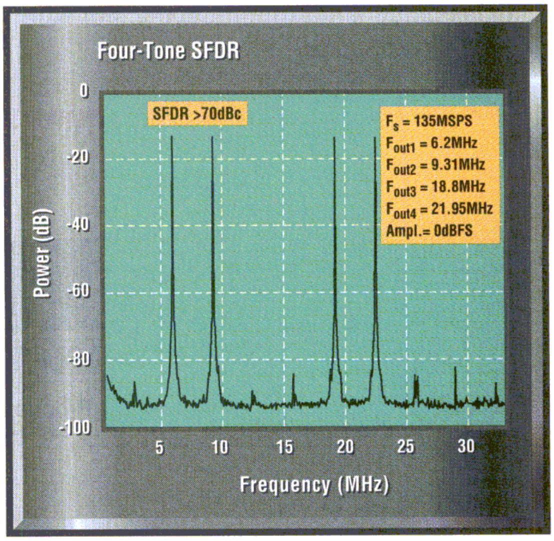

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CIRCLE NO. 254

The baton is being passed to advanced processing technologies as wireless communications moves toward the third generation.

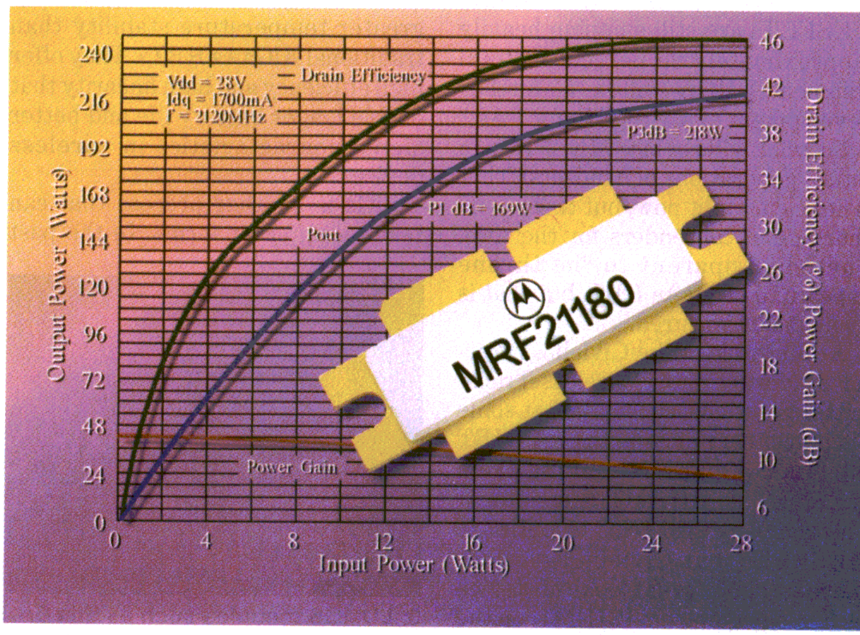
Wireless Semi Technology Heads Into New Territory

GENE HEFTMAN
Senior Editor

ONLY a few years ago, most of the semiconductors found in wireless equipment—in portable phones and the infrastructure—were fabricated from gallium arsenide (GaAs) using metal-semiconductor field-effect transistors (MESFETs). But as the wireless world turns toward third-generation (3G) systems, the underlying structure of its semiconductor technology is also in transition as a broad array of new processes make their way onto the stage. The potential candidates read similar to a bowl of alphabet soup with tongue-twisting names ranging from heterojunction bipolar transistors (HBTs), pseudomorphic high-electron-mobility transistors (PHEMTs), and silicon germanium (SiGe) to laterally diffused metal-oxide semiconductor (LDMOS).

To make things even more confusing, there are processes within processes. For example, HBTs can come in a number of versions including GaAs, SiGe, and a newcomer from ANADIGICS (Warren, NJ) known as indium gallium phosphide (InGaP), which is said to be an improvement over traditional aluminum GaAs (AlGaAs) HBTs. How it will all turn out is anyone's guess because semiconductor technology does not drive the wireless business. Rather, the features that consumers will demand from 3G technology such as data capability, Internet access, video, and others will push device manufacturers to develop processes that can meet the requirements of this rapidly expanding communications medium.

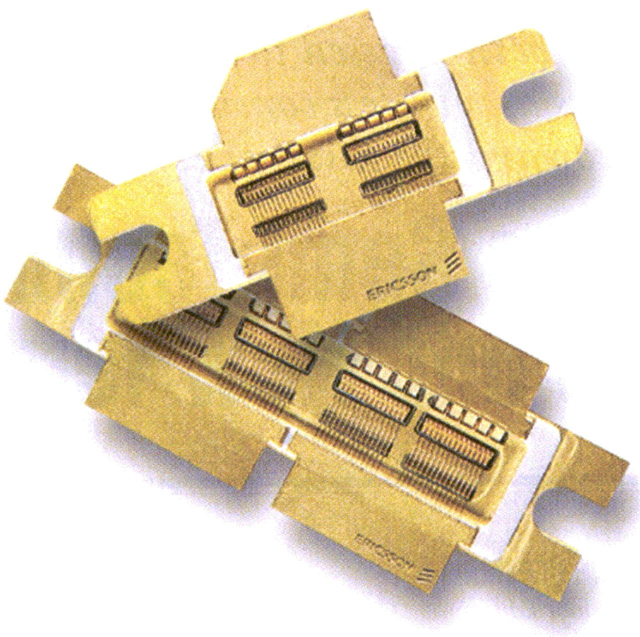
The semiconductor technologies used for portable phones and the wireless infrastructure differ dramatically simply because the power requirements for each type of equipment are miles apart. The situation on the infrastructure side is a bit clearer than it is for handsets. Over the past two years, LDMOS has made major inroads on GaAs and Si bipolar and seems poised to keep its momentum. That assessment is echoed by managers from major power integrated-circuit (IC) makers. Lynelle McKay, Director of Operations, RF Operations, Wireless Infrastructure of Motorola (Phoenix, AZ) says "LDMOS is the dominant base-station power technology, and Motorola is working on its fifth gen-



1. LDMOS transistor power-handling capability for base-station applications gets higher with each new device introduction. Motorola's latest, the MRF21180 is rated for 220 W with a P1dB of 170 W, and a drain efficiency of 42 percent.

eration." And Gary Lopes, Director of Engineering at Ericsson RF Power Products (Morgan Hill, CA), reports that "There is more focus on LDMOS in base-station applications and we have no new bipolar designs in the next year." John Quinn, Vice President at newcomer UltraRF (Sunnyvale, CA), says, "LDMOS is the technology of choice by amplifier manufacturers, but there is competition above 2.3 GHz from GaAs amplifier modules."

On the portable phone side, the picture is more clouded. MESFET power amplifiers (PAs) still rule the roost but, by all accounts, their life is on the endangered list. A recent report on wireless power devices by Allied Business Intelligence, an Oyster Bay, NY market-research firm, predicted that MESFET market share is expected to fall from approximately 40 percent in 1999 to slightly more than 10 percent by 2004. Another tip off comes from Barak Maoz, Vice President of the Wireless Segment at ANADIGICS, who says, "While MESFETs are still shipping heavily (approximately 90 percent of total shipments), there will be no further development of MESFETs for PAs."

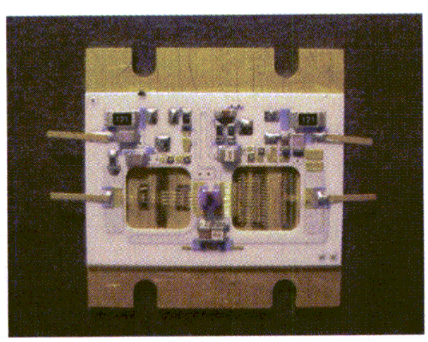


2. Four active devices make up the Ericsson PTF 10120 LDMOS FET, with the four connected in push-pull mode. This enhancement-mode device is rated for 120-W peak equivalent power (PEP).

So what will replace MESFETs in future generations of handsets? That is not clear right now, but there is no shortage of contenders for the title. The heir apparent to the throne appears to be GaAs HBT, but that is by no means ensured.

At Motorola, Mike Civiello, Director of Marketing of Wireless Transmitter Solutions, concedes that some big players are moving to GaAs HBT but his company is moving from MESFET to PHEMT technology. The company still tries to optimize MESFETs, but Civiello says that there is not a great deal of device technology going on. PHEMT, on the other hand, has promise because, he says, "We're moving to E-Mode or Enhancement-mode GaAs FETs. These are single-supply devices that have very low off-state leakage current." ANADIGICS, on the other hand, recently opened a new 6-in. (15.24-cm) wafer fab that will manufacture HBTs based on the InGaP process for PA applications. The company says that InGaP HBTs exhibit higher current gain and greater temperature stability than traditional AlGaAs HBTs. They offer high efficiency and high linearity that enable longer battery life and better signal characteristics in wireless devices.

SiGe is another process that can take on different forms. In the past year, there have been HBT and bipolar versions developed for the handset market. Companies are now working on merging SiGe with their bipolar-complementary-metal-oxide-semiconductor (BiCMOS) and CMOS processes. Lucent Technologies, for example, has developed a 0.25-μm technology that requires only four more masks than its CMOS process and one more mask than its BiCMOS process. The devices are said to have cutoff frequencies greater than 70 GHz. Low-power versions, operating at 450 μA, have cutoff frequencies of more than 60 GHz. Part of the process is the ability to include high-Q inductors with a Q factor greater than 15.



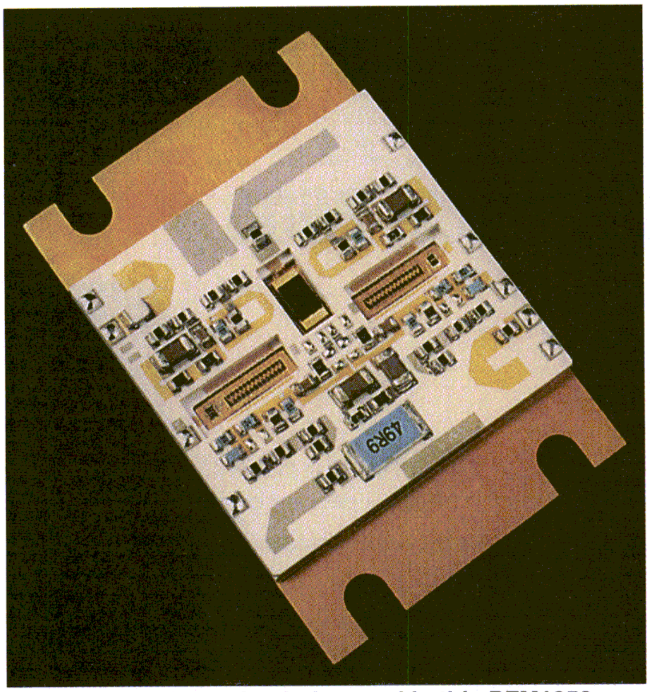
3. LDMOS modules such as this Ericsson device incorporate matched input and output circuitry to ease the design task and conserve circuit-board area.

Similar SiGe development as part of a BiCMOS process is going on at Motorola and ANADIGICS. According to Motorola's Civiello, "SiGe is targeted at small-signal applications in handsets such as synthesizers, transmitter drivers, and receivers. SiGe has difficulties with voltage breakdown, and the things you need to do to make an RF IC for good performance are not the same as to make a good PA." This view is echoed by ANADIGICS' Maoz, who says, "It's not for PAs yet. It has promise, but it's probably a few years out." CMOS has some possibilities, he believes, but not for power amplification. A good RF CMOS process could handle every function in a handset except for the PA and switch.

Within the last few weeks, Texas Instruments (Dallas, TX) announced an SiGe BiCMOS process for the cellular and personal-communications-services (PCS) markets. The +3-VDC, 0.35-μm process has an peak f_T of greater than 50 GHz, with an optimum f_T of 25 GHz. At this level, the process takes less than 20 μA, which will allow it to operate at very low power levels. Products based on the technology are due to appear in the third quarter.

If there is one thing that base-sta-

tion power-system designers can count on, it is that every few months there will be an LDMOS device with greater power-handling capability than previously. The latest example comes from Motorola with its introduction of the MRF21180 and MRF21125 devices (Fig. 1). The 21180 is a 220-W (P3dB) push-pull transistor while the 21125 is a 155-W (P3dB) single-ended version. Both are N-channel enhancement-mode parts for amplification of base-station signal based on the yet-to-be-finalized 3G standard (IMT2000/UMTS), and are designed for either single- or multicarrier-operation. They are intended for operation over the range of 2110 to 2170 MHz. Both come with internal matching circuits at the input (I) and output (O) to simplify the designer's task. A key aspect of these power devices is that they are optimized for handling the high peak-to-average digital modulation signals used in wideband code-division multiple access (WCDMA), a leading candidate for at least some of the wireless systems proposed under IMT2000.



4. Modular construction is featured in this PFM1950, a 50-W, LDMOS PA module from UltraRF. This is the first in a series of 50-Ω-matched modules for PCS base stations.

Most LDMOS transistors must be customized for the particular air-interface standard that they will operate in. "Customers want specs and testing of the part in the application," says Nagaraj Dixit, Applications Engineer in Motorola's RF Operations supporting Asia and Japan. And his colleague, Lynelle McKay, adds, "At the circuit level, changes are made to optimize the perfor-

mance for a CDMA versus the Global System for Mobile Communications (GSM) or Enhanced Data for GSM Evolution (EDGE)." An LDMOS design is dependent on peak/average requirements. In CDMA, this is typically 8 to 10 dB while for time-division multiple access (TDMA), the number is 3 to 4 dB.

Power LDMOS devices are a mainstay at Ericsson RF Power where the power-handling capability has been rising steadily. One of the newer offerings is the PTF10120, an enhancement-mode type rated for a peak equivalent power (PEP) of 120 W (Fig. 2). It is intended for 1.8-to-2.0-GHz communications systems and offers a typical power gain of 10.5 dB when biased class AB with a +28-VDC drain supply. The 10120 is an example of the modularity associated with high-power LDMOS devices. This device contains four active chips assembled in a push-pull configuration. Each chip has 28 cells having 60 0.8-μm gate lengths. The I and O are internally matched to optimize broadband performance. According to Ericsson's Gary Lopes, "Many of our high-powered devices employ not only input matching but output matching as well to make them more user friendly."

I/O matched power devices not only ease the designer's task, they lower the passive component count in the base-station amplifiers, an important consideration in the cost and maintenance of the facility. With on-chip matching components (Fig. 3), the physical size of amplifiers is reduced, allowing them to be installed on poles. Lopes believes that PA technology is heading in the direction of new die structures, higher power, higher gain, and improved efficiency and linearity—two of key parameters of users.

Another economic consideration for the modular approach comes from UltraRF's John Quinn who says, "Modules give you a lower dollar/Watt cost than can be achieved with discrete devices." His company, recently spun off from well-known linear-power-supply manufacturer Spectrian Corp. (Sunnyvale, CA), offers LDMOS and bipolar transistors for the PA market. Modules to be arriving soon contain a number of active devices and use a 10-layer ceramic substrate with internal matching. The device is mounted on a metal carrier, which for low power is copper (Cu), and for higher power, Cu-tungsten (CuW) [Fig. 4]. Products due out soon include a number of band-specific modules made for 1930-to-1990-MHz PCS and 1800-to-1880-MHz digital-communications-systems (DCS) wireless phone markets.

One process with the potential to have a significant impact on the future of power-semiconductor design is Si carbide (SiC). Most of the development work is being carried on by Cree, Inc. (Durham, NC), and it may be some day before it becomes available for widespread use. But the

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potential is enormous because SiC has a much wider energy bandgap than GaAs and Si, can withstand much higher electric fields, and is a better thermal conductor than the two traditional processes. Translating these attributes into electrical characteristics for a power transistor, an SiC device will be able to run at very high temperatures, operate at high voltages even when densely packed, and be able to dissipate large amounts of power while operating at high temperatures. The jury is out about when these devices will be ready for mainstream applications.

In the fast-paced, often confusing world of semiconductor technologies for portable phones, GaAs HBT technology appears to have taken the early lead in the race for new PA designs. But only a few years ago, many questioned the suitability of the process for the extremely price-sensitive handset market. The main sticking points were the price of the starting material and processing costs compared with other technologies.

Barak Maoz of ANADIGICS concedes that starting material costs of HBTs are higher, but he believes that those costs are starting to come down. Second, an HBT device can be made smaller than say, a MESFET, which helps to offset some of the cost disadvantage. Another potential benefit is the use of multichip modules to pack more functions in a single device. The real advantage of HBTs over MESFETs is how the devices are used in a phone. With an HBT, a couple of external components that are rather expensive can be eliminated. One is a converter that turns a positive battery voltage into a negative value to operate a MESFET. A second is a supply-side switch needed to disconnect a MESFET from the battery so it does not draw current while the phone is inactive. HBTs do not need the switch because the base voltage can be pulled down to shut the devices off. Further, the support circuitry for HBTs is less expensive than that for MESFETs. Combining these three elements—elimination of two costly parts and less-expensive support circuitry—brings HBTs close to cost equilibrium with MESFETs.

The latest in HBT technology comes from ANADIGICS in the form of an InGaP HBT intended for PA applications in handsets. The device operates from a +3-VDC supply and is supplied as a modular product with all necessary passive components integrated into the package, including 50-Ω matching.

GaAs HBTs, while valued for their power-handling capability, can also be used in a number of signal-handling applications. RF Micro Devices (Greensboro, NC) just announced the RF2365, a +3-VDC low-noise amplifier (LNA) for a range of wireless applications including DCS GSM, PCS CDMA, PCS TDMA, and 2.4-GHz radio systems. Noise-figure

specifications are 1.6 dB at 1850 MHz and 1.75 dB at 2450 MHz. It is intended for the front end of digital cellular systems operating at DCS/PCS frequencies. Also available from the company is the RF2459 GaAs HBT designed as a downconverter for the 1500-to-2400-MHz band. Applications are as a general-purpose downconverter, in a microcell PCS base station, and in battery-powered equipment. The device is configured as a double-balanced Gilbert-cell mixer with a balanced intermediate-frequency (IF) input.

HBT PAs are in the offing from Raytheon RF Components (Lexington, MA) which has a line of devices for handsets covering the frequency range of 800 to 1900 MHz. The devices are housed in leadless ceramic carriers and include I and O matching circuitry for 50-Ω operation. The RMPA0950-78 is a matched PA module for Advanced Mobile Phone Service (AMPS)/CDMA with a power-added efficiency (PAE) of 60 percent in AMPS mode and 35 percent in CDMA. The RMPA1950-78 is also a matched PA module designed for the PCS band with a PAE of 35 percent.

As noted previously, not all semiconductor manufacturers are taking the HBT path. Motorola has come out with a dual-band TDMA PA using its GaAs PHEMT process. The two amplifier chains in MRFIC1856 are designed to operate in the frequency ranges of 824 to 849 MHz for TDMA and AMPS handsets, and 1850 to 1910 MHz for PCS TDMA phones. A special 20-pin package with a backside metal contact enables good electrical and thermal performance through the solderable metal contact. A second dual-band GaAs PA for GSM/DCS phones was announced recently to handle all the performance requirements of the GSM market. The MRFIC1859 integrates 900- and 1800-MHz PAs on the same chip. It includes an on-chip voltage generator that eliminates the need for a negative supply voltage. The device is designed specifically for use in 1-W DCS-1800 phones, but can deliver a peak power output of 2 W.

Despite the doubts by some industry insiders, Stanford Microdevices (Sunnyvale, CA) has come out with a SiGe chip family aimed at low-power applications in portable equipment. The SGA-6486, for example, is rated for +20-dBm power output over a range of 0 to 2400 MHz. The third-order intercept is +34.8 dBm and f_T is 65 MHz. Small-signal gain is specified at 16.7 dB at 2000 MHz. The device is matched internally to 50 Ω and is designed to run from a +5.2-VDC supply. Typical applications include buffer amplifier for oscillators, final PAs (low-power circuits) and infrared (IR)/IF buffer amplifiers. The company also announced a family of LDMOS transistors for linear applications such as CDMA and WCDMA. The devices are tuned and characterized for communications frequencies from 800 to 2400 MHz. ••

Cable and antenna analyzer upgraded

The B-series upgrade to the Site Master cable and antenna analyzer features several improvements over its A-series predecessors, including superior rejection of interference signals and improved distance-to-fault measurement capabilities. The handheld device is used to test antenna systems for applications such as cellular, data, paging personal communications services (PCS), specialized mobile radio (SMR)/enhanced SMR (ESMR), and microwave communications. Other advancements include higher dynamic range and greater storage capacity for test setups and measurements. It is immune to on-channel interfering signals as strong as +13 dBm, and its directivity is better than 42 dB. The analyzer captures 517 data points and boasts a distance-to-fault measurement capability of more than twice that of its competitors. It also has a high-resolution (640 × 480) full-VGA display, filed-replaceable nickel-metal-hydride (NiMH) battery, and enough memory to store 200 measurements. **Anritsu Co., Microwave Measurement Div., 490 Jarvis Dr., Morgan Hill, CA 95037-2809; (800) 267-4878, Internet: http://www.global.anritsu.com.**

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Software aids carrier power measurements

The PasCom software is designed to be used with the company's Selective Power Meters to help earth-station operators monitor transponder and carrier power parameters. The Windows-compatible software uses an Explorer-style graphical user interface (GUI) and features drag-and-drop spectrum analysis. It can display information in tabular or graphical form, and can automatically calculate outage and availability figures. Measurement data can be archived for subsequent analysis. In addition, the operator can set parameters to accept or reject any new or unrecognized carriers. This allows the system to trigger an alarm automatically when it detects unauthorized carrier usage and, thus, helps combat carrier piracy. **Pascall Electronics Ltd., Westridge Business Park, Cothey Way, Ryde, Isle of Wight PO33 1QT, United Kingdom; +(44) 198 381-7300, FAX: +(44) 198 356-4708, Internet: http://www.intelek.plc.uk.**

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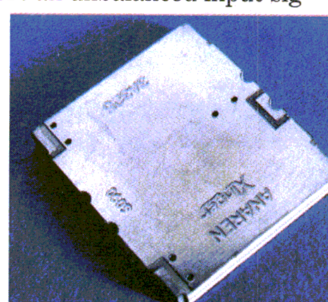
TV broadcast baluns cover 470 to 860 MHz

A family of Xinger television-broadcast baluns is designed for broadcast power amplifiers (PAs) and covers frequencies from 470 to 860 MHz. The baluns convert an unbalanced input signal into a balanced output signal, which is then fed into a push-pull transistor pair. This allows the transistors to work in parallel at high efficiency and with low impedance. The baluns have a maximum insertion loss of 0.35 dB and an input return loss of better than 10 dB. They also feature a maximum amplitude balance of ±0.4 dB and a phase balance of better than ±5 deg. The Xinger baluns are smaller than conventional printed microstrip baluns and are surface mountable. **Anaren Microwave, Inc., 6635 Kirkville Rd., East Syracuse, NY 13057; (800) 411-6596, FAX: (315) 432-9121, Internet: http://www.anaren.com.**



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SiGe power amplifier boasts high efficiency

Model T0930 silicon-germanium (SiGe) power amplifier (PA) boasts a power-added efficiency (PAE) of 50 percent and an efficiency of more than 45 percent over its full supply-voltage range of +1.8 to +3.6 VDC. It is meant to replace the gallium-arsenide (GaAs) devices traditionally used in 900-MHz two-way pagers, personal digital assistants (PDAs), meter-reader transceivers, driver amplifiers, and industrial-scientific-medical (ISM) phones. Nominal output power is 1 W at +2.4 VDC and 2 W at +3.2 VDC. It is capable of delivering continuous-wave (CW) output power to +33 dBm (2 W). **Temic Semiconductor GmbH, P.O.B 3535, D-74025 Heilbronn, Germany; +(49) 713 167-2594, FAX: +(49) 713 167-2423, Internet: http://www.temicsemi.com.**

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More Data Available On Phones

Point-and-click may replace yakkety-yak on the next versions of wireless phones as third-generation (3G) communications turns handsets into data terminals that can surf the web, deliver TV-quality images, even permit stock-market trades and restaurant reservations. Some industry pundits believe that Internet-type data access in the wireless world will surpass conventional voice communications as the primary use of portable phones. Indeed, approximately half of the wireless phones to be produced next year are supposed to be equipped for web surfing. At the extreme edge, there are predictions that the words "phone" and "telephone" may not be used at all for portable communication products since the Internet and e-commerce capture the public's need for information, entertainment, and business transactions.

There is some truth to this scenario, especially in Japan, where a variety of 3G services are available to subscribers. TV and newspaper reports from Japan—although anecdotal—reveal that some users already consider their phones primarily as data terminals rather than telephones. Telephone executives, however, are confident that old-fashioned gab will remain the largest part of the communications business as many wired users switch over to wireless in the next few years. While there is recognition that data usage is inevitable, some of them forecast that as much as 50 percent of all voice traffic will be over wireless in the next five years compared with 6 to 7 percent two years ago.

It is not only the wireless phone that is under siege. Personal computers (PCs), which now are the chief way to access the Internet, could feel the heat from wireless devices known as web pads and Internet appliances that act as portable computers but without a clumsy keyboard. These devices will permit a user to shop, transact personal business, and download information while on the move. So the future of personal computing will be tied to mobile access to the Internet and other information sources rather than the user sitting at a desk at home or at work. This is called the anywhere, anytime paradigm, with major equipment manufacturers, such as Ericsson, Nokia, Lucent Technologies, and others, developing products that tout mobile data capabilities rather than the traditional voice features.

Manufacturers of PCs and wireless phones must realize that the Internet, and not hardware types, will come to dominate personal communications. And they will have to accommodate the desires of users with whatever products provide the optimum solution for information on the go. ••

Contracts

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

Electro-Radiation, Inc. (ERI)—Announced the receipt of a US Army "Fast Track" Small Business Innovative Research Phase II contract valued at approximately $1.5 million. The 24-month program will adapt portions of ERI's commercial-off-the-shelf (COTS) and proven technology to specific RAH-66 Comanche helicopter requirements.

SpectraPoint Wireless LLC and REMEC, Inc.—Announced a multiyear Strategic Supply Agreement where REMEC will produce broadband wireless RF transport equipment for SpectraPoint Wireless for use in local-multipoint-distribution-services (LMDS) systems. The agreement provides for nominally $150 million in revenue and is the vehicle for the execution of Purchase Orders for SpectraPoint to purchase RF electronic equipment from REMEC.

Glenayre Technologies, Inc.—Has received a multimillion-dollar, two-year agreement to provide the Glenayre enhanced-services solution to OnePoint Communications, a provider of communications services to residents of US apartment communities.

Microwave & Video Systems, Inc. (MVS)—Has received the first phase of a government contract for microwave integrated assemblies, switch/filter banks, and positive-intrinsic-negative (PIN)-diode switches. Contract enhancements and options would add more than $2 million to the basic contract, which is scheduled to run for three years.

Spire Corp.—Has been awarded a $600,000 contract from NASA's John H. Glenn Research Center in Cleveland, OH to develop advanced, high-speed photodetectors that will be needed for next-generation fiber-optic communications system.

Fresh Starts

Stanford Microdevices—Announced a new organizational structure centered around the strategic business units focused on specific markets that serve the wireless and wired infrastructure equipment suppliers and, in addition, the formation of a business unit with a strong focus on the market for broadband, high-linearity standard products.

Burle Industries, Inc. and Galileo Corp.—Completed the purchase by Burle Industries of Galileo's Scientific Detector and Spectroscopy Products (SDP) business located in Sturbridge, MA. The SDP business includes the manufacture of Channeltron@single-channel detectors, microchannel plate-based detectors, flexible fiber-optic light guides, glass-coated wire, and remote spectroscopy products.

Gabriel Electronics, Inc. and K-C Marketers, Inc.—Have entered into an agreement for K-C Marketers to represent Gabriel's wireless broadband point-to-point and point-to-multipoint antenna products, transmission-line systems, and pressurization equipment.

Harris Corp.—Has been selected by Batelco, a communications provider in the Arabian Gulf region, to provide an advanced network-management system to support Batelco's expanding network and customer base. Harris will provide consolidated fault and configuration management for Batelco's entire telecom infrastructure, service provisioning, and performance management supplied by ADC Metrica, one of Harris' technology partners.

Silicon Wave, Inc.—Has formed a partnership and signed a development contract with Tokyo-based Taiyo Yuden Co. Ltd., the world's fastest-growing module maker. Under the agreement, Taiyo Yuden will incorporate Silicon Wave RF integrated circuits (RF ICs) into modules for use in communications and computing products.

TRW and RF Micro Devices—Jointly announced the expansion of their strategic relationship through several agreements that license RF Micro Devices to use TRW's patented gallium-arsenide heterojunction-bipolar-transistor (GaAs HBT) technology to manufacture products for commercial coaxial and other non-fiber wire applications, including the broadband wired market.

Ansoft Corp.—Has completed a software-licensing agreement with Thomson-CSF. Under the agreement, Ansoft will be the worldwide supplier of electromagnetic (EM)-analysis software to Thomson-CSF.

Andrew Corp.—Has acquired Conifer Corp., a privately owned company that designs and manufactures multichannel-multipoint-distribution-services (MMDS) subscriber products, wireless local-area-network (WLAN) equipment, and direct-broadcast-satellite (DBS) accessories.

Southwest Microwave, Inc., Microwave Products Division—Has appointed GLX, Inc. to provide coverage in Colorado, Utah, Wyoming, Montana, and Idaho. Also, the company has appointed McBride Scientific Sales, Inc. to provide coverage in Texas, Louisiana, Oklahoma, and Arkansas for Southwest Microwave, Inc.'s line of performance microwave connectors and adapters.

Qualcomm, Inc.—Announced that Sanyo Electric Co. Ltd. has selected Qualcomm CDMA Technologies' sixth-generation MSM3100™ chip set and system software for its next-generation code-division-multiple-access (CDMA) handsets.

CTS Corp.—Revealed that its February 26, 1999 acquisition of Motorola's Components Products Division, renamed CTS Wireless Components, has created significant growth opportunities for CTS in the communications equipment market. CTS will expand its product line to include RF integrated modules. RF integrated modules will best benefit highly complex and miniaturized radios including multiband and multimode cellular phones.

ANADIGICS—Has announced that it will open its third remote design center and that the facility's start-up leadership is in place. The new remote design center will be located in Thames Valley, United Kingdom.

Hittite Microwave Corp.—Norman G. Hildreth Jr. to director of marketing and business development; formerly director of sales. Also, Stephen G. Daly to director of sales; formerly principal sales and application engineer.

Narda Microwave-East—Nicholas Bainlardi to regional sales manager; formerly engineering manager at Tern Technology, Inc.


BAINLARDI


CAMMARATA

TriPoint Global Communications, Inc., CSA Wireless—Vincent D. Cammarata to chief technical officer; formerly regional RF engineering manager for Sprint PCS.

Anritsu Corp.—Bill Lovelace to general manager of the International Sales Management Center (ISMC); formerly president of the North American Region Operation (NARO) and assistant general manager of ISMC. Also, Phil Bowen to president of NARO; formerly general manager of NARO.

@Road, Inc.—Shirish Puranik to vice president of software development; formerly director of product development for the data-base group at Oracle Corp. Also, Mike Walker to vice president of engineering; formerly vice president of engineering for Silicon Wireless. In addition, Tom Allen to vice president of information technology; formerly vice president of network engineering at Visa International. And, Dave Manovich to vice president of sales; formerly provided strategic-management consulting at Union Atlantic.

Millitech Corp.—Mervyn N. FitzGerald to senior vice president of operations; formerly vice president of operations and customer service for the Broadband Wireless Access division of Nortel Networks.

Leitch Technology Corp.—John A. MacDonald to president and chief executive officer; formerly president and chief operating officer of Bell Canada. Also, Thomas M. Jordan to senior vice president of strategic relations; formerly vice president of sales and marketing for the US.

Philips Semiconductors—Thierry Laurent to managing director of the Telecom Terminals Business Unit; formerly senior vice president and general manager of the Communications Product Group at VLSI Technology.

Cascade Microtech, Inc.—Craig M. Swanson to chief financial officer; formerly vice president of finance for Protocol Systems.

Oak Frequency Control Group (OFCG)—John A. Kohler to president of OFCG North America; formerly group director of Global RF/Coax and Antenna at AMP, Inc.

Ohmite Manufacturing Co.—Al Kirwan to national distribution manager; formerly worked for Ohmite's Commercial Products Group (CPG).

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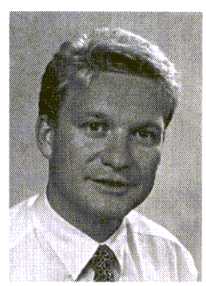
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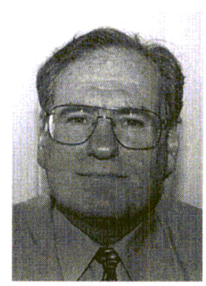
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Advanced Microtek Ltd.—Roger Tucker to managing director; formerly employed in the radar division with Thorn-EMI Electronics.

IPC—Denny McGuirk to president; formerly worked as executive director at the National Fluid Power Association.

Electro-Radiation, Inc. (ERI)—Anthony J. Rubinich to vice president and program director; formerly director of program management at Curtiss-Wright Flight Systems, Inc. (CWFS).

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University of Wisconsin-Madison
Madison, WI 53706
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e-mail: Custserv@epd.engr.wisc.edu
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Analyze ADC errors through a unified model

Models of analog-to-digital converters (ADCs) can aid in allowing a device's error characteristics to be analyzed and corrected more easily than without models. Most of the ADC models developed to date are related to the conversion method, the topology, or a specific type of error. Now, Pasquale Arpaia of the Dipartimento di Ingegneria Elettrica, Università di Napoli Federico II, 80125 Napoli, Italy, Pasquale Daponte of the Facoltà di Ingegneria, Università del Sannino at Benevento, Italy, and Linus Michaeli of the Department of Electronics and Multimedial Telecommunications, Technical University of Kosice at Kosice (Slovak Republic) propose a unified behavioral model for the three most popular ADC architectures—integrating, successive approximation, and flash. The effects of the main error sources are analyzed in terms of integral nonlinearity (INL) and differential nonlinearity (DNL). Although further investigations are needed, partial success has been achieved with a unified model of INL. See "Influence of the Architecture on ADC Error Modeling," *IEEE Transactions on Instrumentation and Measurement*, Vol. 48, No. 5, October 1999, p. 944.

Multipath propagation delivers unsuspected benefits

Multipath effects in wireless communications channels usually have a negative connotation when signal fading is caused by the destructive addition of the multipath components. However, there is a constructive aspect to multipath according to the research of Apostolis K. Salkintzis and P. Takis Mathiopoulos of the Department of Electrical and Computer Engineering at the University of British Columbia (Vancouver, BC, Canada). They show that the constructive mechanism is more dominant than the destructive—its average amplitude is greater than that of the destructive and its occurrence has greater probability—which results in a higher power level at the receiver. This idea can be exploited for designing receivers with greater performance. Formulas are derived showing why constructive addition is greater than destructive. See "On the Combining of Multipath Signals in Narrowband Rayleigh Fading Channels," *IEEE Transactions on Broadcasting*, Vol. 45, No. 2, June 1999, p. 192.

Broadbanding microstrip antennas opens applications

Microstrip antennas are attractive because of their low profile, light weight, small volume, and simple architecture. Their narrow bandwidth, however, restricts applications in which they can be used. Research done by Zhang-Fa Liu, Pang-Shyan Kooi *et al.* of the Communications and Microwave Division, Department of Electrical Engineering, National University of Singapore (Singapore) has resulted in a microstrip antenna architecture having an impedance bandwidth of up to 25.7 percent, compared with about 5 percent for a conventional microstrip antenna. The authors provide the design procedures together with the nonlinear design equations needed to realize a two-layer wideband microstrip patch antenna. To verify the procedure and equations, an L-band antenna was constructed that displayed a 25.7-percent bandwidth. The equations can be revised and extended to the design of other types of wideband multilayered structures. See "A Method for Designing Broad-Band Microstrip Antennas in Multilayered Planar Structures," *IEEE Transactions on Antennas and Propagation*, Vol. 47, No. 9, September 1999, p. 1416.

InP HEMTs lag behind their GaAs counterparts

Indium-phosphide (InP) high-electron mobility transistors (HEMTs) could be an alternative to gallium-arsenide (GaAs) pseudomorphic HEMTs (PHEMTs) at millimeter-wave frequencies if not for their lower power output compared with GaAs. The fact is, InP HEMTs exhibit greater power-added efficiency (PAE) and higher gain per stage, but their lower output power is too much of a drawback. The reason for InP's poor power showing, according to research performed by J.A. del Alamo and M.H. Somerville of the Massachusetts Institute of Technology (Cambridge, MA), is an off-state breakdown voltage (BV_{off}) about +2 to +3 VDC below that of GaAs and significantly worse on-state breakdown voltage (BV_{on}) compared with GaAs. The authors contend that careful management of impact ionization in the channel of InP devices could improve both breakdown voltages and that the power output could even exceed that of GaAs. This would give InP a significant advantage in applications in which a large number of transmitters are integrated together in a small volume. See "Breakdown in Millimeter-Wave Power InP HEMT's: a Comparison with GaAs PHEMTs," *IEEE Journal of Solid-State Circuits*, Vol. 34, No. 9, September 1999, p. 1204.

Balance Trade-Offs In GaAs FET LNAs

Through careful design techniques, it is possible to achieve very-low noise figures in LNAs without sacrificing VSWR performance.

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LOW-NOISE amplifiers (LNAs) play dominant roles in RF/microwave receiver systems. They determine the system sensitivity, which, in turn, dictates the complexity and ultimate system cost. Increasing demand for improved communications system performance is placing more stringent requirements on LNAs. With the advancement in gallium-arsenide (GaAs) semiconductor technology, state-of-the-art pseudomorphic-high-electron-mobility-transistor (PHEMT) devices are available that can achieve 0.5 dB or better noise figure at 900 MHz (the cellular band). These devices are available in chip form or are housed in miniature SOT packages. LNAs made of these devices require optimum design techniques and careful considerations of performance trade-offs in order to meet the systems requirements. This article briefly presents a design technique used for several LNAs from M/A-COM (Lowell, MA): the model AM40-0032, an ultra-low-noise PHEMT amplifier for use from 1710 to 1910 MHz, and the soon-to-be-released model AM40-0032, a low-cost version of the AM40-0023.

GaAs PHEMT devices are commonly used in the microwave frequencies where silicon (Si)-bipolar transistors cannot provide the required performance in gain and noise. However, 0.5-μm PHEMTs work very well at frequencies of 2 GHz and below. For optimum noise performance, the design should conjugately match the forward reflection coefficient (S_{11}) of the field-effect transistor (FET) to the source or generator's reflection coefficient, Γ_{opt}. Γ_{opt} is the generator's reflection coefficient under a condition of minimum FET noise figure. Equation 1 shows the definition of Γ_{opt}. The noise parameter Y_{opt} is obtained by measuring the FET in a 50-Ω system with very low-loss precision impedance tuners at the input and output ports. The FET is biased for minimum noise, the output tuner is adjusted for maximum gain, and the input tuner is set for minimum noise figure. The input tuner is then measured for Y_{opt} and Γ_{opt} is calculated from the result of that measurement.

Table 1: Typical parameters shown for the NE33200 (at +25°C, Vds=+2 VDC, and Ids= 10mA)

Frequency (GHz)	Nfopt (dB)	Gain (dB)	Γ_{opt} MAG	Γ_{opt} ANG	Rn/50
1.0	0.29	21.3	0.82	8	0.39
2.0	0.31	18.3	0.81	17	0.36

The manufacturer of the transistor device usually performs this measurement.

To achieve maximum power gain, the load impedance is then conjugately matched to the FET output (in this case, the drain reverse reflection coefficient, S_{22}'). S_{22}' is the output reflection coefficient of the PHEMT including the input-matching network at the gate. This optimum noise-figure match at frequencies of 2 GHz or lower generally produces a high input VSWR. The ideal situation would be to have the Γ_{opt} be equal or close to the complex conjugate of S_{11}. Under that condition, minimum noise figure and VSWR can be achieved. However, that condition typically occurs above 2 GHz and tends to diverge at lower frequencies where S_{11} approaches 1.

Table 2: S_{11} values shown for magnitude and phase with source inductance to ground

Frequency (GHz)	S_{11}	
	Magnitude	Angle
1.0	0.87	−7.8
2.0	0.85	−15.5

$$\Gamma_{opt} = (Y_o - Y_{opt})/(Y_o + Y_{opt}) \quad (1)$$

where:

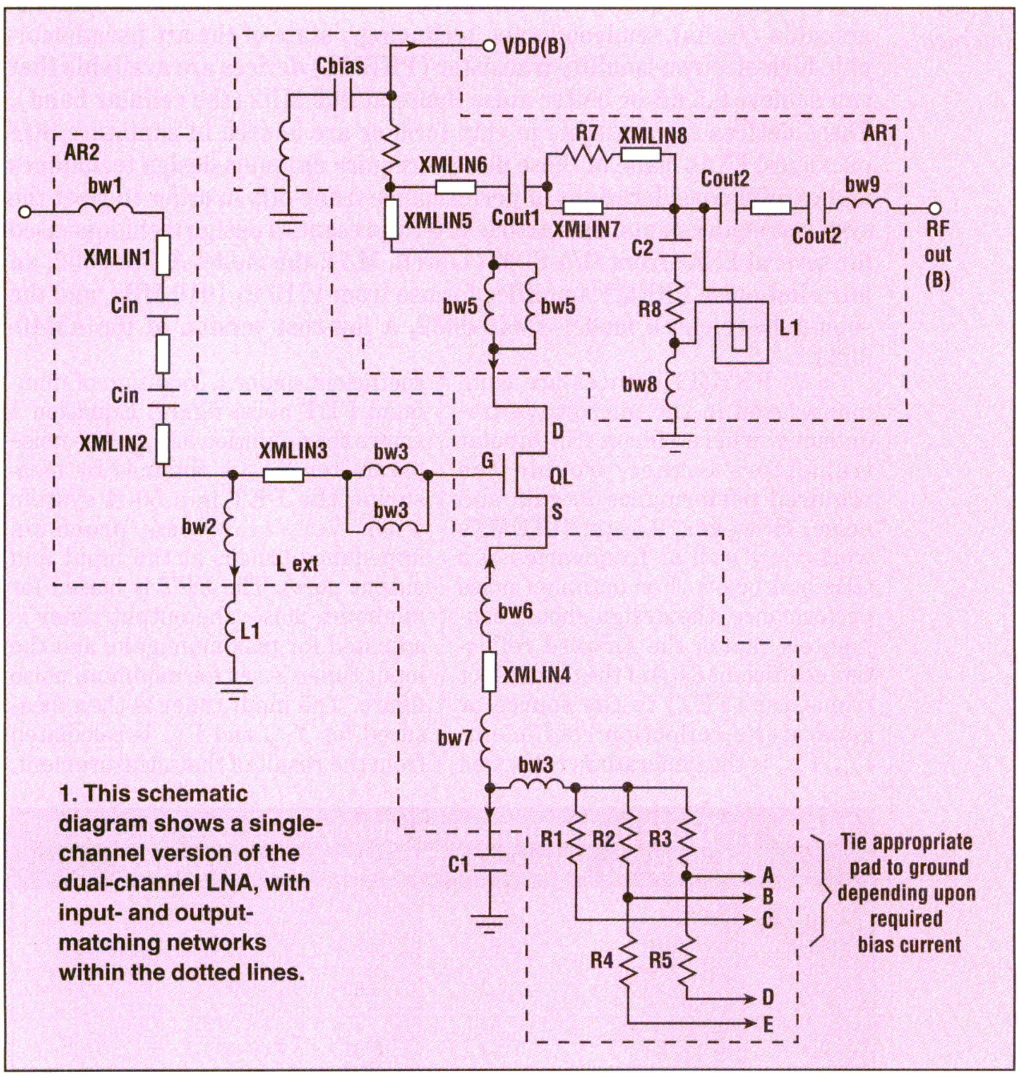
Ω_{opt} = the source reflection coefficient with device biased for optimum noise figure,

Y_o = the source admittance. In a 50-Ω system, this value would be 1/50, and

Y_{opt} = the source admittance with the device biased for optimum noise figure.

To solve the high-VSWR problem, an inductive negative feedback can be incorporated from the FET s source to ground. The advantages of feedback are unconditional stability and convergence of the Γ_{opt} and S_{11} conjugate. The drawback is a decrease in gain, but higher inherent gain of FET devices at lower frequencies makes the sacrifice insignificant.

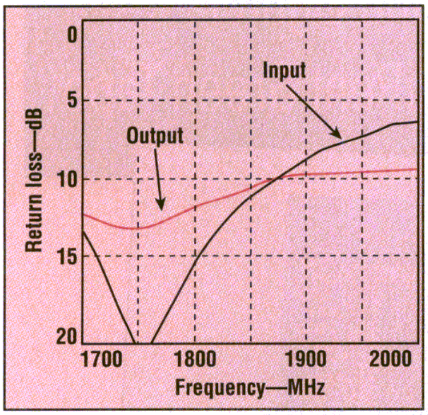
Figure 1 shows a single-channel schematic representation of the LNA. The dotted-line blocks are the input- and output-matching networks fabricated on glass monolithic microwave integrated circuits (MMICs). The inductive feedback for the amplifier is made up of bond wires bw6 and bw7 (bond wire from the source to the glass MMIC, and the bond wire to capacitor C1 and the RF ground), and microstrip line XMLIN4. The device is a super low-noise PHEMT chip from NEC (model NE33200). Table 1 shows the device parameters at 1 and 2 GHz under the selected bias conditions.



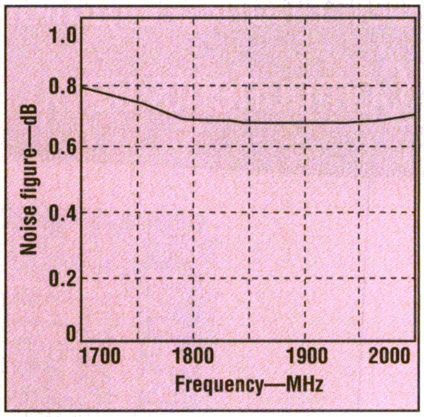
1. This schematic diagram shows a single-channel version of the dual-channel LNA, with input- and output-matching networks within the dotted lines.

Using EEsof Libra computer-aided-engineering (CAE) software from Agilent Technologies (Santa Rosa, CA), the device is mapped for S_{11}' with source inductance to ground. The results are shown in Table 2.

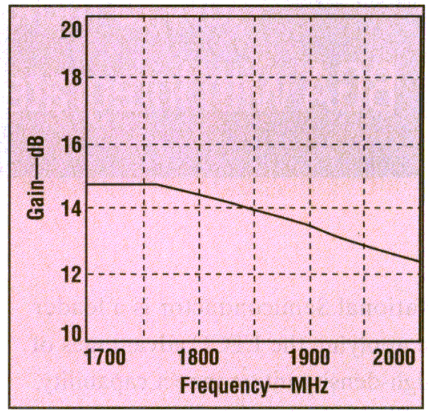
By inserting the inductance from source to ground on the device, the magnitude of S_{11}' changes from 1.00 and 0.99 to 0.87 and 0.85. These values are close to the magnitude of Γ_{opt} at 1 and 2 GHz. The angle of S_{11}' also remains close to Γ_{opt}. Under these conditions, conjugately matching S_{11}' of the device to Γ_{opt} produces minimum noise figure and VSWR.



2. The input and output return loss are shown as functions of frequency.



3. The noise figure of the LNA is shown as a function of frequency.



4. The gain for the LNA was measured as a function of frequency.

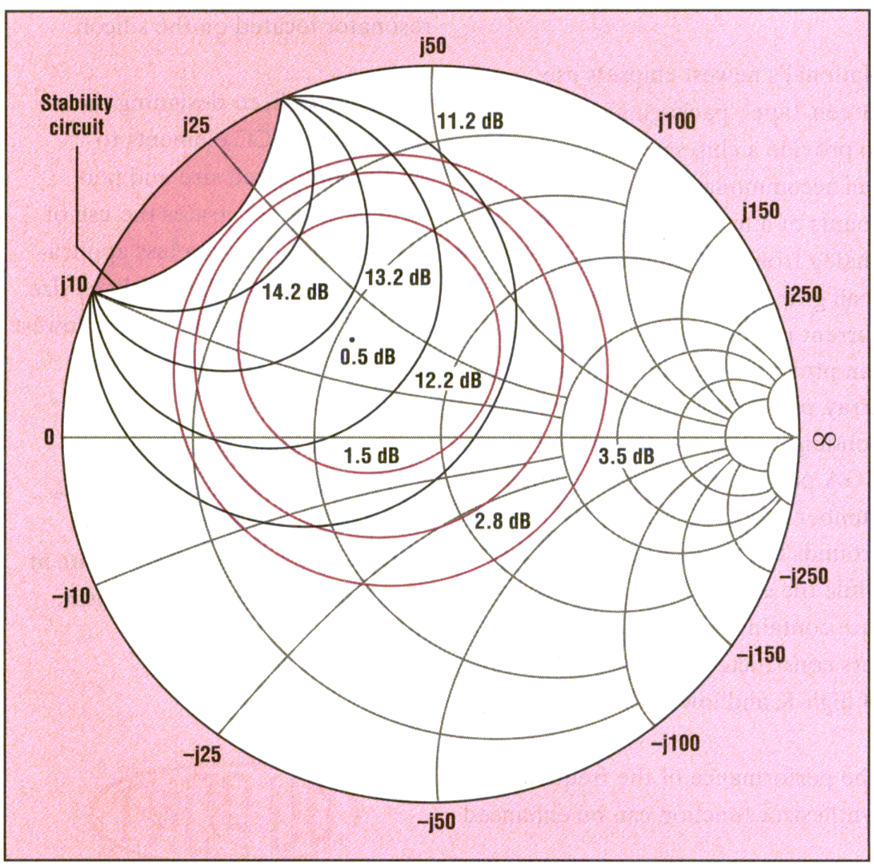
To obtain maximum unconditional power gain, the 50-Ω load impedance is matched to the conjugate of S_{22}'. The final measured results of the design are shown in Figs. 2 to 4. The midband noise figure is less than 0.7 dB, while the VSWR and gain are less than 2.0:1 and 14 dB, respectively.

The FET is powered with a self-biasing scheme that requires only a single positive drain voltage. The drain voltage (V_d) is controlled by a resistor in series with the external bias voltage, while the drain-to-source current (I_{ds}) is controlled by selectable source resistors. This simple biasing scheme is compromised by having to include multiple resistors to accommodate FET lot-to-lot performance variations, in order to keep I_{ds} at an optimum operating level. The variation in the pinchoff voltage (V_p) and saturated drain current (I_{dss}) in different lots of FETs requires a different value of resistance to achieve optimum I_{ds}. Equation 2 shows the relationship of these parameters and how to arrive at the proper resistor value:

$$R_s = [V_p \times [1-(I_d / I_{dss})^{0.5}] / I_d \quad (2)$$

where:

R_s = the source resistor,
V_p = the pinchoff voltage,
I_d = the operating drain current, and
I_{dss} = the saturated drain current.



5. These constant gain and noise-figure circles were measured with inductive feedback for the NE33200 low-noise transistor.

OPTIMIZING LNAs

This design example shows the compromises taken to achieve the necessary performance. In general, LNA designs at all frequencies can be optimized by using device parameters with circuit feedback to generate constant gain and noise-figure circles on the impedance plane.[1,2] The derivations of the centers and radii of these circles are shown in Ref. 2. Figure 5 shows the constant gain and noise-figure circles of the NE33200 with inductive feedback from the FET's source to ground. By selecting the intersections of these circles closer to the unity circle, optimum noise figure, gain, and VSWR can be achieved. ●●

References

1. K. Kurokawa, *IEEE Transactions on Microwave Theory & Techniques*, March 1965.
2. Fukui, H., "Available Power Gain, Noise Figure and Noise Measure of Two Ports and Their Graphical Representation," *IEEE Transactions On Communications Technology*, Vol. CT-13, No. 2, pp. 137-142.

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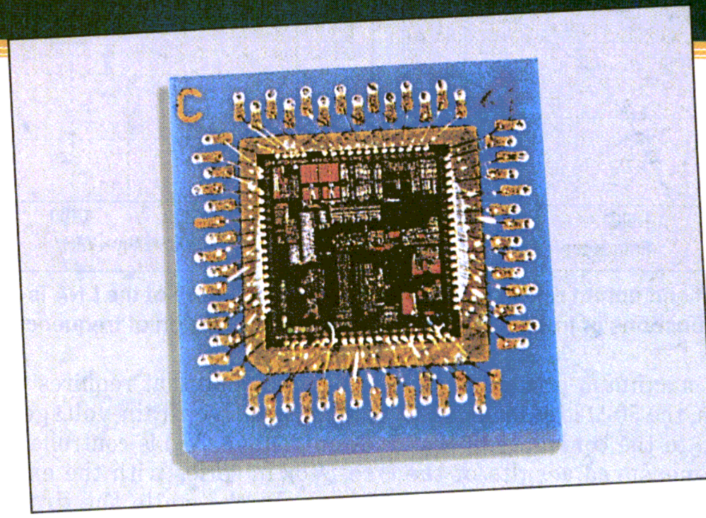

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Achieve Phasor Rotation With Analog Switches

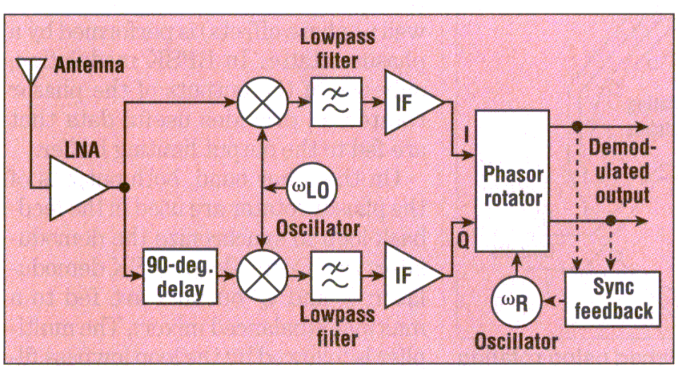
An unusual approach to phasor rotation can decrease complexity and reduce costs in analog and digital radios.

Matjaz Vidmar
Sergeja Masere 21, 5000 Nova Gorcia, Slovenia; FAX: (386) 61 176-8424, e-mail: s53mv@uni-mb.si.

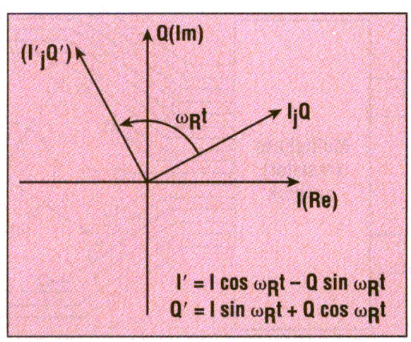
PHASOR rotation is a rather common signal-processing step in radio transmitters and receivers. But the traditional implementation of phasor rotation using four multiplications or four balanced mixers can be complex and costly. This article presents a simpler and less-expensive approach using rotating electronic switches. It explores the application of the rotating-switch approach in analog circuits such as a Weaver single-sideband (SSB) radio, and digital circuits such as a Costas-loop demodulator. A practical design of a Costas-loop, binary-phase-shift-keying (BPSK) demodulator using rotating analog switches—along with its measured performance—is presented at the end of the article.

Zero-intermediate-frequency (zero-IF) receivers and direct-modulation transmitters are becoming increasingly popular. Figure 1 shows the basic zero-IF receiver design. Since there is no RF filtering, two mixers operating in quadrature and two subsequent IF chains are required to process the in-phase (I) and quadrature (Q) signals. Although the I and Q IF signals are already in the same frequency range as the baseband modulation signal, an additional signal-processing step is usually required to obtain the baseband signal. For example, in a Weaver (analog voice) SSB receiver, the LO operates in the center of the RF signal passband, and the I and Q IF signals have to be shifted in frequency by approximately 1.4 kHz and combined to recover the original voice signal. In a digital-data receiver, the frequency error of the LO has to be removed. The LO itself generally cannot be synchronized directly to the incoming signal because it is very difficult to design a voltage-controlled oscillator (VCO) that has low phase noise, high stability, and fast tuning response.

In any case, the additional signal-processing step required to obtain the baseband signal is best described as phasor rotation, which is shown in Fig. 2. The I and Q IF signals represent the two components of a phasor. Phasor rotation can be performed in positive and negative directions. And since a frequency shift corresponds to phasor rotation,

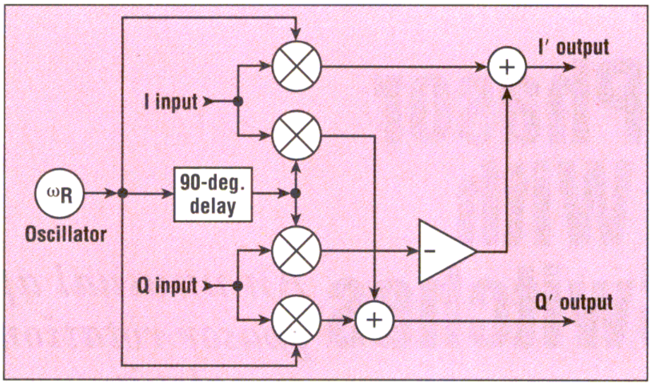


1. This block diagram shows a zero-IF receiver design.



2. This graph shows a basic phasor-rotation operation.

forward or backward phasor rotation corresponds to positive or negative frequency shifts. The output of a phasor rotator is itself a phasor, represented by the signals I' and Q'. In the case of a digital-data receiver, the phasor is rotated to compensate for the frequency offsets of transmitter and receiver. The I' and Q' signals directly represent components of quadrature-amplitude-modulation (QAM) or quadrature-phase-shift-keying (QPSK) modulation.



3. This diagram shows phasor rotation using four multipliers (mixers).

Communication engineers usually translate the phasor-rotation formulas from Fig. 2 directly into the circuit diagram shown on Fig. 3. But this requires either precious central-processing-unit (CPU) time (to perform four multiplications) or the expense of additional analog hardware (four mixers). A simpler, less-expensive solution would be highly desirable. Fortunately, power engineers have been working with phasor rotation since the invention of multiphase AC power more than a century ago, so taking a cue from their experience might prove useful.

When two quadrature signals (I and Q) are available, an arbitrary phasor can be generated as a linear combination of the I and Q signals. A multiphase system with a large number of phases can be generated by a simple linear network of resistors driven by a four-phase system that includes both polarities of the I and Q signals. Selecting the desired phase provides an arbitrary phase shift. The phasors are rotated by a simple rotating switch, as shown in Fig. 4. Of course, two switches with their sliding contacts offset by 90 deg. are required to obtain the I' and Q' output signals.

In most zero-IF receivers, phasor rotation does not need to be very accurate. A phasor rotator with 32, 16, or even 8 steps might be sufficient. Similar constraints also apply to phasor rotation in direct-modulation transmitters. A limited number of phases makes the multiphase network and rotating switches a practical solution. Of course, no mechanical rotating parts are used inside radio receivers and transmitters. Instead, they use simple and efficient complementary-metal-oxide-semiconductor (CMOS) analog switches driven by digital counters.

SSB DEMODULATOR

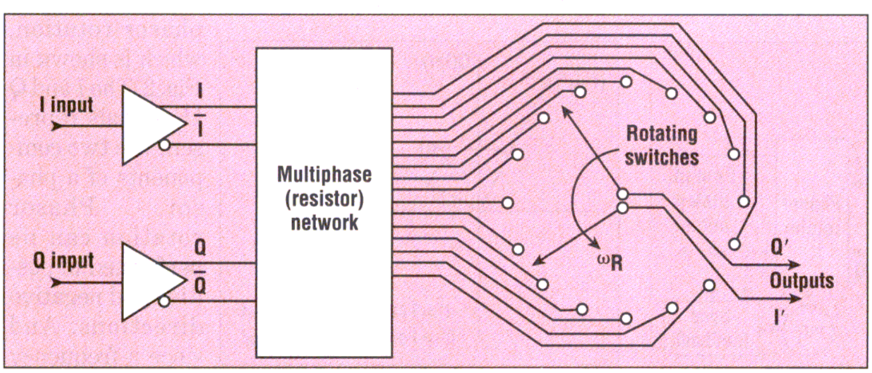
The simplest practical application of the theory mentioned is a Weaver (analog voice) SSB demodulator. A rotating-switch Weaver SSB demodulator, along with the corresponding I/Q IF chain and audio amplifier, are shown in Fig. 5. In a Weaver SSB receiver, both I and Q IF channels are limited to approximately 1.2 kHz. In a practical receiver, the two I and Q IF channels require a common automatic-gain control (AGC). A Weaver SSB demodulator converts and combines the two I and Q channels of 1.2-kHz bandwidth each to yield the original 200-to-2600-Hz voice channel.

In the case of a Weaver SSB demodulator, a single output of the phasor rotator is used, requiring a single rotating switch. Unwanted conversion products are easy to filter away using an eight-phase system. Thus, the entire demodulator only requires a network of 12 resistors and an eight-position switch. The unwanted conversion products are filtered away with a simple 3-kHz, lowpass filter in front of the audio amplifier. A simple CMOS analog multiplexer (CD4051) can be used as an audio rotating switch. The direction of the switch rotation selects the demodulation of either the upper or lower sideband. It makes sense to drive the CD4051 switch with a bidirectional counter from the same logic family, such as the CD4029. The UP/DOWN input of the bidirectional counter then becomes the upper sideband/lower sideband (USB/LSB) control. The clock input requires eight times the switch-rotation frequency of 1.4 kHz, or 11.2 kHz.

BPSK DEMODULATOR

The Costas loop is a very popular BPSK demodulator (Fig. 6). In a practical zero-IF BPSK receiver, a common AGC is required for I and Q IF channels. Of course, the IF bandwidth should match the actual data rate. The signal demodulation (correction for the transmitter and receiver frequency as well as phase offsets) is performed by a phasor rotator. In BPSK modulation, only one of the outputs of the phasor rotator, I', provides useful data that are fed to the output limiting buffer.

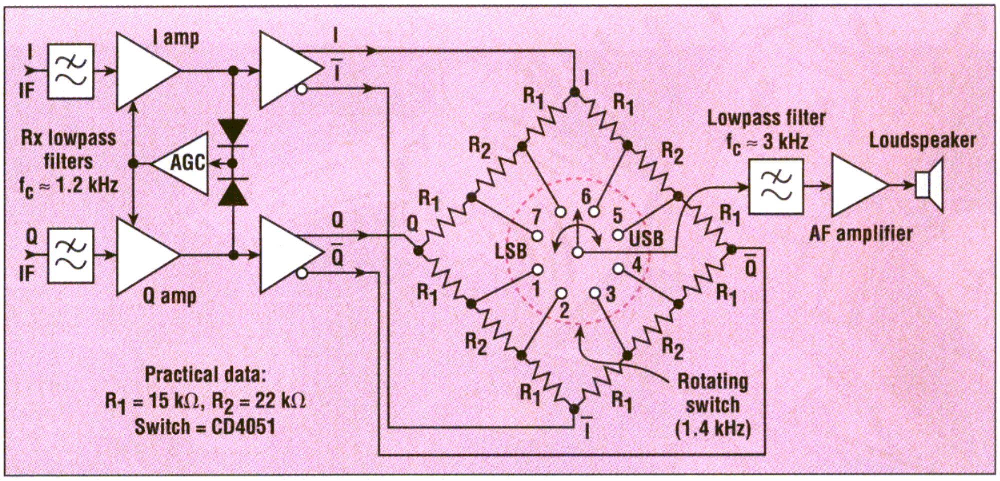
On the other hand, both outputs of the phasor rotator are used in the feedback loop to synchronize the demodulator. In a Costas-loop BPSK demodulator, I' and Q' outputs are fed to a multiplier (balanced mixer). The multiplier is followed by the loop lowpass filter. The output of this filter steers the



4. This diagram shows how phasor rotation can be accomplished using rotating switches.

VCO frequency to compensate for the transmitter and receiver frequency and phase offsets. Due to the analog multiplier, the loop gain is proportional to the square of the input signal. An efficient AGC action is therefore required for proper demodulation.

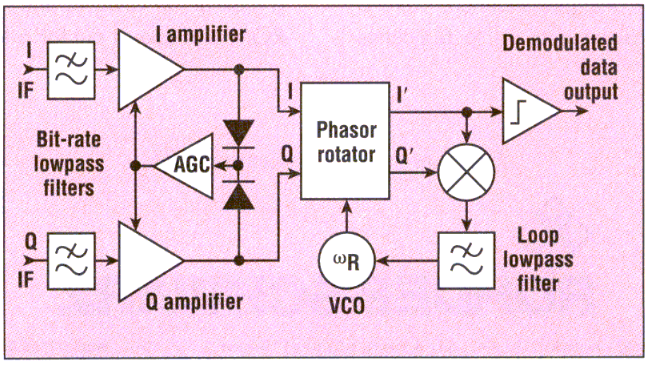
A Costas-loop BPSK demodulator also requires a more accurate phasor rotation than a Weaver SSB demodulator. Since phase errors are converted directly into a decrease of the signal-to-noise ratio (SNR), a phasor rotator with 16 or 32 steps is required in a Costas-loop BPSK demodulator. Increasing the number of steps above 32 does not bring any significant advantage (maximum 0.17 dB). The phasor rotator for a BPSK demodulator can therefore be built with a reasonable number of resistors and analog switches.



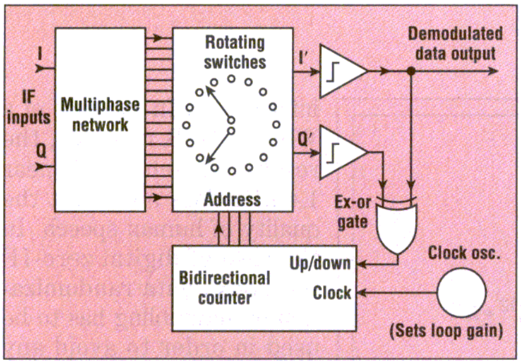
5. A diagram of a Weaver SSB (voice) demodulator can be seen.

The described Costas-loop demodulator still contains a few critical analog components, such as the AGC, the multiplier, and the VCO. It is not easy to build an analog VCO that provides positive and negative frequencies to the phasor rotator, since the polarity of the transmitter and receiver frequency offsets may be arbitrary. A simple all-digital solution to the problems previously mentioned is shown in Fig. 7. First, I' and Q' outputs of the phasor rotator are limited to logical levels. Although some noise is introduced in this way, the operation of the AGC is much less critical. Second, the analog multiplier can be replaced by a simple exclusive-OR gate. Finally, a completely digital solution is used for the VCO. The digital VCO is built around a bidirectional counter. The VCO control input is actually the up/down control of the counter, while the counter itself is clocked with a constant frequency. The VCO frequency depends on the duty cycle of the signal present at the up/down input. If the duty cycle is larger than 50 percent, the counter will count mainly forward, thus producing a positive frequency. If the duty cycle is less than 50 percent, the counter will count mainly backward, thus producing a negative frequency. If the duty cycle of the up/down input is exactly 50 percent, the VCO frequency is zero. The loop gain is set by the counter clock frequency and the number of stages inside the counter. The high-order bits of the counter can therefore be used to drive the address inputs of the analog switches directly. If the up/down input is driven directly by the exclusive-OR gate, the whole circuit only behaves as a first-order, phase-locked loop (PLL). But some averaging is provided by the low-order bits of the bidirectional counter.

A practical BPSK demodulator for a 1.2288-Mb/s digital radio was designed according to the principles described. The detailed circuit diagram of the BPSK demodulator is shown in Fig. 8. The entire demodulator is designed around standard 74HC-series building blocks (silicon-gate CMOS logic) and other simple components. The demodulator requires a four-phase IF input (I and Q of both polarities). The amplitude of each of the four input signals is approximately +2 VDC peak-to-peak centered around the +2.5-VDC reference. A resistor network is used to obtain a 16-phase system from the four-phase input. Two 74HC4067 16-input multiplexers are used as the rotating switches. The two switches are followed by simple lowpass filters to remove the switching transients, while two LM311 comparators limit the I' and Q' signals to CMOS-logic levels. The loop-error signal is obtained with an exclusive-OR gate (pins 1, 2, and 3 of 74HC86). The error signal is then "cleaned" in a two-stage shift register (74HC74) to avoid metastable problems in the following bidirectional counter. The remaining three gates of the 74HC86 are tied in parallel to send the output data over a 75-Ω cable. The digital VCO includes an 8-b, bidirectional binary counter built from two 74HC191 up/downcounters for a total count of 256. The counter is clocked at 6144 kHz, resulting in a VCO frequency range of ± 24 kHz. The latter figures represent the maximum carrier-frequency offset that can be corrected by the described BPSK demodulator. The upper four bits (from the second 74HC191) are used to steer the two 74HC4067 analog switches. The lower four bits (from the first 74HC191) are not used externally. The entire BPSK demodulator requires a single supply voltage of +5 VDC. The four ana-



6. This diagram shows a Costas-loop BPSK demodulator.



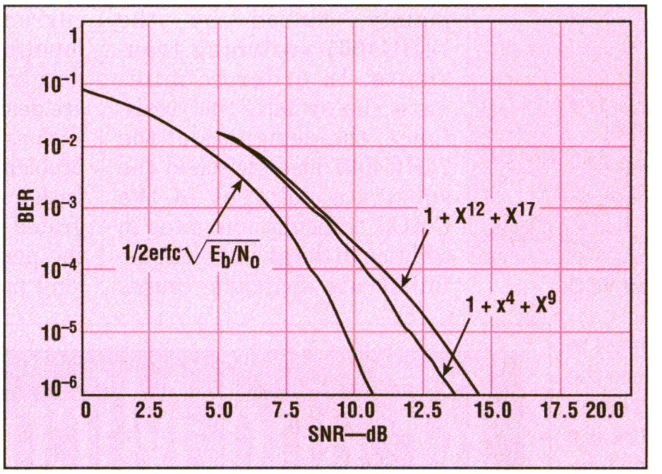
7. This Costas loop uses a digital VCO.

log inputs are therefore referenced to half this value, or +2.5 VDC. Since the remaining circuits of the BPSK receiver are running on +12 VDC, there is a 7805 regulator built into the demodulator unit.

MEASURED PERFORMANCE

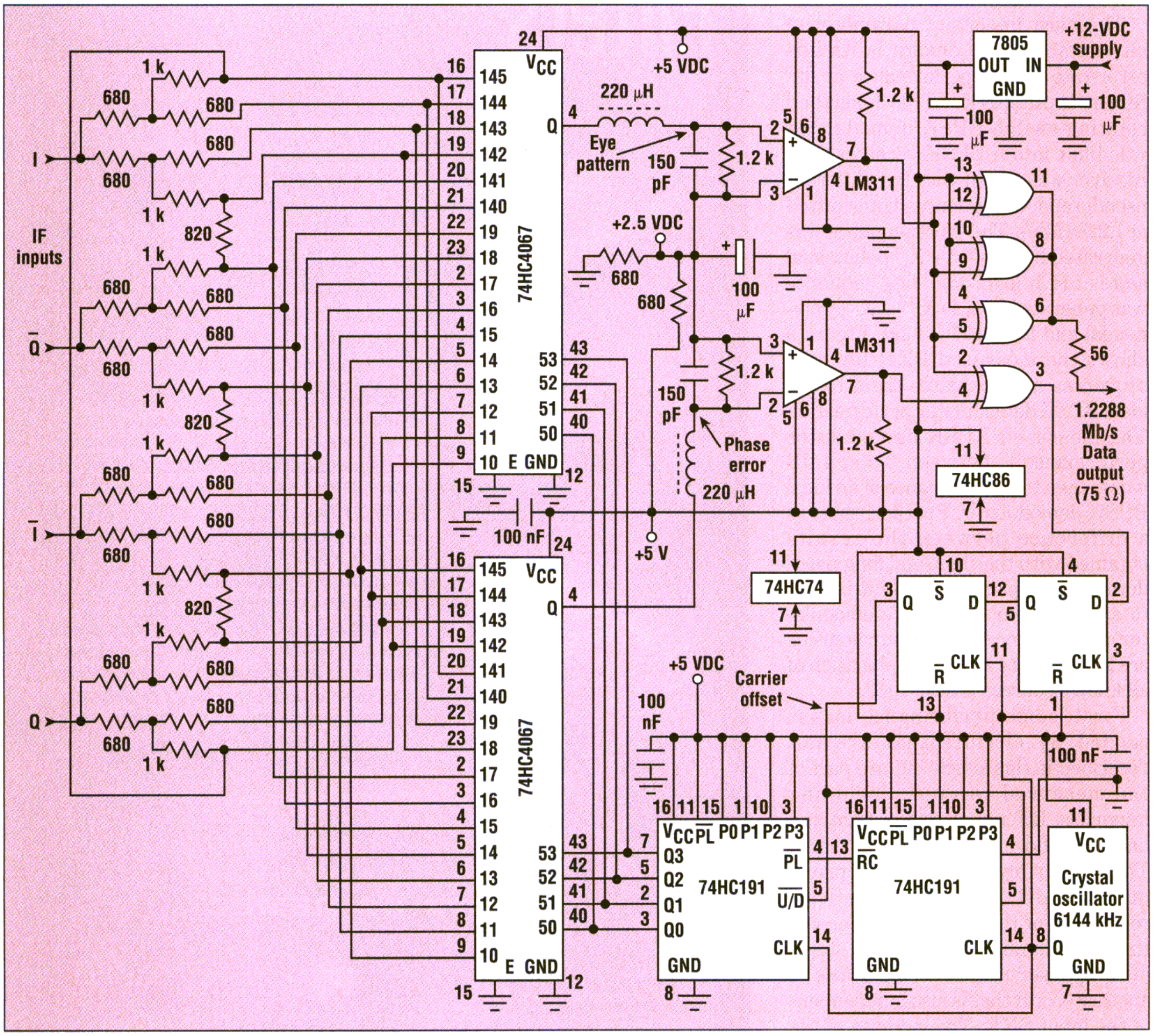
The most important parameter of any demodulator design is its bit-error-rate (BER) performance at different input SNRs. The described rotating-switch BPSK demodulator was built into a BPSK receiver. The receiver was tested with two different pseudorandom sequences transmitted at 1.2288 Mb/s. The short sequence was generated by a $1+X^4+X^9$ polynomial and is 511 b long. The long sequence was generated by a $1+X^{12}+X^{17}$ polynomial and is 131071 b long. Figure 9 shows the measured BER for both sequences and compares them to the ideal BPSK demodulator performance. The measured BPSK demodulator performance is between 2 and 4 dB worse than the performance of an ideal BPSK demodulator. Further, there is a difference between the results obtained with the short and long pseudorandom sequences. While it is easy to explain some sources of demodulator-performance loss, other sources are less obvious. A short explanation of all known sources is described.

First, some performance loss is caused by circuits that are not described in this article but are part of the measured receiver, including imperfect RF mixer quadrature and balance, and imperfect IF filtering. These probably account for approximately 1 dB of performance loss, regardless of the input SNR. Second, the finite number of phases (16 steps) of the phasor rotator bring a loss of another 0.7 dB that is also independent of the input SNR. The increase of the demodulator loss at high SNRs is mainly caused by the 74HC4067 switching transients. In order to minimize the switch's "on" resistance, the designers of the 74HC4067 also switched the substrate voltages of the CMOS transmission gates in addition to the gate voltages. Substrate switching causes current drain from the analog input/output (I/O) pins. Unfortunately, most CMOS analog switches are designed with substrate switching such as the 74HC4067. Of course, the problem that is described can be solved with a custom-designed CMOS integrated circuit (IC). The difference in the performance between short and long pseudorandom sequences also increases at better SNRs. The difference is not caused by the described BPSK demodulator. The real causes are the corresponding I and Q IF amplifiers. In zero-IF receivers, the IF amplifiers ideally should be DC coupled. In practice, this is almost impossible for several reasons, and AC-coupled IF amplifiers are used instead. When using zero-IF receivers with AC-coupled IF amplifiers, the information encoding should be selected so that the distortion introduced by the AC coupling does not corrupt the



9. This graph shows the measured performance of the BPSK demodulator.

information. In the case of a Weaver SSB receiver, the resulting notch that is near 1.4 kHz does not affect the quality of human speech. In the case of digital zero-IF receivers, data randomization or scrambling has to be used in order to avoid any strong discrete spectral lines in the signal spectrum. Further, the lower frequency limit of the IF amplifiers has to be kept low enough (approximately 1 kHz in the described 1.2288-Mb/s receiver) so that the resulting distortion can be tolerated.••



8. This schematic describes a practical 1.2288-Mb/s BPSK demodulator.

Directional Detector Chip Monitors Wireless Handsets

Low-cost monolithic couplers and Schottky diodes team for power monitoring and control applications in cellular and PCS handsets.

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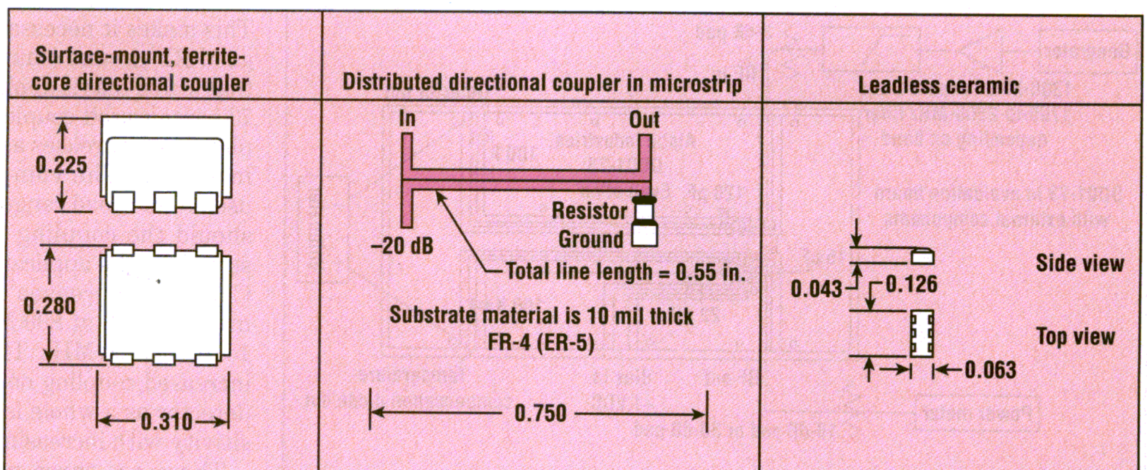
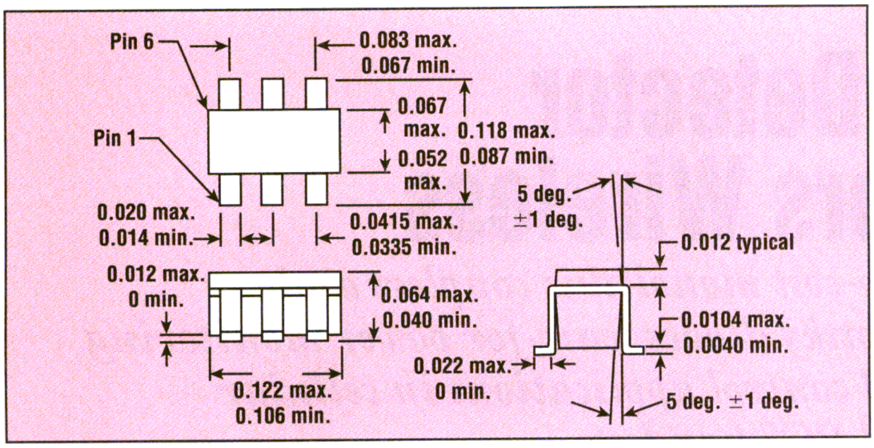
PRACTICAL transmitters require some method of controlling output power. Some form of a closed-loop control system is often used for accurate power control. The power-control system usually adds a detector, coupler, an attenuator, and digital logic or analog control circuitry to a transmitter's power amplifier (PA). Traditionally, directional couplers and detectors have been implemented with ferrite cores or microstrip printed circuits. But a more compact alternative is a device that combines these functions monolithically, with construction similar to a multilayer chip capacitor.

The latest generation of monolithic directional couplers offers considerable advantages in size compared to traditional components (Fig. 1). By using mature gallium-arsenide (GaAs) processes, the engineers at Alpha Industries (Woburn, MA) have developed lines of low-loss couplers characterized by high repeatability.[1,2] The components, which are housed in standard SOT-6 packages, are suitable for cellular applications to 1 GHz and personal-communications-services (PCS) applications to 2 GHz. As an example, an outline drawing of an SOT-6 directional coupler is shown in Fig. 2.

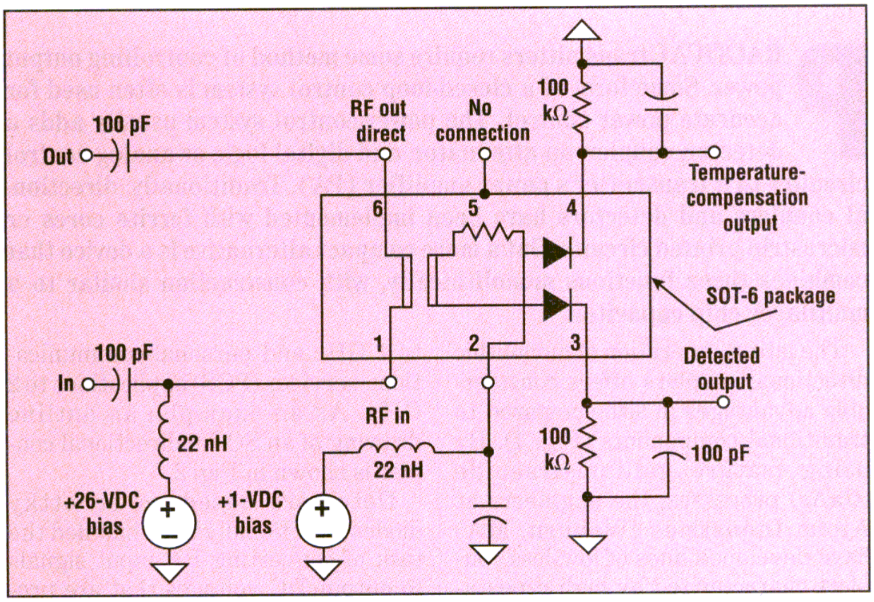
Detectors based on Schottky diodes traditionally have handled the task of converting RF input signals to output DC voltages that are proportional to the power or voltage of the RF input signal. But similar to the monolithic couplers, a shift has



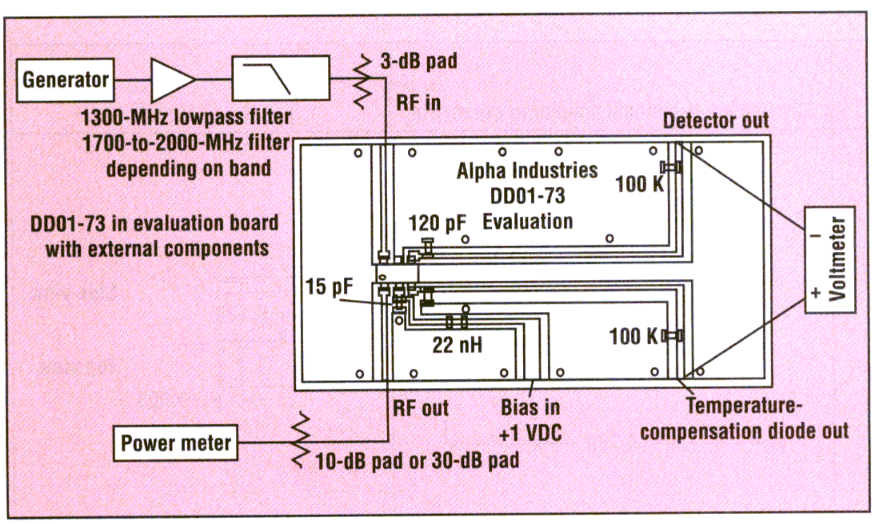
1. This comparison shows the difference in size between ferrite-core, microstrip, and leadless ceramic 20-dB directional couplers.



2. This outline drawing shows a monolithically fabricated SOT-63 directional coupler.



3. The DD01-73 combines a passive directional coupler with two Schottky diodes in an SOT-6 surface-mount package.



4. The surface-mount directional detector was mounted in a test fixture for characterization.

occurred in the packaging of Schottky diodes, from expensive ceramic packages to low-cost, tiny plastic surface-mount packages.

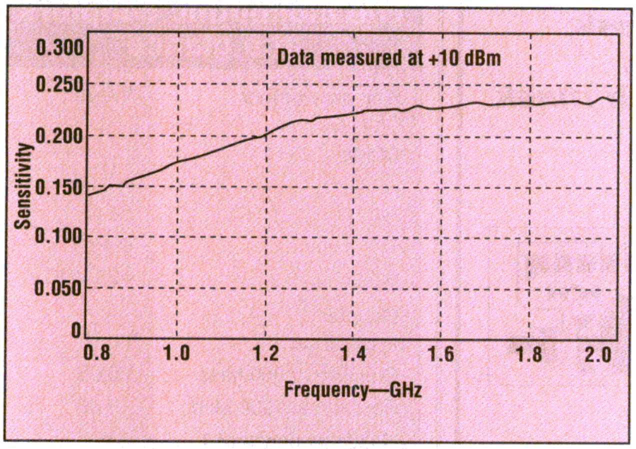
DIODE DETECTORS

Together, directional couplers and Schottky diodes can be combined to form a diode detector for measuring the RF power output of an amplifier without disturbing or attenuating that power significantly. With separate surface-mount directional couplers and Schottky diodes, any further size reduction for a directional detector would require the integration of the components. With the exception of highly integrated systems on a chip, these two devices have generally been procured separately and combined together on a printed-circuit board (PCB). But the DD01-73 from Alpha Industries represents the latest stage in integrating the coupling and diode-detection functions. This GaAs integrated circuit (IC) marries a passive directional coupler and two Schottky barrier diodes on a single GaAs chip that can be housed in a six-lead SOT-6 surface-mount plastic package (Fig. 3). A development program is also underway for mounting the same coupler/detector IC in a smaller, SC-70 package.

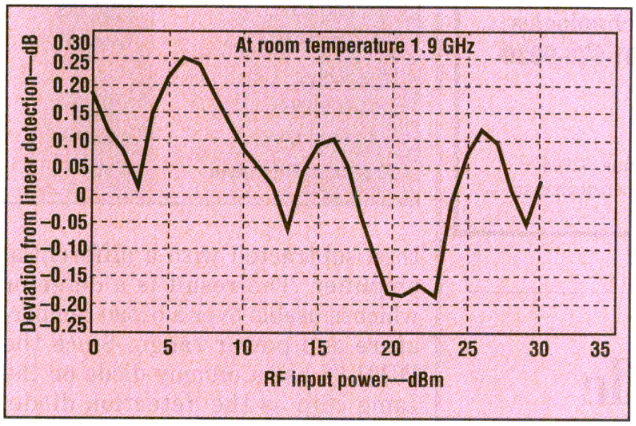
FLEXIBLE OPERATION

The DD01-73 monolithic directional detector can be used with or without DC bias. The ground circuit for the coupler's isolation termination doubles as the bias injection point. This makes it necessary to provide a good RF ground (shunt capacitor) at that lead if bias is applied. The DD01-73's integrated coupler exhibits coupling that increases as the frequency increases. The component can be used for a broad range of frequencies should the coupling value be well-suited for the application of interest. Coupling increases from approximately 20 dB at 800 MHz to roughly 15 dB at 1900 MHz. The result of this increased coupling on the directional detector as a whole is increased sensitivity with increasing frequency.

There are three main regions of operation of a diode detector—the square-law region, the linear region,



5. These measurements show sensitivity as a function of frequency.



6. These measurements reveal errors in linear detection as a function of input power.

and the saturated region.[3] The square-law region is named because it produces an output voltage that is proportional to the square of the input voltage and, consequently, is directly proportional to the input power. Unbiased diode detectors operate in this region for low power levels, typically below 0 dBm, and for only a limited dynamic range (10 dB). As power is increased or if bias is used, the detector falls into a region known as the linear region. Here the output voltage is directly proportional the input RF voltage. The DD01-73 is characterized for, but not limited to, use in the linear region. The saturated region refers to the mode of operation where the RF input voltage peaks approach the reverse breakdown voltage of the diode. At this point, the output voltage levels compress and then finally limit to a maximum value.

The power level where a detector changes from the square-law region to the linear region depends on the diode's saturation current (I_s), from the well-known diode equation:

$$I_d = I_s(\exp\{V_j/[k(t/q)n]\} - 1)$$

Smaller saturation currents are typically indicative of high-level diodes. High-level diodes are named this because a higher voltage is required to forward bias them than diodes with larger saturation currents. The saturation current of a diode is determined by its geometry, doping levels and, in the case of Schottky diodes, the barrier metal. GaAs diodes are considered high-level devices. The GaAs diodes in the design of the DD01-73 have a saturation current, I_s, of 3.5 pA. The lowest level zero-bias diodes (ZBDs) have saturation currents on the order of microamperes. Without matching circuitry, GaAs diodes may be used as square-law detectors down to 0 dBm. ZBDs may be used down to −10 dBm. Since this square-law region has approximately 10-dB dynamic range, the easiest way to make a sensitive diode detector work well over a broad range of input power is to bias the detector diodes. Biasing the diode extends the linear region down to much lower power.

The problem with bias, however, is the drift of detector output voltage versus temperature. The common solution to this has been to include a "dummy" diode that is physically located near the detecting diode. The "dummy" diode is supplied with the same bias current as the detecting diode and their output voltages are

The monolithic couplers at a glance

Key parameters	Value
Diode	
I_s	3.5 pA
R_s	60 Ω
C_j	0.4 pF
B_v	+25 VDC
Coupler	
Impedance	50 Ω
Coupling at 800 MHz	20 dB
Coupling at 1900 MHz	15 dB
Maximum power (2:1 VSWR max.)	4 W
Through loss	
At 800 MHz	0.20 dB
At 1900 MHz	0.45 dB
Directivity	
At 800 MHz	13 dB
At 1900 MHz	20 dB
Maximum RF power	4 W

then subtracted with a differential amplifier. The result is a detector which is usable over a broad temperature and power range. Since the DD01-73 has a dummy diode on the same chip as the detection diode, excellent temperature tracking is achieved. Measurements made on the DD01-73 with temperature compensation show the change in detected voltage from room temperature (approximately +25 to +85°C at 0-dBm input to be only 1 dB at 1.7 GHz. The same situation without temperature compensation yields 7 dB of error for the monolithic detector.

The DD01-73 was mounted in a test fixture (Fig. 4) and characterized. Some of the important parameters are outlined in the table. Figure 5 offers plot of sensitivity versus frequency while Fig. 6 shows the errors in linear detection versus input power. (Additional information on the DD01-73 can be found by visiting the company's website, at http://www.alphind.com.) ●●

References

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2. Jack Browne, "MMIC Techniques Yield Tiny Couplers And Power Dividers," *Microwaves & RF*, Vol. 36, No. 2, February 1997, pp. 155-156.
3. Application notes on Schottky diodes, 1996 catalog, Alpha Industries, Woburn, MA.


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Model PLL Dynamics And Phase-Noise Performance

Model PLL Dynamics, Part 2

By understanding the basic sources of phase noise, it is possible to accurately model a PLL with the help of commercial CAE programs.

Eric Drucker

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

PHASE-LOCKED loops (PLLs) and their importance to modern communications were detailed in the first part of this article series (see *Microwaves & RF,* November 1999, p. 69). In that article, a basic PLL model was presented. The loop dynamics were modeled to determine the open- and closed-loop transfer functions and the various loop parameters (such as open-loop gain-crossover, phase margin, closed-loop bandwidth, etc.). An example was presented to illustrate these concepts. In Part 2, the basics of phase noise will be reviewed. Using the mathematical-analysis software MathCAD, along with the previous example, it will be possible to show how the various noise sources in a PLL can be modeled. A novel method using PSPICE to model the noise sources will also be shown.

An amplitude and phase-modulated sinusoidal signal can be written as follows:

$$V(t) = V_0 \cdot [1 + v_{am}(t)] \cdot \{sin[2\pi f_c t + \theta(t)]\} \quad (6)$$

where:

V_0 = the amplitude,

f_c = the carrier frequency,

$v_{am}(t)$ = the amplitude-modulation (AM) component, and

$\theta(t)$ = the phase-modulation (PM) component.

For the purposes of this discussion, the AM component will be disregarded.

$$V(t) = V_0 \cdot \{sin[2\pi f_c t + \theta(t)]\} \quad (7)$$

Modeling a sinusoidal-angle modulation with a rate of f_m, yields:

$$\theta(t) = \frac{\Delta f}{f_m} \cdot sin(2\pi f_m t);\ \beta = \frac{\Delta f}{f_m} \quad (8)$$

$$V(t) = V_o \cdot [sin(2\pi f_c t) + \beta sin(2\pi f_m t)] \quad (9)$$

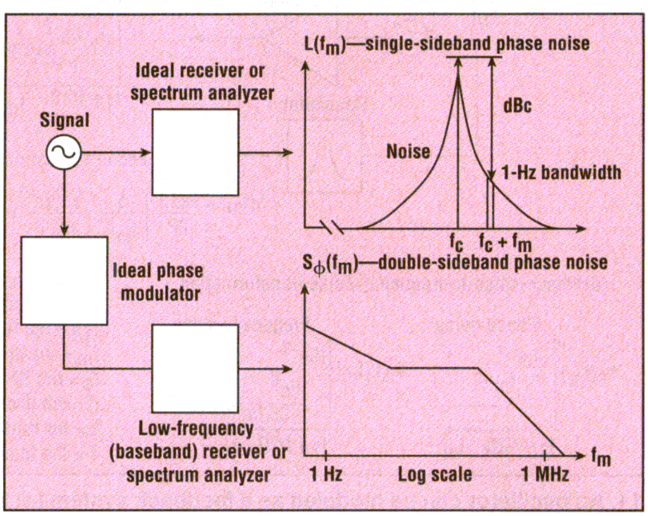
For $\beta \leqslant 1$ (small-angle modulation) and applying trigonometric identities:

$$V(t) = V_o \cdot [sin(2\pi f_c t) + \frac{\beta}{2}\{sin[2\pi(f_c + f_m)] - sin[2\pi(f_c - f_m)]\}] \quad (10)$$

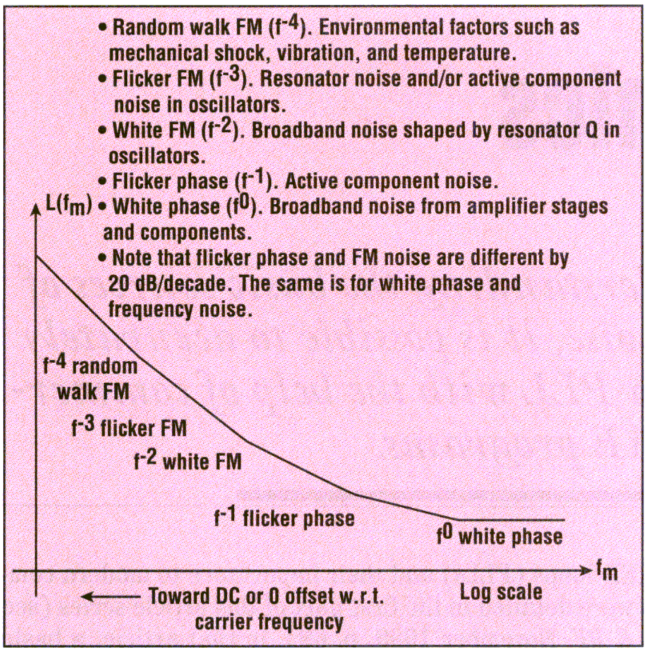
where:

f_m = the modulation frequency,

Δf = the peak frequency-modulation



12. A spectrum analyzer can be used to evaluate single-sideband (SSB) or double-sideband phase noise.



13. Phase noise consists of several components, including random-walk FM, flicker-noise FM, white-noise FM, flicker phase noise, and white phase noise.

Noise Equations for Noise Sources

VCO: Noise: Floor = -155 dB/Hz; ~ -110 dB/Hz @ 10 kHz, 1/f corner @ ~5 kHz

$k_{0\_vco} := 10^{-15.5}$ $k_{2\_vco} := 10^{-3}$ $k_{3\_vco} := 10^{0.7}$

$S_{\phi\_vco}(f) := \frac{k_{2\_vco}}{f^2} + k_{0\_vco} + \frac{k_{3\_vco}}{f^3}$ $dB_{vco_i} := 10 \cdot \log(S_{\phi\_vco}(f_i))$

$dB_{30\_vco} := 10 \cdot \log\left(\frac{k_{3\_vco}}{f^3}\right)$ $dB_{20\_vco} := 10 \cdot \log\left(\frac{k_{2\_vco}}{f^2}\right)$

$10 \cdot \log(S_{\phi\_vco}(10^3)) = -82.21$ $10 \cdot \log(S_{\phi\_vco}(5 \cdot 10^3)) = -100.964$

$10 \cdot \log(S_{\phi\_vco}(10 \cdot 10^3)) = -108.236$ $10 \cdot \log(S_{\phi\_vco}(10^7)) = -154.865$ VCO Noise at 1, 5, 10 KHz, & 10 MHz Offset

Divider: Noise: Floor = -155 dB/Hz, 1/f corner @ 1 kHz

$k_{1\_md} := 10^{-12.5}$ $k_{0\_md} := 10^{-15.5}$

$S_{\phi\_md}(f) := \frac{k_{1\_md}}{f} + k_{0\_md}$ $dB_{md_i} := 10 \cdot \log(S_{\phi\_md}(f_i))$

$10 \cdot \log(S_{\phi\_md}(10^3)) = -151.99$ $10 \cdot \log(S_{\phi\_md}(10^7)) = -155$ Divider Noise at 1 KHz, & 10 MHz Offset

Reference Noise from Manufacturers Data Sheet

$k_{0\_ref} := 10^{-15.8}$ $k_{1\_ref} := 10^{-12.7}$ $k_{2\_ref} := 10^{-9.85}$ $k_{3\_ref} := 10^{-7.82}$

$S_{\phi\_ref}(f) := \frac{k_{1\_ref}}{f} + k_{0\_ref} + \frac{k_{2\_ref}}{f^2} + \frac{k_{3\_ref}}{f^3}$ $dB_{ref_i} := 10 \cdot \log(S_{\phi\_ref}(f_i))$

$10 \cdot \log(S_{\phi\_ref}(10^1)) = -107.807$ $10 \cdot \log(S_{\phi\_ref}(10^2)) = -135.029$

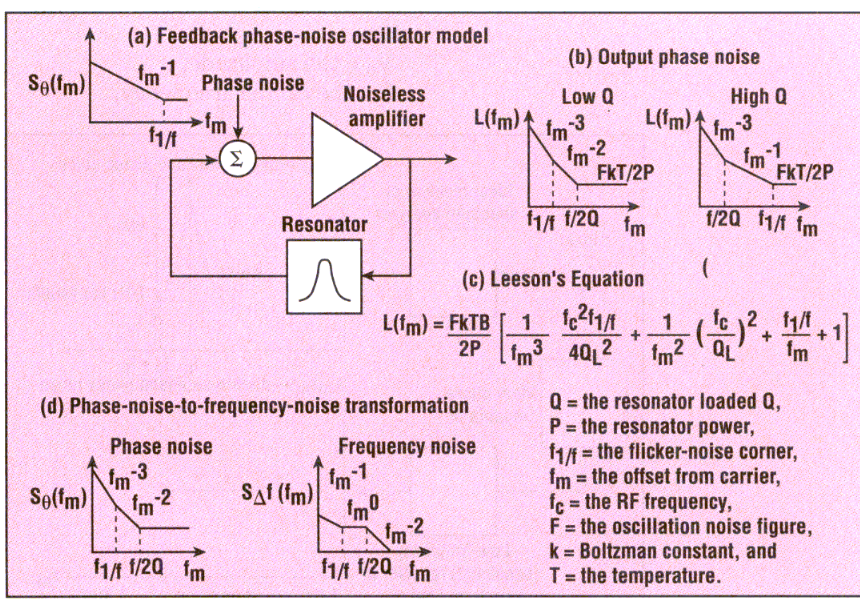
$10 \cdot \log(S_{\phi\_ref}(10^3)) = -152.887$ $10 \cdot \log(S_{\phi\_ref}(10^4)) = -157.45$ Reference Noise at 10, 100 Hz and 1, 10 KHz Offset

(FM) deviation, and
β = the modulation index.

This shows that a small-angle deviation gives rise to sidebands on each side of the carrier at an amplitude of $\beta/2$. Extending this idea, noise can be treated as an infinite number of single FM sidebands.

The distribution of these noise sidebands as a function of offset frequency can be expressed in different ways (Fig. 12). One way is to use an ideal receiver or spectrum analyzer at RF with a 1-Hz resolution-bandwidth filter. The total power of the signal would first be measured and, since the noise is small, this is essentially equal to the carrier power. Then the receiver would be tuned to a particular offset (f_m) from the carrier, and the phase-noise power is measured. The ratio of these two measurements, expressed in decibels, is the normalized power-spectral density (PSD) in one sideband referred to the carrier at a frequency offset f_m. This is known as the single-sideband (SSB) phase noise, $L(f_m)$, with units of Hertz^{-1} or expressed in decibels relative to the carrier level as dBc/Hz. A plot of the phase noise as function of the offset is commonly shown in data sheets in order to characterize oscillators and frequency synthesizers.

Another method is to demodulate the signal with an ideal phase demodulator. The output of the phase demodulator is the baseband phase noise and can be analyzed with a low-frequency spectrum analyzer, again with a 1-Hz resolution-bandwidth filter. The resulting plot as a function of baseband or offset frequency is the double-sided phase-noise spectrum, $S_\phi(f_m)$, expressed in $(\text{radians})^2/\text{Hz}$. The double-sideband phase noise is twice that of (or 3 dB more than) the SSB phase noise. One can integrate the area under the double-sideband phase-noise curve, over a specific bandwidth (f_1 to f_2) to obtain the root-mean-square (RMS) phase noise and, by extension, the RMS frequency noise. From the RMS phase or frequency noise, the

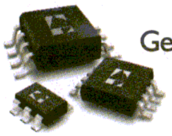


14. An oscillator can be modeled as a feedback system (a) that produces noise as a function of oscillator Q (b). The phase noise can be described in terms of Leeson's equation (c) and transformed to frequency noise (d).

When it comes to building a better

SWITCH,

a larger wafer can make a lot of cents.



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SPDT reflective switches

	Freq. (GHz)	Isolation (dB)	Insertion Loss (dB)	P1dB (dBm)	IP3 (dBm)	Package
AWS5501	DC-3.0	23	0.45	35	55	SOIC 8 pin
AWS5502	DC-2.5	20	0.45	28	45	SOT-6
AWS5503	DC-3.0	22	0.45	35	55	MSOP 8 pin
AWS5504	DC-2.0	17	0.4	38	55	SOT-6
AWS5506	DC-2.5	20	0.45	28	45	SOT-6

Note: specs typical at 900 MHz

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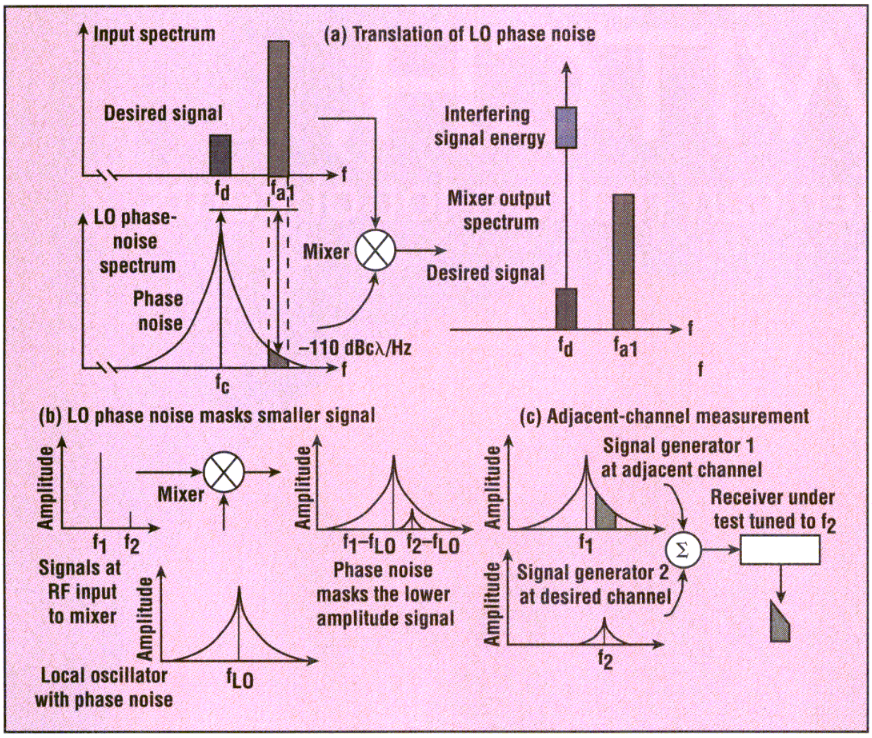
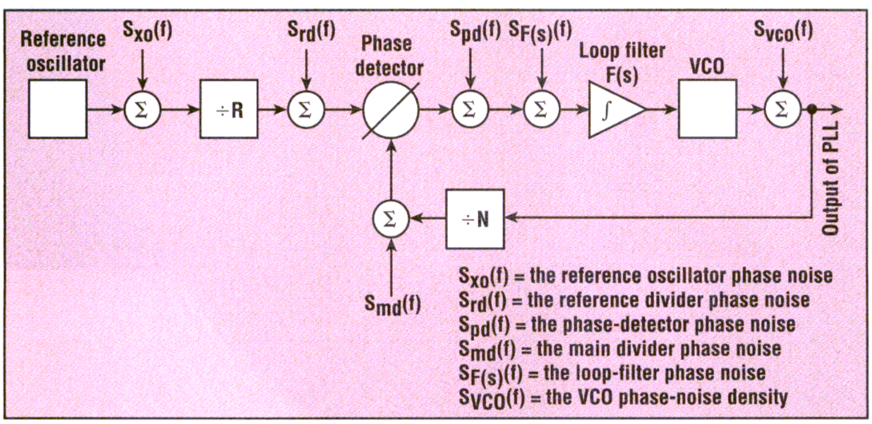
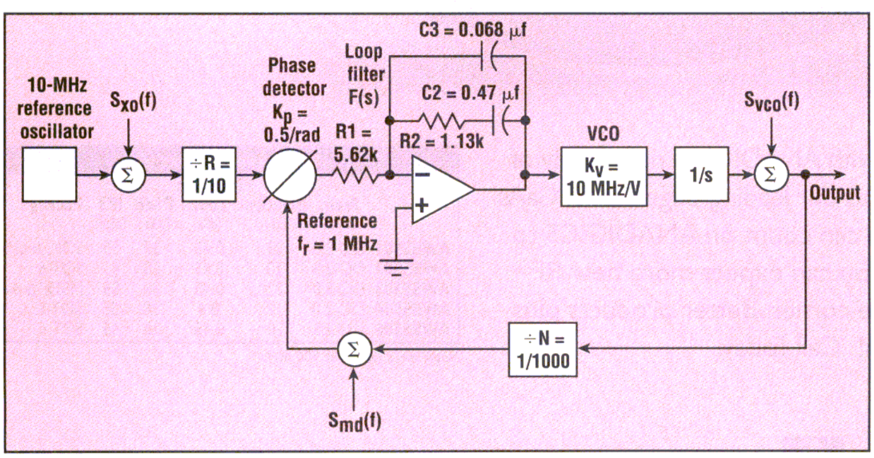


15. An LO's phase noise can affect a receiver's adjacent-channel rejection (a) by masking low-level signals (b). Adjacent-channel performance can be evaluated with a pair of signal generators (c).



16. A variety of noise sources affect the performance of a PLL.



17. In this example, the PLL is assumed to have only three noise sources—VCO noise, reference oscillator noise, and main divider noise.

RMS time jitter can be computed. When looking at phase-noise plots, it should be noted that a "zero" frequency offset or "DC" is the carrier.

The integrated phase noise in terms of RMS radians can be expressed as eq. 11:

$$\Delta\theta_{RMS} = \left[\int_{f_1}^{f_2} S\phi(f_m) df_m \right]^{0.5} \qquad (11)$$

while the integrated frequency noise in terms of RMS Hz can be expressed as eq. 12:

$$\Delta f_{RMS} = \left[\int_{f_1}^{f_2} S\phi(f_m) \cdot f_m{}^2 df_m \right]^{0.5} \qquad (12)$$

When the phase noise is plotted in decibels versus log frequency, various regions of the phase-noise curve can be identified. These regions have slopes of 0, 1/f (−10 dB/decade), $1/f^2$ (−20 dB/decade), $1/f^3$ (−30 dB/decade), etc. (Fig. 13).

The flat or zero-slope region corresponds to white phase noise of thermal origin. This thermal, resistive, or Johnson noise has a Gaussian amplitude distribution, constant with frequency. Amplifier noise figure is a manifestation of this thermal noise.

Close to "DC" or a zero-frequency offset, there is a region of 1/f, or flicker noise. This is believed to come from irregularities in the semiconductor structure. Typically, the 1/f corner is between 1 and 10 kHz.

Frequency dividers and amplifiers exhibit only 0 and 1/f slope regions. Oscillators can have all regions. The general phase-noise equation shown below (with the m subscript dropped) is typically used for double-sideband noise $[S_\phi(f_m)]$ expressed in power:

$$S\phi(f) = k_o + \frac{k_1}{f} + \frac{k_2}{f^2} + \frac{k_3}{f^3} + \frac{k_4}{f^4} \qquad (13)$$

An oscillator can be modeled as a feedback system consisting of a resonator and a noiseless amplifier. Phase noise is injected at the input to the amplifier (Fig. 14a). This injected phase noise consists of a flat region and a 1/f region. This noise is shaped by the resonator, which has a −20-dB/decade

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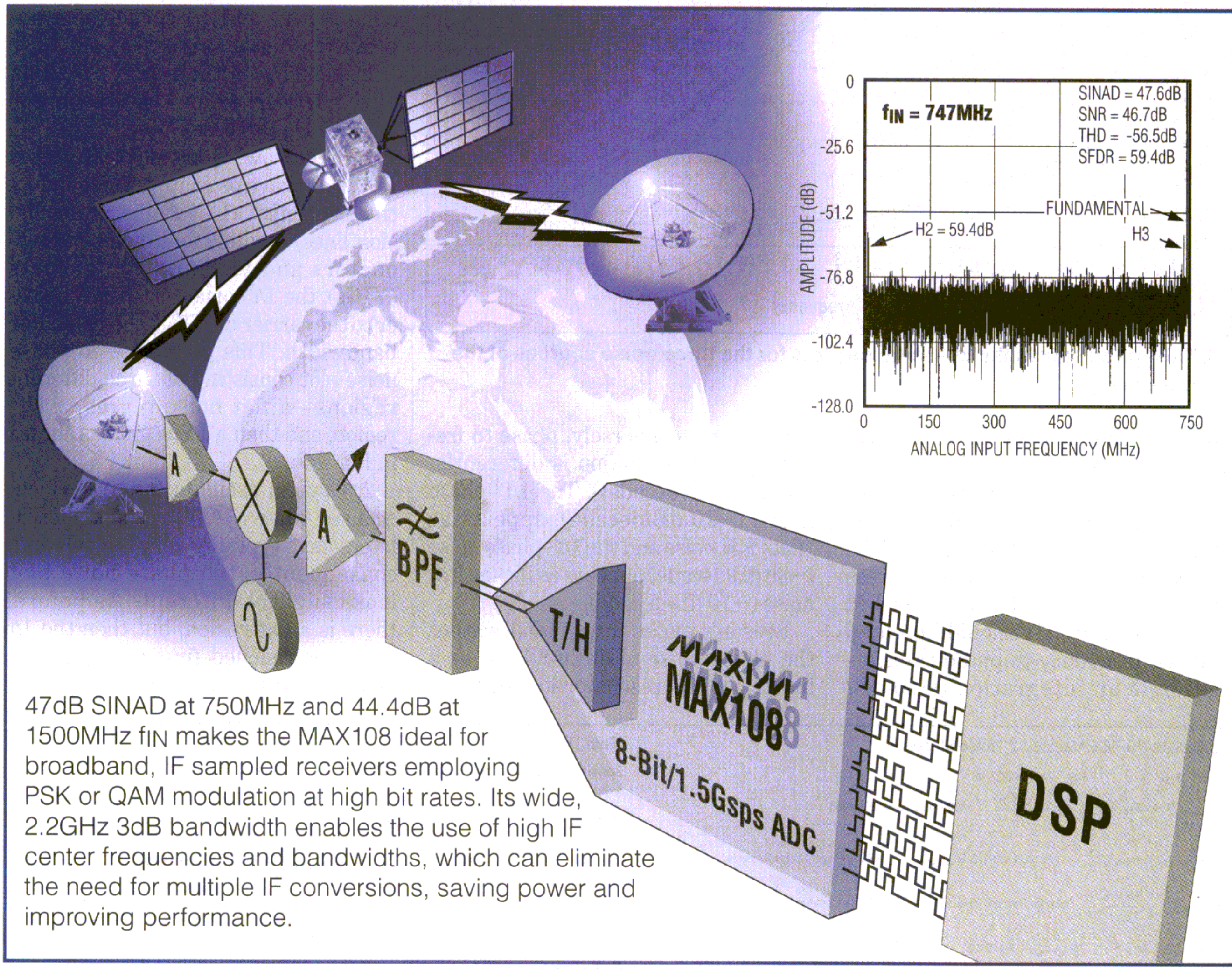

47dB SINAD at 750MHz and 44.4dB at 1500MHz f_{IN} makes the MAX108 ideal for broadband, IF sampled receivers employing PSK or QAM modulation at high bit rates. Its wide, 2.2GHz 3dB bandwidth enables the use of high IF center frequencies and bandwidths, which can eliminate the need for multiple IF conversions, saving power and improving performance.

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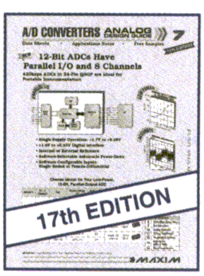

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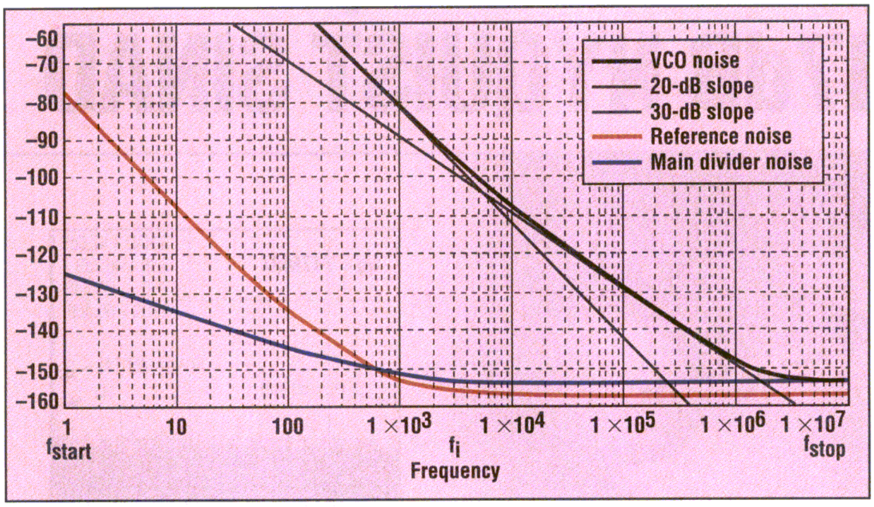

18. This graph shows the phase-noise plots for the three noise sources of the example in Fig. 17.

slope on either side of the center frequency. This flat phase noise becomes $1/f^2$ in nature and the 1/f phase noise becomes $1/f^3$ in nature, within the resonator bandwidth, at the output of the oscillator (Fig. 14b). One could express this noise in terms of frequency or FM noise (Fig. 14d). [It should be noted that the frequency-to-phase transformation is an integration (−20 dB/decade) and, conversely, phase to frequency transformation is differentiation (+20 dB/decade)]. The $1/f^2$ phase noise, (−20 dB/decade), appears as "flat" FM noise and the $1/f^3$ phase noise (−30 dB/decade) appears as flicker FM noise (−10 dB/decade).

Leeson's model (Fig. 14c) describes the phase noise of an oscillator as a function of the resonator-loaded quality factor (Q) or bandwidth, oscillation frequency, noise figure, power, and offset from the carrier. The phase noise improves with higher Q (narrower resonator bandwidth) and power. A lower noise figure and 1/f corner also improves the noise.

Depending on the relative position of the 1/f corner and the resonator bandwidth, two cases arise:

1. In the "low" Q case, typical of most RF/microwave oscillators using inductive-capacitive (LC) tanks, transmission lines, ceramic and dielectric resonators and yttrium-iron garnets (YIGs), the 1/f corner is "inside" (closer to the carrier or "DC") the resonator bandwidth. This gives rise to phase noise plot consisting of three different regions—a flat noise region, a $1/f^2$ region, and then a $1/f^3$ region as shown in Fig. 14b.

2. Crystal oscillators and surface-acoustic-wave (SAW) resonator oscillators have typically very high Qs and, consequently, the phase-noise plot looks slightly different. As before, there is a flat region, but then the 1/f corner is reached first and then the phase noise increases at a rate of 10

Calculate PSPICE Values for Noise Models

$k := 1.38\cdot10^{-23}$ $\quad T := 300$ $\quad q := 1.602\cdot10^{-19}$

$$S_{\phi\_vco}(f) := \frac{k_{2\_vco}}{f^2} + k_{0\_vco} + \frac{k_{3\_vco}}{f^3}$$ VCO Noise Power Equation

$e_{n0}^2 = 4\cdot k\cdot T\cdot B\cdot R_{k0\_vco}$ $\quad k_{0\_vco} = 4\cdot k\cdot T\cdot B\cdot R_{k0\_vco}$ $\quad e_n^2$ Represents Noise Power k_{0_vco} Term

$$R_{k0\_vco} := \frac{1}{4}\cdot\frac{k_{0\_vco}}{(k\cdot T)}$$ Solve for Resistor Values for VCO Flat Noise Component (k_{0_vco})

$R_{k0\_vco} = 1.91\cdot10^4$ $\quad 2\cdot R_{k0\_vco} = 3.819\cdot10^4$

$$e_{n2}^2 = \left(\frac{K_{i2\_vco}}{f}\right)^2\cdot e_n^2$$ Integrating Flat Noise Power to Produce $1/f^2$ Noise Power

$$e_{n2}^2 = \frac{\left(K_{i2\_vco}\cdot\sqrt{4\cdot k\cdot T\cdot B\cdot R_{k2\_vco}}\right)^2}{f^2} \qquad \left(K_{i2}\cdot\sqrt{4\cdot k\cdot T\cdot B\cdot R_{k2\_vco}}\right)^2 = k_{2\_vco}$$ Numerator Equals k_{2_vco}

$K_{i2\_vco} := 10^5$ integrator Gain for VCO $1/f^2$ Noise Component

$$R_{k2\_vco} := \frac{1}{4}\cdot\frac{k_{2\_vco}}{(k\cdot T)\cdot K_{i2\_vco}^2}$$ Solve for Resistor Values for VCO $1/f^2$ (k_{2_vco}) Noise Component

$R_{k2\_vco} = 6.039\cdot10^6$ $\quad 2\cdot R_{k2\_vco} = 1.208\cdot10^7$

$$e_{n1}^2 = \left(\frac{k\cdot T}{q\cdot I_D}\right)^2\cdot\frac{K_f\cdot I_D}{f}$$ 1/f Noise from PSPICE Diode Noise Model

$K_{i3\_vco} := 10^7$ $\quad I_D := 10^{-3}$ integrator Gain for VCO $1/f^3$ Noise Component & Diode Bias Current

$$e_{n3}^2 = \left[\left(\frac{K_{i3\_vco}}{f}\right)^2\cdot\left(\frac{k\cdot T}{q\cdot I_D}\right)^2\cdot\frac{K_f\cdot I_D}{f}\right] = \frac{K_{i3\_vco}^2\cdot\left(\frac{k\cdot T}{q\cdot I_D}\right)^2\cdot\left(K_{f3\_vco}\cdot I_D\right)}{f^3}$$

$$K_{i3\_vco}^2\cdot\left(\frac{k\cdot T}{q\cdot I_D}\right)^2\cdot\left(K_{f3\_vco}\cdot I_D\right) = k_{3\_vco}$$ Numerator Equals k_{3_vco}

Solve for K_f in Diode Model for VCO $1/f^3$ (k_{3_vco}) Noise Component

$$K_{f3\_vco} := \frac{k_{3\_vco}}{\left(K_{i3\_vco}^2\cdot k^2\cdot T^2\right)}\cdot q^2\cdot I_D \qquad K_{f3\_vco} = 7.505\cdot10^{-14}$$

Main Divider

Solve for Resistor Values for Main Divider Flat (k_{0_md}) Noise Component

$$R_{k0\_md} := \frac{1}{4}\cdot\frac{k_{0\_md}}{(k\cdot T)} \qquad R_{k0\_md} = 1.91\cdot10^4 \qquad 2\cdot R_{k0\_md} = 3.819\cdot10^4$$

Solve for K_f in Diode Model for Main Divider 1/f (k_{1_md}) Noise Component

$$\left(\frac{k\cdot T}{q\cdot I_D}\right)^2\cdot\left(K_{f1\_md}\cdot I_D\right) = k_{1\_md} \qquad K_{f1\_md} := \frac{k_{1\_md}}{\left(k^2\cdot T^2\right)}\cdot q^2\cdot I_D \qquad K_{f1\_md} = 4.735\cdot10^{-13}$$

Reference

Solve for Resistor Values for Reference Flat (k_{0_ref}) Noise Component

$$R_{k0\_ref} := \frac{1}{4}\cdot\frac{k_{0\_ref}}{(k\cdot T)} \qquad R_{k0\_ref} = 9.571\cdot10^3 \qquad 2\cdot R_{k0\_ref} = 1.914\cdot10^4$$

Solve for K_f in Diode Model for Reference 1/f Noise (k_{1_ref}) Component

$$K_{f1\_ref} := \frac{k_{1\_ref}}{\left(k^2\cdot T^2\right)}\cdot q^2\cdot I_D \qquad K_{f1\_ref} = 2.988\cdot10^{-13}$$

$K_{i2\_ref} := 10^2$ integrator Gain for Reference $1/f^2$ Noise

Solve for Resistor Values for Reference $1/f^2$ (k_{2_ref}) Noise Component

$$R_{k2\_ref} := \frac{1}{4}\cdot\frac{k_{2\_ref}}{(k\cdot T)\cdot K_{i2\_ref}^2} \qquad R_{k2\_ref} = 8.53\cdot10^5 \qquad 2\cdot R_{k2\_ref} = 1.706\cdot10^6$$

$K_{i3\_ref} := 10^2$ integrator Gain for Reference $1/f^3$ Noise

Solve for K_f in Diode Model for Reference $1/f^3$ (k_{3_ref}) Noise Component

$$K_{f3\_vco} := \frac{k_{3\_vco}}{\left(K_{i3\_vco}^2\cdot k^2\cdot T^2\right)}\cdot q^2\cdot I_D \qquad K_{f3\_vco} = 7.505\cdot10^{-14}$$

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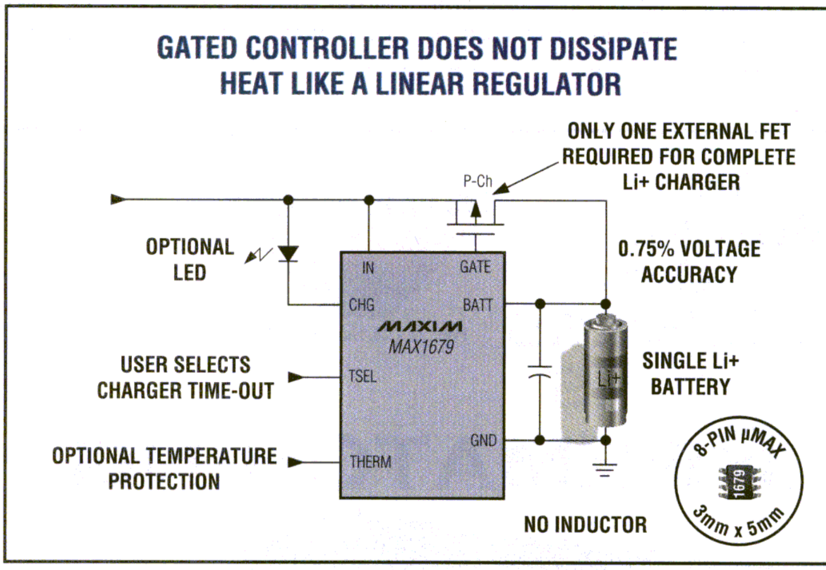

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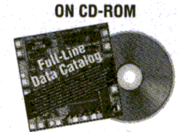

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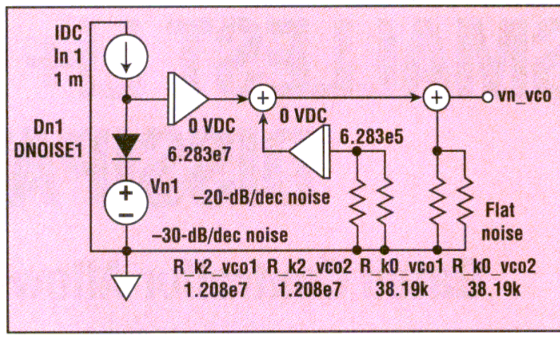
dB/decade. Since the resonator has very high Q, this implies a very narrow bandwidth—less than the flicker corner. Therefore, the region closest to the carrier is the $1/f^3$ region. Even the explanation is somewhat simplified since $1/f^4$ sloped regions have been observed very close to the carrier.

If a PLL is used as a local oscillator (LO) in a receiver, the phase noise of the LO can degrade the adjacent-channel rejection of the receiver by a process known as reciprocal mixing. In Fig. 15a, an input spectrum consisting of the desired signal and a adjacent-channel signal is mixed with an LO. If the LO is only a pure sinusoid, the intermediate-frequency (IF) output of the mixer would just be a shifted replica of the input spectrum. Of course, the IF would have a filter to reject the adjacent channels. The LO phase noise will mix with the unwanted signals in adjacent channels, producing energy that appears in the IF passband, coincident with the desired signal. An example, using the Global System For Mobile Communications (GSM), assumes that the first interfering signal is spaced 600 kHz away and that the detection bandwidth is 200 kHz. If the average LO phase noise is -110 dBc/(Hz)$^{0.5}$ 600 kHz away, the total noise power in the 200-kHz channel with respect to the carrier is:



19. The phase noise of a VCO can be modeled using a simple schematic diagram in PSPICE software.

$$P = \frac{-110\ dBc/\sqrt{Hz}}{\sqrt{200\ kHz}} =$$
$$-110 + 10\log(200\ kHz) =$$
$$-57\ dBc$$

This approximates the noise as being flat across the 200-kHz-wide channel. If the desired signal is -100 dBm and the undesired signal is -60 dBm, the signal-to-noise ratio (SNR) is [-100 dBm $-$ 0 dBc (LO carrier)] $-$ [-60 dBm -57 dBc (LO noise)] = 17 dB. This procedure can be repeated with additional interfering signals to determine the worst-case phase-noise requirement.

In a spectrum analyzer (Fig. 15b), when an attempt is made to resolve two closely spaced signals with widely differing amplitudes, the phase noise of the LO masks the weaker of the two signals. To measure the adjacent-channel selectivity of a receiver, two signal generators are used (Fig. 15c). The amplitude of the in-channel generator is set at the desired sensitivity level and the amplitude of the adjacent or off-channel generator is increased until the sensitivity decreases by a known amount. The phase noise from the adja-

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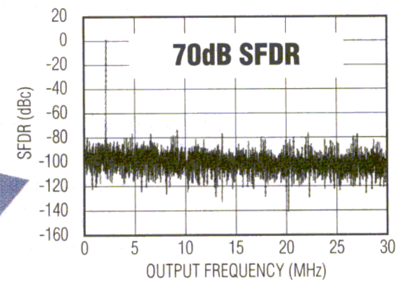

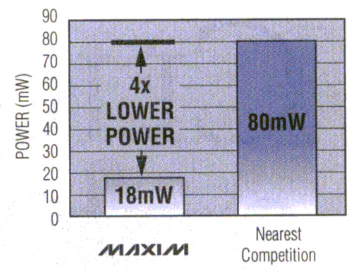

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MAX5182/MAX5185	10	2 (Alternate Phase)	70	N/A	N/A	IOUT/VOUT
MAX5186/MAX5189	8	2 (Simultaneous)	58	±1	±0.2	IOUT/VOUT
MAX5187/MAX5190	8	1	60	N/A	N/A	IOUT/VOUT
MAX5188/MAX5191	8	2 (Alternate Phase)	58	N/A	N/A	IOUT/VOUT

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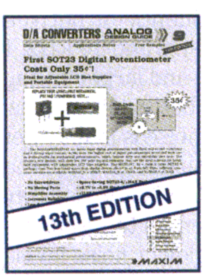

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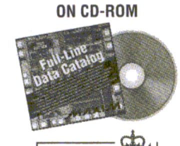

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cent-channel generator that spills into the desired channel could cause the receive selectivity to be worse than expected. The phase noise of the off-channel generator will be measured rather than the in-channel signal. Phase noise can also have an adverse effect in clock-recovery systems, radar, and digital communications systems.

All of the elements in the PLL contribute to the overall phase noise. The noise mechanisms for crystal and RF/microwave oscillators were previously discussed. Another significant source of noise in PLLs is the dividers and phase detectors. There are a number of different types of phase detectors, including mixers or multipliers, sample-and-hold devices, digital exclusive OR gates, and the most common, digital flip-flop phase detectors. The flip-flip phase detector, in addition to providing phase information, also has an intrinsic mechanism to provide proper steering so the loop can achieve lock.

The main disadvantage of the flip-flop phase detector is that it suffers a nonlinear region or "dead zone" close to a zero phase offset. However, various design techniques can mitigate this problem. Other types of phase detectors only provide phase information, and additional circuitry is necessary for steering and acquisition.

LOGIC DEVICES USED FOR PHASE DETECTORS AND DIVIDERS INTRODUCE PHASE NOISE, AND DIFFERENT LOGIC FAMILIES HAVE DIFFERENT PHASE-NOISE CHARACTERISTICS.

The diode mixer phase detector has the best phase noise and is used in critical applications. Logic devices used for phase detectors and dividers introduce phase noise and different logic families have different phase-noise characteristics as a function of the operating frequencies. ECL has a noise floor of approximately −145 to −150 dBc/(Hz)$^{0.5}$, while advanced complementary-metal-oxide-semiconductor (CMOS) logic has a noise floor between −155 and −165 dBc/(Hz)$^{0.5}$ depending on the input (I) and output (O) operating frequency. The 1/f or flicker corner is at an offset frequency of between several hundred Hertz and 10 kHz. In general, faster logic and larger voltage swings give rise to better phase noise. Assuming that the noise is mainly generated in the transition region between logic 0 and 1, with the faster rise time, then less time is spent in the transition region, resulting in lower noise levels.

Since advanced CMOS logic has a +5-VDC swing versus approximately 800 mV for ECL, this would explain the noise improvement when using advanced CMOS logic. It is difficult to measure the phase noise of dividers and

(concluded on p. 117)

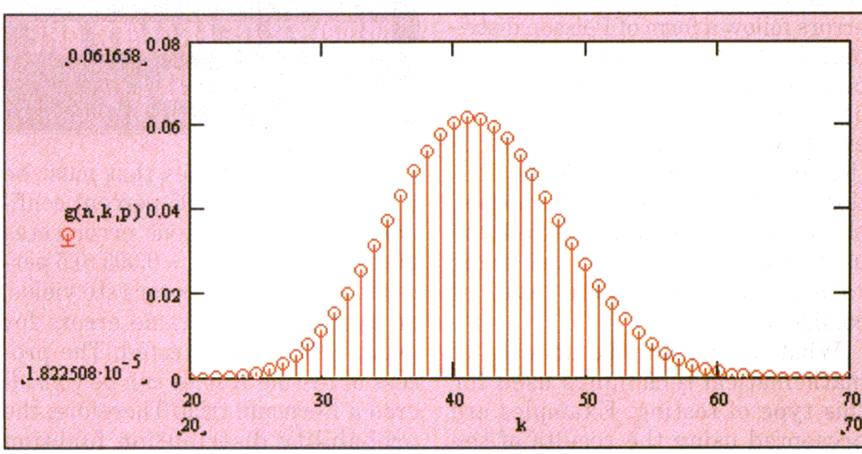
Increase Throughput With Confidence Interval Testing

Statistical analysis can be useful in reducing the total test time to quality production-line wireless units for shipping.

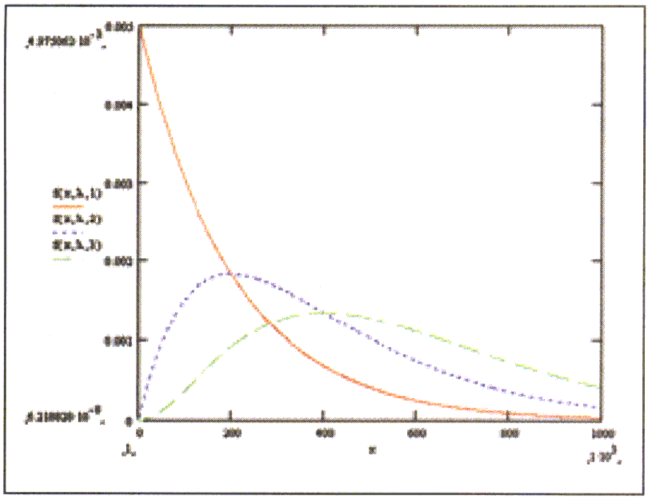
Thomas Yeager
Product Manager
Agilent Technologies, Spokane Div., 1620 Signal Dr., Spokane, WA 99220; (509) 921-4615, FAX: (509) 921-4305, Internet: http://www.agilent-tech.com.

CELLULAR-TELEPHONE handset manufacturers rely on frame-error-rate (FER) testing to ensure the quality of their products. Of course, the speed at which the test can be performed has a direct impact on testing throughput. In general, a trade-off can be made between test time and measurement variance. High variance negatively impacts yield, and attempts to improve variability usually lead to slower measurement times. What follows is a description of the theory behind "confidence interval testing." This concept can be used to adaptively adjust measurement time in order to maintain a desired confidence factor in the measured results. When confidence interval testing is used, test times will be optimized and rework will be minimized, allowing the cellular-telephone manufacturer to minimize capital equipment cost and operating expenses.

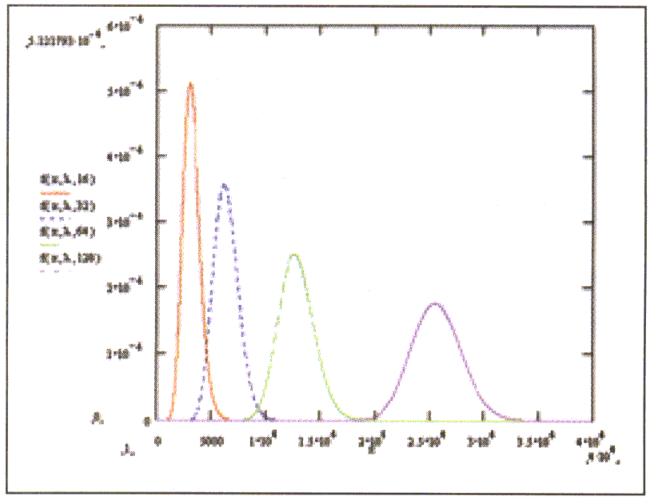
The performance of a digital receiver is typically specified in terms of its bit-error rate (BER), word-error rate (WER), or FER. The error-rate performance is typically characterized with respect to signal-to-noise ratio (SNR) commonly expressed in terms of energy per bit to noise power spectral density (E_b/N_o). FER testing is often performed instead of BER testing because knowledge of the actual pattern is not necessary at the receiver. Frame errors are determined by recalculating the cyclic redundancy code (CRC) and comparing it to that received from the transmitter. The FER is more encompassing than



1. This plot shows the Poisson probability distribution for p = 0.005 and n = 8368 frames.



2. These plots show the m-Erlang distribution for λ = 0.005 as a function of trial size, for m = 1, 2, and 3.



3. These plots show the m-Erlang distribution for λ = 0.005 as a function of trial size, for m = 16, 32, 64, and 128.

BER since it exercises a receiver's error-correction algorithms.

Of course, the goal of any of these receiver tests is to determine, with a particular degree of confidence, that a unit under test exceeds a specified performance criterion. If the measured error-rate performance exceeds the specified limits by a sufficient amount, testing can be terminated early. But if the measured error rate hovers near the specified limit, long test times may be needed to conclusively determine that the unit under test passes or fails its performance test.

If the test strategy can involve fixing the level of confidence, then test time may be adapted, based on the observed FER, so that a minimal number of frames are tested. The overlying assumption of this confidence interval testing is that frame errors follow a form of Poisson distribution. The distribution function can be developed by considering the sum of inter-arrival times of Poisson events with an exponential distribution. With knowledge of this distribution function, test results from short observation times will be able to be used to determine the actual error rate with a particular level of confidence.

What follows are the underlying mathematical techniques used for this type of testing. Examples are presented using the results of the derivation. To find results quickly, a methodology will be developed to determine the results in terms of an χ^2 distribution. This is consistent with the technique provided in the code-division-multiple-access (CDMA) standard (IS-95) and facilitates finding the results using easy-to-find lookup tables. Several tables of χ^2 distributions are provided in the Appendix.

The CDMA international standard includes a listing of fixed-length test sizes to serve as a guide to determine the number of frames that must be tested to achieve a 95-percent confidence level for various error rates (Table 1). When λ_{lim} = 0.005 (0.5 percent), a 0.5-percent error rate yields, on average, 41.84 frame errors for every 8368 frames tested. The process of testing frames can be considered a Bernoulli trial. Therefore, the probability distribution function (PDF) can be described by replacing p with λ_{lim} in eq. 1:

IF THE TEST STRATEGY CAN INVOLVE FIXING THE LEVEL OF CONFIDENCE, THEN TEST TIME MAY BE ADAPTED, BASED ON THE OBSERVED FER, SO THAT A MINIMAL NUMBER OF FRAMES ARE TESTED.

$$P[Errors = k] = \binom{n}{k} p^k (1-p)^{n-k}$$
$$= [n!/k!(n-k)!]p^k(1-p)^{n-k} \quad (1)$$

If the probability of an error is very small (p < 1), n tends to be large, then n! is troublesomely large. Under these circumstances, the Poisson distribution can be used to approximate the distribution of eq. 2:

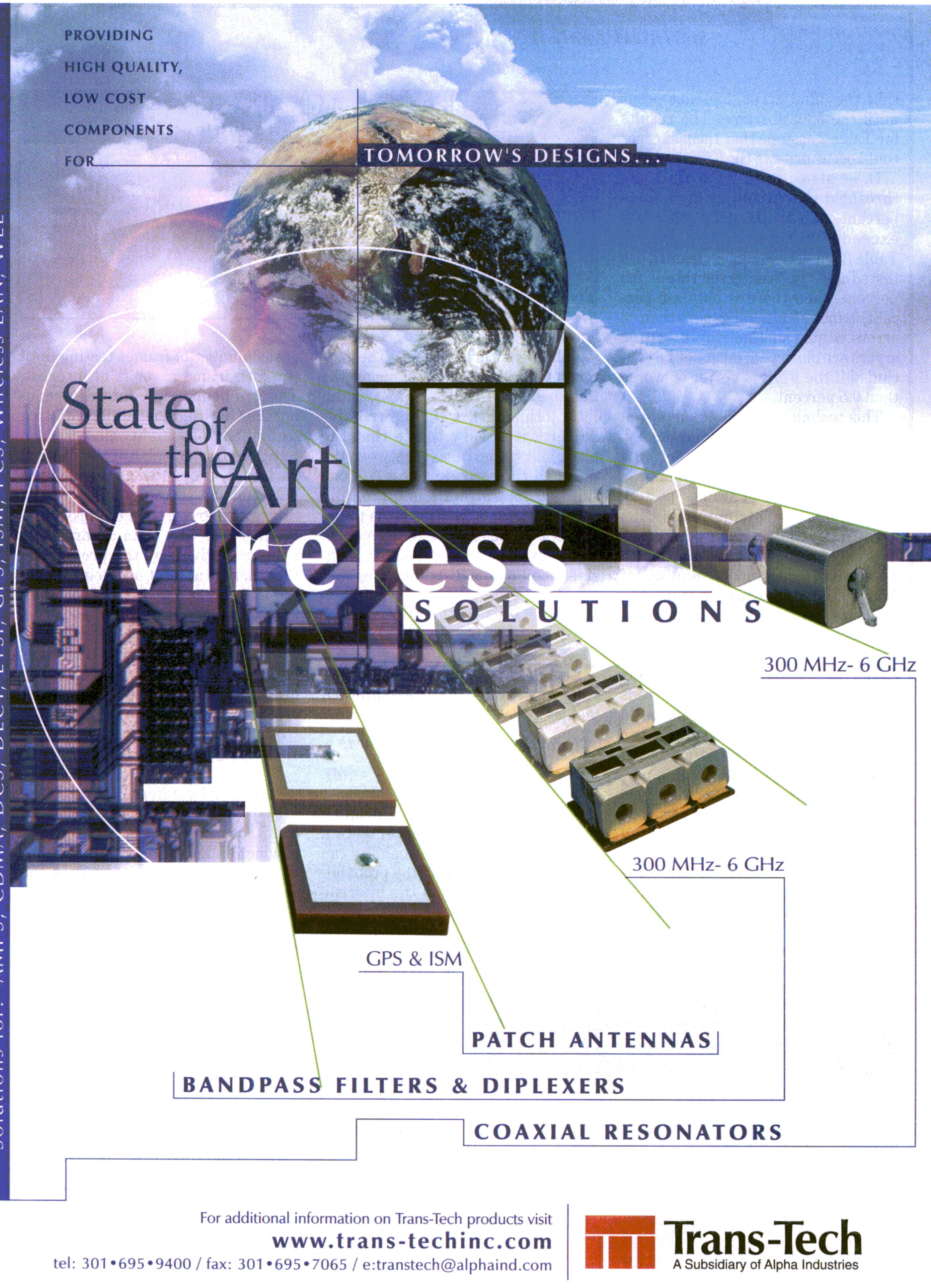
$$f(n,k,p) = [n!/k!(n-k)!]p^k(1-p)^{n-k}$$
$$\cong g(n,k,p) = e^{-np}[(np)^k/k!] \quad (2)$$

The Poisson distribution function is plotted in Fig. 1. To be 95-percent confident that the error rate is less than 0.5 percent, Table 1 states that the measured error rate must be less than or equal to 32. This is valid if the following holds true:

$$\sum_{k=0}^{K_{max}} P\{n_X = k\} < 1 - C \quad (3)$$

$$\sum_{k=0}^{K_{max}} [(np)^k/k!]e^{-np} < 1 - C \quad (4)$$

The data in Table 2 show that if 8368 frames are tested and the true error rate is 0.5 percent, then there is only a 5-percent chance of obtaining fewer than 32 errors. The apparent discrepancy between Table 1 stating K_{max} = 32 errors and this result of 31 errors arises from the details of the technique applied. The procedure involves testing frames until 32 errors are observed. This implies



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that the 32nd error does not occur until the 8368th frame. The results for Table 2 assume that the last error could occur prior to the 8368th frame.

It is also possible to find the threshold that would make one 95-percent confident that the error rate exceeds 0.5 percent. The data in Table 3 show that if 8368 frames are tested and the true error rate is 0.5 percent, then there is only a 5-percent chance of obtaining more than 53 errors. Stated another way, if 32 to 52 errors are observed, it cannot be concluded if the FER is better or worse than 0.5 percent.

This technique could be employed to set the test time dynamically. When a frame error is detected, a new distribution could be computed with the number of frames tested. The computed distribution is then used to determine if the confidence limits are satisfied. After observing a number of errors, the number of frames tested information is used to calculate the distribution of the number of errors. This approach, although yielding an answer, is a rather backward process. An alternative approach would be preferred.

A different approach would be to generate a function that describes the probability distribution of the number of frames tested until a particular number of errors occurs. This approach is more straightforward as the random variable is not tested against itself. Notice that up to this point the total number of successes have been considered in a series of trials without being concerned about the pattern of events. Consider the experiment of flipping a coin until a head (h) is observed and then observing the next three events. The pattern of the three flips following the first head will be {ttt}, {tth}, {thh}, {htt}, {hth}, {hht}, or {hhh}. From this pattern, it is easy to see that the probability of no tails (t) between the first and the second head is 0.5, one tail is 0.25, two tails is 0.125, and three tails is 0.0625. This leads to the concept of "arrival time of events" and is the basis of this section. Consider the Poisson distribution again:

$$f(n,k,p) = [n!/k!(n-k)!]p^k(1-p)^{n-k}$$

$$\cong g(n,k,p) = e^{-np}[(np)^k/k!] \quad (5)$$

Letting $p = \lambda_{lim}$ to maintain consistent notation with the standard yields:

$$P\{k \text{ frame errors in } n \text{ frames tested}\}$$

$$= e^{-\lambda_{lim}n}[(\lambda_{lim}n)^k/k!] \quad (6)$$

Next, consider the distance between frame error events in a Poisson process. This probability can be found by considering the probability of no events within the interval, hence:

P[t > 1] = P[no frame errors in n frames tested] = P{k = 0 in n frames tested}

P[no frame errors in n frames tested] = $e^{-\lambda\, lim}$

The inter-arrival distance distribution in a Poisson process is thus described by an exponential distribution. In this case, the distribution function of the number of frames testing until the mth frame error is of interest. This can be found by first determining the CDF and then differentiating to find the probability distribution function. In particular, if the random variable X is denoted as the distance between the starting time and the first frame error, then the probability that X is less than a particular value n is given by the probability that there is at least one point within the interval of interest. As the number of frames tested becomes large, the discrete index n can be replaced by x and integrals and derivatives become a good approximation to the discrete math, hence:

$$F_x(x) = P\{X \le x\} = P\{k \ge 1\}$$

$$= 1 - P\{k = 0\} = 1 - e^{-\lambda x} \quad (7)$$

It follows by differentiating that:

$$f_x(x) = [dF_x(x)/dx] = \lambda e^{-\lambda x} \quad (8)$$

which is an exponential distribution. Again, there is interest in the distribution of the number of frames until the mth error. The distribution can be found by considering the CDF:[1]

$$F_m(x) = 1 - \sum_{k=0}^{m-1} P\{m_x = k\}$$

$$= 1 - \sum_{k=0}^{m-1}[(\lambda x)^k/k!]e^{-\lambda x} \quad (9)$$

After differentiation, this becomes:

$$f_x(x) = [\lambda e^{-\lambda x}(\lambda x)^{m-1}/(m-1)!] \quad (10)$$

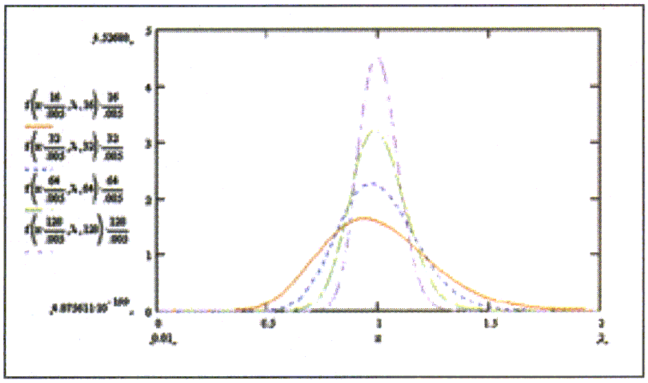
which is an m-Erlang distribution. This result may also be obtained by considering the characteristic equation. The characteristic equation of a single exponential random variable

Table 1: Fixed-length test sizes (trials for 95-percent confidence)

K_{max} (Errors)	λ_{lim} (Error-rate limit)			
	0.005	0.010	0.050	General
32	8368	4184	837	$41.84/\lambda_{lim}$
64	15540	7770	1554	$77.70/\lambda_{lim}$
128	29432	14716	2943	$147.16/\lambda_{lim}$
256	56575	28287	5657	$282.87/\lambda_{lim}$

Table 2: Lower values of the summation for various values of K_{max} for n = 8368 and p = 0.005

K_{max}	$\sum_{k=0}^{K_{max}}[(np)^k/k!]e^{-np}$
30	0.035
31	0.04996
32	0.070
33	0.095
34	0.126
35	0.164



4. These normalized m-Erlang distribution curves are plotted for m = 16, 32, 64, and 128.

Table 3: Upper values of the summation for various values of K_{max} for n = 8368 and p = 0.005

K_{max}	$\sum_{k=0}^{K_{max}} [(np)^k / k!]e^{-np}]$
50	0.907
51	0.929
52	0.946
53	0.960
54	0.971
55	0.979

is:

$$\Phi_x(\omega) = [\lambda / (\lambda - j\omega)] \quad (11)$$

The characteristic equation of the sum of n exponentially distributed random variables is:

$$\Phi_{S_n}(\omega) = [\lambda / (\lambda + j\omega)]^m \quad (12)$$

By studying published distribution tables, it can be shown that this is an m-Erlang distribution.[2]

Now that the mathematical tools have been developed, it is time to apply the results to a specific example. Consider the possibility of k frame errors out of n frames tested with a target FER of λ. It follows that the Poisson distribution describes the density function:

$$g(n,k,\lambda) = e^{-n\lambda}[(n\lambda)^k / k!] \quad (13)$$

It may be interesting to analyze a few simple cases. One trivial question is this: How likely is it that the first trial will be error free if the average error rate is known to be 0.5 percent? By setting λ = 0.005, k = 0, and n = 1, it can be found that there is a 99.5-percent chance of no error and 0.5-percent change of an error. If testing begins and an error is observed in the first frame, it is possible to be 99.5-percent confident that the long-term error rate exceeds the target. If, however, no error occurs, there can only be 0.5-percent confidence that the long-term error rate is within the target specification. If it is desirable to be 95-percent confident that the long-term error rate is less than 0.5 percent, then how many consecutive error-free frames must be tested? Setting k = 0 and solving for the equation just shown yields:

$$n = [ln(\alpha) / \lambda] \quad (14)$$

where:

α = the (1 –) confidence factor). Solving for this, it can be found that n = 599 error-free frames must pass before 95-percent confidence is achieved. In general, after a number of frames are tested, it is desirable to know with a level of confidence that less than a certain number of frames (K) should be in error. This can be found by performing the following test:

$$\alpha < \sum_{k=0}^{k} e^{-n\lambda}[(n\lambda)^k / k!] \quad (15)$$

Another approach is to identify the distribution of the distance between errors. Again, using a target error of 0.5 percent, the expected time between errors is 1/0.005 = 200 frames. The distribution of the distance between errors can be found by letting λ = 0.005 in the Poisson PDF:

$$f_x(x) = \lambda e^{-\lambda x} \quad (16)$$

The expected value can be found by evaluating the following expression:

$$E(f_x(x)) = \int_{-\infty}^{+\infty} x\lambda e^{-\lambda x} u(x)dn =$$

$$(\lambda e^{-\lambda x} / \lambda^2)(-\lambda x - 1)\Big|_0^{\infty} = \frac{1}{\lambda} \quad (17)$$

It was discovered earlier that the distribution of the distance between a particular number of frame errors is m-Erlang in nature. Several m-Erlang distributions are plotted in Fig. 2 for m = 1, 2, and 3. The m-Erlang distribution is plotted again in Fig. 3 for m = 16, 32, 64, and 128 for comparison. It appears that the distribution spreads as the number of errors increases. Remembering that the sample variance is inversely proportional to the number of samples when a normal population is sampled, this seems counter-intuitive. This apparent disparity is a result of the data not being displayed as a percentage. For comparison, a family of normalized curves is shown in Fig. 4. It is interesting to analyze the curve in Fig. 2. The solid curve represents the time between two adjacent errors. If the integral is evaluated:

$$\int_0^{z} f_x(x)dx = \int_0^{z} \lambda e^{-\lambda x}dx =$$

$$-e^{-\lambda x}\Big|_0^{z} = 1 - e^{-\lambda z} \quad (18)$$

it is possible find the 95-percent confidence limit, $0.95 = 1 - e^{-\lambda z}$, or ln(0.05) = –λz for λ = 0.005, then Z = 599 frames. This means that 599 frames must pass prior to a frame error. This corresponds to an average FER of 0.17 percent, which is unlikely. If a unit under test had this much margin, it is either over-designed or the target specifications may need to be tightened to gain a competitive advantage. The more likely scenario is that another frame error would occur prior to measuring 599 frames. If this were to happen, then the distribution would follow the short dashed line. The probability density for two errors is given by evaluating:

Table 4: Lookup values of χ^2 distributions

	$F\chi^2$								
n	**0.005**	**0.010**	**0.050**	**0.100**	**0.500**	**0.900**	**0.950**	**0.990**	**0.995**
32	15.1	16.4	20.1	22.3	31.3	42.6	46.2	53.5	56.3
64	38.6	40.6	46.6	50.0	63.3	78.9	83.7	93.2	96.9
128	90.5	93.7	102.9	108.0	127.3	148.9	155.4	168.1	173.0
256	201.5	206.3	220.0	227.5	255.3	285.4	294.3	311.6	318.0
512	433.3	440.5	460.5	471.4	511.3	553.4	565.7	589.4	598.2
1024	911.2	921.7	950.7	966.5	1023	1082	1100	1132	1144

$$f_x(x) = [\lambda e^{-\lambda x}(\lambda x)^{m-1}/(m-1)!] \qquad (19)$$

with m = 2, which yields:

$$f_x(x) = \lambda^2 x e^{-\lambda x} \qquad (20)$$

Using the procedure described previously to find the 95-percent confidence point, it follows that:

$$\int_0^z f_x(x)dx = \lambda^2 \int_0^z xe^{-\lambda x}dx = \lambda^2 (e^{-\lambda x}/\lambda^2)(\lambda x - 1)\Big|_0^z \qquad (21)$$

This integral is more difficult to evaluate than the case when m = 0. For this reason, when m > 0, it is convenient to use the techniques described in the Appendix so that χ^2 tables can be used to find the results. Following this technique, the value for $\alpha = 0.05$ with 2 degrees of freedom is 9.488. Dividing this by 2λ where $\lambda = 0.005$ yields 949 frames. It follows directly that one error may occur at frame 950 and one more may occur somewhere in the middle of the test. This corresponds to an average error rate of 0.2 percent. Since there is no knowledge of the errors prior to starting the test, it can be assumed as the worst-case scenario that there was one error just prior to starting the test.

The failing case can be analyzed in a similar manner. Consider how close two frame errors would have to occur to obtain 95-percent confidence of a failure. Evaluating the m = 0 integral reveals that if only 10 passing frames fall between 2 failing frames, then it is possible to be 95-percent confident that the device under test has failed to meet the target specification. It would probably be overly pessimistic when observing the first error to assume that an error occurred in the frame prior to the start of the test. If these curves are normalized, it is easier to see that the distribution as a percentage tightens as the number of frames tested increases.

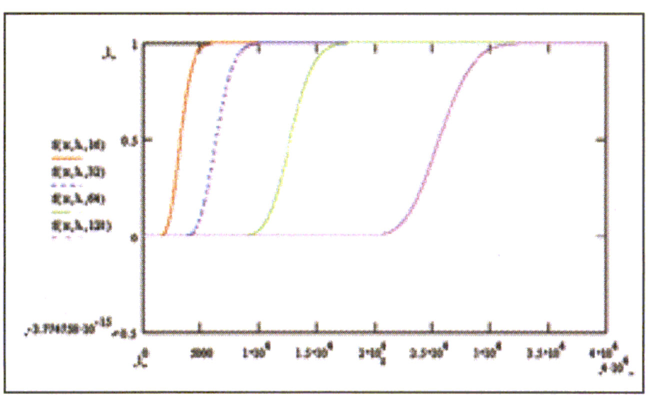
Several strategies can be used to implement confidence interval testing. One strategy is to make use of the CDF derived previously:

$$F_n(x) = 1 - \sum_{k=0}^{n-1} P\{n_x = k\} = 1 - \sum_{k=0}^{n-1} [(\lambda x)^k / k!]e^{-\lambda x} \qquad (22)$$

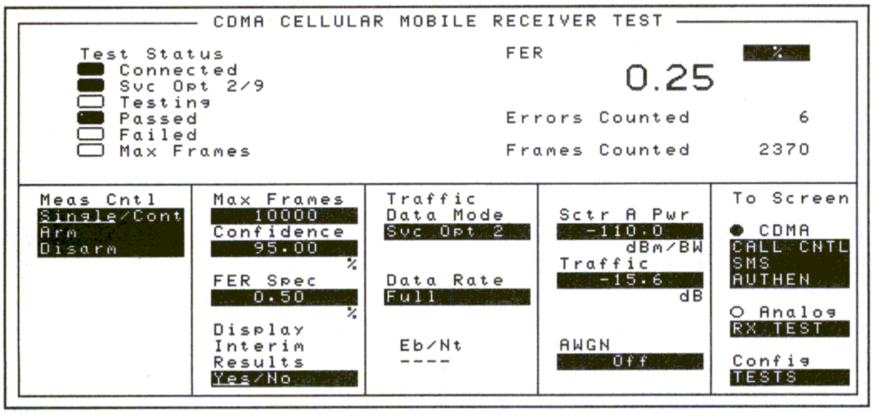
To terminate a test early with a passing grade, tally the number of frame errors (n) and the number of frames tested (x). Each time a frame error occurs, a new CDF curve is computed. When the CDF exceeds the target confidence factor, the test may be terminated. The CDF for several levels of frame errors is plotted in Fig. 5.

To terminate a test early with a failure, tally the frame errors (n) and the number of frames tested. The x-axis begins when the first frame is observed. A new m-Erlang CDF curve is plotted each time a frame error occurs with (m) equal to the number of errors following the first error. When a frame error occurs, the CDF curve is checked to determine if the CDF is less than (1 – confidence). If the value is less than (1 – confidence), then it can be said that the frequency of errors occurs at too high of a rate to be confident that the test passes the target FER specifications.

It is apparent that a solid knowledge of probability and statistics is necessary to develop confidence interval testing algorithms. The algorithms must be documented and maintained throughout the life of the production test system. Fortunately, test equipment is available that eliminates this burden by providing confidence interval testing as an integral part of the measurement capability. Figure 6 shows how the task of confidence interval testing can be easily managed on the receiver test screen of the 8924C CDMA mobile-station test set from Agilent Technologies (Palo Alto, CA), a complete measurement system to verify the performance of dual-mode CDMA mobile phones. On this screen, the user enters the maximum number of frames to test, confidence level, and specified FER. The measurement terminates early when the unit under test passes or fails with the confidence factor specified. The screen shows that this particular unit under test passed a 0.5-percent FER specification with 95-percent confidence after 2370 frames were tested. This corresponds to a measured FER of 0.25 percent. The results can be verified by following the procedure outlined in the Appendix. Following this



5. These cumulative m-Erlang distribution curves are plotted for m = 16, 32, 64, and 128.



6. This screen from the Agilent Technologies' model 8924C CDMA cellular mobile-receiver test system shows how confidence interval testing can be performed routinely and automatically.

procedure, it can be found that the minimum number of frames to satisfy the confidence criteria is 2103:

$$N_{max} = \frac{\chi^2(0.95, 2\times 6)}{2\times 0.005} = \frac{\chi(95,12)}{0.01} = \frac{28.3}{0.01} = 2103 \qquad (23)$$

When confidence interval testing is employed, test times can be minimized in order to maximum throughput of a production-line test system. Two techniques were presented to determine the number of frames needed. The first technique was a direct result of applying the Poisson distribution function. The second approach was derived using Poisson inter-arrival times. The later method used an m-Erlang distribution. The m-Erlang distribution was then related to the χ^2 distribution so that lookup tables can be referenced (as presented in the Appendix).

The 8924C CDMA mobile-station test set performs call-processing verification according to a variety of protocols, including IS-95, IS-95A, and Advanced Mobile Phone Service (AMPS). The test system is designed to evaluate dual-mode CDMA mobile telephones operating from 500 to 1000 MHz, but can be equipped with a PCS interface for testing handsets from 1700 to 2000 MHz.

The 8924C is essentially a calibrated, high-performance CDMA base station. The test system can verify not only the parametric characteristics of CDMA telephones, but also the functional aspects of telephone performance. The system incorporates an accurate CDMA source with 1-Hz frequency resolution for receiv-

er testing (to check the channel's signals and noise) and a sensitive receiver for transmitter testing (for evaluating a channel's signal power, adjacent-channels' power, rho, frequnecy error, modulation phase and amplitude errors, as well as carrier feedthrough).●●

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2. Alberto Leon-Garcia, *Probability and Random Processes for Electrical Engineering*, 2nd Ed., Addison Wesley, Cambridge, MA, Table 3.2.
3. Kishor, Shridhardhai, and Trivedi, *Probability & Statistics With Reliability, Queuing and Computer Science Applications*, Prentice-Hall, Saddle River, NJ, Problem 3-1 and Section 10.2.3.3.
4. Athanasios Papoulis, *Probability & Statistics*, Prentice-Hall, Saddle River, NJ.

APPENDIX: Transforming m-Erlang distributions to χ^2 distributions

A χ^2 distribution is also a special case of the Γ distribution and tables (Table 4) are readily available in the back of many probability and statistics books. It can be shown[3] that if X has an m-Erlang distribution with parameter λ, then $Y = 2\lambda X$ has a χ^2 distribution with 2k degrees of freedom. It follows from the linear transformation of random variables that:

$$F_Y(y) = P\{Y \le y\} =$$

$$P\{X \le x\} = F_x\left(\frac{y}{2\lambda}\right) \qquad (24)$$

$$f_Y(y) = \frac{d}{dy}F_x\left(\frac{y}{2\lambda}\right) = \frac{1}{2\lambda}f_x\left(\frac{y}{2\lambda}\right) \quad (25)$$

and it follows that:

$$f_Y(y) = \frac{\lambda^n\left(\frac{1}{2\lambda}\right)^{m-1} \times (y)^{m-1}e^{-y/2}}{2\lambda(m-1)!} \quad (26)$$

If $\lambda = 0.5$ and $m = k/2$, a positive integer, then using the relationship:

$$\Gamma(n) = (n-1)!$$

$$\textit{if } n \textit{ is an integer, } n > 0 \qquad (27)$$

makes it possible to write the m-Erlang distribution as:

$$f_Y(y) = \frac{x^{(k-2)/2}e^{-y/2}}{2^{k/2}\Gamma(k/2)} \qquad (28)$$

where:

$k = 2m$. This is a χ^2 random variable with k degrees of freedom. This can also be derived by manipulating the characteristic functions of the two probability densities. The characteristic function of the m-Erlang and χ^2 distribution are, respectively:

$$\Phi_x(\omega) = \left(\frac{\lambda}{\lambda - j\omega}\right)^m$$

$$\Phi_Y(\omega) = \left(\frac{1}{1-j2\omega}\right)^{k/2} \qquad (29)$$

Once again, let Y = 2X. Then, from the well-known coordinate scaling of Fourier transforms:

$$F.T.\{f(\alpha t)\} = \frac{1}{|\alpha|}F\left(\frac{\omega}{\alpha}\right) \qquad (30)$$

Applying this to the two density functions, it is possible to show that:

$$f_Y(y) = \frac{1}{2\lambda}f_x\left(\frac{y}{2\lambda}\right) \qquad (31)$$

Then:

$$F.T.\{f_Y(y)\} = \frac{1}{2\lambda}F.T\left\{f_x\left(\frac{y}{2\lambda}\right)\right\} \quad (32)$$

Taking the Fourier transform of both sides yields:

$$\Phi_Y(\omega) = \frac{1}{2\lambda}|2\lambda|\Phi_X(2\lambda\omega) \qquad (33)$$

Revealing that:

$$\Phi_X(\bullet) = [\lambda/(\lambda - j\bullet)]^m \qquad (34)$$

Then the function becomes:

$$\Phi_Y(\omega) = \frac{1}{2\lambda}|2\lambda|\Phi_X(2\lambda\omega) \qquad (35)$$

and

$$\Phi_Y(\omega) =$$

$$(1/2\lambda)\,|2\lambda|\,[\lambda/(\lambda - j2\lambda\omega)]^m$$

$$= [1/(1-j2\omega)]^m \qquad (36)$$

which is an m-Erlang random variable.

The χ^2 distribution:

$$F(x^2) = \int_0^{x^2} \frac{1}{2^{\frac{n}{2}}\Gamma\left(\frac{n}{2}\right)} x^{\frac{n-2}{2}} e^{-\frac{x}{2}} dx \quad (37)$$

can be transformed by describing the gamma function[4] by:

$$\Gamma(\alpha) = \int_0^{\infty} y^{a-1}e^{-y}dy \qquad (38)$$

by replacing α by $\alpha + 1$ and integrating by parts, to obtain:

$$\Gamma(\alpha + 1) =$$

$$\alpha\int_0^{\infty} y^{a-1}e^{-y}dy = \alpha\Gamma(\alpha) \qquad (39)$$

Then, if $\alpha = n$, an integer, and knowing that $\Gamma(1) = 1$, it means that

$$\Gamma(n+1) = n\Gamma(n) =$$

$$n(n-1)\cdots\Gamma(n) = n! \qquad (40)$$

and

$$N_{max} = \frac{[\chi^2(1-C, 2K_{max})]}{2\lambda_{lim}} \qquad (41)$$

For example, if $\lambda_{lim} = 0.005$ and 32 errors are observed, what is the minimum number of frames that must be tested for a 95-percent confidence so that the unit under test meets its specifications?

The answer can be found by substituting these values into eq. 41:

$N_{max} = \chi^2(1 - 0.95, 2 \times 32)/(2 \times 0.005) = \chi(0.005, 64)/0.01 = 46.6/0.01 = 4660$

$N_{max} = \{\chi^2(1 - 0.95, 2 \times 32)/(2 \times 0.005) = \chi(95, 64)/0.01 = 83.7/0.01 = 8370$

If 32 errors occur prior to frame number 4660, then it is possible to say with 95-percent confidence that the unit does not meet the specification. If 32 errors occur, then the test can be stopped at frame number 8370 with 95-percent confidence so that the unit's performance exceeds the specifications.

Thick-Film Fabrication Yields Thin-Film Performance

Advanced thick-film fabrication techniques yield good line resolution and low loss.

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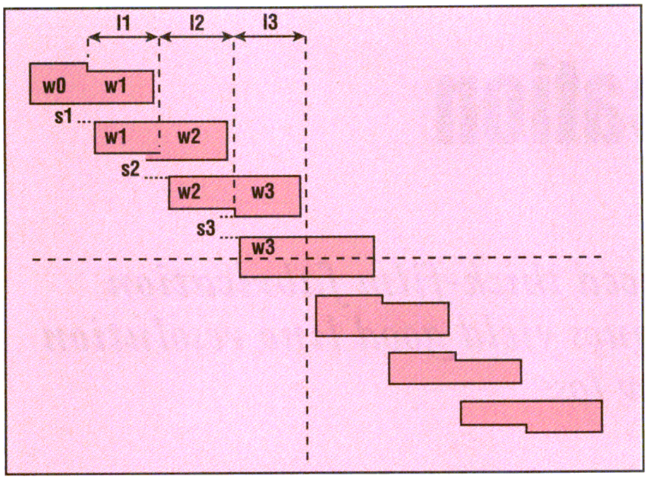
INCREASED use of microwave hybrid circuits for wireless communication systems has led to a search for new technologies offering advanced circuit functions at low cost. Characteristics such as high electrical conductivity, fine-line and space resolution, well-defined conductor edges, nearly vertical walls, and smooth upper surfaces are essential for achieving low losses in microwave structures at frequencies above 1 GHz.[1,2,3] Until now, thin-film technology has dominated the microwave market because traditional thick-film technology yielded poor line resolution and high losses. But rapid development in novel thick-film materials and advanced thick-film circuit-patterning techniques has brought improvements that allow current thick-film technology to reach beyond its previous limitations. Advanced thick-film techniques such as photolithography over fired layer, and photoimaging over photosensitive dried layer, yield conductor strips with resolution and edge definition comparable to thin-film technology. Thick-film technology allows designers to combine microwave and digital functions on common high-thermal-conductivity alumina substrates and to incorporate capacitors and laser-trimmable thick-film resistors into the main microwave structures. The self-smoothing tendency of thick films at the substrate interface permits the use of less-expensive, 96-percent alumina substrates. Additionally, thick-film technology provides significant advantages such as low cost and feasibility for mass production.

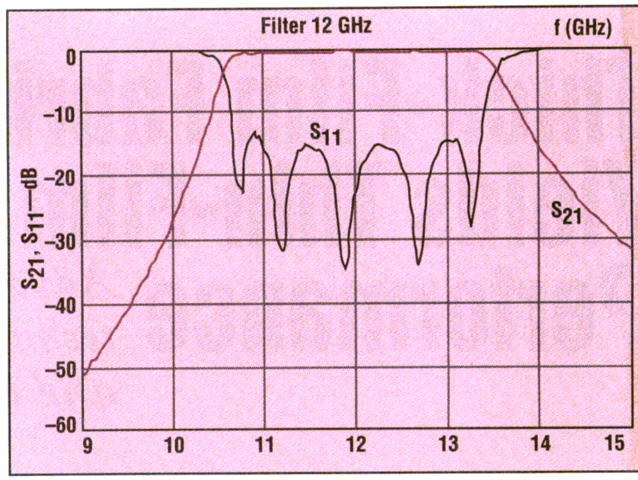
Table 1: Geometrical dimensions of filters

	F061	F122	F141
W_0	0.635	0.614	0.615
W_1	0.479	0.338	0.511
W_2	0.738	0.594	0.612
W_3	0.645	0.568	0.536
S_1	0.140	0.100	0.070
S_2	0.515	0.250	0.400
S_3	0.808	0.410	0.575
1_1	4.787	2.348	1.909
1_2	4.654	2.238	1.872
1_3	4.693	2.255	1.892

This article discusses the application of advanced thick-film techniques—such as etching of fired thick films and photoimaging of photosensitive thick films—to manufacture microwave hybrid circuits that operate from 6 to 14 GHz. The authors constructed models of bandpass, edge-coupled microstrip filters with center frequencies of 6, 12, and 14 GHz, using bandpass microwave filters and 50-Ω microstrip circuits as test patterns. For metallization, they used novel thick-film materials, including etchable gold (Au) QG150 and photosensitive Fodel Ag 4065. Substrates were formed from co-



1. This chart shows the basic layout of the microstrip, edge-coupled filter structure.



2. This graph shows the computed transmission and reflection characteristics of Filter F122 (12 GHz).

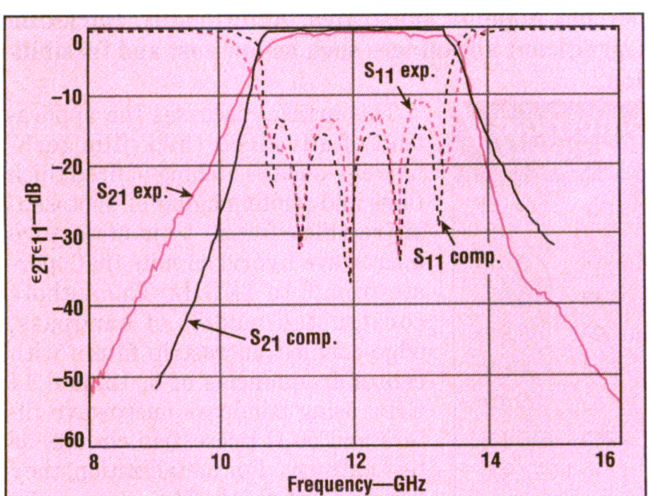
fired alumina containing 96-percent Al_2O_3, with a thickness of 0.635 mm and dielectric constant $\epsilon_r = 9.5$.

Insertion and return losses of filters were measured using vector network analysis, and then compared to the computed characteristics. Additionally, attenuation of microstrips was measured to 20 GHz. Physical analyses such as cross-sectioning, roughness, and edge definition of etched and photoimaged thick films were performed with the requirements of microwave applications in mind.

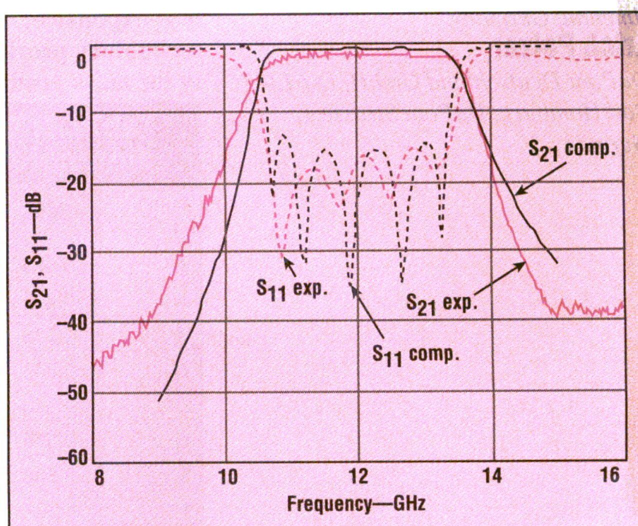
Figure 1 shows a basic layout of the filter models, while Table 1 contains detailed geometrical dimensions for each filter. Transmission and reflection characteristics were computed using the Touchstone program. On the basis of these characteristics, electrical parameters of filters—including the center frequency, fractional bandwidth, attenuation, and reflection in the passband, ripple in the passband, and shape factor—were determined and collected, as shown in Table 2. Figure 2 presents an example of the computed transmission and reflection characteristics for a filter with a center frequency of 12 GHz. A microstrip of characteristic impedance equal to 50 Ω, designed as a straight line 59.8 mm long and 0.635 mm wide, was deposited on the 96-percent Al_2O_3 substrate.

Thick-film conductor layers for microwave applications should have high conductivity, homogeneity, fine-line resolution, fine-edge definition, rectangular cross-sections, smooth upper and lower surfaces, and a thickness of several skin depths.

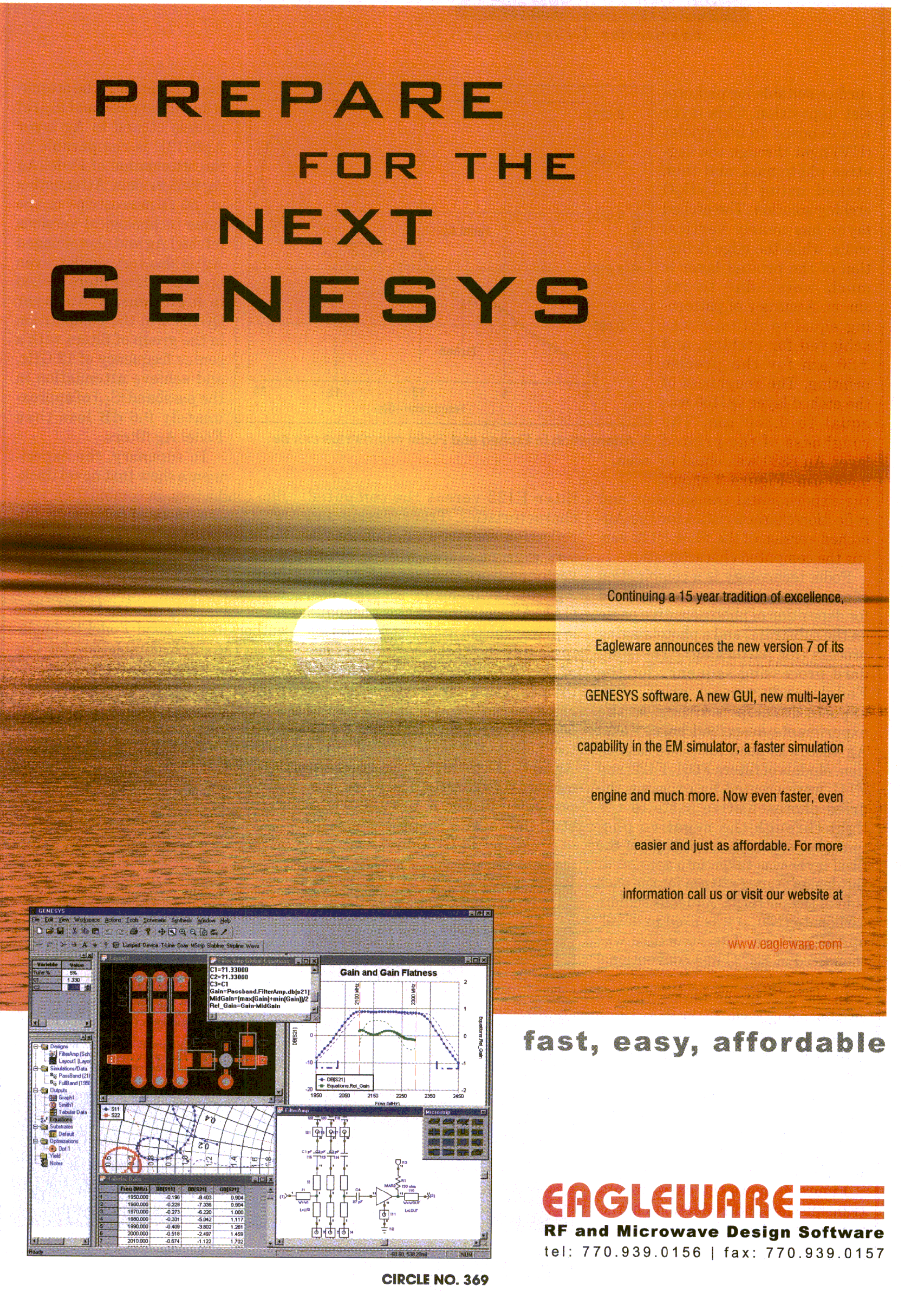
Etching of a fired thick-film layer combines screen printing for paste deposition on the substrate with photolithography for patterning. In the experiments described here, high-density Au layer QG150 was etched to obtain an F122 filter structure. The QG150 paste is one of a new generation of thick-film conductors based on Au powders containing very-small, tightly distributed spherical particles of submicron dimensions and a new vehicle system.[5] The fired thick-film layer that is blank-screen printed using this paste is characterized by high conductivity, high density, and a smooth upper



3. This graph shows the experimental and computed characteristics of the etched high-density Au QG150 filter F122 (12 GHz, 25-percent bandwidth).



4. Graph showing the experimental and computed characteristics of the Ag4065 Fodel version of filter F122.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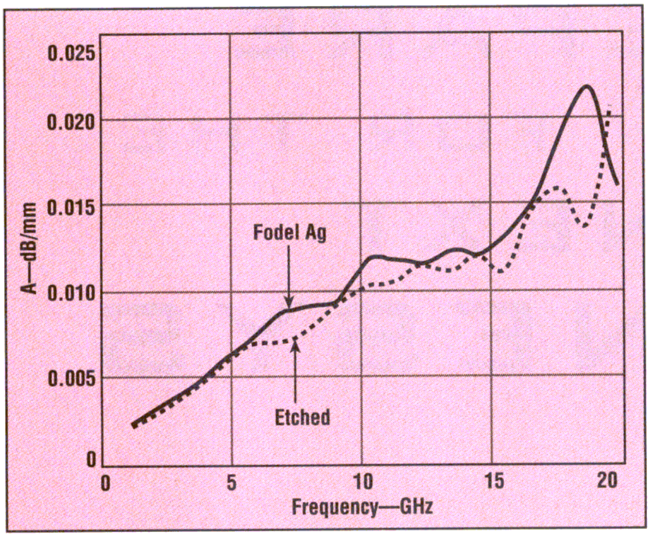
surface suitable for photoresist deposition. This layer was exposed to ultraviolet (UV) light through the negative photomask and then etched using $KJ/J_2/H_2O$ etching solution. The etched layer has nearly vertical walls, while the edge definition of the printed layer is much worse due to ink slump. Accuracy of patterning equal to ±10 μm was achieved for etching, and ±20 μm for the precise printing. The roughness of the etched layer QG150 was equal to 0.350 μm. The roughness of the printed layer Au 8880 was equal to 0.500 μm. Figure 3 shows the experimental transmission and reflection characteristics for the Au-etched version of the filter F122 versus the computed characteristics.



5. Attenuation in Etched and Fodel microstrips can be seen.

Fodel technology is a combination of conventional screen printing (used for deposition of photosensitive paste on the substrate) and a photoimaging process (for patterning).[6] The standard processing sequence for the Fodel conductor is: print, dry, expose, develop, and fire. In the experiments described here, Fodel Ag 4065 was used for paste deposition. Models of filters F061, F122, and F141 were created by exposing the dried photosensitive paste to UV light through the negative photomasks, where the shrinkage of the fired layer was taken into account at the design stage (+20 μm for the each side of the pattern). Aqueous sodium carbonate (NaC_3) was used to develop the pattern. The structure was then fired at 850°C in a conventional thick-film furnace.

In the part of the filter structure created using Fodel technology there were nearly vertical walls and a characteristic edge curl, and the gap s_1 was equal to 248 μm. The roughness of the fired Fodel layer was equal to 0.270 μm. Accuracy equaling ±10 μm of patterning was achieved. The shrinkage of the fired Fodel layer was observed at the level of 30 μm.

Figure 4 shows the experimental transmission and reflection characteristics for the Fodel version of the filter F122 versus the computed characteristics. Transmission and reflection characteristics of the filters were measured with an 8020C vector network analyzer (VNA) from Agilent Technologies (Santa Rosa, CA). Attenuation in the passband $|S_{21}|$ was chosen as the basic criterion to classify the technological versions of the filter. To extend the area of investigation, microstrips with a characteristic impedance equal to 50 Ω were fabricated using Fodel technology and etching of the fired Ag layer. Then their attenuation was measured to 20 GHz. The results are presented in Fig. 5. For the microstrip filters with center frequencies of 6, 12, and 14 GHz investigated in the experiments, the attenuation in the passband $|S_{21}|$ of models etched in Ag layer Ag9912F is comparable to the attenuation of Fodel Ag Ag4065 models. Attenuation of 50-Ω microstrips in the same technological versions (etched Ag and photoimaged Ag) is also very similar, even up to 20 GHz. Filters etched in high-density Au layer QG150 win the competition in the group of filters with a center frequency of 12 GHz, and achieve attenuation in the passband $|S_{21}|$ of approximately 0.6 dB less than Fodel Ag filters.

Table 2: Filter parameters determined on the basis of computed characteristics

Filter	F061	F122	F141
f_o(GHz)	6 GHz	12 GHz	14 GHz
3-dB bandwidth (percentage)	13	25	22
$\|S_{21}\|$ passband (dB)	1.00	0.56	0.65
$\|S_{11}\|$ passband (dB)	>25	>13.59	>12.8
Ripple (dB)	0.20	0.24	0.38
Shape factor	1.75	1.67	1.68

In summary, the experiments show that novel thick-film conductors, in combination with advanced methods of patterning, fulfill microwave requirements. The advanced thick-film techniques used here are capable of meeting the circuit-fabrication requirements for microwave structures to 14 GHz. The authors recommend Fodel technology as the most suitable one for this frequency range. Based on the comparable results, Fodel is less expensive, more user-friendly, and safer for the environment than etching. Etching of high-density Au is recommended for special, very-precise microwave applications. ••

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Studying Biomedical Issues Of High-Frequency Radiation

A review of electromagnetic radiation and exposure levels can improve one's understanding of RF/microwave safety and exposure limits.

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STORIES about potential health risks from (and occasionally about medical applications of) electromagnetic (EM) fields can be found everywhere from technical journals to supermarket tabloids. In the 1970s and 1980s, microwave radar and communication antennas received a lot of alarming press and more recently, the EM fields from cellular telephones have come under heavy scrutiny.[1] This primer is intended to provide a basic framework for tackling bioelectromagnetic issues. It is by no means an all-encompassing study of the biomedical effects of EM radiation, but is meant to include only those ideas which may one day be useful for performing quick calculations of EM radiation and safety. The interested reader can, of course, consult the references for additional details or data as there is no shortage of specialized material (see, for example, references 2 and 3).

The EM spectrum can be approximately divided into two broad classes: ionizing radiation [ultraviolet (UV) and higher frequencies including X-rays and gamma rays] and non-ionizing radiation [visible light and lower frequencies including infrared (IR), microwaves, and power frequencies]. This division occurs because EM energy of a particular wavelength, λ, (or equivalent frequency, $f = c/\lambda$, where c = the speed of light), is associated with a specific photon energy, E, given by:

$$E \text{ (in electron volts)} = 1.24 \times 10^{-6}/\lambda \text{ (in meters)}$$

If the photon energy exceeds approximately 10 eV (corresponding to a wavelength of 0.124 μm which falls in the UV region), the photon can ionize tissues (i.e., break molecular bonds) without (necessarily) producing any heating. This ionization, particularly if it involves DNA molecules (which store genetic information and are located in the nuclei of tissue cells), can result in serious irreversible damage. In contrast, the energy of a 1-GHz photon is a mere fraction of the thermal kinetic energy of a tissue molecule, preventing it from breaking even the weakest bonds.[4] Of course, this does not eliminate the possibility that low-intensity non-ionizing radiation can somehow directly alter tissue molecules, but neither does it suggest a clear-cut interaction mechanism along the lines of X-rays.[5]

It is natural to wonder what happens as one increases the intensity of non-ionizing waves. The photon energy is, of course, independent of the intensity. However, as the power density (W/m^2) is increased (by increasing the number of photons incident per unit area per unit time), a non-ionizing wave (such as microwave energy) can cause significant heating (the familiar microwave-oven situation), which can, in turn, damage tissues. The main point to remember is that in terms of these thermal effects, high-intensity microwaves are not fundamentally different from other more conventional sources of heat. For example, therapeutic hyperthermia (controlled heating in a hospital setting) can be produced by a hot water bath, ultrasound, or RF/microwave energy.

It should be noted that people are surrounded by natural sources of EM radiation. The power density from the sun is 100 mW/cm^2 on a clear day,[6] but in the microwave range, it is approximately 90 dB below[7] the current 10-mW/cm^2 safety limit (for X-band and above) prescribed by the current IEEE standard C95.1-1991. In addition, the human body emits (as part of the tail of its black-body radiation) microwave radiation at a level of approximately:

0.3 μW/cm^2.[7]

There are many man-made sources of EM exposure.[8-10] In the RF/microwave frequency range, the incidental exposure (from telecommunications, broadcasting, and radar equipment) usually occurs in a far-field situation. Therefore, the incident power density is the parameter commonly measured. According to a survey conducted by the Environmental Protection Agency (EPA), the median exposure level in an urban environment is estimated to be 0.005 μW/cm^2. The typical exposure level to police radar is approximately 1 μW/cm^2. A 3-W mobile-telephone antenna produces local power densities that are approximately 0.1 to 0.3 mW/cm^2 inside a car.

To understand the biomedical effects of EM radiation, it is helpful to know the EM properties of biological materials.[11,12] The magnetic permeability, μ, is still μ_0 for a vacuum. At 1 GHz, typical values for muscle tissue (high water content) are approximately 50 for the relative permittivity or dielectric constant, ϵ_r, and approximately 1 S/m for the conductivity, σ. These values imply a loss tangent of approximately 0.4. (Low-water-content fat and bones have lower values of ϵr and σ.) Therefore, the overall body behaves similar to a lossy dielectric medium in the RF and microwave frequency range. For an incident plane wave, the reflection coefficient at the air-body interface is approximately 0.8 in the gigahertz range with a depth of penetration ("skin depth") of approximately 2 cm at the microwave-oven frequency 2.45 GHz.[12] The absorbed power density (W/m ^{3}) can be determined from the usual relationship of $0.5\ \sigma\|E_{\text{int}}\|^2$ and the corresponding SAR in W/kg is then $0.5\ (\sigma/\rho)\|E_{\text{int}}\|^2$, where ρ is the local density (kg/m^3).

The interaction of EM fields with biological organisms varies between animal species as well as between animals and humans.[13] For example, if resonant absorption of RF/microwave energy is of interest, it should be noted that the peak absorption for human beings is approximately 70 MHz, while for rats the resonant frequency is around 1 GHz.[5]

Epidemiology—the study of patterns and possible causes of diseases in human populations—has been used for statistical correlations between EM fields and cancer. Even when a "statistically significant" (95-percent confidence level) association is noted, supplemental data are generally needed from laboratory studies to conclude that a cause-effect relationship exists.[14]

There have been few epidemiological studies on the biological effects of microwave radiation. A large occupational study in the 1970s of US Navy enlisted men who were in technical job classifications found no consistent excess cancer mortality or acute morbidity that could be attributed to microwave (radar) occupational exposure. Another "cohort" study involved employees of the US embassy in Moscow who were exposed to very-low levels of microwave radiation (up to 18 μW/cm^2 during long periods (50 to 70 s) of irradiation of the embassy building. In this case, too, no excess of cancer cases or deaths (compared with cases among unexposed embassy personnel in other east European embassies) could be attributed to the microwave radiation.

IEEE standard C95.1-1991 provides recommendations on how to prevent harmful effects in human beings who are exposed to EM fields in the frequency range of 3 kHz to 300 GHz. At frequencies below 100 MHz, the maximum permissible exposure limits are specified separately in terms of electric (E) and magnetic (H) field intensities, while above 100 MHz, the standard specifies the limits on the incident power density (ranging from 0.2 mW/cm^2 at 100 MHz to 10 mW/cm^2 at 300 GHz for the general public in an uncontrolled environment).[9,15]

In the frequency range of 100 kHz to 6 GHz, which includes the resonance range for human beings of approximately 70 MHz, the C95.1 standard uses the SAR as the key parameter of concern and generally includes a safety factor of 10 or better in setting the maximum limits. Below 100 kHz, the data base on electrostimulation of biological tissue plays the dominant role and the primary parameter is the internal current density. The recommendations at 300 GHz are consistent with the existing safe exposure limits in the IR range (starting at 300 GHz).

The Federal Communications Commission (FCC) guidelines for human exposure to RF fields from cellular and personal-communications-services (PCS) radio transmitters are similar (though not identical) to the IEEE standard. For specific details, the reader is referred to the following FCC document at its website: www.fcc.gov/oet/rfsafety/. ••

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Chip Solutions Aim At Low-Cost WLANs

The need for mobility and flexibility in office environments fuels the growth of data wireless-network applications at 2.4 GHz.

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DAWN PRIOR
Editorial Assistant

WIRELESS local-area networks (WLANs) have been slow in gaining acceptance for a number of reasons. For one thing, they face competition from a well-established copper (Cu)-cable infrastructure that has been relatively low in cost and reliable in operation. The lack of a well-defined standard, until fairly recently, has also made potential users leery of incompatibility among different software and hardware systems. But with the general acceptance of the IEEE's 802.11 specification as an industry standard, the growth of WLANs should parallel that of other emerging wireless applications, including Bluetooth and HomeRF.

WLAN systems generally operate in the unlicensed 2.400-to-2.483-GHz industrial-scientific-medical (ISM) band, although some higher-speed systems, such as HiperLAN, operate in the 5-GHz band. The idea for setting aside spectrum for wireless data networking was initially promoted by Apple Computer and later taken up by additional computer, telephone, and private-branch-exchange (PBX) manufacturers.

A WLAN can serve as a stand-alone network or augment a conventional wired Ethernet system. The protocol layers in a wireless network include the physical (PHY) first layer. The PHY layer transmits bits over a Cu wire in a conventional Ethernet system and through the air in a WLAN system. The second layer is the data-link layer, with two sublayers—the medium-access-control (MAC) sublayer and the logical-link-control (LLC) sublayer. The third layer is the network layer. The fourth layer is the Transport layer, which is responsible for reliable transmission of data. The fifth layer is the Session layer. The sixth layer is the Presentation layer, and the seventh layer is the Application layer.

In order not to interfere with other users in the unlicensed ISM band, WLAN systems use either frequency-hopping or direct-sequence, spread-spectrum (DSSS) techniques. In a frequency-hopping system, portions of the information to be transmitted are sent in different frequency slots. The transmitter sends information on one channel for a fixed time and then hops to another channel. The receiver is synchronized with the transmitter and hops in the same sequence.

In a DSSS system, a data stream is multiplied by a spreading code at a bit or chip rate that is many times that of the transmitted data. The result is a spectrum spreading of the transmitted information, with very low spectral energy over any particular communications band. Due to this spectral spreading, there is little change of interfering with other occupants of the bandwidth. The DSSS signal is decorrelated with the identical spreading sequence at the input of the receiver, in order to recover the desired transmitted data.

Within each cell in a WLAN is an access point known as a mobile support station (MSS) or base station that is connected to some wired network. Mobile users are known as mobile hosts. Uplink communications occur from the mobile user to the MSS while downlink communications occur in the reverse direction.

Suppliers of WLAN systems have attempted to provide increasing data rates. Most baseline systems offer a data rate of 1 Mb/s (often with a fall-back rate of half that speed under noisy or multipath signal conditions). Newer systems, however, have pushed the data rate of WLAN systems to 10 Mb/s and beyond, with HiperLAN systems in Europe offering data rates to 24 Mb/s at transmission frequencies of 5.8 GHz.

The RangeLAN2 system from Proxim, Inc. (Sunnyvale, CA), for example, operates at 2.4 GHz with frequency-hopping, spread-spectrum (FHSS) technology. Built into computer PCMCIA cards, the WLAN system is noteworthy for its low current consumption of 150 mA or less and its ability to transmit with 100-mW power (from a +5-VDC supply) at 2.4 GHz. The operating range with a dipole antenna (with approximately 1-dB gain) is about 500 ft. in typical office environments and around 1000 ft. in open spaces. Using

a snap-on antenna (with 0-dB gain), the operating range is approximately 400 ft. in an office environment and about 700 ft. in an open space. Since their earliest products, Proxim has worked closely with integrated-circuit (IC) supplier Stanford Telecom (Sunnyvale, CA) to develop high-performance, cost-effective solutions for WLAN systems.

Another leading supplier of WLAN end products is Lucent Technologies, which is also a supplier of WLAN ICs, such as baseband controllers. Lucent's WaveLAN family includes a new WaveLAN Turbo system capable of providing 11-Mb/s operation while maintaining compatibility with earlier 2-Mb/s WaveLAN versions. The firm recently developed a technology called direct-sequence/pulse-position modulation (DS/PPM) which supports data rates of 10 Mb/s at 2.4 GHz.

In the latter part of 1999, Symbol Technologies (Bohemia, NY) and Compaq Computer announced a joint effort for Symbol Technologies to deliver networking solutions to complement Compaq's WLAN products. Under the agreement, Compaq will refer its business customers to Symbol Technologies for implementation and maintenance of IEEE 802.11 WLAN solutions tailored to specific customer needs, especially in the enterprise marketplace and including products sold by Compaq Computer for 11-Mb/s WLAN systems.

Although IC manufacturers were hesitant to support WLAN development some years ago (without a clear-cut standard), the emergence of the IEEE 802.11 standard has made the WLAN market appear more attractive. Numerous manufacturers offer individual ICs and/or chip sets for WLAN applications at 2.4 GHz, ranging from RF amplification and frequency translation to baseband controllers. The SX040 series of WavePlex programmable complementary-metal-oxide-semiconductor (CMOS) ICs from American Microsystems, Inc., (Pocatello, ID), for example, is designed for DSSS transceivers. They can send and receive data rates to 2 Mb/s with chipping rates near 64 MHz. The ICs include the SX041 transmitter, the SX042 receiver, the SX043 baseband controller, and the SX045 transceiver. In addition to AMI, Cylink offers a number of ICs for DSSS WLAN applications, with baseband modems, receivers, and transmitters for data rates from 160 kb/s to 6 Mb/s.

One of the best known WLAN chip sets, the PRISM collection from Intersil Corp. (Melbourne, FL), is now in its second generation as the PRISM II chip set. Employing advanced silicon-germanium (SiGe) process technology where appropriate, the PRISM II chip set offers a top data rate of 11 Mb/s compared to 2 Mb/s for standard 802.11 PRISM I operation and 4 Mb/s for nonstandard PRISM I operation. Complementary-code-keying (CCK) technology, which is the basis for the IEEE 802.11 high-rate (HR) draft standard, is one of three modulation schemes that PRISM II supports. The PRISM II chip set can "downshift" to lower data rates (1.0, 2.0, or 5.5 Mb/s) to maintain the integrity of the wireless link or operate with legacy 802.11-compliant systems operating at only 1 or 2 Mb/s.

The PRISM II chip set includes the HFA3983 power amplifier (PA) and detector, the HFA3683 RF-to-intermediate-frequency (IF) converter, the HFA3783 in-phase/quadrature (I/Q) modulator/demodulator and frequency synthesizer, the HFA3861 DSSS baseband processor, as well as the HFA3841 MAC. The SiGe process is used for the first three ICs. The HFA3861 features a rake receiver and provides an antenna diversity option which enables the baseband signal to sample the inputs of two separate antennas and select the one with the better signal. The HFA3861, which features variable data rates of 1.0, 2.0, 5.5, and 11 Mb/s, can serve as a full- or half-duplex packet baseband transceiver for other wireless applications.

Motorola (Phoenix, AZ) offers a wide range of monolithic devices for 2.4-GHz WLANs, including the MC12179 phase-locked-loop (PLL) single frequency synthesizer IC and the MC12210 low-voltage 2.5-GHz frequency synthesizer/prescaler IC. The PCNET-Mobile controller from Advanced Micro Devices is a single-

chip MAC that implements the 802.11 WLAN protocol and Xircom's Netwave protocol. Digital Ocean has also developed an MAC chip known as the SETH chip that implements the 802.11 WLAN protocol.

Rockwell's Wireless Communications Division supplies numerous chips in support of DSSS networks and cordless telephones at 900 MHz, including the RDSSS9M baseband-device set, which includes a voice codec and an application-specific IC (ASIC) with onboard DSSS modem, an ADPCM engine, a controller, random-access memory (RAM), and read-only memory (ROM). The company also offers the R900DC direct-conversion transceiver module [optimized for time-division-duplex (TDD) operation] which provides a connection to the baseband modem and controller for receive (Rx) and transmit (Tx) data and control.

Celeritek, Inc. (Santa Clara, CA) supplies several GaAs ICs suitable for WLAN and general use from 2.4 to 2.5 GHz. The CMM2321 PA, for example, provides +20.5-dBm output power at 1-dB compression from 2.4 to 2.5 GHz with 20-dB gain. It draws 675 mA from a +4.8-VDC supply. The firm's model CCV2501 is an integrated frequency converter with an RF range of 2.4 to 2.5 GHz and an intermediate-frequency (IF) range of 100 to 500 MHz. It features 5-dB noise figure and 6-dB conversion gain while drawing 50 mA from a +5-VDC supply. The company's CAS2402 is a GaAs monolithic microwave IC (MMIC) that combines a 2.3-to-2.5-GHz amplifier with 18-dB gain and +23-dBm output power and a high-speed switch.

GEC Plessey Semiconductors offers a variety of ICs for WLAN applications, including the DE6003 2.4-to-2.5-GHz frequency-hopping transceiver and the WL100 LAN interface circuit which is designed to work with the DE6003 in WLAN applications.

GEC Plessey Semiconductors also offers a wide range of application notes for selecting WLAN antennas, performing measurements, and considering the effects of interference on WLAN system performance.

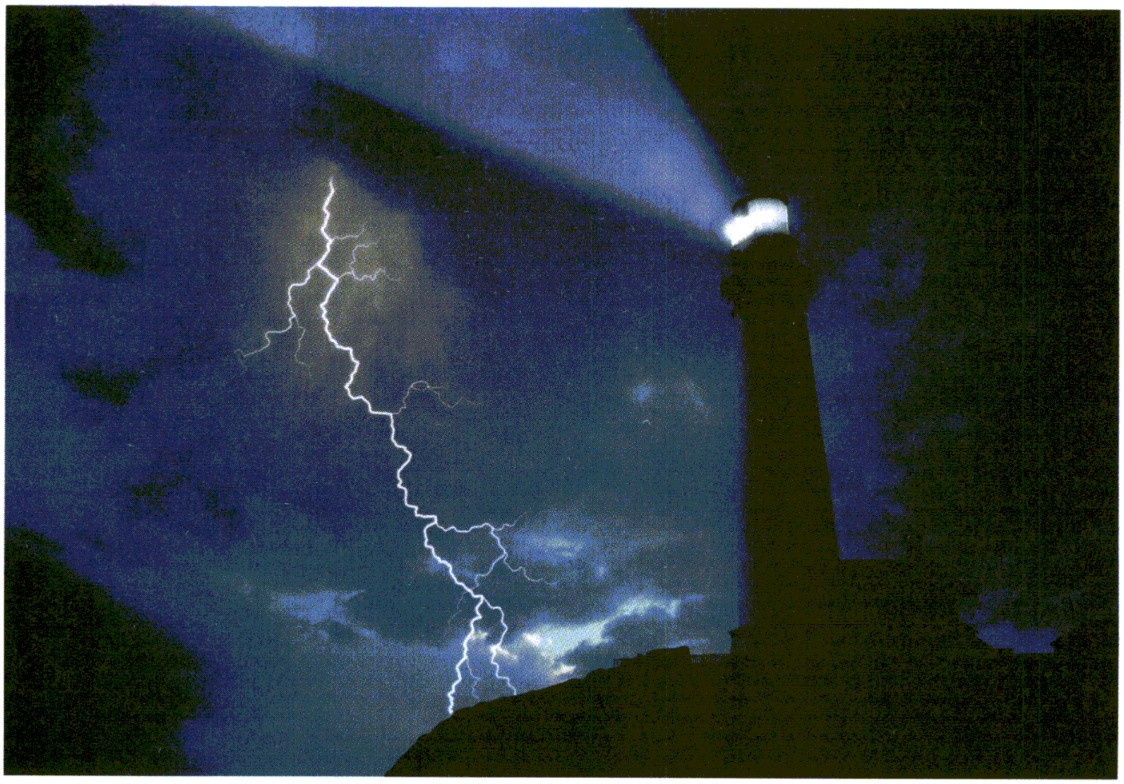
TriQuint Semiconductor (Hillsboro, OR) features the TQ9207, an integrated 100-mW amplifier/switch for use in the 2.4-to-2.5-GHz WLAN band. It provides the Tx/Rx amplifier function, using a low-noise amplifier (LNA) in the receive path with 3.5-dB typical noise figure and a Class A medium-power (150-mW) amplifier (suitable for digital modulation schemes) in the transmit path with +21-dBm 1-dB compression point and 15-dB gain. The IC is designed for a typical +5-VDC power supply and typically draws 190-mA current. The TQ9207 is supplied in a 24-pin SSOP housing that is ideal for PCMCIA cards.

The Raytheon Commercial Electronics Division of Raytheon Co. (Lexington, MA) has been involved in systems-level WLAN product development for several years, but also offers several gallium-arsenide (GaAs) MMICs suitable for data-communications applications at 2.4 GHz. For example, model RMPA2450-53 is a PA IC that provides 30-dB typical gain from 2.4 to 2.5 GHz. The amplifier may be biased for linear Class AB or Class F use. It delivers +31-dBm output power at 1-dB compression when running with a +7-VDC supply and +28-dBm output power when running from a +5-VDC supply. It achieves 35-percent power-added efficiency (PAE). Fabricated with the company's 0.25-μm power pseudomorphic-high-electron-mobility-transistor (PHEMT) process, the amplifier's on-chip matching components support operation in a 50-Ω system without external matching elements.

For designers of WLAN systems who need slightly more gain and less output power, the firm also offers model RMPA2451-53, (with a +5-VDC supply) and 33-percent associated PAE. As with the model RMPA2450-53, the RMPA2451-53 can be biased for either Class AB or Class F use. Both amplifiers are supplied in compact surface-mount housings.

Philips Semiconductors has developed a line of Si MMICs for applications beyond 2.4 GHz, including the BGA2001 SOT-packaged amplifier. The device provides 19-dB gain and 1.5-dB noise figure at 2 GHz while

When you need Clarity and Power



Stanford Microdevices is the clear choice.

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Stanford Microdevices, Inc. (SMI) is a leading supplier of RF integrated circuits for the wireless and wired telecommunications markets and a supplier of choice of OEMs worldwide. Stanford Microdevices continues to be on the industry's leading edge because of our superior quality, outstanding value and innovative technological advances. SMI provides the tools to create wireless communications equipment that is smaller, lighter, more powerful at market leading prices.

Our SGA 6000 series of silicon germanium MMIC amplifiers offers the high intercept point, high efficiency and high integration level at high output power desired, while providing the low noise figure and low power consumption needed for all wireless applications.

SGA-6386 has 1dB compressed output power of +20dBm, output third-order intercept point of +36dBm and 15.5dB of gain at 900MHz. Pricing on the SGA-6386 is $1.21 in quantities of 10,000 pieces with availability from stock to eight weeks.

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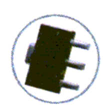
SPECIFICATION MATRIX

	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:


SOT-89 package

- Cascadable 50Ω
- Single voltage supply
- High output intercept
- +20dBm 1dB Compression Point
- Low current draw
- Low noise figure


86 package

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drawing only 4-mA current from a +2.5-VDC supply.

The firm provides an application note ("2.4 GHz low noise amplifier with the BGF480W") on its website (at http://www.philips.com) that instructs how to build an LNA for WLAN applications using a wideband transistor. The final amplifier delivers small-signal gain of 9 dB at 2.4 GHz with an associated noise figure of 3.5 dB and input third-order intercept point (IP3) of +15 dBm.

The LMX3162 radio-transceiver IC from National Semiconductor (Santa Clara, CA), while nominally for Bluetooth and HomeRF applications, is also suitable for WLAN applications at 2.4 GHz. The IC contains all of the Tx and Rx functions needed for a complete radio front end, including a PLL, frequency doubler, an LNA, a 2.4-GHz mixer, limiting amplifier, and an IF amplifier.

Analog Devices (Wilmington, MA) offers the ADF4118 frequency synthesizer for applications to 2.8 GHz, including LO generation. The IC consists of a low-noise phase/frequency detector, a precision charge pump, a programmable reference divider, programmable A and B counters, and a dual-modulus prescaler. It operates from +2.7- to +5.5-VDC supplies. In fact, a complete frequency synthesizer can be implemented if the ADF4118 is used with an external loop filter and a voltage-controlled oscillator (VCO).

Siemens (Munich, Germany) supplies a number of devices that can be used in WLAN designs, including the PMB 22xx series of vector modulators, the PMB 230x series of PLLs, and the PMB 231x series of prescalers.

M/A-COM Semiconductors (Lowell, MA) offers numerous amplifier/switch ICs for use from 2.4 to 2.5 GHz, including the model AM55-0001, a PA/switch combination capable of +22-dBm saturated output power. The IC, which is supplied in a QSOP-24 package, achieves 26.5-dB amplifier gain, with 1.2-dB switch loss and 15-dB typical switch isolation. The firm's model AM55-0003 is a PA/switch designed for linear applications at 2.4 GHz. Supplied in a QSOP-28 package, it delivers +24.5-dBm output power and 28-dB amplifier gain, with 1.2-dB switch loss and 12-dB switch isolation at 2.4 GHz. For designers seeking additional switch products, Alpha Industries (Woburn, MA) provides a wide range of single-pole, single-throw (SPST) switch ICs and single-pole, double-throw (SPDT) units for use in the 2.4-GHz WLAN band, in addition to a variety of monolithic passive components, such as couplers, attenuators, and power dividers.

Agilent Technologies (Palo Alto, CA) provides helpful information on the use of their INA-12063 IC amplifier for 2.4-GHz WLAN applications in the form of application note 1147 (available from the company's website at http://www.aglient.com). Since a user can set the device current of the INA-12063 with an external resistor, it can be readily optimized for WLAN applications requiring minimum current drain during transmissions. The INA-12063 can operate with current as low as 1 to 2 mA, although the current can be increased to 8 mA to provide a 1-dB compressed output power of approximately +6 dBm at a nominal supply voltage of +3 VDC (the device can operate with +1.5 to +5.0 VDC). The bias current is regulated through a patented active circuit, a 10:1 current mirror.

The note shows how to assemble a sample amplifier for 2.4-GHz use, running on +3 VDC and 5-mA current. The circuit employs a series (4.5-nH) chip inductor to achieve 50-Ω output matching. A 4.7-nH inductor was used as the nearest-valued standard component. The other components that are needed in the circuit are DC blocking and bypass capacitors, an RF choke, and the current-setting resistor (a 4.4-kΩ component as calculated by CAE software).

For more information on WLANs, the Wireless LAN Alliance (WLANA) is a nonprofit consortium of WLAN vendors established to help educate the marketplace about WLANs and their applications. Sponsor members include 3 Com, Aironet, BreezeCom, Intersil, Lucent Technologies, Nokia, Nortel Networks, and Symbol Technologies. The WLANA website can be found at http://www.wlana.com.

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(continued from p. 82)
phase detectors due to the low noise levels involved. Some manufacturers of single-chip PLLs provide noise-floor results.

Other sources of noise in the PLL include operational amplifiers (opamps) used for loop filters and noise from power supplies. These noise sources are shown in Fig. 16. The loop operates on these various noise sources. It is possible to write the Laplace transfer function, substituting $s = j2\pi f$ from each of these noise sources to the output. The magnitude of the transfer function squared multiplied by the phase-noise equation of the source, expressed in power, provides the output phase-noise power of that source at the output of the loop. By superposition, it is possible to power sum the effects of all the individual noise sources in order to produce a composite output phase-noise curve.

NOISE CAN BE MODELED USING PSPICE ELEMENTS, BUT IT IS NECESSARY TO TRANSFORM THE NOISE EQUATION FOR THE VARIOUS NOISE SOURCES IN THE PLL.

MODELING NOISE

Using the previous example (Fig. 17), assume that the PLL has only three noise sources—voltage-controlled-oscillator (VCO) noise, reference-oscillator noise, and main-divider noise. It is desirable to express these noise sources in terms of double-sideband phase-noise power [(radians^2/Hz]. The VCO has a noise floor of −155 dB/Hz, a 1/f corner frequency of 5 kHz, and a specified noise of approximately −110 dB/Hz at an offset frequency of 10 kHz. The coefficients of the phase-noise equation (equation 13) were manually adjusted in the MathCAD program to yield the specified phase noise at the particular offsets from the carrier. These are shown in the accompanying sidebar (see "Noise Equations for Noise Sources") on page 74 with some representative values for the noise at particular offsets.

The main divider has a floor of −155 dBc/_Hz and a 1/f or flicker corner of 1 kHz. The 10-MHz reference noise was obtained from the data sheet for a commercial 10-MHz temperature-compensated crystal oscillator (TCXO). The four coefficients in the reference-noise equation were experimentally determined using asymptotic lines with 0, −10-, −20-, and −30-dB/decade slopes in MathCAD to best approximate the noise plot from the manufacturer's data sheet. The phase-noise plots, in decibels, of these three sources are displayed in Fig. 18.

In Part 1, PSPICE was used to model the loop dynamics of a PLL. Noise can also be modeled by using PSPICE elements. First, it is necessary to transform the noise equation for the various noise sources in the PLL (VCO noise, reference noise, etc.), into PSPICE elements. Resistors produce "flat" noise and the diode model in PSPICE has a term for 1/f or flicker noise (−10 dB/decade). By integrating the flat resistor noise, it is possible to produce $1/f^2$ (−20 dB/decade) noise and by integrating the diode noise (1/f), it is possible to produce $1/f^3$ (−30-dB/decade) noise. Each term in the noise equation can be represented by one of the PSPICE elements mentioned, or a combination of elements. This concept will be illustrated by modeling the VCO noise in the example (Fig. 19).

Next month, the third and final installment of this article series on modeling PLL dynamics and noise will show how to combine the different noise sources to review what has been learned and to show how to integrate the knowledge of phase-noise contributors to produce a final phase-noise curve for the example loop.••

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Learn more about DWDM optical systems

Dense-wavelength-division multiplex (DWDM) is a powerful multiplexing technique for transferring several channels of high-data-rate signals. The method makes use of available fiber-optic infrastructures by multiplexing several frequencies in the 200-THz range. But in order to make best use of DWDM techniques, optical systems based on these methods must be properly characterized. For those interesting in learning more about measurements on DWDM systems, Application Note 70, "Modern Test Solutions Make Your DWDM System Crystal Clear" is available from Wavetek Wandel Goltermann (Eningen, Germany).

The application note provides the following example of a DWDM system. The system features four 2.5-Gb/s signals that are fed to the four optical transmitter modules in cable-television (CATV) terminal equipment. The optical output signals are converted using wavelength transponders to define wavelengths in the 1550-nm window when required, allowing standard 1310- and 1550-nm transmitter modules to be used. An optical wavelength division multiplexer is then used to combine the four optical signals, which are then fed to an optical fiber amplifier (OFA).

The application note describes the nature of DWDM signals and how to analyze them, including the type of test equipment currently available for characterizing DWDM signals and systems. Details are given on how to measure the different optical parameters of a DWDM system, including channel wavelengths and spacings, signal-to-noise ratio (SNR), and signal quality or bit-error rate (BER). The note also provides information on the firm's WG CATS software, which can be used with the company's test equipment to automate measurements on DWDM signals and systems.

Copies of the eight-page application note are free, from: **Wandel & Goltermann GmbH & Co., Marketing International, Postfach 1262, D-72795 Eningen, Germany; (49) 7121-86-1616, FAX: (49) 7121-86-1333, e-mail: info@wago.de, Internet: http://www.wg.com.**

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Understanding HFC technology

Hybrid-fiber-coax (HFC) technology enables cable-television (CATV) operators to offer the two-way communications essential for Internet access, interactive data services, and telephone services along CATV lines. For those seeking a simple but informative overview of HFC technology, a white paper, "Hybrid Fiber-Coax Technology Overview," is available from the website maintained by Analog Devices (Norwood, MA), at http://www.analog.com.

The white paper answers some of the more basic questions regarding HFC, including what is the capacity of a cable system with HFC. In terms of capacity, a cable link provides about 735 MHz of usable bandwidth. This is divided in a HFC system into a downstream (to the home) band and an upstream (to the hub) band. The downstream band typically occupies 50 to 750 MHz while the upstream band occupies from 5 to 40 MHz.

According to Bell Atlantic, a full HFC system could provide a number of services, including telephone service, as many as 37 analog television channels, as many as 188 digital television channels, over 400 digital point cast channels (with customer-requested programming delivered in time slots according to the customer), and high-speed two-way data services. In order to make these services a reality, however, requires clever engineering at the integrated-circuit (IC) stage. The white paper details a chip set for HFC applications based on discrete-wavelet-multitone (DWMT) technology for high performance and resistance to ingress noise on the upstream cable-network band. The chip set is designed to reliably and efficiently use the available cable-communications infrastructure by employing a wavelet-transform-based multicarrier modulation capable of superior channel isolation. DWMT divides the channel bandwidth into a large number of narrowband subchannels, and adaptively optimizes the number of digital bits per second that can be transmitted over each subchannel.

The literature provides a cost estimate of HFC technology, as well as a comparison to asymmetric-digital-subscriber-line (ADSL) modulation technology. Copies of the white paper are free, and can be downloaded from the company's website, at: **Analog Devices, One Technology Way, P.O. Box 9106, Norwood, MA 02062-9106; (781) 329-4700, (800) 262-5643, FAX: (781) 326-8703, Internet: http://www.analog.com.**

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InGaP/GaAs Provides High-Linearity HBTs

These wideband amplifiers benefit from advanced processing technology to deliver high gain over long operating lifetimes.

Jack Browne
Publisher/Editor

DEVICE technology has grown in leaps and bound over the last decade. Heterojunction-bipolar-transistor (HBT)-based RF integrated circuits (RF ICs) have gained wide acceptance among major wireless and broadband-communications equipment suppliers as the preferred technology for applications where high performance, high linearity, and competitive pricing are important. These RF IC products include power amplifiers (PAs) for cellular and personal-communications-services (PCS) handsets, driver amplifiers for cellular/PCS base stations, as well as cable-television (CATV)/fiber-cable line-driver amplifiers. In those applications, HBT-based products have overtaken many incumbent products based on gallium-arsenide (GaAs) metal-epitaxial-semiconductor field-effect transistor (MESFET) and silicon (Si) bipolar transistor technologies by providing high-performance, cost-effective solutions. The latest process advancement in HBT technology, indium-gallium-phosphide (InGaP) emitters on GaAs substrates, is the basis for a new line of high-linearity gain blocks from Stanford Microdevices (Sunnyvale, CA) for applications to 8 GHz.

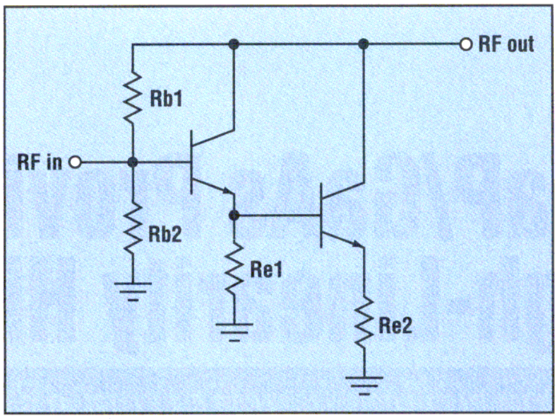
MATCHING TECHNOLOGY TO THE APPLICATION

Stanford Microdevices is a firm that tries to make the best use of available device technology. Founded in 1992, the "fabless" company makes use of outside silicon (Si) and gallium-arsenide (GaAs) foundries, matching process technologies from foundries such as Temic Semiconductor and TRW to particular designs. As a result, the company is not tied to any one technology, but can use its staff of skilled circuit designers to extract the best features of each technology.

The firm offers a wide range of semiconductor device technologies, including GaAs heterojunction bipolar transistor (HBT), indium-gallium-phosphide (InGaP) HBT, Si-germanium (SiGe) HBT, and Si laterally diffused metal-oxide silicon (LDMOS). Sales are made through large distributors, such as Avnet and Richardson, with knowledgeable support staffs and the capability to deliver products quickly.

Stanford Microdevices has its corporate headquarters in Sunnyvale, CA, which also houses final test, quality assurance, operations, applications engineering, and reliability engineering. The company also boasts facilities in Dallas, TX (for standard product development), Long Beach, CA (for wireless infrastructure product development and millimeter-wave product development), and Ottawa, Ontario, Canada (for next-generation wireless infrastructure product development). The company pursues opportunities in most wireless and wired infrastructure and communications-access markets, including cable modems, line amplifiers, digital cable television (CATV), microwave radios, wireless local-area networks (WLANs), fixed wide-area networks (WANs), cellular/personal-communications-services (PCS), and cordless/wireless-local-loop (WLL) markets.

The first available products using the InGaP/InGaAs HBT technology are high-linearity gain blocks with as much as 19-dB gain at 2 GHz. At 2 GHz, they achieve better than +17-dBm output power at 1-dB compression, with impressive dynamic range. The first pair of NGA amplifiers (Table 1) boast output third-order intercept points (IP3) of +38 dBm and better. In spite of the high linearity, these versatile amplifiers consume only approximately 80-mA current for single voltage supplies between +4 and +5 VDC.



1. This simplified circuit schematic diagram shows the Darlington amplifier configuration of an NGA series amplifier.

COMPARING PROCESSES

Although much has been written recently about inroads made by Si device technology in the form of Si germanium (SiGe), GaAs still enjoys considerable advantages over Si in terms of carrier mobility and, hence, gain at high frequencies. Although Si wafers are still larger [to 12 in. (30.48 cm) in diameter] and less expensive than GaAs [3 to 6 in. (7.62 × 15.24 cm) in diameter], GaAs is still the preferred technology when high performance and the semi-insulating properties of a GaAs substrate (for fabrication of passive circuit elements) are required (Table 2).

For power density, bipolar transistors have an inherent advantage over devices due to the vertical nature of current flow in a bipolar. FET-based devices, including high-electron mobility transistors (HEMTs), typically require a device that is two to three times larger than a bipolar for the same output power at a particular frequency, although the GaAs devices ultimately provide more gain at higher frequencies (above 2 GHz approximately). HEMT devices are slightly better than MESFETs in terms of power density because of the higher current density in the HEMT channel.

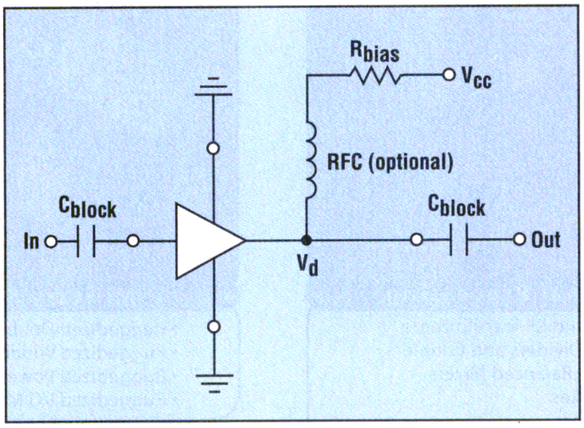
For high-linearity applications, a good figure of merit is the difference between the output third-order intercept (TOI) and the output 1-dB compression point (OIP3-P1dB). This figure of merit is a measure of the amplifier's efficiency. Typical values for MESFET and HEMT devices are between 9 and 10 dB. GaAs HBTs

typically yield OIP3-P1dB values of more than 15 dB. This excellent figure of merit has been attributed to the cancellation of base-emitter heterojunction nonlinearities, which results in a suppression of intermodulation (IM) products for multi-tone signals. Earlier published results for SiGe HBTs have indicated an OIP3-P1dB figure of merit of approximately 9 dB, but recent measurements at Stanford Microdevices have shown numbers comparable to the performance of GaAs HBTs—more than 14 dB.

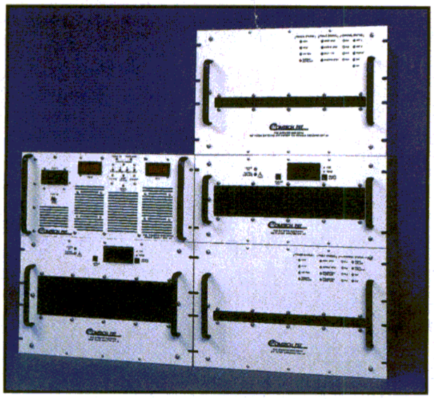
For devices that operate under large-signal conditions, high device breakdown is required. GaAs devices typically have higher breakdown voltage than Si devices for comparable doping concentrations. This is the result of the wider bandgap for GaAs compared to Si. Breakdown voltage is less of a concern for PAs used in handsets and portable devices where the trend has been toward +3-VDC operation. But for base stations, where the supply rail is typically greater than +10 VDC, a higher operating bias point is preferred due to current-resistance (IR) losses associated with high-current, low-voltage operation.

In contrast to Si and GaAs bipolars that operate from a single voltage supply, GaAs MESFET and HEMT power devices are typically depletion-mode devices that require a negative supply voltage (often generated with a separate IC) to control channel current. They also employ a drain switch to achieve acceptable off-state leakage current. Enhancement-mode MESFETs can be fabricated, but these are generally not viable due to the low-channel current and gate turn-on voltage. Enhancement-mode HEMTs have been used for single-supply operation, although they suffer low-power density. This results in output devices of more than 20 mm in total periphery for a typical +3.5-VDC handset PA.

Most of the GaAs HBTs produced to date have employed aluminum (Al)GaAs emitters on GaAs substrate. There are several benefits to the use of InGaP emitters, however. Traditional



2. This bias configuration is recommended for the NGA series of InGaP/GaAs Darlington amplifiers.



MODEL BHED2758-1000

Model BHED2758-1000 operates over the 20 to 110 MHz band at 1000 watts CW power output and 100 to 500 MHz at 400 watts CW power output. Harmonic rejection filters are included to reduce harmonics to -50 dBc. These filters are switched in 50 microseconds in VHF and 20 milliseconds in UHF. A single broad band 20 to 500 MHz solid state T/R switch as well as high speed noise quieting is included.

Size is 19"W x 26"D x 55"H and the operating voltage is 208 VAC, 50/60 Hz.

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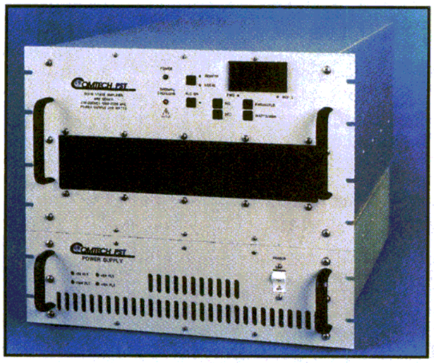

Model ARD1929-250 is a solid state Class "A" RF power amplifier system with a power output of 250 watts at 1 dB compressed (400 watts saturated). Gain is 53 dB minimum and noise figure is 10 dB nominal.

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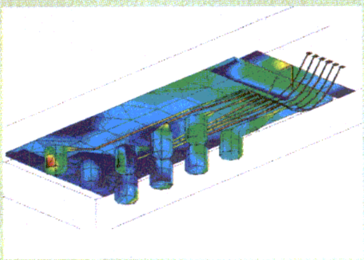
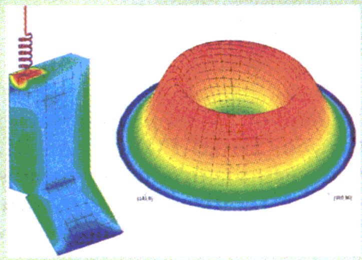
- RLC parameter extraction compatible with SPICE
- 2D and 3D display of current distribution, radiation patterns and near field
- Calculation of antenna and scattering parameters including directivity, efficiency and RCS
- Current and near field animation

IE3D Simulation Examples and Display

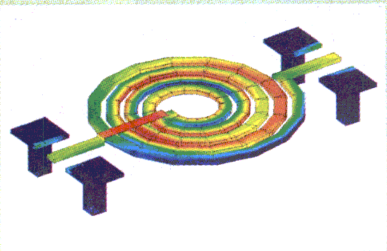
The current distribution on an AMKOR SuperBGA model at 1GHz created by the IE3D simulator



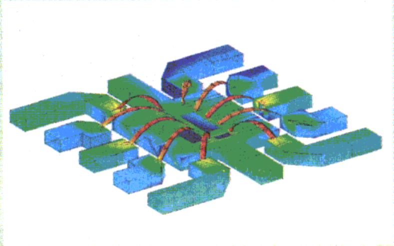
The current distribution and radiation pattern of a handset antenna modeled on IE3D



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

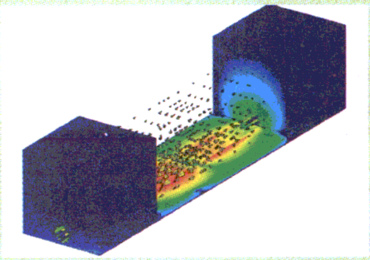


IE3D modeling of an IC Packaging with Leads and Wire Bonds

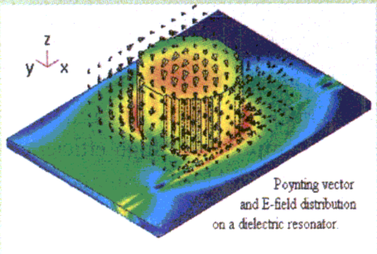


FIDELITY Examples

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



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



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CIRCLE NO. 243

Table 1: The NGA amplifiers at a glance

Parameter	NGA-489	NGA-589
Frequency	0.1 to 8.0 GHz	0.1 to 6.0 GHz
Gain	14.5 dB at 2 GHz	19 dB at 1 GHz
Output IP3	+39.5 dBm at 2 GHz	+38 dBm at 1 GHz
Noise figure	4.5 dB	4.5 dB
Output power	+17.5 dBm at 2 GHz	+19.0 dBm at 1 GHz
Supply voltage	+4.2 VDC	+5.0 VDC
Supply current	80 mA	80 mA

metal-organic-chemical-vapor-deposition (MOCVD)-grown GaAs HBTs with AlGaAs emitters exhibit low activation energy (less than 1.0 eV) and low mean time to failure (MTTF) of less than 10^5 hours when operating under high-current-density conditions (greater than 25 kA/cm^2), although HBTs with AlGaAs emitters formed by molecular-beam epitaxy (MBE) have shown higher activation energies.[1,2] InGaP/GaAs HBT devices fabricated for Stanford Microdevices have shown MTTFs of typically more than 10^7 hours at a junction temperature of +125°C, which is consistent with other published results for InGaP/GaAs HBTs.

LEVERAGING TECHNOLOGY

The InGaP/GaAs process technology is the basis for a new line of high-linearity gain blocks from Stanford Microdevices for applications to 8 GHz. (For more on Stanford Microdevices, see the sidebar). The NGA amplifiers (Table 1) include the NGA-489 for use from 0.1 to 8.0 GHz and the model NGA-589 for use from 0.1 to 6.0 GHz. The amplifiers boast more than 14-dB gain at 2 GHz with output TOI points (IP3s) in excess of +39 dBm at 2 GHz.

Featuring 50-Ω input and output impedances for broadband cascaded applications, the amplifiers are available in Micro-X-86 and SOT-89 packages. Broadband performance is achieved with resistive rather than reactive impedance matching. From the schematic representation (Fig. 1), the second emitter resistor, R_{E2}, established the trade-off between gain and bandwidth, allowing a designer to choose the 3-dB rolloff frequency at the expense of gain. The first emitter resistor, R_{E1}, determines the current balance between the first and second stages. The input impedance can be lowered to 50 Ω by setting the equivalent parallel resistance of R_{E1} and R_{E2}. Overall amplifier efficiency has been optimized by balancing the current through the first and second stages so that both stages experience full current or voltage swing at the 1-dB compression point. Linearity and gain rolloff were optimized through the use of through-wafer via holes and biasing each device in its best current density region.

In addition to these three amplifiers, the company is planning to expand the InGaP/GaAs product family with additional gain blocks by the second quarter of this year (Table 3). The lower-gain amplifiers employ the same Darlington topology as the

Table 2: Comparing RF IC technologies

Parameter	Si BJT	SiGe HBT	GaAs FET	GaAs HEMT	GaAs HBT
High gain	–	+	+	+	+
High-power density	+	+	–	0	+
High PAE	–	+	0	+	+
High OIP3-P1dB	0	+	0	0	+
High breakdown	0	0	+	+	+
Single supply	Yes	Yes	No	No	Yes

Note: (–) denotes fair, (0) denotes good, and (+) indicates excellent.

Table 3: NGA amplifiers to be announced

Model	Frequency range (GHz)	Gain (dB)	Output power at 1-dB comp.	Third-order intercept (dBm)
NGA-100	0.1 to 8.0	12 to 2 GHz	+15 dBm at 2 GHz	+32 at 2 GHz
NGA-200	0.1 to 6.0	15 to 2 GHz	+15 dBm at 2 GHz	+32 at 2 GHz
NGA-300	0.1 to 3.0	21 at 1 GHz	+15 dBm at 3 GHz	+32 at 3 GHz
NGA-600	0.1 to 6.5	12 at 2 GHz	+19 dBm at 2 GHz	+36 at 2 GHz
NGA-700	0.1 to 2.0	24 at 2 GHz	+15 dBm at 2 GHz	+36 at 2 GHz
NGA-800	0.1 to 2.0	24 at 2 GHz	+19 dBm at 2 GHz	+36 at 2 GHz
NGA-900	0.1 to 2.0	27 at 2 GHz	+15 dBm at 2 GHz	+32 at 2 GHz
NGA-1000	0.1 to 2.0	27 at 2 GHz	+19 dBm at 2 GHz	+36 at 2 GHz

NGA-486, NGA-489, and NGA-589, while the higher-gain devices were designed using current-feedback topologies and on-chip multistage-cascaded Darlington amplifiers. Since the higher-gain units are more prone to oscillation, ground inductance must be minimized to prevent unwanted ground loops. A typical biasing scheme is shown in Fig. 2. The supply-dropping resistor should be selected to account for the correct voltage at the chip for different supply voltages. A drop of +2 to +3 VDC across this bias resistor is recommended to compensate for β, operating temperature, and supply-voltage variations.

AMPLIFIERS TO COME

The additional NGA amplifiers shown in Table 3 cover a wide range of applications, with frequency coverage from 0.1 to 8.0 GHz, and choices of either +32- or +36-dBm output third-order-intercept (TOI) performance. Supplied in Micro-X-86 or SOT-23-5 housings, a total of eight new NGA amplifiers will be available in the second quarter of this year, including the SOT-23-5-packaged NGA-800, which achieves 24-dB gain at 2 GHz with +19-dBm output power and +36-dBm IP3. The NGA-800 is designed for applications from 0.1 to 2.0 GHz. Additional models include the SOT-23-5-packaged model NGA-1000, with 27-dB gain, +19-dBm output power, and +36-dBm IP3 at 2 GHz, and the Micro-X-86 housed model NGA-600, with 12-dB gain, +19-dBm output power, and +36-dBm IP3 at 2 GHz. The NGA-1000 is designed for applications from 0.1 to 2.0 GHz while the NGA-600 is aimed at applications from 0.1 to 6.5 GHz.

Since GaAs has a much higher (factor of 3) thermal resistance than Si, the thermal management of GaAs HBTs is critical for reliable operation. The NGA amplifiers feature very-conservative nominal current density of approximately 25 kA/cm^2. In addition, single-emitter-finger devices were used, spread apart with a very conservative pitch between devices. When running at a nominal +5 VDC, 80 mA, and a lead temperature of +85°C, the NGA-589 achieves a junction temperature of +120°C, resulting in an MTTF of more than 1 million hours.

The NGA amplifiers feature very flat gain with frequency. The NGA-489, for example, achieves ±0.5-dB gain flatness from 0.1 to 4.0 GHz. The NGA-589 achieves ±1.0-dB gain flatness from 0.1 to 3.0 GHz. The reverse isolation for the NGA amplifiers is typically 18 dB, while the group delay at 1.9 GHz is typically 90 ps or less. The NGA amplifiers are rated for operating temperatures from −45 to +85°C.

Given their frequency ranges, compact packing, and performance (gain and linearity), the NGA amplifiers should find application for a range of markets, including cellular, PCS, and the emerging variety of Bluetooth applications. **Stanford Microdevices, Inc., 522 Almanor Ave., Sunnyvale, CA 94086; (800) 764-6642, (408) 616-5400, Internet: http://www.stanfordmicro.com.**

CIRCLE NO. 51 or visit www.mwrf.com

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Noise Generator Checks High-Data-Rate Satcom

This precision noise generator is well-suited for emulating noise in K-band LEOS communications links and signal paths in digital microwave radios.

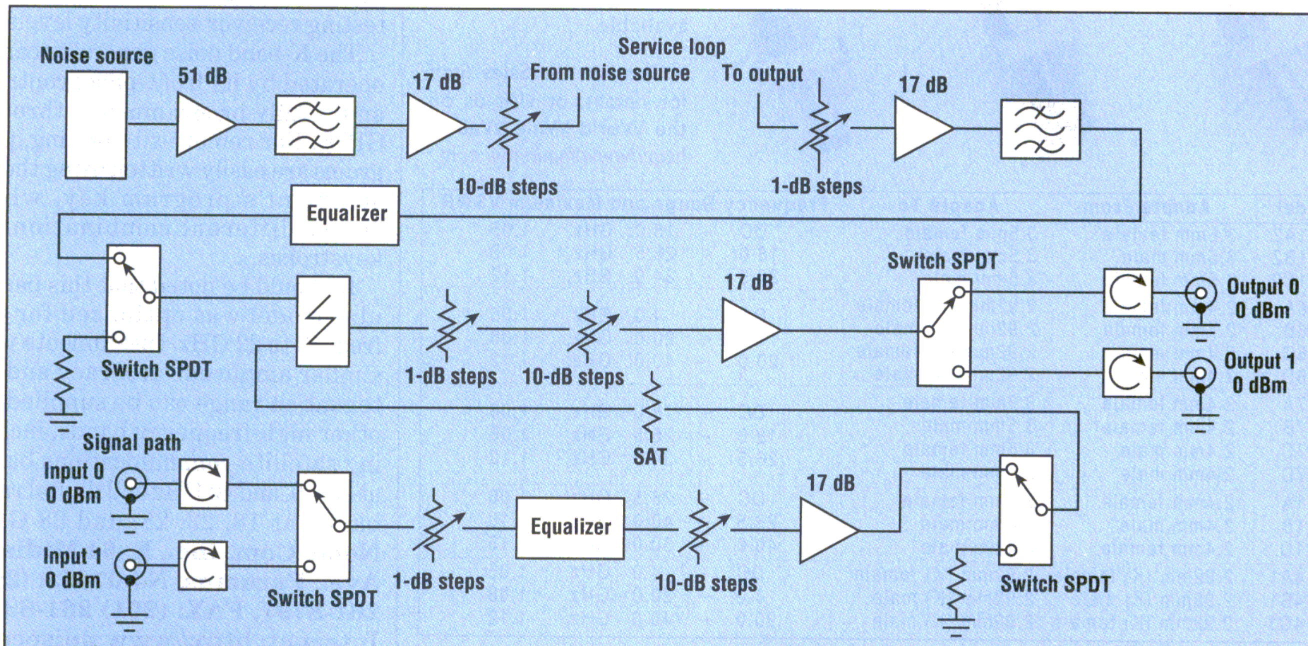
JACK BROWNE

Publisher/Editor

HIGH-DATA-RATE communications systems require innovative test solutions. The UFX series K-band noise generator from Noise Com, Inc. (Paramus, NJ) is one of these solutions—a programmable noise source that delivers ruler-flat output levels from 18 to 22 GHz. Designed specifically for evaluating the wide-bandwidth, high-data-rate links found in low-earth-orbit-satellite (LEOS) communications systems, the instrument maintains an output flatness of better than 0.5 dB peak-to-peak across its 4-GHz bandwidth, with maximum output power of 0 dBm.

The UFX series K-band noise generator contains an amplified noise source that can be mixed at different levels with one of two input signals (see figure). The instrument's internal noise source is boosted by a solid-state amplifier that is optimized through filtering and equalization to provide an output with true Gaussian amplitude distribution. The generous use of step attenuation throughout the UFX series K-band noise generator results in a total attenuation range of 0 to 127 dB in 1-dB steps (0.1-dB steps are available as an option). The internal attenuators are carefully selected to maintain amplitude flatness. For example, attenuators following the internal noise source provide a total attenuation range of 0 to 60 dB, adjustable in steps of 1 dB. The attenuators feature an accuracy of better than 0.25 dB and a repeatability of better than 0.1 dB.

The UFX series K-band noise generator provides two input signal



The UFX series K-band noise generator provides precisely controlled noise levels for testing commercial and military satellite-communications systems from 18 to 22 GHz.

ports for connection of sources at levels up to 0 dBm. Although lacking the total amount of gain applied to the internal noise source, these signal paths are constructed with the same precision as the noise path, achieving amplitude-ripple performance of better than 0.5 dB peak-to-peak over the 4-GHz output bandwidth. As with the internal noise source, signal sources applied to the two signal-path inputs can be controlled with 0-to-60-dB attenuation, adjustable in 1-dB steps. As with the noise path, the signal-path attenuators offer an accuracy of better than 0.25 dB and a repeatability of better than 0.1 dB. To maintain amplitude accuracy, a signal combiner (for blending internal noise with external signals) within the UFX series K-band nois generator exhibits ±0.5-dB flatness

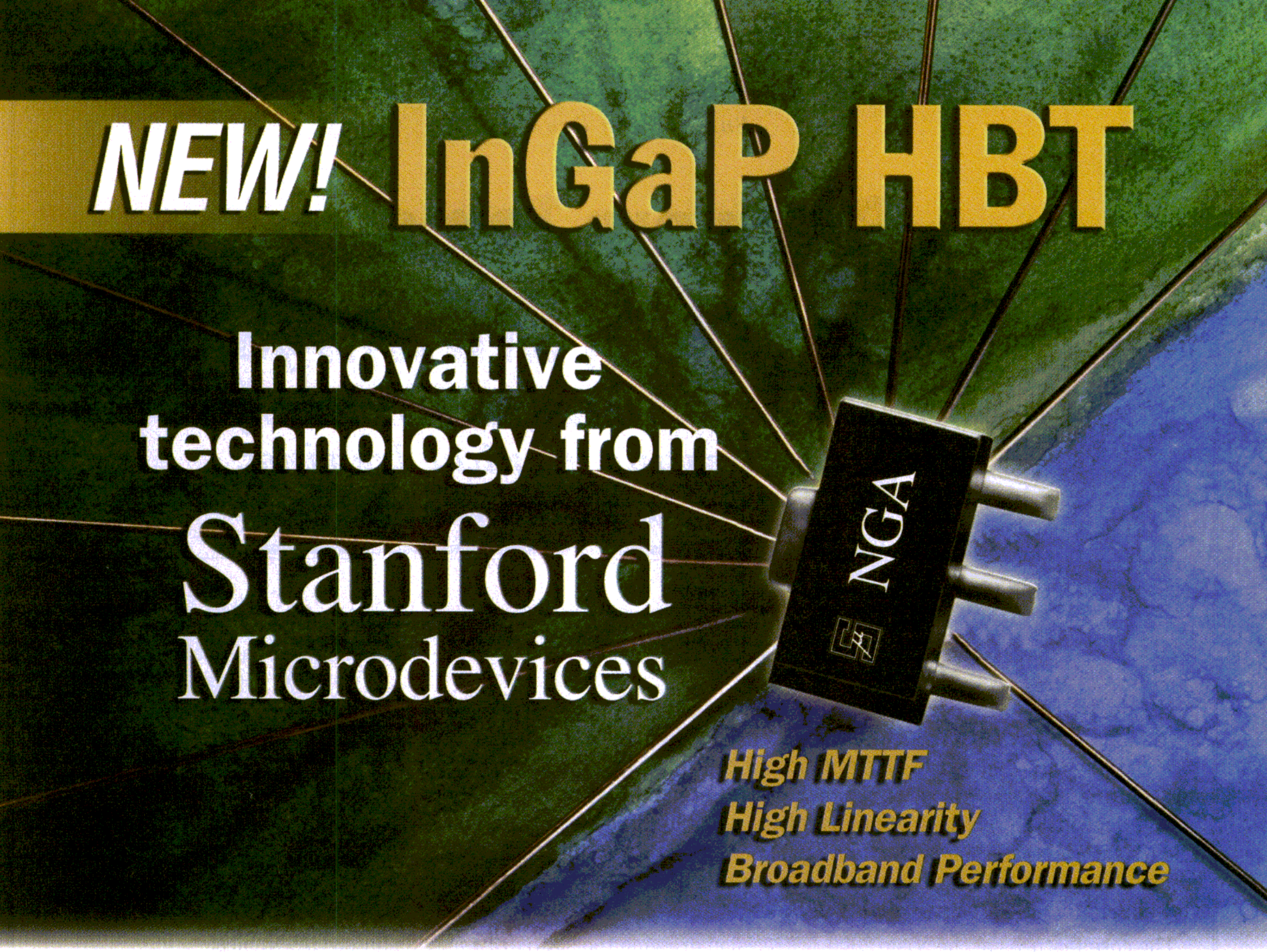
Variable attenuation in the outpu path accounts for at least 20-dB at tenuation, settable in 1-dB maximur steps. As with the other attenuator used in the UFX series K-band nois generator, the attenuation accurac in this output path is outstanding, a 0.25 dB or better. The repeatabilit of the attenuators in the output pat is 0.1 dB or better. The output port can be switched between an activ mode, where noise/signal power i available, or a standby mode, wher the noise/signal power is terminate into an RF load.

The UFX series K-band noise ger erator provides two output port with at least 30-dB isolation betwee them. Output signals may consist c noise alone or signals mixed wit precise amounts of noise. These sig nals are used with a bit-error-rat (BER) tester, to determine thresh old levels for performance in higl frequency digital terrestrial ar satellite-communications system By controlling the mix of noise ar signal levels, different signal-t noise ratios (SNRs), carrier-to-nois ratios (CNRs), or bit-energy-t noise (E_b/N_o) ratios can be esta lished to exercise a communicatio link under degraded-signal cond tions. The instrument is also ideal f testing receiver-sensitivity levels.

The K-band noise generator can l operated by its front-panel contro and is fully programmable throug GPIB. For remote-site testing, pr grams are easily written using the i strument's program key, whic stores different combinations keystrokes.

It should be noted that this parti ular model was optimized for u from 18 to 22 GHz. Instruments wi similar amplitude accuracy and a tenuation range can be supplied f other high-frequency bands, inclu ing satellite-communications ban at 28, 30, and 40 GHz and digital rad bands at 18, 23, 28, and 38 GH **Noise Com, Inc., E. 64 Midlar Ave., Paramus, NJ 07652; (20 261-8797, FAX: (201) 261-833 Internet: http://www.noisecor com.**

CIRCLE NO. 52 or visit www.mwrf.c



A new generation of HBT MMIC Amplifiers

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NGA-489 DC-8 GHz

Designed with InGaP process technology for greater reliability, this Darlington configured, high gain, heterojunction bipolar transistor MMIC amplifier offers value and performance for all wireless and broadband communication applications. Outstanding features are:

- 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:

- Cascadable 50Ω: 1.5:1 VSWR
- Low thermal resistance package
- High linearity
- High gain
- High P1dB

Stanford Microdevices is a world leader in meeting the need for high performance components at market leading prices. Fabless technology allows us to manufacture over 200 different products covering the following applications: Cellular, PCS, Wireless LAN, Wireless Internet, GSM and ISM. InGaP represents the latest in design and innovation for swiftly growing wireless communication technologies. Founded in 1992, Stanford Microdevices has grown to become the favorite of OEMs worldwide.

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CIRCLE NO. 379

Software Tames Engineering Drawings

This software is an essential addition for anyone involved in creating technical drawings, documents, and presentations.

ALAN ("PETE") CONRAD

Special Projects Editor

ENGINEERING presentations are often judged by their graphics quality. Even the most cynical and hardened audience can be brought to life by the precise use of graphics and the creative use of color. And, although a number of packages are available for developing presentation drawings, only the Technical Edition of Visio 2000 from Visio Corp. (Seattle, WA) combines powerful drawing capabilities with extreme ease of use. Visio 2000 lets users create a full range of drawings, from simple flowcharts to multiple-page, highly detailed technical diagrams.

The core features of the Visio 2000 program include SmartShapes® technology, intelligent diagramming, and an open architecture. With little or almost no training, any or more than 4000 SmartShapes symbols can be dragged and dropped upon a computer page to produce detailed technical drawings. Unlike static computer-aided-design (CAD) building blocks, SmartShapes objects feature built-in information ("intelligence") about the real-world objects they represent.

The process of using Visio 2000 begins with a simple concept called drag-and-drop drawing. To draw anything, from an organization chart to an electrical schematic, an operator simply drags predrawn SmartShapes symbols from task-specific stencils onto a drawing page. Unlike clip art, the SmartShapes symbols are programmed to behave intelligently, eliminating the need for tedious fine-tuning with drawing tools. As drawings are developed, the SmartShapes resize without distortion, shifting text to appropriate locations, and displaying or hiding segments as appropriate to the multiple dimensions of a design. Precision drawing tools can draw and align shapes using shape-extension lines that provide visual feedback.

Smart connector lines automatically connect symbols and shapes. When shapes are re-positioned on the drawing page, the connecting lines stretch, contract, change angles, and remain connected no matter what the re-positioning distance or angle is. The connector lines can even re-route themselves around shapes or add line jumps where lines intersect. With an intelligent layer control, users can create drawings with multiple layers while providing complete control over which drawing elements are visible, editable, and printable at any particular time.

Many of Visio 2000 shapes represent real-world objects that have important information associated with them, such as part numbers, prices, available quantities, dimensions, materials, or tensile strength. With custom property data stored in shapes, any diagram can become a powerful tool for analysis and communication. With Visio 2000, it is possible to automatically generate bills of materials, parts lists, and other reports from data linked to the shapes in a diagram.

Visio 2000 Technical Edition works with solutions from leading third parties such as CADcenter, Cadis, DataViews, Digital Tools, Echelon, ICARUS, Lomasoft, Siebe Environmental Controls, Technical Toolboxes, and Visimation. Users can import DWG files from Autodesk AutoCAD (v2.5-AutoCAD 2000) and IntelliCAD 98, as well as Bentley Systems Microstation (DGN) files.

Visio 2000 Technical Edition exports drawings in an enhanced Vector Markup Language (VML) format, an emerging standard for exchanging, editing, and delivering drawings on the Internet. VML-format graphics download more quickly and enable an Internet audience to pan, zoom, and dynamically navigate through multiple-page diagrams in Microsoft Internet Explorer 5.0 plus export files as AutoCAD DWG or DXF files.

The program runs on any IBM personal computer (PC) or compatible machine with Microsoft Windows 95, Windows 98, or Windows NT 4.0. The computer should be equipped with a Pentium processor operating at 166 MHz or faster, 16 Mb of random-access memory (RAM) for Windows 95/98 and 24 Mb of RAM for Windows NT, and 125 MB of available hard-disk space for a typical installation. P&A: $399 (new) and $199 (upgrade to older version); stock. **Visio Corp., 2211 Elliot Ave., Seattle, WA 98121; (206) 956-6000, FAX: (206) 956-6001, Internet: http://www. visio.com.**

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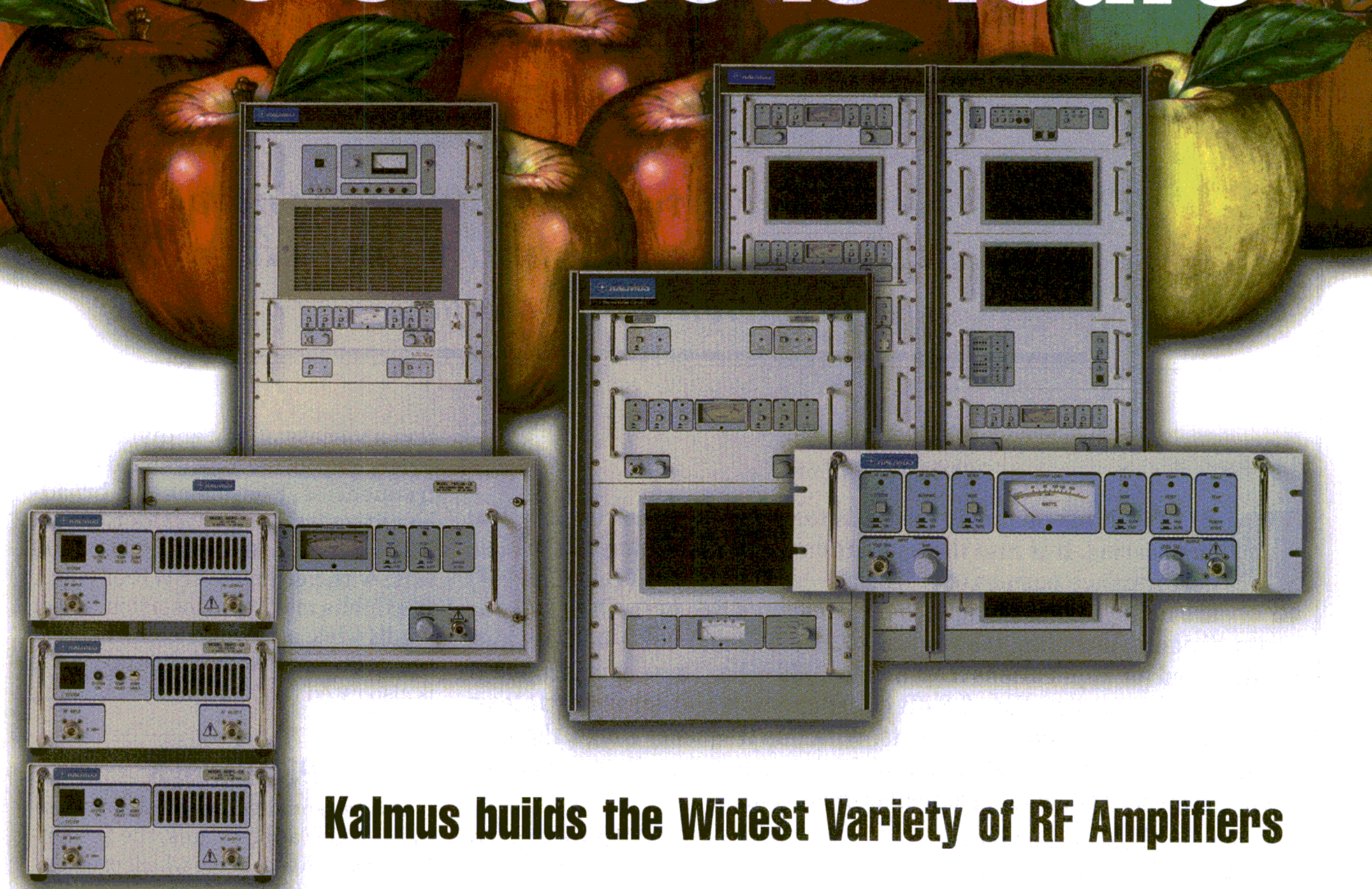
The Choice is Yours



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CIRCLE NO. 272

Power Dividers Hold Tight Tolerances

These two-way power dividers/combiners achieve excellent electrical performance as a result of precision machining.

JACK BROWNE
Publisher/Editor

PRECISION machining is the basis for many advances in electrical engineering. It is also the foundation upon which many of the high-performance components from Meca Electronics (Denville, NJ) are built. A good example is the company's line of low-loss two-way power dividers, including the models 802-2-0.900 and 802-4-0.900, which are designed for applications from 800 to 1000 MHz. These compact power dividers suffer no more than 0.2-dB insertion loss while delivering at least 30-dB isolation between output ports. With this frequency range, these dividers are ideal for separating and combining signals in cellular communications systems while adding minimum distortion.

The model 802-2-0.900 power divider is supplied with stainless-steel SMA connectors while the model 802-4-0.900 is equipped with nickel (Ni)-plated brass female type-N connectors (see table). The type-N unit measures 2.0 × 2.0 × 0.75 in. (5.08 × 5.08 × 1.905 cm) while the SMA unit measures 2.00 × 1.50 × 0.440 in. (5.08 × 3.81 × 1.12 cm).

The power dividers are designed with the help of computer-aided-engineering (CAE) software and computer-layout tools, but it is the firm's expertise in microwave laminates that contributes to the high performance. The production control and tight circuit tolerances of these components are evident in the lack of unbalance to be found in the output-signal paths. Both power-divider models achieve impressive worst-case amplitude unbalance of 0.1 dB and worst-case phase unbalance of 1 deg. from 800 to 1000 MHz.

Whether using SMA or type-N connectors, the power dividers each feature gold (Au)-plated beryllium-copper (BeCu) connector pins, Teflon insulation within the connectors, and rugged aluminum (Al) housings for the divider circuitry. In both types of power divider, the connectors meet the mating characteristic requirements of MIL-STD-348A. The power dividers are rated for operating temperatures from −55 to +70°C.

In addition to these cellular-band power dividers, the company also offers a variety of power dividers and other components for lower and higher frequencies. For example, a broadband two-way power divider, model 802-4-2, operates from 500 to 1000 MHz with maximum VSWR of 1.10:1. It exhibits maximum insertion loss of 0.2 dB while achieving minimum isolation of 20 dB between output ports. It is also rated for 10-W power-handling capability.

PCS DIVIDER

For personal-communications-services (PCS) applications, the firm offers model 802-4-1.850, which provides maximum VSWR of 1.15:1 from 1.7 to 2.0 GHz with maximum insertion loss of 0.25 dB and minimum isolation of 25 dB with type-N connectors. The model 802-2-1.850 is available for PCS applications requiring SMA connectors. The PCS model measures 1.50 × 2.0 × 0.75 in. (3.81 × 5.08 × 1.905 cm). In addition, the company offers a wide range of two-way, three-way, four-way, six-way, and eight-way power dividers for frequencies from 0.5 to 4.2 GHz (with SMA female connectors or type-N female connectors), as well as broad lines of microwave attenuators, couplers, and terminations. **Meca Electronics, Inc., 459 East Main St., Denville, NJ 07834; (973) 625-0661, FAX: (973) 625-1258, e-mail: sales@e-meca. com, Internet: http://www.e-meca.com.**

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The power dividers at a glance

Model	802-2-0.900	802-4-0.900
Frequency range	800 to 1000 MHz	800 to 1000 MHz
Maximum insertion loss	0.20 dB	0.20 dB
Minimum isolation	30 dB	30 dB
Maximum VSWR	1.10:1	1.10:1
Amplitude imbalance	0.1 dB	0.1 dB
Phase imbalance	1 deg.	1 deg.
Power-handling capability	10 W CW	10 W CW
Connector type	SMA female	Type-N female

Attenuators Combine Speed And Bandwidth

These switched-bit attenuators provide an optimum solution when bandwidth and switching speed are the critical performance characteristics.

Matt Jacobs

Senior Engineer

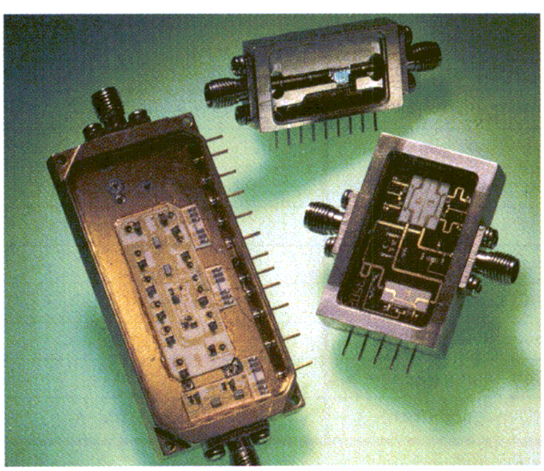
Switch Products, L-3 Communications,

Narda Microwave-East, 435 Moreland Rd., Hauppauge, NY 11747; (516) 231-1700, FAX: (516) 231-1711, Internet: http://www.L-3com.com.

BROAD operating bandwidth and fast switching speed are directly conflicting specifications when applied to microwave attenuators. Most high-frequency attenuators deliver one parameter or the other. However, digitally controlled switched-bit attenuators can satisfy both requirements. Switched-bit attenuators also feature good attenuation accuracy with well-controlled attenuation flatness. The DA series of switched-bit attenuators from L3 Communications Narda Microwave-East (Hauppauge, NY) exploits the inherent characteristics of switched-bit attenuators to deliver a switching speed of 25 ns or less for operating bandwidths as broad as 2 to 18 GHz.

The DA series of switched-bit attenuators (see figure) is available with operating frequencies from 100 to 1000 MHz, 2 to 6 GHz, 2 to 18 GHz, and 6 to 18 GHz with as much as 63.75-dB total attenuation range. The attenuation level is controlled by transistor-transistor-logic (TTL) signals while power requirements are either +5 VDC or −12 VDC depending on the model. The switching speed (defined by the transition from 50-percent TTL to 10- or 90-percent RF) is uniformly fast throughout the line, ranging from 15 to 100 ns. The rise and fall times are as fast as 15 ns. Monotonicity (i.e., an attenuator's ability to maintain a consistent relationship between changes in control input and changes in attenuation) is guaranteed for all units.

Digitally controlled attenuators differ from their voltage-controlled counterparts by the way they achieve attenuation. In a digitally controlled attenuator, fixed attenuator pads are switched in and out of the circuit with positive-intrinsic-negative (PIN)-diode single-pole, double-throw (SPDT) switches. One control bit is used per attenuator pad, so that the attenuation step size is determined by the lowest attenuator pad value. The total attenuation range is the sum of all attenuator pads.



The DA series of switched attenuators achieves high amplitude accuracy with fast switching speeds over broad bandwidths to 18 GHz.

The DA series of attenuators employs one or more pairs of SPDT switches depending on the model, with a low-loss connection between one pair of outputs, and a fixed attenuator between the other outputs. The diodes are switched between their forward-biased and reverse-biased states, which gives the attenuator higher switching speed. The attenuators achieve low, consistent VSWR performance of 1.80:1 or less throughout their operating range, and their power-handling ability is also high due to the properties of PIN-diode switches.

The DA series of switched-bit attenuators is available in dual-in-line-package (DIP) enclosures or housings with coaxial connectors, and can be tailored to meet specific requirements for attenuation range, power handling, control scheme, and other parameters. P&A: stock to 12 wks. **L-3 Communications, Narda Microwave-East, 435 Moreland Rd., Hauppauge, NY 11747; (516) 231-1700, FAX: (516) 231-1711, Internet: http://www.L-3com.com.**

CIRCLE NO. 55 or visit www.mwrf.com

Receiver ICs

TDMA Receiver ICs Serve PCS

These monolithic GaAs receiver ICs conserve battery power and PC-board real estate in cellular/PCS phones.

DON KELLER
Senior Editor

SMALL cellular-telephone handsets with long talk times are in demand. To meet this demand, TriQuint Semiconductor, Inc. (Hillsboro, OR) has developed two gallium-arsenide (GaAs), integrated-circuit (IC), time-division-multiple-access (TDMA)-mode receivers. The model TQ5122 cellular frequency-band receiver, and the model TQ5622 personal-communication-services (PCS)-band receiver, are complementary devices designed for manufacturers of IS-136, TDMA, or equivalent wireless handsets.

Each receiver operates from a single +2.8-VDC power supply and includes a power-down or "sleep" mode to extend standby and talk times in wireless applications. Each device typically draws 12 mA in the on state and less than 100 μA in the sleep mode. Each includes a low-noise amplifier (LNA), a mixer, a local-oscillator (LO) buffer, and an intermediate-frequency (IF) buffer (see figure). They are designed to minimize the number of external bypass and matching elements to keep board space and cost to a minimum.

The model TQ5122's LNA and mixer are matched to the 869-to-894-MHz cellular band. The mixer uses a high-side LO frequency between 954 and 1044 MHz, and the IF has a usable frequency range of 85 to 150 MHz. Typical gain is 18.5 dB, with a maximum gain variation of ±2 dB from −40 to +85°C. Typical noise figure is 2.7 dB, and third-order intercept point (IP3) is −8.5 dBm. The minimum return loss at the LNA's input (with external matching) and output, and at the mixer's RF and LO inputs, is 10 dB. The minimum LO-to-LNA RF-input isolation is 35 dB. After IF matching, the LO-to-IF isolation is 40 dB.

The model TQ5622's LNA and mixer are matched to the 1930-to-1990-MHz PCS band. The mixer uses a high-side LO frequency between 2015 and 2140 MHz, and the IF has a usable frequency range of 85 to 150 MHz. Typical gain is 17.5 dB, with a maximum gain variation of ±2 dB from −40 to +85°C. Typical noise figure is 2.8 dB, and IP3 is −9 dBm. The minimum return loss at the LNA's input (with external matching) and output, and at the mixer's RF and LO inputs, is 10 dB. The minimum LO-to-LNA RF-input isolation is 35 dB. After IF matching, the LO-to-IF isolation is 40 dB, and the RF-to-IF isolation is 20 dB.

In terms of maximum ratings, both ICs can dissipate up to 500-mW power. They can tolerate power-supply voltages to +5 VDC and operating temperatures from −55 to +100°C.

Both ICs are packaged in small, inexpensive JEDEC QSOP-16 plastic packages and are shipped in tape-and-reel format. **TriQuint Semiconductor, Inc., 2300 N.E. Brookwood Pkwy., Hillsboro, OR 97124; (503) 615-9000, FAX: (503) 615-8900, Internet: http://www.triquint.com.**

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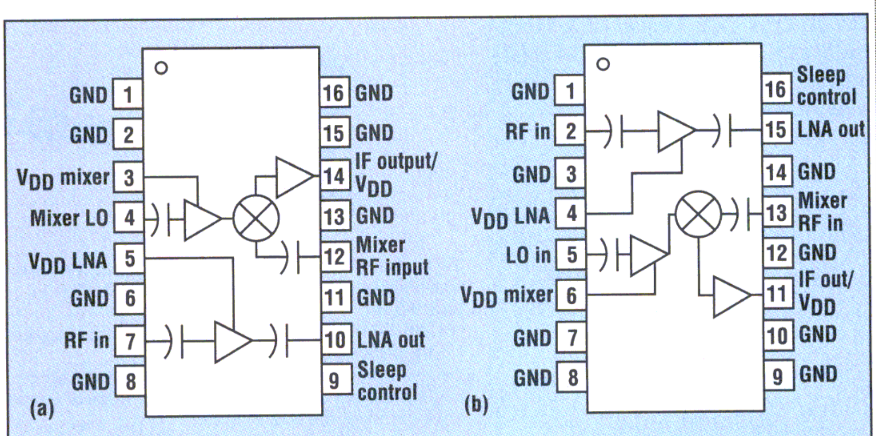

These internal block diagrams show the LNA, mixer, LO buffer, and IF buffer in the model TQ5122 (a), and the TQ5622 (b).

Field-Strength Analyzer Undergoes Upgrade

Fortified with improved internal shielding, this flexible handheld analyzer is ideal for scanning mobile communications systems as well as EMC precompliance testing.

JACK BROWNE
Publisher/Editor

FIELD-STRENGTH measurements can be used to pinpoint electromagnetic (EM) leaks in broadcast systems or areas of weak coverage in mobile communications systems. These measurements are made with a portable analyzer such as the 3201 from Protek (Northvale, NJ), which has been recently upgraded for enhanced spurious rejection and improved sensitivity of low-level signals. The 3201 RF field-strength analyzer provides a frequency range of 100 kHz to 2060 MHz and can measure signals at levels as low as −118 dBm.

The 3201 reads the levels of different types of communications signals with settings for narrowband frequency modulation (FM), wideband FM, amplitude modulation (AM), and single sideband (SSB). The 3201 achieves a level measurement range of −118 to −68 dBm from 1 to 2060 MHz in narrowband FM mode and −108 to −58 dBm from 10 to 2060 MHz for wideband FM, AM, and SSB measurements. The 3201 can scan signals in frequency steps from 5 to 9995 kHz, adjustable in 5- and 6.25-kHz steps. The analyzer features a reference oscillator with a frequency accuracy of ±3 PPM resulting in a frequency marker accuracy of ±25 PPM.

The 3201 is the successor to the company's 3200 RF field-strength analyzer. Due to the need for wide dynamic range in field-strength testing, the 3200 was redesigned and reinforced with additional shielding and gasketing materials to reduce internally generated spurious products by approximately 95 percent. Another problem in the 3200 was distortion from the analyzer's second local oscillator (LO), which was limiting the overall measurement dynamic range. To improve dynamic range, the LO level was reduced and gain was added, through the addition of a gallium-arsenide (GaAs) monolithic-microwave-integrated-circuit (MMIC) amplifier in the analyzer's input BNC connector.

The 3201 RF field-strength analyzer is a compact stand-alone unit measuring only 9 × 4 × 1.77 in. (22.86 × 10.16 × 4.4958 cm) and weighing only 1.4 lb. (0.63 kg). It runs from 6 AA nickel-cadmium (NiCd) batteries, a +12-VDC vehicle adapter, or an AC adapter (which is also used to charge the NiCd batteries).

The analyzer offers several frequency-scan modes, including manual mode, channel mode (where a set of memorized frequencies is scanned), and a search scan. The scanning rate is 12.5 channels/s. Sweep modes include single, normal (where the analyzer performs a continuous sweep until the squelch level is reached), and free-running modes. Once captured, signals can be displayed on a conventional spectrum-analyzer-type screen, in bar-graph form, or in numerical form using the built-in frequency counter. Up to 160 bar graphs can be shown per display. The integrated frequency counter operates from 9 to 2060 MHz with 1-kHz resolution and 50-PPM ± 1 count accuracy.

The 3201 RF field-strength analyzer is shipped with a Windows-based program that provides control of and data acquisition (DAQ) from the 3201 when running on a personal computer (PC). The software is provided on a single 3.5-in. (8.89-cm) floppy disk and requires only approximately 580 kb of hard-disk memory. The 3201 connects to the computer via an RS-232C cable (supplied with the analyzer). The software generates a computer-screen display that mimics the CRT of the 3201, and features a wide range of drop-down menu choices that allow operators to set system and measurement parameters, display settings, reference level, as well as attenuation.

The 3201 is also shipped with the rechargeable NiCd batteries, a quick connect/disconnect whip antenna, an RS-232C cable, the AC-to-DC adapter and battery charger, simple operating manuals for the 3201 and the PC software, a leather storage case, an earphone, and a carrying strap. For listening to demodulated signals, the 3201 includes a speaker and amplifier capable of 120-mW audio power. **Protek, 154 Veterans Dr., Northvale, NJ 07647; (201) 767-7242, FAX: (201) 767-7343, e-mail: hcprotek@hcprotek.com, Internet: http://www.hcprotek.com.**

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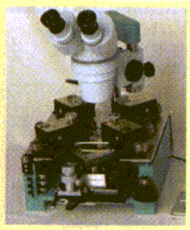

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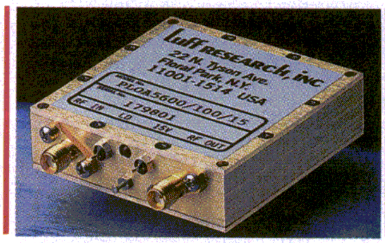

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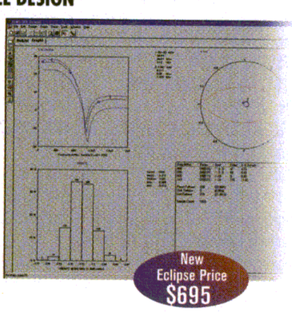

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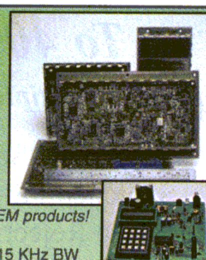

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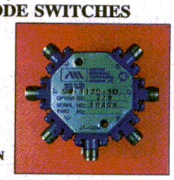

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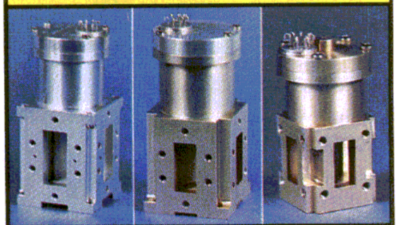

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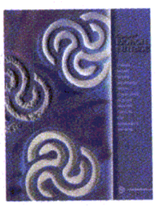

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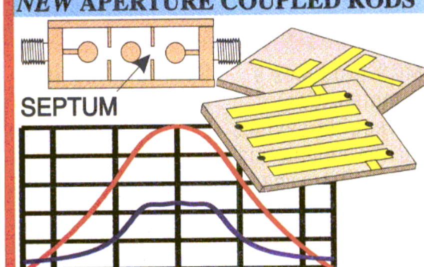

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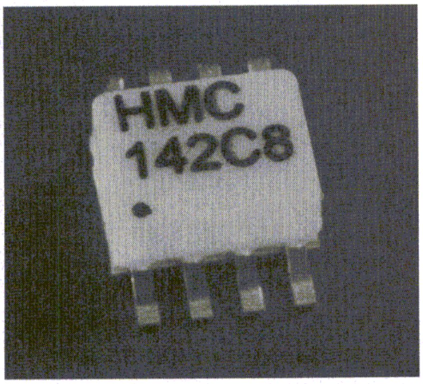
Double-balanced mixer spans 6 to 15 GHz

The model HMC142C8 miniature gallium-arsenide (GaAs) monolithic-microwave-integrated-circuit (MMIC) double-balanced mixer operates from 6 to 15 GHz and can be used as an upconverter or downconverter. It is housed in a non-hermetic, surface-mount package and requires no external components. This passive diode/balun mixer boasts high dynamic range and a third-order intercept of +20 dBm. Isolation between the local oscillator (LO) and RF, and between LO and intermediate frequency (IF), is greater than 30 dB. Conversion loss is 8.5 dB. The mixer is ideal for point-to-point microwave radio and very-small-aperture-terminal (VSAT) ground-equipment applications where small size and surface-mount compatibility are important. **Hittite Microwave Corp., 12 Elizabeth Dr., Chelmsford, MA 01824; (978) 250-3343, FAX: (978) 250-3373, Internet: http://www.hittite.com.**



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RAM-based chip set serves GSM handsets

The model AD20msp430 SoftFone chip set is a set of random-access-memory (RAM)-based integrated circuits (ICs) that allow Global System for Mobile Communications (GSM) cellular-phone manufacturers to customize features and options entirely in software, and to support an entire family of low-end and high-end handsets simply by loading different software versions. The programmability of the chip set can also be used by network operators to add or remove features over the airwaves. The chip set is comprised of two ICs—the AD6522 digital baseband processor and the AD6521 baseband converter. **Analog Devices, Inc., 804 Woburn St., Wilmington, MA 01887; (800) 262-5643, FAX: (781) 937-1021, Internet: http://www.analog.com.**

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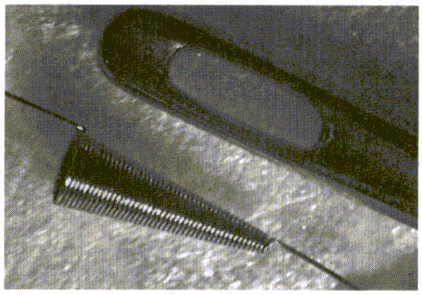
Miniature conical inductor covers 10 MHz to 40 GHz

A conical inductor only slightly larger than the eye of a needle operates without resonance over a bandwidth of 10 MHz to 40 GHz when connected across a +50-VDC microstrip. The broadband response of the coil is attributed to precision winding, gold (Au) plating, and powdered-iron (Fe) filling. The inductor is available in values of 0.84, 6.5, and 8.0 mH. They



are capable of handling 250-mA current. Typical insertion loss across the band is 0.3 dB. Due to its wide bandwidth, the inductor can be employed in numerous RF and microwave applications. **Piconics, Inc., 26 Cummings Rd., Tyngsboro, MA 01879-1406; (978) 649-7501, FAX: (978) 649-9643, Internet: http:// www.piconics. com.**

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Variable attenuators handle 8-kW peak power

A series of continuously variable attenuators includes seven models, each of which covers one of the following frequency bands: 0.925 to 2 GHz, 2 to 4 GHz, 4 to 8 GHz, 8 to 12.4 GHz, 8 to 18 GHz, and 12.4 to 18 GHz. The entire series offers a total frequency coverage of 0.925 to 18 GHz. They are available with attenuation ranges from 10 to 20 dB, and many of these models boast very-flat attenuation characteristics as a function of frequency. For example, the 0.925-to-2-GHz model with an attenuation range of 10 dB offers attenuation versus frequency that is flat within ±1 dB. The attenuators handle 50-W average power and 8-kW peak power. In general, the variable attenuators suffer no more than 0.5-dB insertion loss and no more than 1.50:1 VSWR. **ARRA, Inc., 15 Harold Court, Bay Shore, NY 11706-2296; (631) 231-8400, FAX: (631) 434-1116, e-mail: sales@arra.com, Internet: http:// www.arra.com.**

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Kitting system protects small components

The Partfolio portable kitting system stores a wide range of electronic components, prevents loss of expensive parts, and protects sensitive parts against damage from electrostatic discharge (ESD) and electromagnetic interference (EMI). The system is foldable, portable, and boasts quick and easy set up. In a typical application, a parts manager simply loads each kit with the precise number of components required by the entire assembly process. It is designed to replace bins dedicated to one particular electronic component. **Global Kitting Systems, P.O. Box 1757, Roseburg, OR 97470; (800) 663-4591, FAX: (541) 672-8737, Internet: http://www.globalkitting.com.**

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Cable touts thermal stability

A line of semirigid cable for RF and microwave signals claims to eliminate the normal difficulties associated with dielectric expansion during automated soldering processes. The cable uses a low-density dielectric that is composed of microporous polytetraflouroethylene (PTFE) that is said to maintain position even with repeated exposure to typical soldering-operation temperatures. The line of cable is available in coils, sticks, and formed assemblies in diameters ranging from 0.085 to 0.25 in. (0.2159 to 0.635 cm). **Storm Products Co., 116 Shore Dr., Hinsdale, IL 60521; (630) 323-9121, FAX: (630) 323-9398, Internet: http://www.stormproducts.com**

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Contactless phase shifters reach 3 GHz

A new line of contactless phase shifters (CPSs) covers frequencies to 3 GHz for applications requiring phased signals. Standard CPS models KPH900SCL000 and KPH900SCL001 provide a minimum of 90-deg. incremental phase shifting at 2 GHz and a minimum electrical delay of 125 ps. Insertion loss is 0.15 dB from 0 to 1 GHz, 0.25 dB from 1 to 2 GHz, and 0.35 dB from 2 to 3 GHz. VSWR is 1.25:1 from 0 to 3 GHz. Drop-in CPS model KPH200SCL000 provides a minimum of 30-deg. incremental phase shifting at 2 GHz and a minimum electrical delay of 41.7 ps. Miniature CPSs for embedded-system applications provide a minimum of 35-deg. incremental phase shifting at 2 GHz. **KMW, Inc., 13131 E. 166th St., Los Cerritos, CA 90703; (562) 926-2033, FAX: (562) 926-6133, Internet: http://www.kmwinc.com.**

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Base-station antenna offers remote tilt control

The Slant-Pol Teletilt antenna is an 1800-MHz, dual-polarized, cellular base-station antenna whose tilt can be adjusted ±45 deg. on site or remotely through the network links. The manufacturer claims that this adjustability supports better control of the base station's RF energy, thereby reducing distortion of the ground-illumination pattern. The adjustability is also said to provide increased flexibility when selecting a site location, since the cellular base-station antenna can be placed in an available site and then adjusted to suit it. The antenna can be adjusted without powering down the cell site, which allows service to continue while the site is reconfigured. **Deltec Telesystems, P.O. Box 51 123, Tawa, Wellington 6230, New Zealand; +(64) 4 232-0580, FAX: +(64) 4 232-8255, Internet: http://www.deltek-telesystems.com.**

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Amplifier spans 8 to 12 GHz

The model MSH-6642601 amplifier uses gallium-arsenide (GaAs) field-effect transistors (FETs) and hybrid technology to deliver an output power of 1 W at the 1-dB compression point from 8 to 12 GHz. It is ideal for use as a traveling-wave-tube (TWT) driver and for testing platforms for military and commercial applications. The amplifier offers gain of 40 dB, noise figure of 3 dB, and input and output VSWR of 2:1. The device operates from a +15-VDC power supply and includes internal voltage regulation and overvoltage and reverse-polarity protection. **Microwave Solutions, Inc., 3200 Highland Ave., Suite 3A, National City, CA 91950; (619) 474-7500, FAX: (619) 474-7003, Internet: http://www.microwavesolutions. com.**

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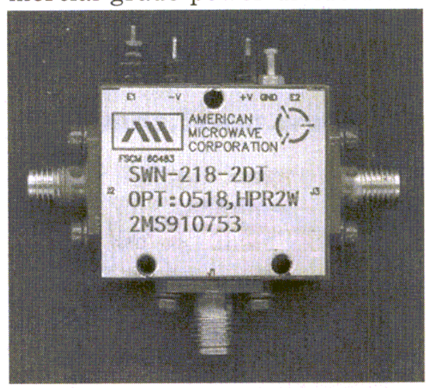
Power dividers suit GSM and AMPS

The model 4C0804 8-way power divider and the model 4A1604 16-way power divider cover 0.8 to 1.0 GHz for signal-distribution and channel-combining applications in Global Systems for Mobile Communications (GSM) and Advanced Mobile Phone Service (AMPS). The in-phase, commercial-grade power dividers are



designed with Wilkinson dividers and offer a minimum of 20-dB isolation and an amplitude balance of better than ±0.15 dB. The 8-way divider has a maximum insertion loss of 0.45 dB and a minimum return loss of 20 dB. **Anaren Microwave, Inc., 6635 Kirkville Rd., East Syracuse, NY 13057; (800) 411-6596, FAX: (315) 432-9121, Internet: http://www.anaren.com.**

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SP2T switch covers 0.5 to 18 GHz

The model SWN-218-2DT absorptive, non-reflective, single-pole, two-throw (SP2T) switch operates from 0.5 to 18 GHz. The switch can handle 2 W of RF power. Its insertion loss is less than 0.85 dB at 500 MHz and less than 2.35 dB at 18 GHz. Isolation is better than 95 dB at 500 MHz and better than 80 dB at 18 GHz. Between ports, its amplitude remains within ±0.5 dB and its phase remains within ±5 deg. from 0.5 to 18 GHz. Typical VSWR is 2:1. The switch has an on delay of less than 65 ns and an off delay of less than 60 ns. **American Microwave Corp., 7311 G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, Internet: http://www.amwave.com.**

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Monolithic ADC conserves power

The model SPT7937 12-b monolithic analog-to-digital converter (ADC) consumes only 170 mW and operates from a single +5-VDC power supply. Its 250-MHz input bandwidth makes it ideal for direct intermediate-frequency (IF) downconversion in wireless communication applications. The ADC has a minimum sampling rate of 28 MSamples/s and incorporates proprietary, parallel, successive-approximation-register (SAR) architecture and complementary-metal-oxide-semiconductor (CMOS) technology. Its on-chip track-and-hold function is said to yield excellent dynamic performance without external components. The ADC operates at temperatures from –40 to +85°C and is available in a 10 × 5.25-mm, 28-lead, shrink-small-outline package (SSOP). **Signal Processing Technologies, Inc., 4755 Forge Rd., Colorado Springs, CO 80907; (719) 528-2300, FAX: (719) 528-2370, Internet: http://www.spt.com.**

CIRCLE NO. 85 or visit www.mwrf.com

MCPA serves base stations

The MCPS 3135 multi-carrier power amplifier (MCPA) boasts compact, modular, configurable design for wireless-base-station applications. The amplifier has an efficiency of greater than 11 percent and can generate 135 W of digital or analog RF power. Its modular design allows up to four amplifiers to be combined in a 32 × 14 × 18-in. (81.28 × 35.56 × 45.72-cm) shelf to generate 500 W. The manufacturer's PowerStack architecture is said to maximize cooling and reliability, and provides "soft-fail" protection in each module. **Spectrian Corp., 350 West Java Dr., Sunnyvale, CA 94089; (408) 745-5400, FAX: (408) 541-0258, Internet: http://www.spectrian.com.**

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Oscilloscope calibration system reaches 2.5 GHz

The model OCS2500 automated oscilloscope calibration system calibrates the accuracy of analog and digital oscilloscopes to 2.5 GHz. The system consists of the company's model SG5050 leveled sine-wave generator, the model CG5011 calibration generator, and the model TM5006A mainframe. The sine-wave generator provides calibrated output voltages ranging from 4.5 mV to +5.5 VDC peak-to-peak into +50 VDC. Absolute accuracy is ±1.5 percent from 10 to 50 kHz, and flatness ranges from ±1.5 to ±4.0 percent over the remainder of the frequency range to 2.5 GHz. The frequency accuracy is ±3 PPM +0.3 Hz below 50 kHz, and ±3 PPM +3.0 Hz from 50 kHz to 2.5 GHz. Second-harmonic distortion is −25 dBc or less, depending on output frequency range and amplitude. The calibration generator features a remote pulse head that provides a pulse rise time of 160 ps. The amplitude mode provides voltage, current, low-edge, high-edge, as well as fast-edge functions. The timing mode provides markers in order to facilitate the calibration of oscilloscope time bases. The trigger rates can be programmed from 100 ns to 5 s. The system's Next-Cal-Date tracking feature lets the user know when the next calibration is due. **TEGAM, Inc., Ten TEGAM Way, Geneva, OH 44041; (440) 466-6100, FAX: (440) 466-6110, Internet: http://www.tegam.com.**

CIRCLE NO. 87 or visit www.mwrf.com

Amplifier suits PCS

A new solid-state amplifier covers the entire personal-communication-services (PCS) band of 1930 to 1990 MHz with an average code-division-multiple-access (CDMA) power level of 15 W. With an operating temperature range of −30 to +65°C, it is designed to mount onto base-station tower tops. The amplifier has circulator protection for transmission into any load mismatch. Monitor lines are provided for overtemperature, overcurrent, VSWR, and device-failure status. The 8 × 8 × 1.25-in. (20.32 × 20.32 × 3.175-cm) amplifier operates from a +24- to +26-VDC power supply. **MPD Technologies, Inc., 49 Wireless Blvd., Hauppauge, NY 11788-3935; (516) 231-1400, FAX: (516) 231-8081, Internet: http://www.mpd.com.**

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Power sensor extends meter range

The model E9304A power sensor, which has a frequency range of 9 kHz to 6 GHz, extends the low-frequency coverage of the company's EPM series E4418B and E4419B power meters. The new sensor provides the low-frequency coverage required for electromagnetic-compatibility (EMC) and electromagnetic-interference (EMI) test applications such as the IEC61000-4-3 radiated immunity test. The sensor has a dynamic range of −60 to +20 dBm and allows engineers to capture a wide range of field strengths produced during testing. The sensor's low-frequency range is also useful for measuring transmitter power and receiver sensitvity during the installation and maintenance of very-low-frequency to high-frequency radios. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844, Internet: http://www.agilent.com.**

CIRCLE NO. 89 or visit www.mwrf.com

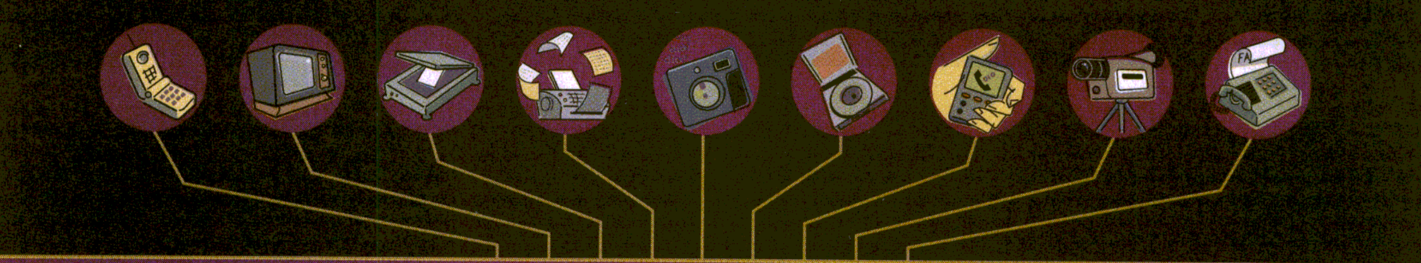
VCO boasts low phase noise

The model CLV0905E voltage-controlled oscillator (VCO) generates frequencies from 896 to 914 MHz with a typical phase noise of −115 dBc/Hz at 10 kHz from the carrier. It attenuates the second harmonic to better than −25 dBc. The VCO has a tuning voltage range from +0.5 to +4.5 VDC and an average tuning sensitivity of 16 MHz/V. It delivers −5 ±2 dBm of output power into a 50-V load. The device draws 11 mA from a +5-VDC power supply and operates at temperatures from −30 to +85°C. It is housed in an industry-standard MINI-14S surface-mount package measuring 0.5 × 0.5 × 0.22 in. (1.27 × 1.27 × 0.5588 cm). **Z-Communications, Inc., 9939 Via Pasar, San Diego, CA 91216; (858) 621-2700, FAX: (858) 621-2722, Internet: http://www.zcomm.com.**

CIRCLE NO. 90 or visit www.mwrf.com

Inductors suit portable applications

The D104C series of shielded, surface-mount, low-profile, fixed inductors is available in inductances from 1 to 47 mH ±20 percent and are designed for portable applications. The maximum current rating for the



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inductors ranges from 9.7 A for the 1 mH model to 1.8 A for the 47 mH model. The inductors have a 10.3 × 10.3-mm footprint and a maximum height of 4.5 mm. **Toko America, Inc., 1250 Feehanville Dr., Mt. Prospect, IL 60056; (847) 297-0070, FAX: (847) 699-1194, Internet: http://www.tokoam.com.**

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Isolator boasts low insertion loss

The model CIC01A1723-01 isolator operates at frequencies from 1700 to 2300 MHz and boasts a maximum insertion loss of 0.3 dB. It provides a minimum isolation of 22 dB and a maximum VSWR of 1.15:1. The isolator's input power is rated at 25 W and it operates at temperatures ranging from 0 to +55°C. It is ideal for protecting an amplifier's output stage. **Narda Microwave-West, 107 Woodmere Rd., Folsom, CA 95630; (916) 351-4500, FAX: (916) 351-4550, Internet: http://www.nardamicrowave.com.**

CIRCLE NO. 92 or visit www.mwrf.com

Medium gain antennas serve wireless applications

The model S57212AMP10SMF DirectLink series of medium-gain antennas is designed for transmission and reception of linearly polarized signals in the 5.725-to-5.875-GHz industrial-scientific-medical (ISM) band. The antennas use a two-element patch array to provide a minimum gain of 12 dBi across the ISM frequency band. Front-to-back ratio is better than 25 dB. Mounting options include a ±30-deg. articulating wall mount, standard wall mount, mast mount, and a universal wall/mast mount that permits as much as ±90 deg. of vertical or horizontal main-lobe steering. The articulating versions of the antennas measure 5.8 × 3.81 × 2.26 in. (14.732 × 9.6774 × 5.7404 cm) and weigh 8 oz. **Cushcraft Corp., 48 Perimeter Rd., Manchester, NH 03103; (603) 627-7877, FAX: (603) 627-1764, Internet: http://www.cushcraft.com.**

CIRCLE NO. 93 or visit www.mwrf.com

Optical spectrum analyzer characterizes UDWDM

The Q8384 optical spectrum analyzer allows designers and manufacturers of optical components, erbium-doped fiber amplifiers (EDFAs), and ultra-dense wavelength-division multiplexing (UDWDM) devices to measure optical signals and characterize transmission systems and components. The analyzer's 10-picometer (pm) resolution bandwidth is said to permit discrimination between closely spaced signals and measure sidebands of modulated optical signals. **Tektronix Measurement Group, P.O. Box 3960, Portland, OR 97208-3960; (800) 426-2200 request code 1176, FAX: (503) 222-1542; Internet: http://www.tektronix.com/measurement.**

CIRCLE NO. 94 or visit www.mwrf.com

EMI shielding

A brochure contains information on products for electromagnetic-interference (EMI)-shielding solutions. Input/output (I/O) panel gaskets, conductive silicones, metal-wire-mesh gaskets, combination gaskets, EMI-shielding tape, and air-vent panels are described. **Advanced Performance Materials, Inc.; (314) 344-9300, FAX: (314) 344-9333, Internet: http://www.apm-emi.com.**

CIRCLE NO. 62 or visit www.mwrf.com

Solid-state amplifiers

A 72-page catalog covers high-power, drop-in, broadband, communication, low-noise, special-product, and stocked amplifiers. Specifications include frequency range, power, IP3, noise figure, VSWR, and gain. Technical information is included, along with outline drawings. **Microwave Solutions, Inc.; (800) 9MSI-AMP, (619) 474-7500, Internet: http://www.microwavesolutions.com.**

CIRCLE NO. 63 or visit www.mwrf.com

SMA connectors

A short-form catalog highlights high-performance SMA, extended-power SMA, and 2.92-mm connectors for microstrip and stripline circuits. Connectors are designed for applications to 40 GHz. **Southwest Microwave, Inc.; (480) 783-0201, FAX: (480) 783-0360, e-mail: connectors@southwestmicrowave.com, Internet: http://www.southwestmicrowave.com.**

CIRCLE NO. 64 or visit www.mwrf.com

Distribution amplifiers

A data sheet contains information on a high-isolation distribution amplifier, a 10-channel distribution amplifier that operates from 1 to 10 MHz. The data sheet offers complete specifications with an outline drawing. **FEI Communications, Inc.; (516) 794-4500, FAX: (516) 794-4340, Internet: http://www.freqelec.com.**

CIRCLE NO. 65 or visit www.mwrf.com

Frequency sources

A 24-page short-form catalog contains a range of microelectronic products, used in a variety of applications. Fixed-frequency oscillators, voltage-controlled oscillators (VCOs), as well as phase-locked products and synthesizers are offered. Surface-acoustic-wave (SAW) products, data-conversion products, and custom microelectronics are offered. Descriptions, specifications, and outline drawings are provided. **Micro Networks Corp.; (508) 852-5400 ext. 3591, FAX: (508) 852-8456, Internet: http://www.micronetworks.com.**

CIRCLE NO. 66 or visit www.mwrf.com

Power conversion

A 12-page reference guide supplies specific definitions from absolute maximum ratings to wide-range input. The application note is designed to aid design engineers in developing effective power-supply solutions. **Powercube; (800) 866-3590, FAX: (800) 866-3589, Internet: http://www.powercube.com.**

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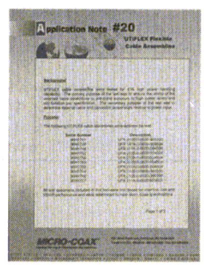

Cable assemblies

An application note explains the performance of cable assemblies at varying levels of continuous-wave (CW) power. The application note details the changes in the cable assemblies' external temperature that are relative to power input. **MICRO-COAX; (800) 223-2629, Internet: http://www.micro-coax.com.**

CIRCLE NO. 68 or visit www.mwrf.com

Antenna development

A guide is presented on antenna systems for developers and manufacturers who are designing portable computing devices using Bluetooth technology. Selecting the right antenna, antenna tuning and testing, as well as antenna choices for Bluetooth products are discussed. Location, orientation, and attachment of the antenna are also offered. **Centurion International, Inc.; (402) 467-4491, FAX: (402) 467-4528, e-mail: sales@centurion.com, Internet: http://www.centurion.com.**

CIRCLE NO. 69 or visit www.mwrf.com

Adhesives and sealants

A four-page application-selector guide focuses on adhesives, sealants, coatings, as well as encapsulants for microelectronics. This guide consists of one- and two-component systems, including epoxies, silicones, acrylics, and latexes. Viscosity, setup times, cure schedules, service-operating temperature ranges, and application recommendations are also listed for 70 different grades. **Master Bond, Inc.; (201) 343-8983, FAX: (201) 343-2132, Internet: http://www.masterbond.com.**

CIRCLE NO. 70 or visit www.mwrf.com

Power meters

Power meters for wireless communications design and test needs are discussed in a brochure. Specifications include maximum peak power-sensor rise time, graphic time-gate setting, direct crest-factor measurement, power versus time graphic display, standard deviation, and sample rate. Features are also listed. **Giga-tronics, Inc.; (800) 726-4442, Internet: http://www.gigatronics.com.**

CIRCLE NO. 71 or visit www.mwrf.com

Test equipment

A catalog covers refurbished electronic test equipment from a variety of manufacturers. Pulse generators, logic analyzers, plotters, meters, RF signal generators, spectrum analyzers, and oscilloscopes are offered. Frequency counters, impedance analyzers, network analyzers, power supplies, audio analyzers, signal generators, and data acquisition (DAQ) are also provided. Pricing information is included. **Test Equipment Connection Corp.; (800) 615-8378, (407) 804-1780, FAX: (800) 819-TEST, (407) 804-1277, Internet: http://www.4testequipment.com.**

CIRCLE NO. 72 or visit www.mwrf.com

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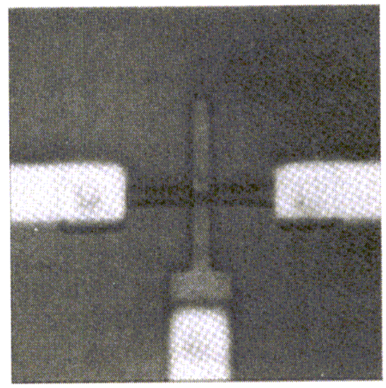

Twenty-two years ago, a researcher at Rockwell Science Center (Thousand Oaks, CA) presented data on the leading edge of gallium-arsenide (GaAs) metal-semiconductor-field-effect-transistor (MESFET) technology. The FETs, with gate dimensions of 0.25 × 0.25 μm and operating frequencies to 12 GHz, were fabricated by electron-beam lithography. The Rockwell research program was aimed at increasing the density of very-large-scale-integration (VLSI) circuits.

Microwaves & RF March Editorial Preview

Issue Theme: Wireless Show Wrapup

News

The Wireless Symposium/Portable By Design Conference & Exhibition enters its eighth year this February with a show slated to draw approximately 10,000 attendees. The March issue of *Microwaves & RF* has traditionally served to wrap up the breaking news, advances in technologies. This year our exclusive news report will be written by Senior Editor Don Keller, who will be attending his first Wireless Symposium.

Design Features

Technical articles in March will highlight the design of components essential to wireless applications. For example, authors from Rockwell Collins will present a low-phase-noise microwave synthesizer that leverages the latest heterojunction-bipolar-transistor (HBT) technology. In addition, authors from Alpha Industries describe a low-cost HBT power amplifier (PA) for personal-communications-services (PCS) frequencies.

Product Technology

March's Product Technology section will boast some exciting products from the snowy Northeast. Learn about a measurement system designed for high-rate differential testing of analog and digital circuits and components. Discover the next generation of impedance-matching tuners for high-power load-pull testing.

CIRCLE NO. 208