A PENTON PUBLICATION NOVEMBER 2000

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FOR DESIGNERS AT HIGHER FREQUENCIES

Microwaves & RF

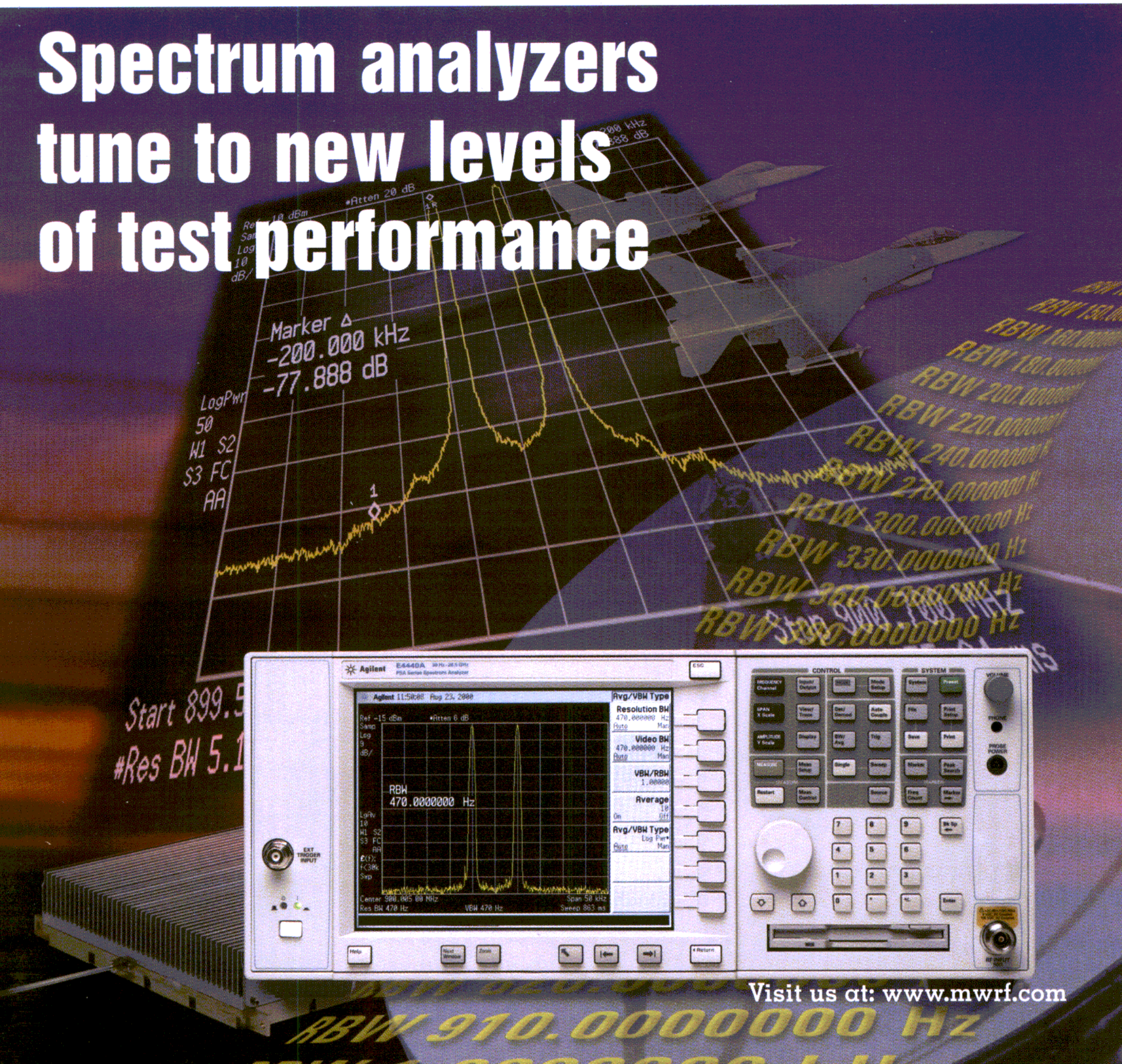
Amplifiers & Oscillators Issue

NEWS
IEDM unveils next-generation devices

DESIGN FEATURE
Design a synthesizer for mobile communications

PRODUCT TECHNOLOGY
Vector modulator transforms analog signal generators

Spectrum analyzers tune to new levels of test performance



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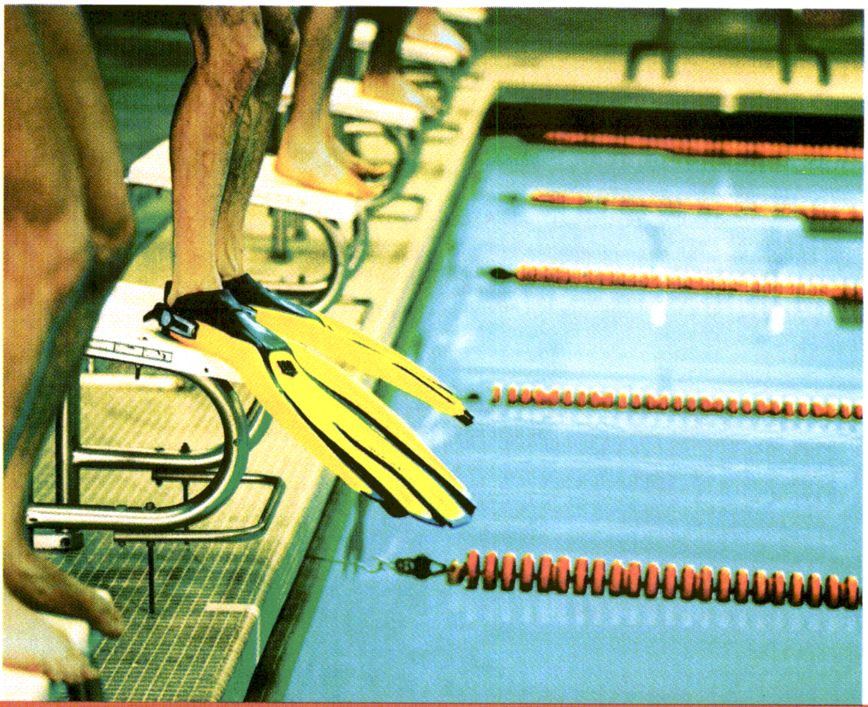

Remember– in the wireless world, there are no rules.

Engineers don't make rules. They figure out ways to stretch them to the point of breaking. The unbelievable becomes the inevitable. As is the case with the new Agilent ADS/89600 VSA integration. For the first time, the capabilities of EDA software are combined with the functionality of a vector signal analyzer.

PC-based VSA software tightly integrates with ADS design tools—eliminating communication gaps in your design process. You use the same interface and algorithms between simulation tools and hardware measurements. So you eliminate disagreements between the two domains at their source. And speed up error detection. Not to mention uncover signal impairments you couldn't see before. All this happens long before anyone even mentions the word "prototype."

In fact, the ADS/89600 VSA integration might even seem like an unfair advantage. That is, if such a thing existed in engineering. See for yourself. For more information and a product demo on CD-ROM, call us or visit our web site.

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CIRCLE NO. 214

VARIABLE GAIN AMPLIFIERS

FEATURES

- Gain can be continuously adjusted (0–15 dB) by applying a 0–10 VDC control voltage
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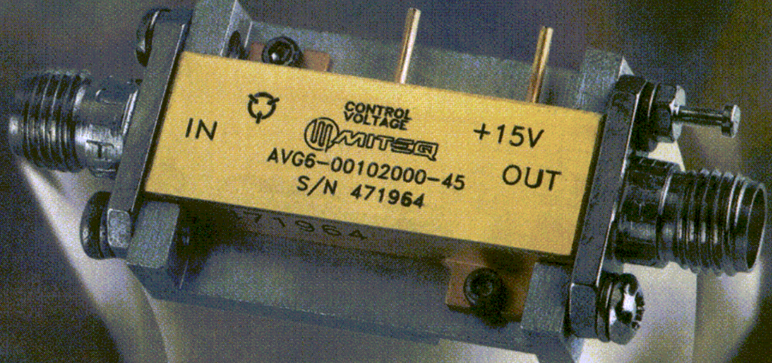
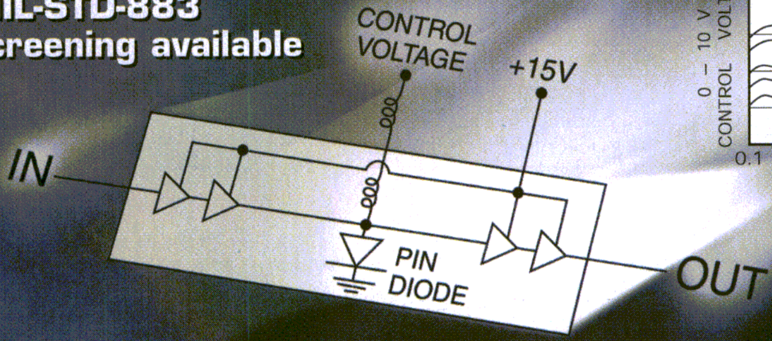
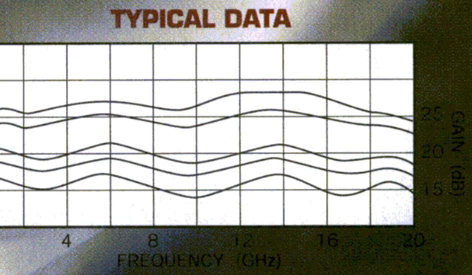
- Hermetically sealed
- Various bandwidth, gain, noise figure and power options available
- MIL-STD-883 screening available





TYPICAL DATA



3 YEAR WARRANTY

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN FLATNESS (dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN/OUT (Max.)	OUTPUT POWER @ 1 dB Comp. (dBm, Min.)	NOM. DC POWER (+15 V, mA)
AVG4-00100400-14	.1–4	28	±1.00	1.4	2.0:1	+10	150
AVG4-00100600-15	.1–6	28	±1.00	1.5	2.0:1	+10	150
AVG4-00100800-18	.1–8	26	±1.50	1.8	2.0:1	+10	175
AVG4-02000800-20	2–8	32	±1.25	2.0	2.0:1	+10	175
AVG5-04000800-12	4–8	30	±1.00	1.2	2.0:1	+10	150
AVG5-00101800-35	.1–18	24	±2.50	3.5*	2.5:1	+10	175
AVG6-00102000-45	.1–20	24	±2.50	4.5*	2.5:1	+10	250
AVG4-06001200-19	6–12	24	±1.50	1.9	2.0:1	+10	175
AVG4-06001800-25	6–18	22	±2.00	2.5	2.3:1	+10	185
AVG6-02001800-40	2–18	25	±2.25	4.0	2.5:1	+10	250

* Noise figure increases below 500 MHz.

Note: All above specifications are with 0 dB attenuation.

For additional information, please contact Naseer Shaikh at (631) 439-9259 or nshaikh@miteq.com

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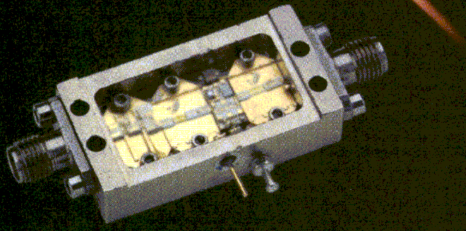
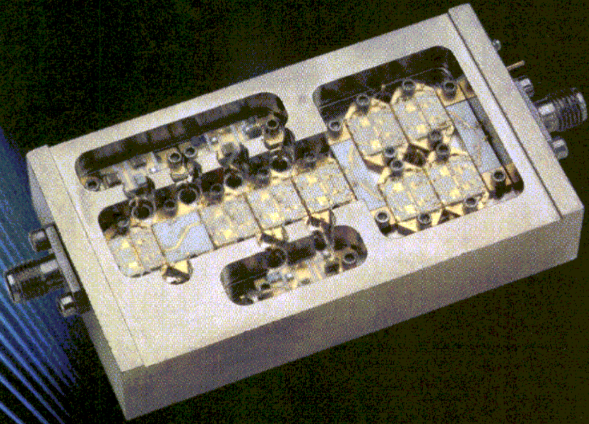
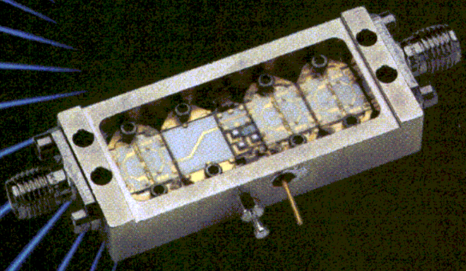
AMPLIFIERS

for every application







ULTRA BROAD BAND

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

MULTI OCTAVE AMPLIFIERS

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

MEDIUM POWER AMPLIFIERS

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

LOW NOISE OCTAVE BAND LNA'S

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

NARROW BAND LNA'S

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

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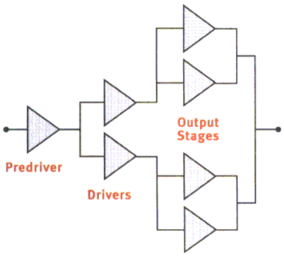

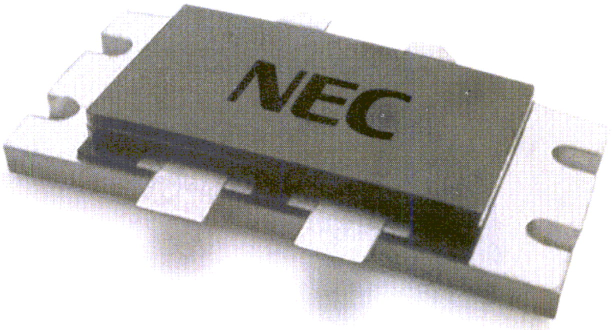

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51.5 dBm P_{OUT}, 9.5 dB Gain
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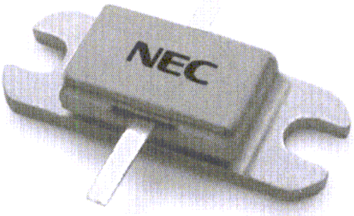

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40% Efficiency
for PCS, IMT-2000, WLL

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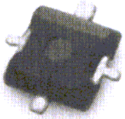
51.5 dBm P_{OUT}, 11 dB Gain
42% Efficiency *for PCS, DCS*



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35 dBm P_{1dB}, 10 dB Gain
50% Efficiency
for PCS, IMT-2000, WLL, WLAN

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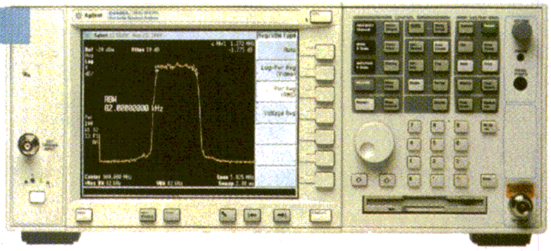
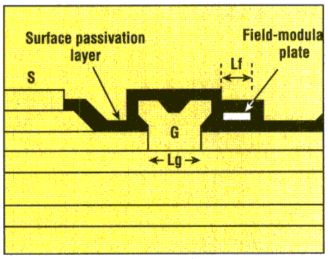
COVER FEATURE



DEPARTMENTS

NEWS



First Glimpse Of Future Devices

SPECIAL REPORT

DESIGN FEATURES

PRODUCT TECHNOLOGY

Penton

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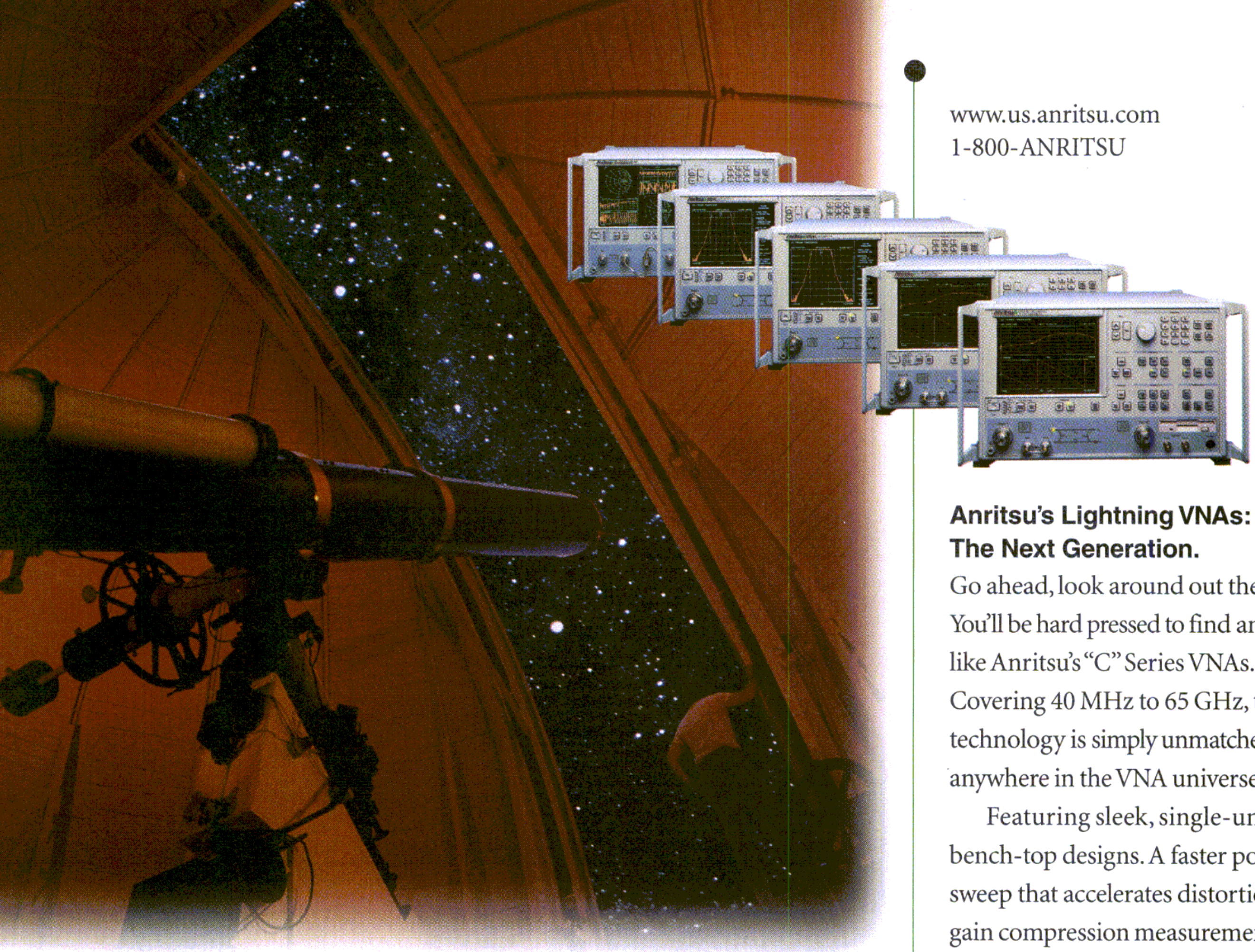

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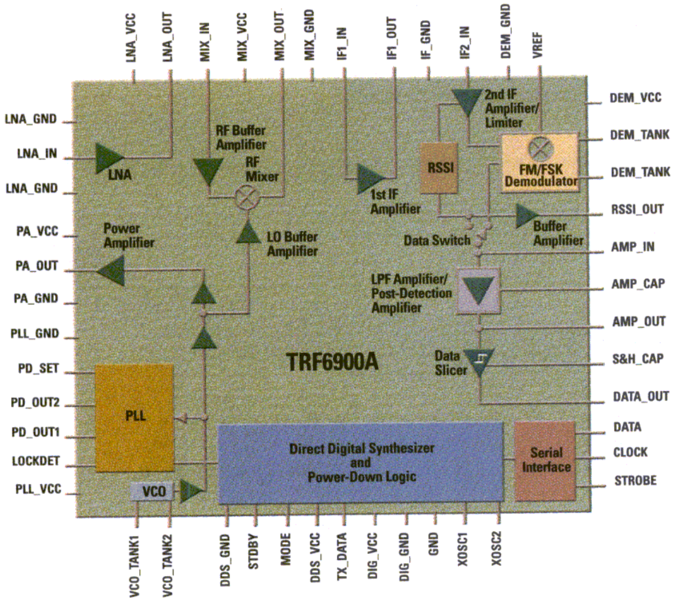
Industry's most integrated RF transceiver and RF transmitter ICs.



- *Single-chip solution with direct digital synthesizer supports wide-band applications with fixed or frequency hopping spread spectrum*
- *Integrated voltage controlled oscillator, phase-locked loop and adjustable power amplifier decreases board space*
- *TRF6900 contains integrated mixer, demodulator, data slicer and Received-Signal-Strength Indicator for interface with a microcontroller or DSP*
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- *30-μs frequency hops*
- *SW programmable frequency settings*
- *2.2-V to 3.6-V operation ensures low power consumption*
- *No external adjustments required*

Part Number	Function	Current Consumption		Output Power	Transmit	DDS Lock Time		Package	Price†
		Standby	*Active**			*Standby To rec*	*Standby To xmit*		
TRF4900	*Transmitter*	*0.5 μA*	*22 mA*	*7 dBm*	*FM/FSK*	*N/A*	*500 μs*	*24-pin TSSOP*	*$2.19*
TRF6900A	*Transceiver*	*0.5 μA*	*21 mA*	*4.5 dBm*	*FM/FSK*	*600 μs*	*500 μs*	*48-pin PQFP*	*$5.34*

†Price is per device in quantities of 1,000.

The ISM-band TRF6900 transceiver and the TRF4900 transmitter from Texas Instruments utilize cutting-edge technology to bring you the integration and low-power operation that your RF designs require. The TRF6900 and the TRF4900 include a fully programmable direct digital synthesizer (DDS) for fast, accurate frequency hopping. The small size, low power consumption and fast standby to transmit operation (500 μs) make these RF components ideal for battery-powered applications. From security systems and toys to utility metering devices and home automation, TI's integrated RF transceiver and RF transmitter ICs are the perfect fit.

48-pin PQFP shown actual size

TRF6900 evaluation module

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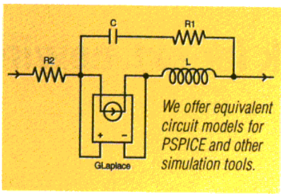

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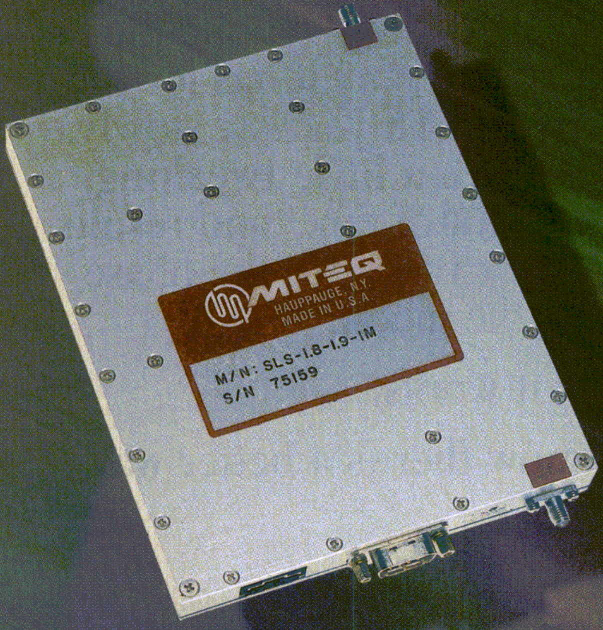


- Operating Bands From 1 – 15 GHz
- Ideal for Wireless Applications

OPTIONS

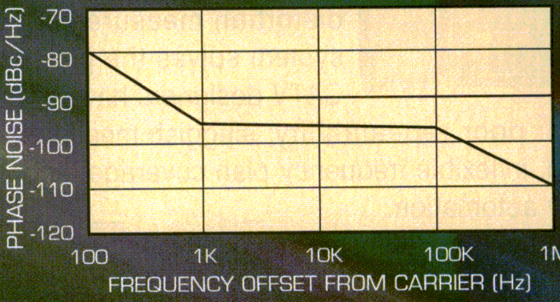
- 12 Volt Operation
- Optimized Bandwidth/Tuning Speed Combination

SPECIFICATIONS

MODEL	SLS SERIES
Frequency	1–15 GHz
Frequency step size	200 kHz to 10 MHz
Tuning range	Up to half octave
Switching speed	500 µs*
Output power	10 dBm min.
Output power variation	±2 dB min.
In band spurs	70 dBc min.
Harmonics	20 dBc
Phase noise	See graph
Reference	Internal or external
External reference Frequency Input power	 5/10 MHz 3 dBm ±3 dB
Frequency control	BCD or binary
DC power requirement	+15 or +12 volts, 200 mA 5.2 volts, 500 mA
Operating temperature	-10 to +60°C
Size	5" x 6.5" x 0.6"

* Acquire time depends on step size (low as 25 µs).

TYPICAL PHASE NOISE AT 2 GHz
(2 MHz Step Size)



3 YEAR WARRANTY

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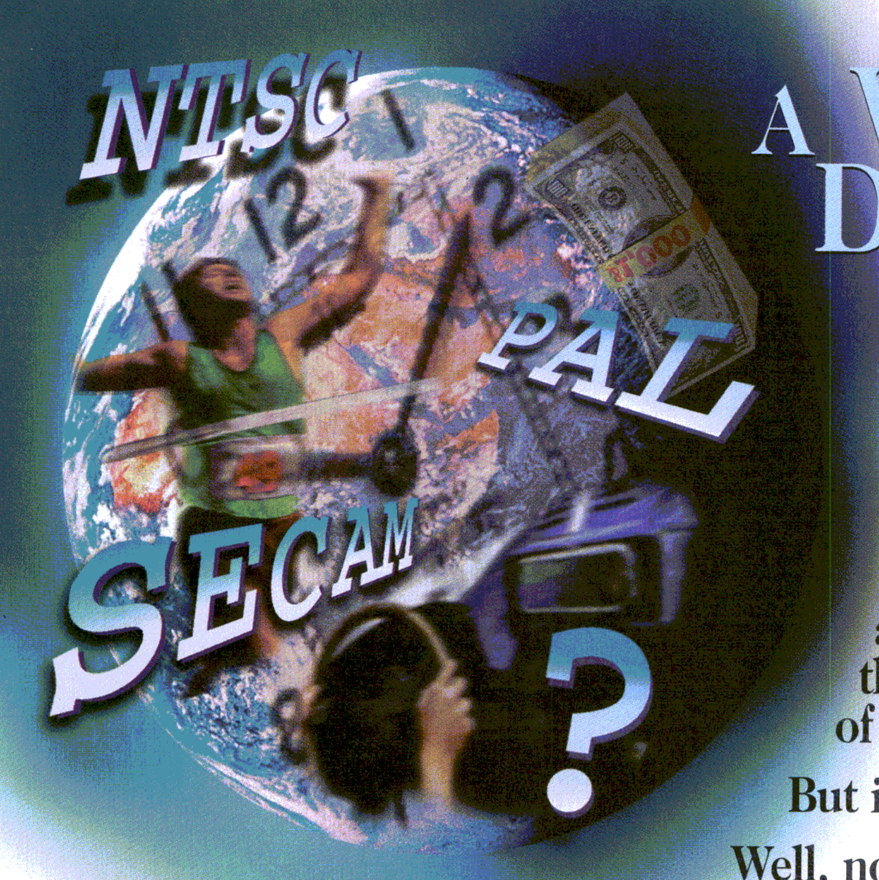

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The CTS performs CTB, CSO, cross-modulation, and C/N measurements in just minutes -- with no operator involvement -- that take our competitors hours with hands-on help. The CTS measurement results will be within ±0.5 dB today and every day. And unlike every other distortion test solution, the CTS measures performance according to virtually any frequency plan - no more creative data interpretation. It's also incredibly easy to use, thanks to its familiar "Windows" operator interface and comprehensive software. Even complex test routines can be run by almost anyone.

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[1] All seven tests (CTB, 4 CSOs, C/N, and XMOD) performed on any nine channels.

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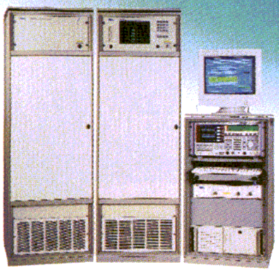

INTERFERING SOURCES

To the editor:

Recently, I read your article in the August 2000 issue on modeling Bluetooth systems (p. 121). I agree with your arguments since I had the exclusive opportunity in 1992 to study the millimeter-wave indoor propagation problems for the Japan Key Technology Center. During six months of an extensive joint research, I wrote a report which was never published whole, but I published some interesting results in 1994.

Since 1992, it took my Japanese colleagues a full seven years to operate a trial system at 60 GHz with the projected data rate over 100 Mb/s. While there is certainly a huge difference between Bluetooth's 11 Mb/s and >100 Mb/s, the propagation problems are similar.

As reported in the *Microwave Journal* in July 2000, in a paper by Buffler and Risman on compatibility issues at 2.4 GHz (p. 126), Bluetooth systems will also face huge interference from Bluetooth neighbors and other sources, such as microwave ovens. I also had an experience with microwave-oven harmonics interfering with C-band and Ku-band satellite receivers (Rxs).

In a room, home, or office, digitally modulated electromagnetic (EM) fields will always create standing-wave maxima and minima. If a person uses a cellular phone or similar device, he or she learns to slightly move away from a "bad" spot if noise interferes with the perceived voice. Machines should learn to behave in a similar manner. Therefore, I suggested in 1992 that "training" protocols be used to determine when valid data may be transmitted without errors. This idea, I think, will prove its preference in Bluetooth systems as it did in the 60-GHz fast data-transmission experiment.

Another idea of mine was that it makes no sense to transmit huge amounts of data through a bad environment. Why not keep the center base station somewhere in the home (e.g., close to a cabinet inlet) and control the data from cordless terminals through narrowband channels for voice and/or display?

Finally, I would like to comment about the concern of "brain damage by cell phones." I wonder why everybody fears a half-watt cell phone while, in many homes, microwave ovens run many times per day at almost 1 kW, and due to the long life of magnetrons, oven doors are often in poor shape and sometimes hundred-watt leaks are found. I have proof because I measured a 240-W CW leak from a new "professional" oven through a "service" opening in the case.

With the number of Bluetooth devices in a home, the power density will be higher than that of a cellular phone. Apparently, nobody is concerned—everybody wants to have more gadgets.

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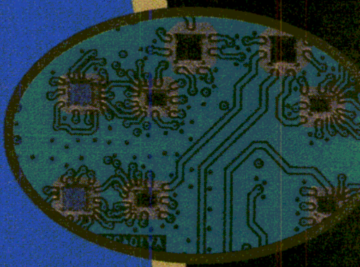

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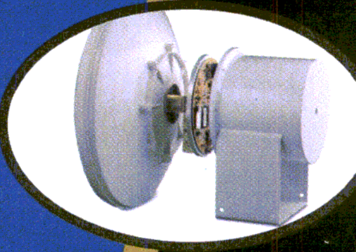

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We have said many times in this column that this is *your* magazine. The magazine represents a covenant of sorts between our editors and you. The subscription form is a way to renew your faith in the quality of information that we try to provide with each issue, to update us on where your career is taking you, and how we might help you with next year's technologies and applications.

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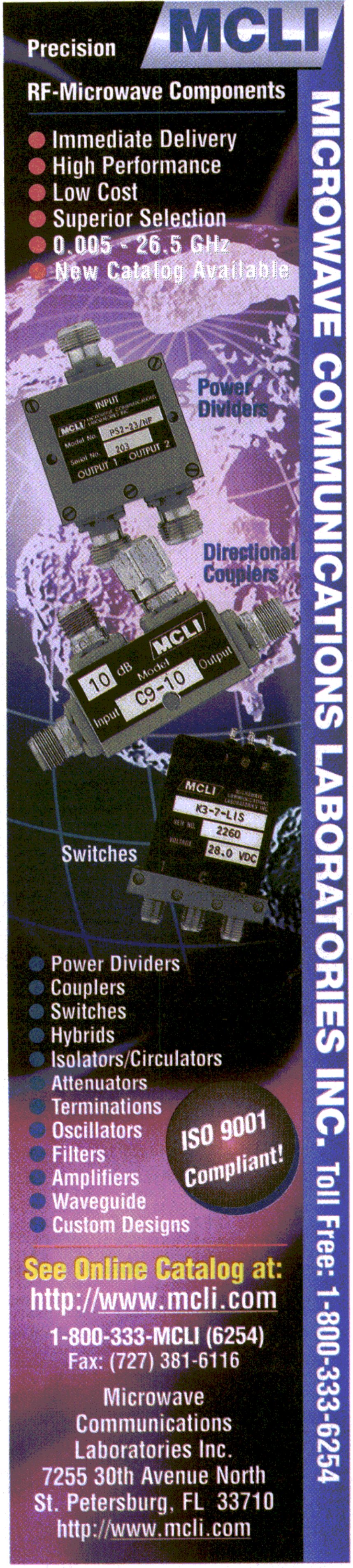
Jack Browne

Publisher/Editor



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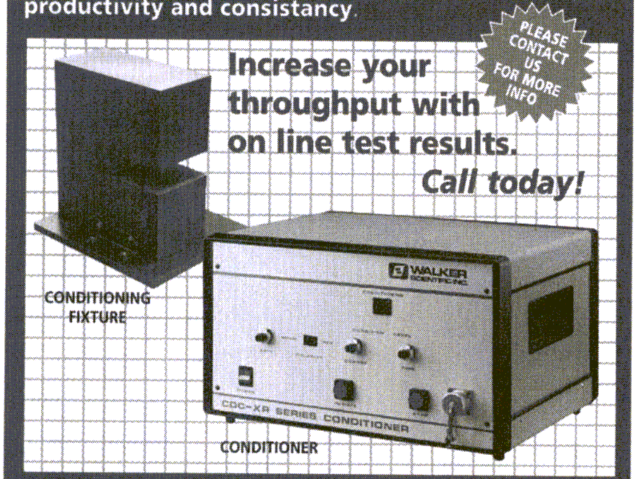
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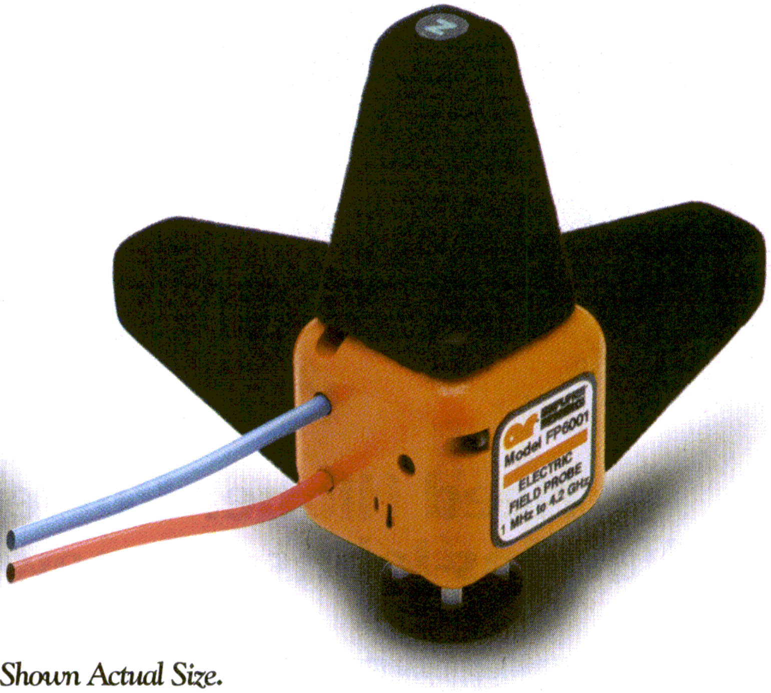

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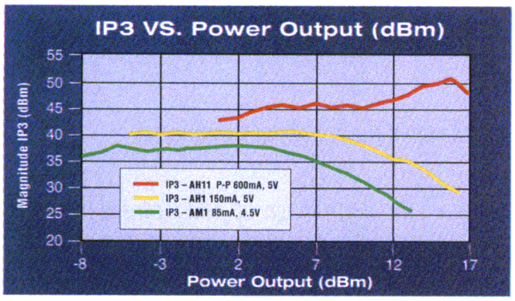

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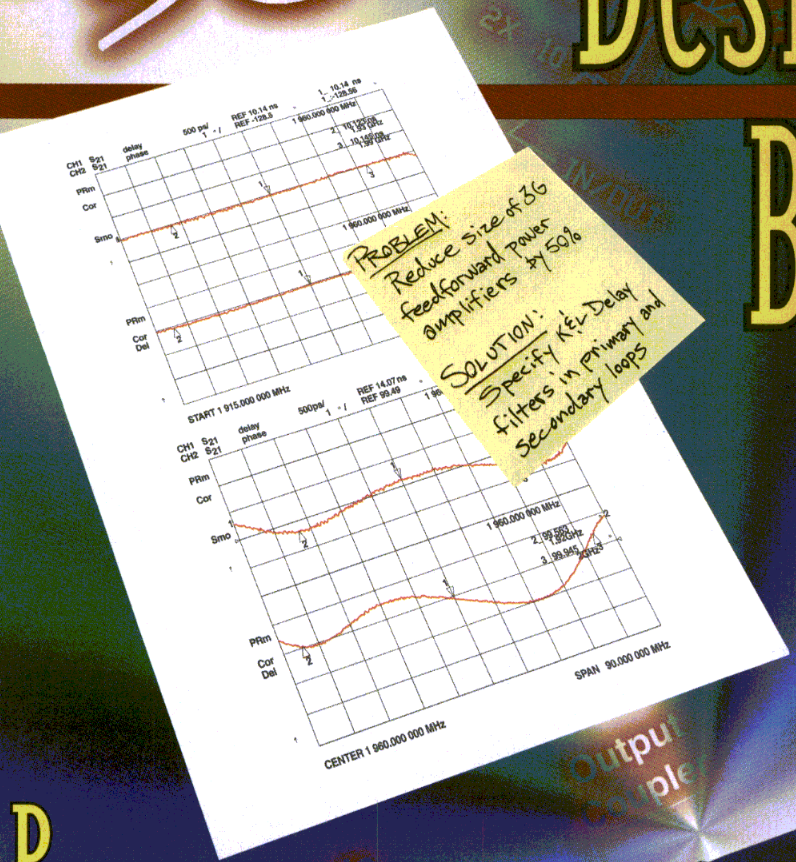

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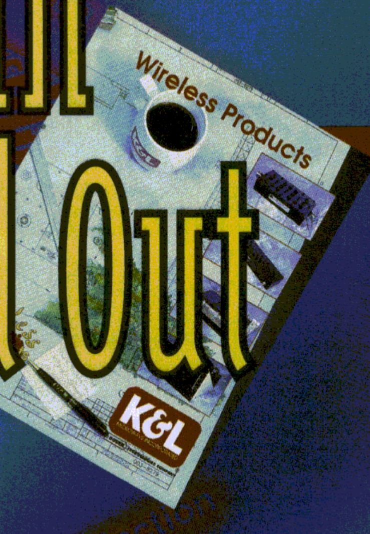

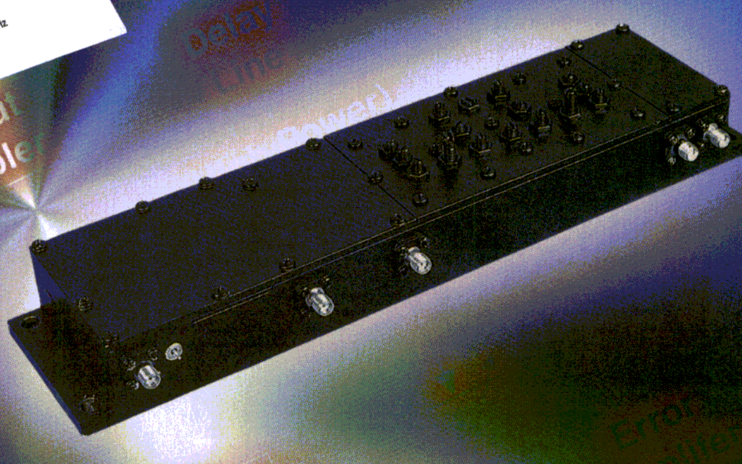

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SiGe Chips Are Commercially Available For Network Communications

SAN FRANCISCO, CA—IBM recently announced the commercial availability of next-generation silicon-germanium (SiGe) chip technology for high-frequency and low-power communications applications. IBM is the first chip manufacturer to offer multiple SiGe technology generations to the communications industry.

IBM's new SiGe technology can be readily integrated with standard complementary-metal-oxide-semiconductor (CMOS) circuitry, enabling circuit designers to incorporate additional logic and functions onto a single, high-performance communications chip. It has been tapped to fuel a variety of advanced customer chips, including radio transceivers and low-noise amplifiers (LNAs) for cellular telephones, Internet appliances, storage devices, and other critical mobile-communications applications. The technology, which is in volume production at IBM's Burlington, VT facility, also enables high-speed components already shipping in IBM's family of Deskstar disk drives.

"The race is on to develop communications chips that meet the exploding performance and connectivity demands of consumers, equipment makers, and service providers," says Dr. Bernard S. Meyerson, IBM Fellow and vice president of IBM's Communications Research and Development Center. "With multiple SiGe technologies, a full suite of design tools, and a significantly expanded R&D operation, we continue to strengthen our ability to help customers satisfy their networking and e-business requirements."

Based on an application-specific-integrated-circuit (ASIC)-compatible, 0.25-μm bipolar-CMOC (BiCMOS) manufacturing process, the new SiGe technology offers 47-GHz performance with a +2.5-VDC power supply. To speed the introduction of devices powered by this technology, IBM is offereing a state-of-the-art design kit based on software developed by Cadence Design Systems, Inc. Cadence and IBM are working closely together to ensure that the design kit and the associated environment for mixed-signal design enables IBM's customers to have a time-to-market advantage.

Measurement Experts Aid In Manufacture Of Baseball Bats

ENGLEWOOD, CO—Science is pushing baseball into new territory. Due to the work of measurement experts such as the father-and-son team of Wesley and Aaron Grenlund at Mu Research, batmakers are receiving new tools to sort good wood from bad wood before beginning the production process. The Grenlunds' research is also creating ways to characterize individual bats and engineer them to the demands of individual hitters.

In the research that they are performing for Johnson-Postman of Tacoma, WA, the Grenlunds apply some of the basic measurement techniques that electrical engineers use every day to characterize product designs. The Grenlunds simulate the system under test with an audio impulse, then use an accelerometer and an Agilent54645A scope to measure the resulting vibrations at various locations up and down the bat.

The Grenlunds knew that this test could save Johnson-Postman significant time and money, since the company's batmakers could tell in advance which wood billets would make good bats. Moreover, by characterizing the structural dynamics of a bat, they could maximize resonant deflections in the sweet spot and minimize them in the handle. The result would be fewer broken bats, cleaner hits, and fewer bone-jarring vibrations sent back into the batter's hands.

Viewed with a regular time-domain scope trace, however, the output signal was often incomprehensible, thanks to the harmonic complexity of the signal moving through the wood.

When the Grenlunds view their test results with a standard scope trace, the time-domain presentation is so complicated that it masks much of the information hidden in the signal. The picture changes immediately, though, when they switch on the Fast Fourier transform (FFT) function in their Agilent 54645A scope to transform the signal into the frequency domain. The spectrum plot clearly shows amplitude deviations at specific frequencies, corresponding to the changes in the structure of the wood (as might be the case with a knot, for instance). As Wesley Grenlund says, "The FFT is a powerful tool for all kinds of engineering problems. Looking at a signal in the frequency domain can quickly uncover important design issues that you could never hope to see in the time domain."

Patent Granted For Microtechnology Process

WEST CALDWELL, NJ—Merrimac Industries, Inc. announced that it has been granted a patent for its Multi-Mix™ Microtechnology from the US Patent and Trademark Office entitled "Method of Making Microwave Multifunction Modules Using Fluoropolymer Composite Substrates."

Multi-Mix Microtechnology is a new and proprietary solution for microwave, multilayer integrated circuits (ICs), and Micro-Multifunction Modules (MMFM™). This process is based on fluoropolymer composite substrates, which are bonded together into a multilayer structure using a fusion bonding process. The fusion process provides a homogeneous dielectric medium for superior electric performance at microwave frequencies. The bonded layers may incorporate embedded semiconductor devices, monolithic microwave integrated circuits (MMICs), etched resistors, active- and passive-circuit elements, and plated-through via holes to form a three-dimensional (3D) subsystem enclosure that requires no further packaging.

Mason N. Carter, Merrimac's chairman and chief executive officer (CEO), says, "There is a growing strategy by wireless companies and original equipment manufacturers (OEMs) across all RF microwave market segments to increase the outsourcing of multifunction modules while they focus on their redefined core competencies. This patent places Merrimac in an excellent position to capitalize on this growing demand for complex multifunction modules.

"One important aspect of this patent is the ability to form cavities for the subsurface attachments of active components such as MMIC devices. The ability to embed active and passive components in a totally enclosed package is proprietary to Multi-Mix," adds Carter.

Upgrades To Offer More Personalized Services To WAP Users

BOULDER, CO—SignalSoft Corp. announced that Siemens AG has selected SignalSoft's location technology to enhance its current deployments of WAP Gateways, improving the variety of services that mobile customers are currently using with their wireless-application-protocol (WAP)-enabled phones. Incorporating location into these WAP services makes them more relevant, practical, and convenient for consumers.

To deliver location-based WAP services, SignalSoft's location software is integrated with Siemens' WAP solution. The software allows mobile network operators to combine the location of its customers with Internet content for a broad range of data services available today. Siemens' WAP solution allows mobile customers to access the vast amount of information available on the Internet in a format that allows this information to be clearly and quickly displayed on a handset.

"SignalSoft recognizes the needs of mobile operators to drive usage and increase customer loyalty, so our products are specifically designed to allow operators to unlock the value of location—a core element of mobile telephony networks," says David Hose, SignalSoft's president and CEO. "Siemens' decision to enhance its WAP Gateways means operators can now personalize their WAP data services with location capability and quickly differentiate their data-service offering."

Industry Celebrates 100 Million Wireless Customers

WASHINGTON, DC—The Cellular Telecommunications Industry Association recently celebrated its 100 millionth subscriber in the US. The industry's growth is not expected to stop, or even slow, anytime soon. The 100 million wireless subscribers in the US are 36 percent of the nation's population. Yet the industry continues to grow at the astounding rate of 67,082 new wireless subscribers every day, or approximately one subscriber every 1.3 seconds. There are numerous choices for customers in today's wireless industry. Eighty-six percent of Americans can choose between three or more wireless carriers, while 60 percent of all Americans have a choice of six or more.

The rate at which cell phones have penetrated the US market is nearly unrivaled. The first cell phone became commercially available in 1983. Yet it took 91 years for landlines to reach 100 million customers and 54 years for televisions to reach 100 million households. Automobiles, meanwhile, took 68 years to reach 100 million homes.

"One hundred million Americans are a great landmark," says Tom Wheeler, president and CEO of CTIA. "But the wireless industry has not lost sight of the fact that while our technology empowers people to be their best, technology is nothing without a person guiding it."

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CIRCLE NO. 252

Large Growth Is Expected For IC Market Over The Next Five Years

OYSTER BAY, NY—Although it is the cellular market that is driving the wireless component market, broadband technologies will also become an important segment of this wireless communication integrated-circuit (IC) market, according to a study from Allied Business Intelligence (ABI). Chips for cellular/personal-communications-services (PCS) applications lead the market today with a 97-percent share, and silicon (Si) remains a dominant technology with an 86-percent share overall. Changes are evident in the Si realm as traditional bipolar-complementary-metal-oxide-semiconductor (BiCMOS), CMOS, silicon-germanium (SiGe), and bipolar processes are being pitted against one another.

According to the report, "Wireless Communications Systems ICs," the former dark horse of the wireless world, multichannel multipoint distribution service (MMDS), is now poised for explosive growth.

"ICs used in MMDS devices account for under 1 percent of the market in 2000, but the value of chips used in the application is expected to show impressive gains over the next five years," says ABI senior analyst, Andy Fuertes, the report's author. "ICs for wireless LANs are also showing considerable growth and, more importantly, the application is rendering itself a test bed for widespread implementation of SiGe chips through Intersil's PRISM solution."

The overwhelming leader in the market today is CMOS, due to its role in the baseband sector. CMOS represents approximately 50 percent of the overall market. However, considerable gains are expected on the part of SiGe, particularly in the intermediate-frequency (IF) sectors of wireless devices. SiGe will be less successful in the power-amplifier (PA) sector, mainly due to the power-added efficiency (PAE) of gallium-arsenide (GaAs) devices. SiGe will be particularly attractive as cellular and other original equipment manufacturers (OEMs) begin to implement Bluetooth and Global Positioning System (GPS) functionality in their devices.

Kudos

SAL, Inc. was recently granted a Japanese patent for SAL's proprietary-alignment patterns and alignment method, an enabler for advanced X-ray stepper performance. This patent covers intellectual property (IP) already protected by American and European patents. The SAL ALX70 automatic alignment system uses an advanced pattern-recognition system and high-speed digital signal processors (DSPs) to quickly and accurately align patterns on X-ray masks with patterns on wafers...Intersil Corp. announced that the Personal Computer Memory Card International Association (PCMCIA) card implementation of its PRISM II reference design has received certification from Microsoft's Windows Hardware Quality Labs (WHQL). The purpose of Microsoft's WHQL is to ensure the compatibility of hardware with Microsoft Operating Systems...Silicon Wave, Inc. has been granted a new position as Associate Member of the Bluetooth Special Interest group (SIG). The Bluetooth SIG recently created Associate Member standing to offer innovative companies, such as Silicon Wave, a greater role in defining the Bluetooth specification. Associate members can participate in working groups only if granted permission from the Bluetooth SIG and Promoter companies...Channel Master, LLC announced International Telecommunications Satellite Organization (INTELSAT) approval of its .96m Receive-Transmit Offset antenna system in Classes I and II and also its 1.2m Receive-Transmit Offset antenna system in Classes I, II, and III. All of these antenna types have been approved under INTELSAT's Standard G for Ku-Band applications...U.S. Wireless Corp. has been added to the Russell 2000® and Russell 3000® Indexes. The Russell 3000 is a measure of the 3000 largest publicly traded US companies in the US stock market, ranked by market capitalization as of May 31. The Russell 2000 measures the performance of small-cap companies in the Russell 3000 Index...RF Micro Devices, Inc. has again been recognized as one of the fastest-growing technology companies in North Carolina. During the past fiscal year, RFMD posted the second-highest five-year growth rate in the state—achieving an increase of 8955 percent. In recognition of that growth, the company has received a "North Carolina Technology Fast 50" award as part of a national program sponsored by Deloitte & Touche, LLP...Ericsson received the 2000 CDMA Industry Achievement Award for Infrastructure Product Innovation at a ceremony held in Hong Kong.

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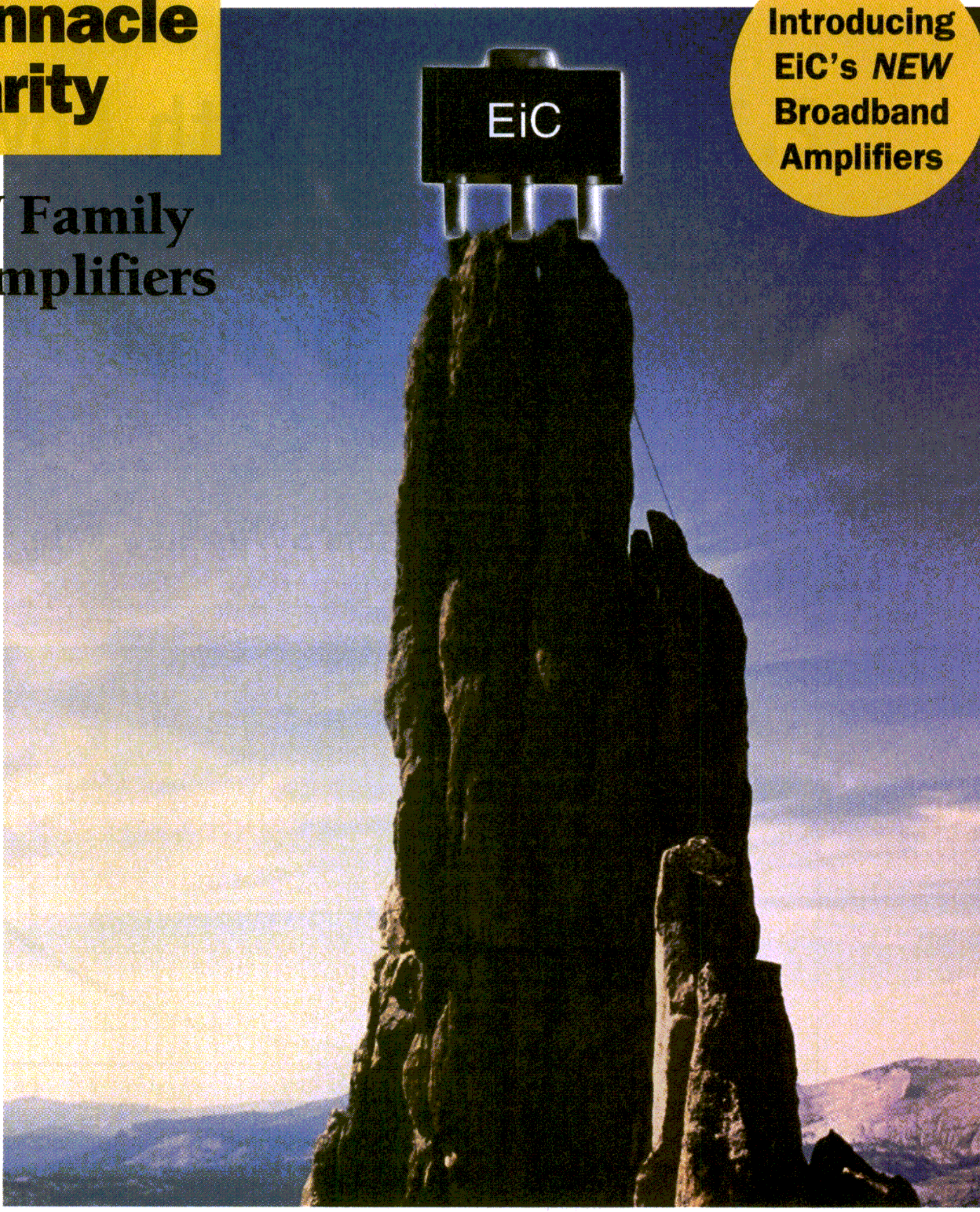

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ECG-004	15dB	12dBm	26dBm	280° C/W	35°C	DC-6 GHz
ECG-002	20dB	15dBm	29dBm	233° C/W	40°C	DC-6 GHz
ECG-006	15dB	15dBm	30dBm	278° C/W	50°C	DC-6 GHz
ECG-003	20dB	23dBm	39dBm	50° C/W	45°C	DC-3 GHz
ECG-008	15dB	23dBm	40dBm	55° C/W	55°C	DC-3 GHz
ECG-009	19dB	24dBm	41dBm	85° C/W	65°C	DC-2 GHz
ECG-011	20dB	8dBm	20dBm	355° C/W	47°C	DC-6 GHz
3V ECG-012	14dB	20dBm	36dBm	120° C/W	45°C	DC-2.5 GHz
EC-1089	15dB	23.5dBm	>42dBm	~ 85°C/W	~ 65°C	DC-2.5 GHz
EC-1019	18.5dB	19dBm	34dBm	120°C/W	40°C	DC - 3 GHz
EC-1078	19.5dB	21dBm	37dBm	120°C/W	60°C	DC - 3 GHz
EC-1119	14.8dB	18.6dBm	36dBm	150°C/W	60°C	DC - 3 GHz

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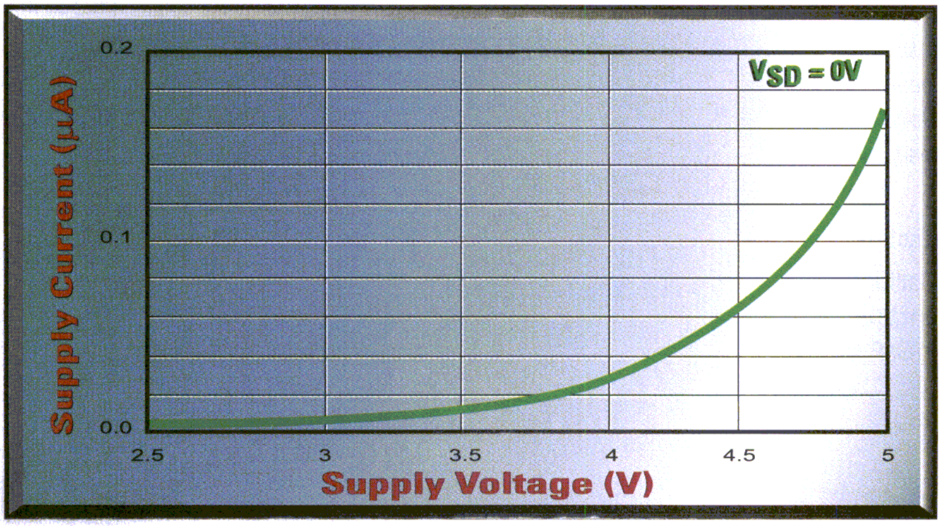

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CIRCLE NO. 278

NEWS

IEDM Preview

The 46th Annual IEDM features an impressive collection of technical papers from around the globe, with presentations on analog and digital devices for next-generation circuits.

IEDM Offers First Glimpse Of Future Devices

JACK BROWNE

Publisher/Editor

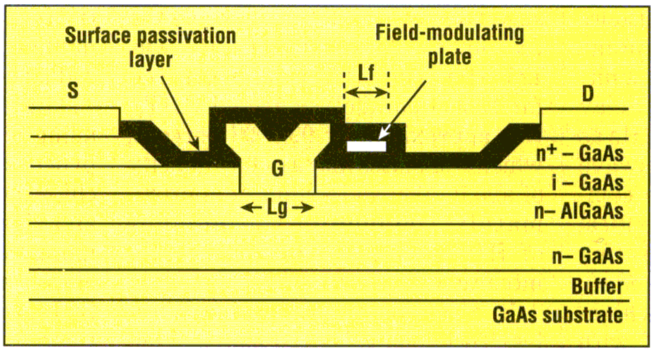
RESEARCH on semiconductor-device technology represents a massive amount of information each year. But if an engineer only had a few days to catch up with the state of the art, there might be no better way to spend that time than at the 46th Annual IEEE International Electron Devices Meeting (IEDM). Scheduled for December 11-13, 2000 at the San Francisco Hilton and Towers (San Francisco, CA), this is the premier conference on semiconductor-device technology, covering everything from advanced analog and digital integrated circuits (ICs) to the latest process and modeling techniques.

Attendees can visit several short courses the day before the formal opening of technical sessions. Short courses scheduled for Sunday, December 10, include a session on "Advanced Interconnects: Design, Process, and Integration," organized by Jim Ryan of IBM (Hopewell Junction, NY). The course description notes that, in order for semiconductor technologies to advance, equal research time must be given to interconnect performance. The course describes the materials, processes, manufacturing techniques, and design trends prevalent in interconnect technology, with a focus on copper (Cu) and low-dielectric-constant systems.

An additional short course, "Technology For The Internet Era," organized by Sam Shichijo of Texas Instruments (Dallas, TX), projects the growth of products designed for fast Internet access. Included in these product offerings are wireless and wired networks, circuit designs for low-power systems, and devices using system-on-a-chip (SoC) technologies.

The IEDM's plenary session includes an interesting talk by David DiVincenzo of the IBM Thomas J. Watson Research Center (Yorktown Heights, NY) on prospects for quantum computing. In light of the continuing need for faster electronic devices and higher-power computers, quantum devices may offer a solution. By making digital bits that obey quantum-mechanical principles, devices and computers could be constructed where digital words are represented by coherent superpositions of discrete quantum states. Not only does this imply an increase in computing speed, but also the possibility of creating efficient algorithms for otherwise intractable problems, such as factoring with prime numbers. This session explores the requirements for the transistors needed in a quantum computer. For the most part, such transistors will be different than any that have come before. A digital bit will be embodied in a single microscopic degree of freedom, such as in the spin of a single trapped electron.

Also of note, an evening panel discussion on Tuesday, December 13th, addresses rapid advances in optical communications and what types of technologies will be suitable for tomorrow's high-speed systems. Entitled "Beyond 40 Gb/s: What Technologies Will Work For These Applications?," the session's notes indicate that there is already talk about 160-Gb/s optical communica-



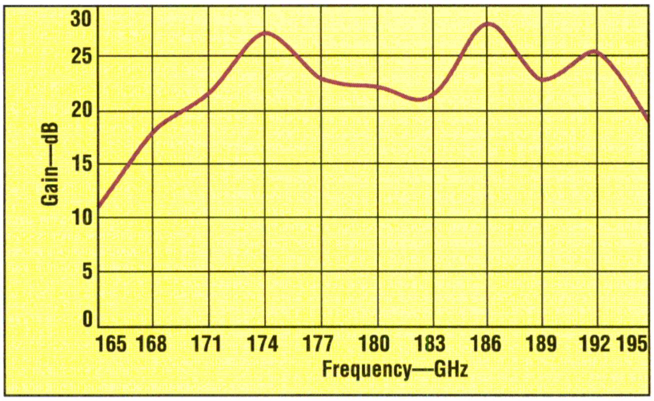
1. The FP-HFET was developed by researchers at NEC Corp. (Shiga, Japan) to achieve high breakdown voltage and high output-power density. The device's saturated output-power density is reported to be 0.67 W/mm of periphery. It employs gate and FP lengths of 1.0 and 0.5 μm, respectively.

tions systems. The panel session will explore the possibilities of using complementary-metal-oxide-semiconductor (CMOS) or silicon-germanium (SiGe) technologies for higher-speed optical technologies, and examine the trade-offs at different speeds, such as 80 and 160 Gb/s.

An additional evening panel session, "Introduction To The Emerging Technologies Session," follows a format of blending fact with opinion. With panel members from such organizations as Cornell University, Fujitsu, IBM, Infineon, and Toshiba, the session intends to expose attendees to "far out" thinking (beyond current industry capabilities) on new technologies, including single-electron devices, alternative memory technologies, molecular electronics, and even new materials, such as Ga nitride (GaN) and Si carbide (SiC).

One of the materials that has caught the attention of the high-frequency industry these past several years is SiGe. The material is highlighted in a presentation by Katsuyoshi Washio and fellow researchers from the Central Research Laboratory of Hitachi Ltd. (Tokyo, Japan). Their presentation, "A 0.2-μm 180-GHz f_{max} 6.7-ps-ECL SOI/HRS Self-Aligned SEG SiGe HBT/CMOS Technology for Microwave and High-Speed Digital Applications," addresses the need for improved yield in SiGe processes. The authors cite SiGe heterojunction bipolar transistors (HBTs) with 5.5-ps emitter-coupled-logic (ECL) gate delay and a cutoff frequency of 122 GHz as promising candidates for microwave and millimeter-wave wireless communications systems and for optical communications systems operating beyond 10 Gb/s. What is needed, however, is the integration of a high-breakdown-voltage SiGe HBT with high-precision resistors, varactor diodes, and other components in a high-yield process.

In answer to this need, the authors have developed a 0.2-μm self-aligned, selective-epitaxial-growth (SEG) SiGe HBT with poly-Si resistors, a four-level metal-layer system with metal-insulator-metal (MIM) capacitors using a Si-on-insulator (SOI) high-resistivity substrate (HRS). The process has already demonstrated SiGe HBTs capable of maximum frequency of oscillation (f_{max}) of 180 GHz and ECL gate delay of 6.7 ps. The process employs shallow-trench and deep-trench isolations. The MIM capacitors were formed with the first and second metal layers, while inductors were formed with the third and fourth metal layers. The breakdown voltage for high-speed HBTs was +2.5 VDC, with a breakdown voltage of +3.9 VDC for HBTs designed for high voltages. The peak cutoff frequencies (f_T) for the high-speed and high-voltage HBTs were 76 and 47 GHz, respectively. The f_{max} for the high-speed device was 180 GHz, while the f_{max} for the high-voltage HBT was 125 GHz. The HBT yield, measured for 10,000 parallel-connected HBTs, was approximately 99.9997 percent, which translates into a defect density of about 2000 and a limit of approximately 100,000 on the number of HBTs that can be practically integrated.



2. Improvements to a InGaAs/InAlAs/InP process enabled the fabrication of a two-stage balanced microstrip MMIC amplifier capable of 10-dB gain and 7-dB noise figure at 170 GHz.

In a session on "Emerging Wireless Technology," N.P. Pham and fellow researchers from the Delft University of Technology (Delft, The Netherlands) explain the use of micromachining techniques to fabricate a highly integrated module with RF Si technology. The researchers' novel micromachining post-processing module is used to structure the Si from a wafer's backside with precise alignment to the frontside to achieve reduced crosstalk through the conductive Si substrate, create an RF ground on the substrate, and produce highly integrated transmission lines and three-dimensional (3D) structures to reduce the overall size of the chip. The micromachining post-process module was used to fabricate microstrip lines with 130-μm spacing between the signal and ground. The transmission lines exhibit residual loss of about 3 dB/cm at 17 GHz on a 3000-Ω-cm Si substrate.

In the same session, Kihong Kim and fellow researchers from the University of Florida (Gainesville, FL) disclose techniques for wireless communications between chips. They cite wireless clock distribution as a potential application, where an external microwave clock transmits a signal at 20 GHz or higher to a grid of integrated clock receivers (Rxs). The signal received by the receiving antenna is divided down to the system-clock frequency and locally distributed using a conventional clock tree structure. The use of the high-frequency clock distribution system reduces the antenna size due to the decreased wavelength. The presenters explain how the same approach can be used to improve on-chip antennas and how the technique might be used for communications in an area around the radius of 2 cm.

The antennas were fabricated using a single aluminum (Al) level process with no passivation. Test chips included linear, meander, loop, and zigzag dipole structures. Since metal structures in the vicinity of the chips can cause antenna losses, an SOI substrate was chosen with continuous buried oxide layer for wave propagation. If antennas are constructed using the lowest-level metal layer in contact with the Si layer, it would be possible to send and receive signals through the buried oxide layer. By fabricating numerous

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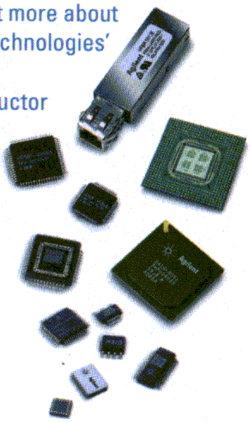

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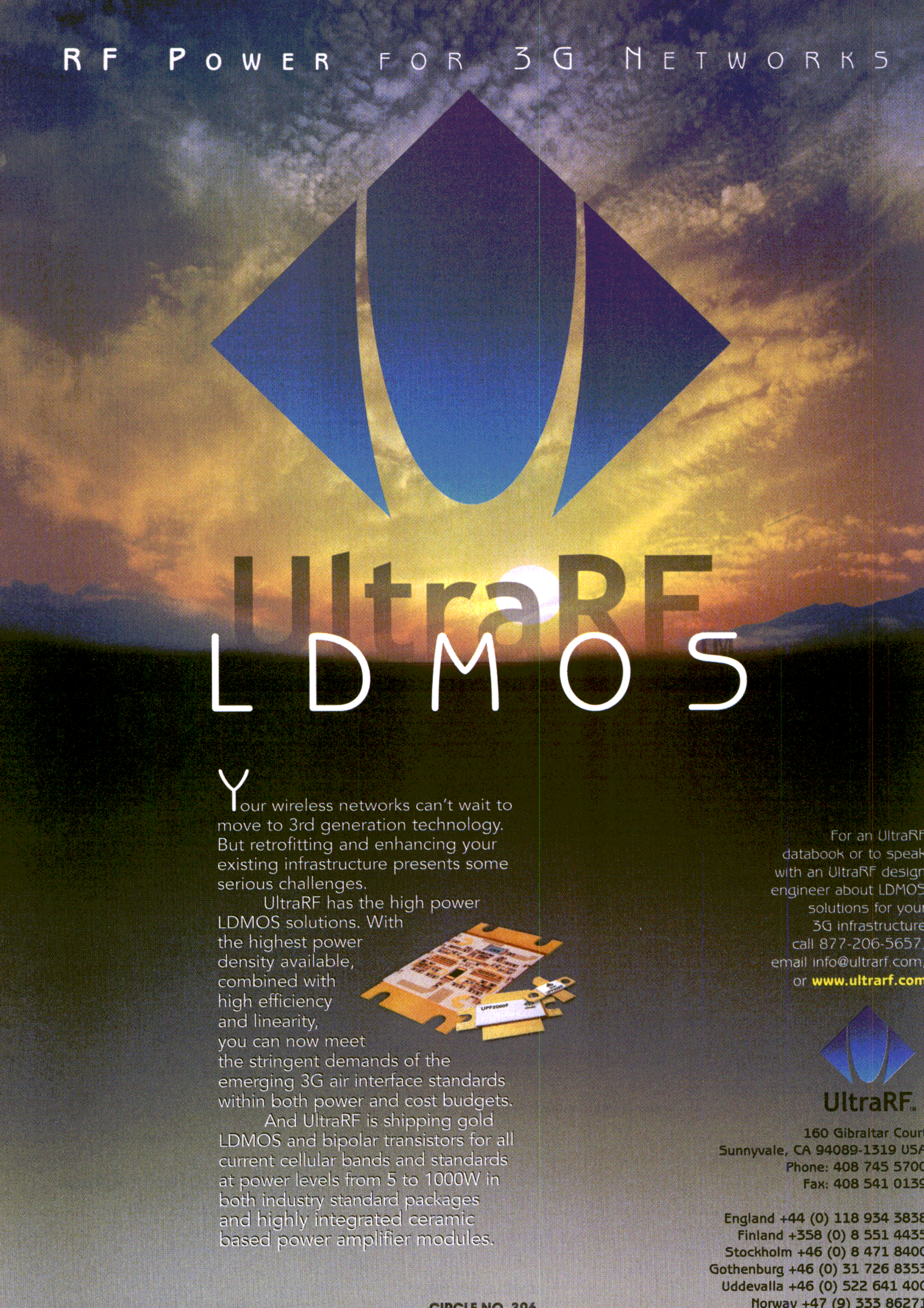

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CIRCLE NO. 306

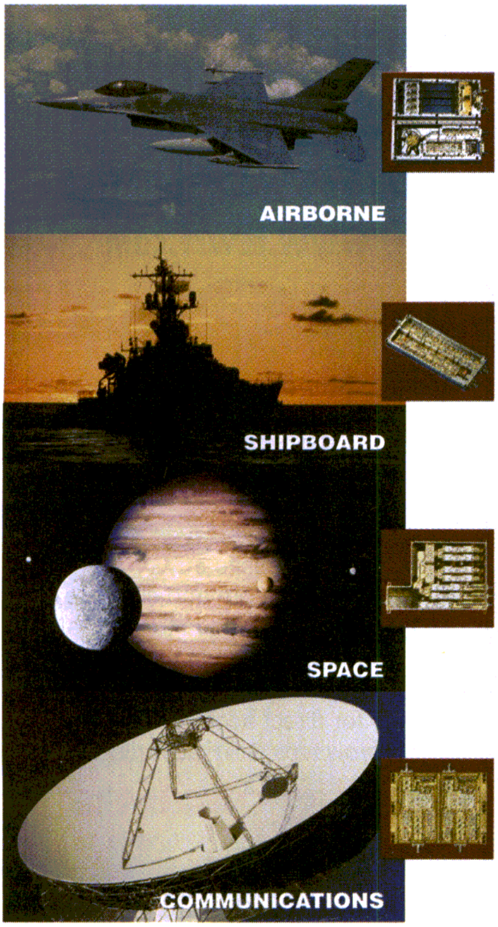
antenna structures, including zigzag dipole antenna pairs, which were chosen for use in a 0.25-µm CMOS 7.4-GHz clock Rx. Since a 2-mm zigzag dipole antenna pair with a 2-cm separation provides –57-dB gain at 17 GHz, and the effects of metal-interference structures might decrease gain by 5 dB, the final gain should be approximately –62 dB. As a result, a clock Rx with a zigzag dipole antenna pair 2 cm away from a transmitter (Tx) with an isotropic loop antenna should be able to lock to a +20-dBm clock signal delivered to the transmitting antenna.

In a session entitled "GaN and RF Power Devices," K. Matsunaga and co-workers from NEC Corp. (Shiga, Japan) describe a low-distortion 230-W GaAs heterostructure-field-effect-transistor (HFET) PA for cellular base stations. The amplifier, which consists of four 86.4-mm-gate-width field-modulating-plate (FP) HFET chips (Fig. 1), provides 230-W output power (+53.6 dBm) at 2.1 GHz with 11-dB linear gain and 42-percent power-added efficiency (PAE) when operating at a drain-bias voltage of +22 VDC. When operating with +46-dBm output power, the same amplifier achieved low adjacent-channel power ratio (ACPR) of –35 dBc with 25-percent PAE. The device features saturated power density of 0.67 W/mm. The presenters note that this is the highest power density ever reported for power FETs delivering more than 200-W output power. When operating under "backed-off" bias conditions, the devices and amplifier show promise for low-distortion, high-voltage applications in cellular base stations. For example, when output power is backed off by 8 dB (from almost +54 dBm to +46 dBm), the amplifier exhibits a linear improvement of approximately 4 dB in ACPR performance, when the drain voltage is increased from +12 to +20 VDC.

In the same session, results are presented on the highest power levels yet achieved from a GaN-based PA. Y.F. Wu and fellow researchers from Cree Lighting (Goleta, CA) offer details on an AlGaN/GaN high-electron-mobility-transistor (HEMT) amplifier capable of 50-W pulsed output power at 6 GHz, or about six to 10 times the output levels possible with GaAs devices of the same size. The HEMT devices were grown on SiC substrates through metal-organic-chemical-vapor deposition (MOCVD).

When biased at +15 VDC, the amplifier delivered 13.2-dB small-signal gain at 5 GHz, which increased to 14.6-dB small-signal gain at bias voltages exceeding +25 VDC. Pulsed measurements were performed with signals having 0.5-µs pulse width and 5-percent duty cycle. The output power at 6 GHz was 35 W when biased at +29 VDC, rising to 49 W with +38-VDC bias. At +39-VDC bias, an output-power level of 51 W was recorded at 6 GHz.

Researchers from Cornell Univer-



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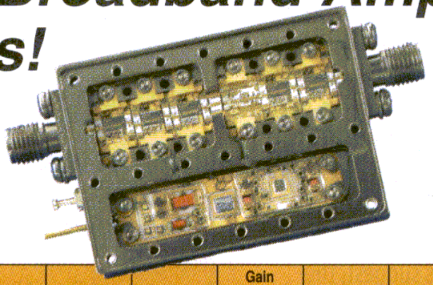

Model Number	Freq. Range (GHz)	Gain dB (min)	Noise Figure dBm (max)	Gain Flatness ±dB	1 dB Comp dBm (min)	VSWR max In/Out	3rd Order Ip dBm (typ)	DC Current +12 to +15 VDC (typ)
AK2018101	2.0–18.0	16	4.0	2.0	+10	2.0:1	+20	100
AK2018102	2.0–18.0	24	4.0	2.0	+10	2.0:1	+20	200
AK2018133	2.0–18.0	35	4.0	2.0	+13	2.0:1	+23	300
AK1018141	1.0–18.0	16	4.0	1.5	+14	2.0:1	+24	100
AK1018142	1.0–18.0	24	4.0	2.0	+14	2.0:1	+24	200
AK2060101	1.0–18.0	35	4.0	1.5	+10	2.0:1	+20	300

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sity (Ithaca, NY) also brought details on AlGaN/GaN HEMTs to the IEDM. E.M. Chunbes and co-workers presented improved performance over previous devices through the use of Si_3N_4 passivation to provide a fixed positive charge. The researchers fabricated metal-insulator-semiconductor (MIS) FETs (MISFETs) on surface-passivated, undoped AlGaN/GaN heterostructures with enhanced performance compared to conventional unpassivated HEMTs. Growing their devices with organometallic vapor-phase-epitaxy (OMVPE) techniques on 2-in. (5.08-cm)-diameter sapphire substrates, the researchers produced passivated, recessed-gate devices with 0.6-µm gate lengths and 100-µm gate widths. Conventional HEMTs were also fabricated on similar heterostructures, but without passivation or the use of recessed gates. Class A power sweeps performed with the passivated devices, using +20, +25, and +28-VDC bias, yielded output-power densities from 2.8 to 4.2 W/mm and PAEs of 35 to 37 percent. The devices yielded 21-W/mm output-power density at 7 GHz, with 289-percent PAE measured with a bias voltage of +15 VDC. These numbers are very close to predicted curves for the passivated devices, in contrast to a measured output-power density of 1.5 W/mm for the unpassivated HEMTs, compared to the 2.1 W/mm from predictions.

Some of the highest-frequency devices reported at the IEDM were presented in a session on "High Speed Compound Semiconductor Devices." R. Lai and co-workers from TRW Semiconductor Products Sector (Redondo Beach, CA) offered details on the development of a indium-phosphide (InP) HEMT amplifier for G-band (140-to-220-GHz) applications. The TRW researchers revealed recent advances in InGaAs/InAlAs/InP HEMT monolithic-microwave-integrated-circuit (MMIC) technology that have made possible the development of a series of G-band amplifiers with outstanding gain characteristics through 220 GHz. In order to develop a successful process for use at G-band, several process enhancements were implemented on the firm's baseline 75-mm-diameter InP HEMT MMIC process. The first enhancement involved the growth and design of pseudomorphic high-indium-composition InGaAs channels and an improved HEMT device structure for higher frequencies. Cutoff frequencies of 300 GHz have already been achieved with In compositions of greater than 70 percent. The second process enhancement involved the reduction of the gate length from 100 to 70 nm. This yielded a 15-to-20-percent improvement in the cutoff frequency and transconductance. The was the development of a 50-µm-thick substrate backside process with very small through-substrate viaholes for grounding. This third enhancement was one that prevented substrate

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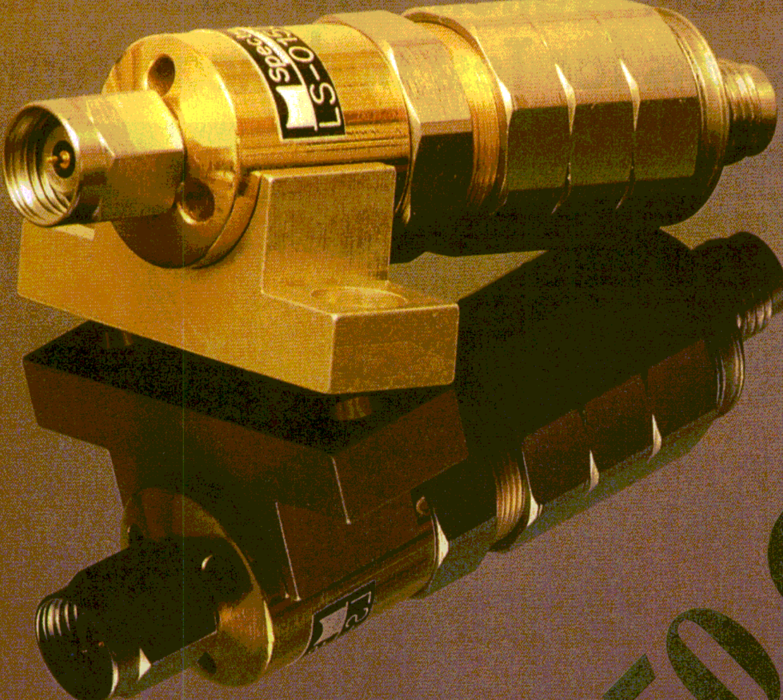

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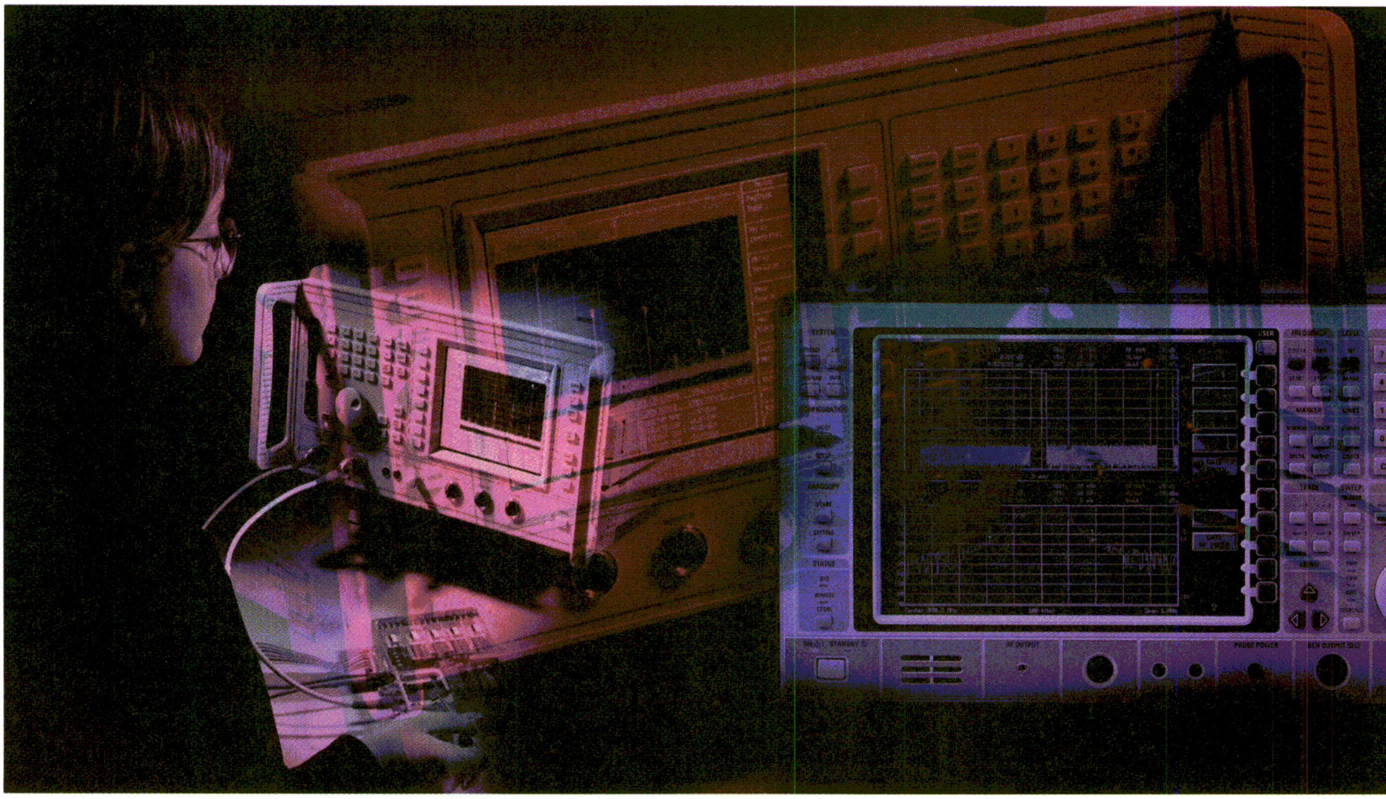

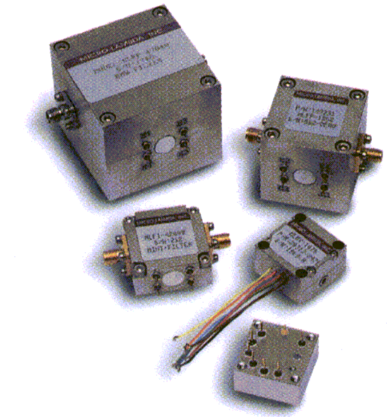

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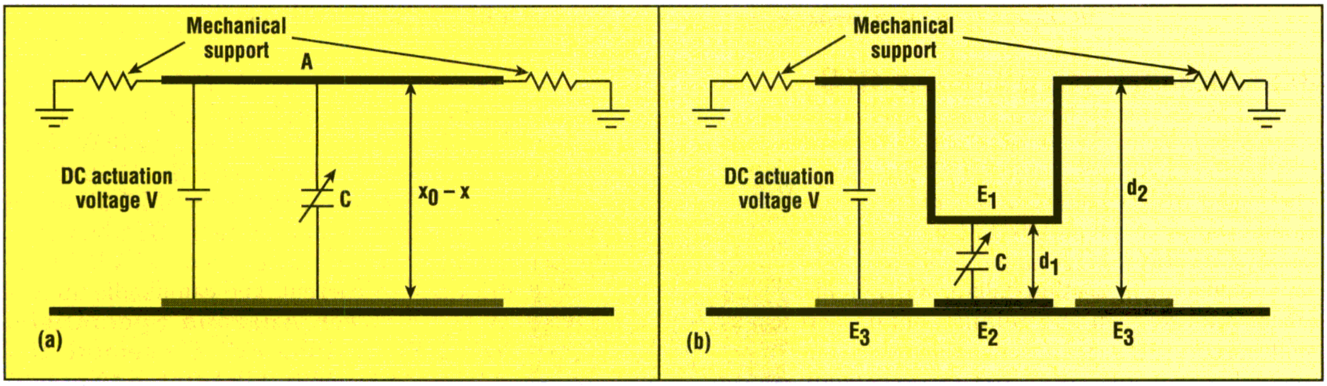

3. A conventional parallel-plate tunable capacitor (a) is limited in tuning range when compared to the novel wide-tuning-range capacitor configuration (b) developed by the University of Illinois researchers.

waveguide-mode propagation and supports minimal device-source inductance to maximize device gain at high frequencies.

Since the process has demonstrated high device available gain of 7 to 8 dB per stage at 200 GHz, it was possible to fabricate complete amplifiers at G-band. A series of amplifiers was fabricated, including a three-stage, single-ended microstrip MMIC amplifier with 10-to-12-dB gain and a noise figure of 5.1 dB at 151 GHz. A two-stage balanced microstrip MMIC amplifier from 165 to 190 GHz demonstrated 10-dB gain and 7-dB noise figure at 170 GHz (Fig. 2). Another three-stage microstrip amplifier yielded 14-dB gain and 7-dB noise figure from 160 to 190 GHz. Two-stage and four-stage amplifiers were cascaded to achieve 20-to-25-dB gain from 170 to 190 GHz, with broadband noise figures of 8 dB or better. A six-stage coplanar-waveguide MMIC amplifier offered 20-dB average gain from 160 to 215 GHz, with a noise figure of 8 dB at 170 GHz. The researchers note that the measured gain of 15 dB at 215 GHz is the highest-gain solid-state amplifier yet demonstrated at that frequency.

In the same session, M.W. Dvorak and fellow students from the Compound Semiconductor Device Laboratory (CSDL) of Simon Fraser University (Burnaby, British Columbia, Canada) presented the school's work on InP-based double-heterostructure bipolar transistors (DHBTs) with high breakdown voltages and cutoff frequencies up to 250 GHz. The presenters note that the process promotes gains in performance and manufacturability by eliminating the need to grow graded emitter-based and base-collector junctions, enables the growth of entirely Al-free transistor structures, and uses carbon (C)-based doping without hydrogen (H)-passivation effects. The presentation includes results on InP/GaAsSb/InP DHBTs implemented within thin 200 to 250 Angstrom strainged $GaAs_{0.6}Sb_{0.4}$ highly doped base layers and a 2000-Anstrom InP collector. The devices exhibit peak cutoff frequencies of 250 GHz and maximum frequency of oscillation of better than 175 GHz with a breakdown voltage of +6 VDC.

Harkening back to the plenary session, a presentation by Yukinori Ono and Yasuo Takahashi of NTT's Basic Research Laboratories (Kanagawa, Japan) explores the workings of a single-electron transistor (SET). Part of a session on nanoelectronics, the researchers note that such a transistor is promising for future large-scale-integration (LSI) circuits due to its potential for very-low power consumption. The researchers constructed an elemental logic circuit on an SOI substrate to represent such a transistor, using two nearly identical single-electron transistors, each with an individual input gate. By sending pass signals to the sources, and complementary signals to the gates, basic logic operations including half-sum and carry-out of the half adder can be achieved at 25 K.

Much has been made in recent years of the possibilities of using microelectromechanical systems (MEMS) in electronic circuits, for example, to tune microminiature filter and resonator structures. In a session on MEMS and biochemical sensors, Jun Zou, Change Liu, and Jose Schutt-Aine from the University of Illinois (Urbana, IL) report on the development of a MEMS-tunable capacitor for wireless communications applications. The most common MEMS tunable capacitor to date has been the parallel-plate type with electrostatic-actuation tuning. Spacing and capacitance values are changed through the DC bias voltage applied across the two plates. However, the tuning range is affected by the "pull-in" effect. When the spacing is reduced to two-thirds of the maximum value, the electrostatic force will exceed the mechanical restoring force and the two plates will snap into contact. Thus, the snap-in effect imposes a theoretical limit of 50 percent on the tuning range of conventional parallel-plate tunable capacitors (Fig. 3a).

However, the University of Illinois team reports on the first parallel-plate tunable-capacitor design to exceed the 50-percent limit imposed by the pull-in effect (Fig. 3b). Through simulation, the presenters report a maximum possible tuning range of 90.5 percent and a nominal capacitance of 0.0328 pF. Measured results show near-ideal capacitance behavior, with a return loss of less than 0.6 dB at 10 GHz.

For more information on the IEDM, please contact Phyllis Mahoney, 2000 IEDM, Suite 400B, 101 Lakeforest Blvd., Gaithersburg, MD 20877; (301) 527-0900 ext. 103, FAX: (301) 527-0994, e-mail: phyllism @widerkehr.com. ••

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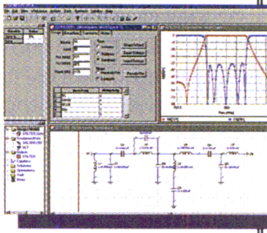
Digital upconverter reaches 200 MSamples/s

The model AD9857 14-b, quadrature digital-upconverter (QDUC) integrated circuit (IC) can sample at rates up to 200 MSamples/s and can serve as a universal in-phase/quadrature (I/Q) modulator and upconverter. The chip accepts complex I/Q input data, samples the data, quadrature-modulates the data onto a carrier, and puts out a frequency-agile modulated carrier in the analog domain. The device moves much of the transmit path from the analog to the digital domain, which supports direct process control, on-the-fly progammability of data rates, precision frequency hopping, and reduced system complexity. **Analog Devices, Inc., Ray Stata Technology Center, 804 Woburn St., Wilmington, MA 01887; (800) 262-5643, FAX: (781) 937-1428, Internet: http://www.analog.com.**

CIRCLE NO. 58 or visit www.mwrf.com.

Software simplifies filter design

The S/FILTER software package allows users to design filters with direct synthesis. In this process, the user inputs the criteria for his or her filter, the program forms a transfer function from the criteria, and it extracts a schematic from the transfer function. The direct-synthesis approach to designing filters differs from the more common look-up approach, which restricts engineers to "canned" filter designs. Direct synthesis allows the user to create customized filters with the exact response desired. The S/FILTER program can extract a schematic in any order from the transfer function, and generate an array of filters for each set of design criteria. Other capabilities include the ability to tune zero locations to optimize a filter design; design classic transfer functions, such as Butterworth, Chebyshev, and Elliptic; asymmetrically place DC and infinite zeros to adjust overall response symmetry; design exact solutions for narrow and wideband filters; create arbitrary stopbands; etc. **Eagleware Corp., 635 Pinnacle Ct., Norcross, GA (678) 291-0995, FAX: (678) 291-0971, Internet: http://www.eagleware.com.**



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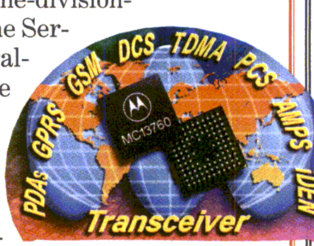
Transceiver IC supports multiple bands, protocols

Model MC13760 software-programmable transceiver integrated circuit (IC) functions over a range of input frequencies in multiband, multimode Global System for Mobile Communications (GSM)/digital communication services (DCS)/time-division-multiple-access (TDMA)/Advanced Mobile Phone Service (AMPS) cellular phones, including general-packet radio service (GPRS). The chip can also be applied to personal digital assistants (PDAs), satellite phones, integrated dispatch enhanced networks (iDENs), and trans-European trunked-radio-system (TETRA) radios. The IC features fractional-N synthesizers, a reconfigurable zero-intermediate-frequency (IF) receiver with programmable bandwidth, receive analog-to-digital conversion (ADC), multirate interface to the baseband digital signal processor (DSP), and full transmit-support circuits. **Motorola Semiconductor Products Sector, P.O. Box 52073, Phoenix, AZ 85072-2073; (480) 413-4991, FAX: (480) 413-6220, Internet: http://www.motorola.com/wireless-semi/.**



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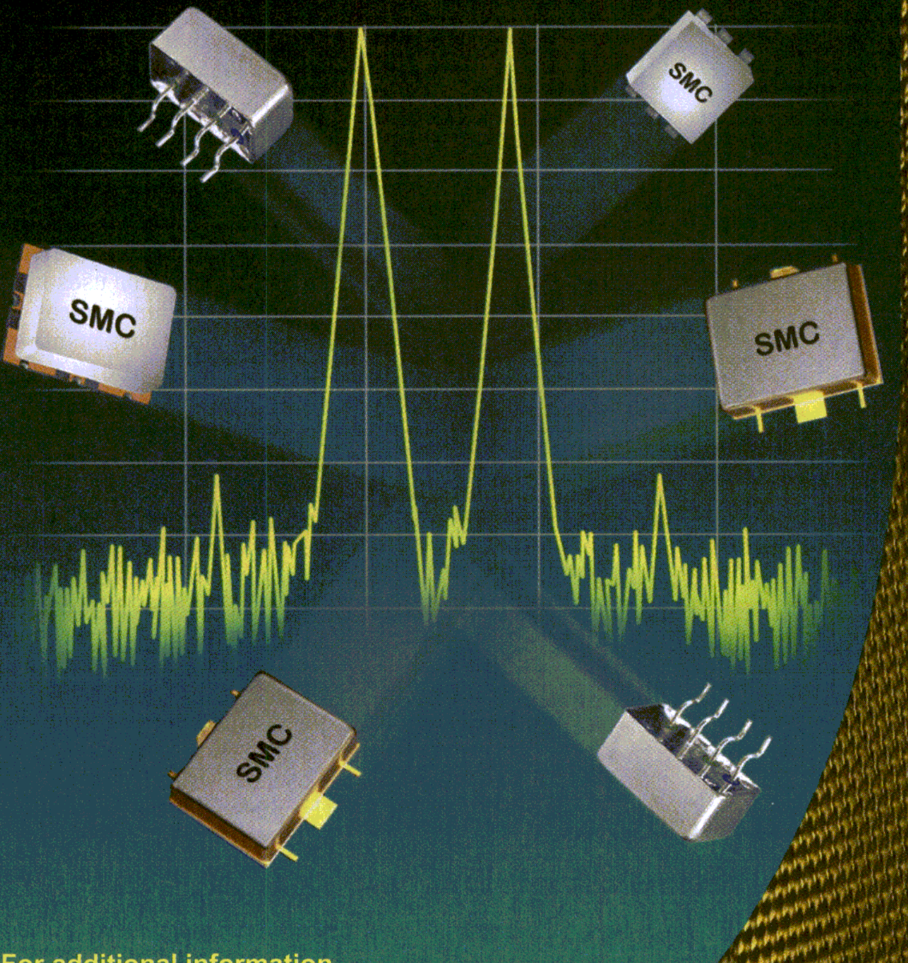

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CIRCLE NO. 305

Cree, Xemod To Jointly Develop SiC

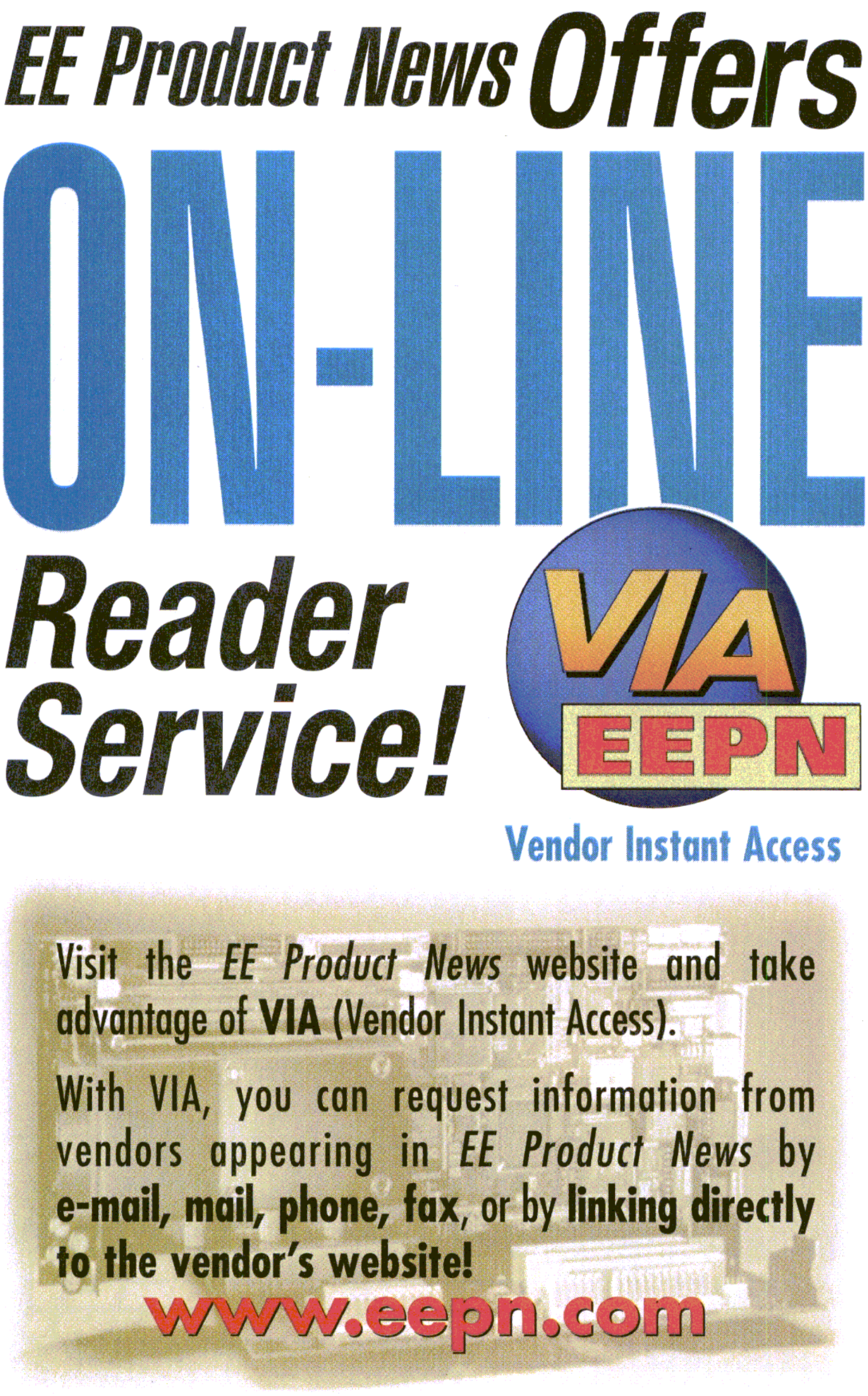
Silicon-carbide (SiC) semiconductor technology seemed to be in limbo over the past year, but its primary developer, Cree, Inc. (Durham, NC), recently teamed up with RF power-module maker Xemod, Inc. (Santa Clara, CA) to jump start the technology and bring it into mainstream wireless communications applications. The companies signed a two-year agreement with Cree, making an $11.3 million equity investment in Xemod as a general investor and a co-developer of SiC-based power amplifiers (PAs).

The companies will split the development task with Cree providing the SiC raw material and contributing its knowledge of the process while Xemod provides its RF and microwave expertise in linear PAs to do the design, layout, and metallization of metal-semiconductor-field-effect-transistor (MESFET) SiC devices. Eventually, Xemod hopes to launch a new generation of products based on the technology.

While SiC presents impressive technical credentials as a PA technology, it is not yet ready for actual hardware deployment in the high-frequency power market. But Richard Clark, vice president of marketing at Xemod, believes that it could be used in the same applications as the current leading wireless-infrastructure PA semiconductor technology, laterally diffused metal-oxide semiconductor (LDMOS). In fact, Clark points out that SiC offers a number of advantages over LDMOS, and even gallium arsenide (GaAs), due its intrinsic physical characteristics.

With respect to LDMOS, SiC is a higher voltage process, being in the +50-VDC range compared to approximately +30 VDC for LDMOS. A benefit of operating at a higher voltage is that a SiC transistor uses higher input and output impedances than an LDMOS device. Clark points out that at power levels above 60 W, the matching required for LDMOS is more difficult than for the higher impedance SiC. Another advantage of SiC is its higher thermal conductivity, which means that SiC transistors can handle greater power than other semiconductor technologies—LDMOS and GaAs—a key criterion for devices to be used in PA modules. These characteristics could result in SiC devices that are smaller and less expensive to manufacture than their counterparts. Finally, SiC can operate at microwave frequencies up to 5 GHz, whereas the present generation of LDMOS goes up only to 2.5 GHz. This could lead to new applications for SiC-based PAs where competing technologies are limited.

Cree presently uses SiC technology in the manufacture of blue and green light-emitting diodes (LEDs) which are found in the instrument panels of automobiles and cellular phones. ●●



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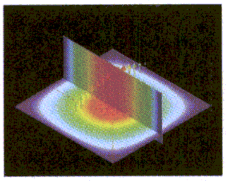

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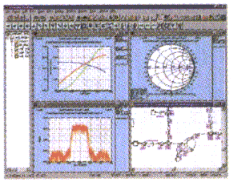

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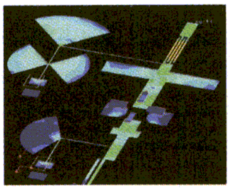

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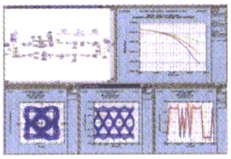

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Application Profile

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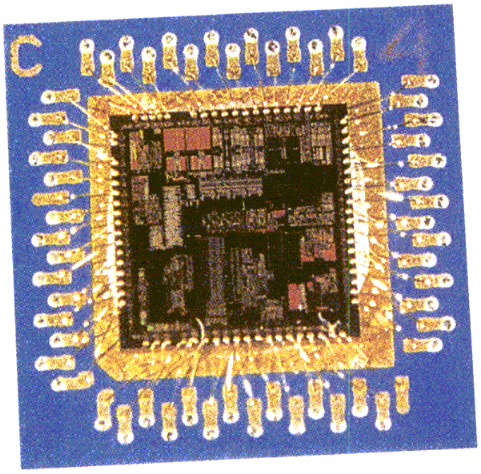

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CIRCLE NO. 223

Contracts

ADC Telecommunications, Inc. and Image Wireless Communications—Signed a $10 million contract extension for ADC's Axity™ Broadband Wireless Access System.

Sanders—Has received a contract from the US Navy for full-rate production of Cooperative OUTBOARD Logistics Update (COBLU) systems. Under the contract, Sanders will provide up to 35 COBLU systems as well as spares, installation, and training support. Total value of the contract, including options, could exceed $250 million. Deliveries for the base-year award of six systems will begin in late 2001.

Cree, Inc.—Has received contracts from Asian customers totaling $21 million for shipments during fiscal year 2001. Included in these contracts are purchase agreements for standard brightness, high brightness, and small-chip silicon-carbide (SiC) light-emitting diodes (LEDs).

CTS Corp.—Announced that it has been awarded a multimillion-dollar contract by one of the world's largest base-station infrastructure equipment manufacturers. The contract is expected to exceed $20 million over the next two years.

Andrew Corp.—Has received an initial $3 million order as part of a $23 million letter of agreement for the development and supply of single-channel personal-communications-services (PCS)/time-division-multiple-access (TDMA) linear power amplifiers (PAs) to Lucent Technologies, Inc.

Motorola—Has been awarded a $28 million Global System for Mobile Communications (GSM) network-expansion contract from Kuwaiti GSM operator Mobile Telecommunications Co. (MTC). Motorola will implement network-expansion and enhancement solutions for the dual-band MTCNet GSM-900/1800 network, increasing capacity by 30 percent.

Fresh Starts

Discovery Semiconductors, Inc.—Has been named to Deloitte & Touche's "Fast 50" Program for New Jersey, a ranking of the 50 fastest-growing technology companies in the area. Rankings are based on the percentage of growth in revenues from 1995 to 1999.

Analog Devices, Inc.—Has entered into an agreement with Caveo Technology, LLC, in which ADI's ADXL202E will be used in conjunction with Caveo's proprietary software in the development of advanced security problems for laptops and other applications. The resulting product, Caveo Anti-Theft™, features the ADXL202E, a micromachined tilt-motion sensor which enables location awareness in hardware. Caveo Anti-Theft is designed for laptop computers and is expected to be available in early 2001.

Vitesse Semiconductor Corp.—Announced that it has completed the acquisition of certain assets of the wide-area-network (WAN) product line of Philips Semiconductors for approximately $30 million in cash. The Philips Semiconductors WAN product line designs, develops, and markets products for WAN access and aggregation. This group brings with it a high-level data-link control (HDLC) controller product, a T1/E1/J1 aggregation product, a DS-3 termination product, as well as several related products that are in advanced stages of development.

Agilent Technologies, Inc.—Announced that it is building a dedicated 6-in. (15.24-cm) wafer-fabrication line in its Newark, CA facility for its innovative film-bulk-acoustic-resonator (FBAR) filter products. FBAR technology enables miniature high-performance filtering products, such as duplexers for code-division-multiple-access (CDMA) mobile phones, operating in the 400-MHz and 10-GHz frequency range.

GHz Technology and Lucent Technologies' Microelectronics Group—Arrived at an agreement to develop and manufacture custom RF power transistors for wireless base stations.

VoiceStream Wireless Corp. and Powertel, Inc.—Have entered into a definitive agreement for VoiceStream Wireless to acquire Powertel. The proposed transaction will substantially expand VoiceStream's all-digital Global System for Mobile Communications (GSM) wireless coverage in the Southeastern US. Powertel owns wireless licenses covering 25 million people and operates a GSM network spanning 12 states in the Southeastern US, in areas where VoiceStream currently does not market its wireless services.

Sanders—Issued the final acceptance of the SAL XRS2000 NanoPulsar System. The system, installed at Sanders' Microwave Electronics Center in Nashua, NH, is the first integrated point-source X-ray lithography system. Sanders has begun production of high-frequency gallium-arsenide (GaAs) integrated circuits (ICs) with the SAL NanoPulsar System.

Advanced Hardware Architectures (AHA)—Opened the company's first United Kingdom design center, located in Southampton, England. Demand for AHA's forward-error-correction (FEC) products based on their patented Turbo Product Code (TPC) technology, spurred the expansion decision.

Texas Instruments, Inc.—Completed its previously announced acquisition of Dot Wireless, Inc. in a stock-for-stock transaction that will expand TI's expertise in digital-signal-processor (DSP)-based solutions for third-generation (3G) wireless handsets and mobile computing devices. Dot Wireless has an advanced 3G system architecture and protocol software for voice and high-speed data applications. Dot Wireless is involved in 3G code-division-multiple-access (CDMA) technology and will support TI's development of technologies for IS-2000 and wideband CDMA (WCDMA).

Enea OSE Systems—Launched the OSE Design Center, which is located in Phoenix, AZ. The OSE Design Center is designed to offer developers working with the OSE™ real-time operating system (RTOS) access to genuine OSE competance and additional resources to get projects started, completed, or developed from start to finish.

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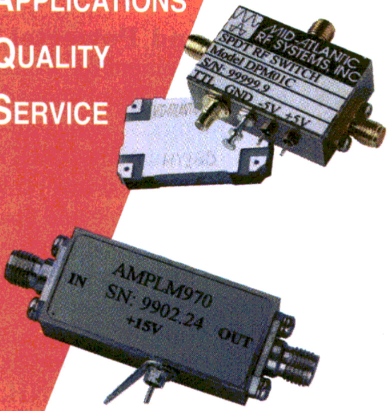

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Flarion Technologies—Matthew (Matt) J. Desch to the board of directors; remains as chairman of Airspan Networks, Inc.

AIR2LAN—James V. Smith to chief operating officer (COO); formerly president of Big Net, Inc.

Wood & Douglas Ltd.—Martin Farrow to business development manager; formerly marketing manager at Yokogawa Marex Technology. Also, Rachael Penfold to broadcast product manager; formerly strategic development manager for Radio Frequency Investigation.

FARROW

PENFOLD

Sanders—Aaron Penkacik to chief technical officer (CTO); remains as vice president for advanced systems and technology.

Microwave Power Devices, Inc.—Thomas Gilboy to vice president and chief financial officer (CFO); formerly CFO of PureTec Corp.

GOAL Semiconductor, Inc.—Nick Tasker to president; formerly executive vice president of marketing and sales. Also, Irving Lustigman to vice president of engineering; formerly vice president of broadband wireless access products.

American Microsystems, Inc. (AMI)—John Murray to regional sales manager for the Southwest territory, including Southern California, Arizona, and New Mexico; formerly test product engineer.

Robinson Nugent, Inc.—James "Chris" Smelcer to director of manufacturing for North America; formerly plant manager at the Dallas, TX manufacturing plant.

IPC—Fern Abrams to director of environmental policy; formerly manager of environmental affairs at the American Trucking Association.

Aegis Broadband, Inc.—D. Colin Boyd to vice president of sales and marketing; formerly vice president of North American sales and worldwide marketing at Harmonic, Inc.

Taconic—Jim O'Keefe to president; formerly vice president and general manager of the Industrial Products Business.

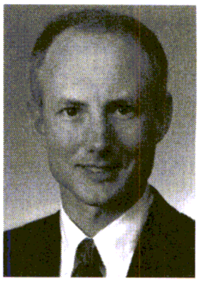
Altera Corp.—Farzad Zarrinfar to vice president of product marketing; formerly director of application-specific-integrated-circuit (ASIC) sales and marketing at Samsung Semiconductor.

SignalSoft Corp.—Brian J. McNiff to vice president of Internet services; formerly senior integration consultant at Little Harbor Capital.

Micro Networks Corp.—Michael J. Ferrantino to chairman; remains as president and chief executive officer (CEO).

EPCOS, Inc.—James Riff to vice president of sales at EPCOS, Inc.; formerly executive vice president responsible for North American operations of the Lapp Group.

RIFF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Ixthos, Inc.—Ed Fenley to director of operations; formerly involved in electronics manufacturing management at AlliedSignal and Raytheon.

Leitch Technology Corp.—Todd Riggs to regional sales manager for the state of Florida; formerly branch manager of the Orlando, FL office for Professional Communication Systems. Also, Neil Karsh to technology deployment manager for audio and DTV at the Northeast sales office; formerly vice president of audio engineering for New York Media Group, Inc. In addition, Jim Lien to regional sales manager for the North Mid-Atlantic region; formerly district sales manager for Snell & Wilcox.

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Wireless Digital Communications
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Center for Professional Development
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(480) 965-1740, FAX: (480) 965-8653
e-mail: asu.cpd@asu.edu
Internet: http://www.eas.asu.edu/cpd

Wireless Communications Standards: Understanding TIA-41
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Fiber Optic Training
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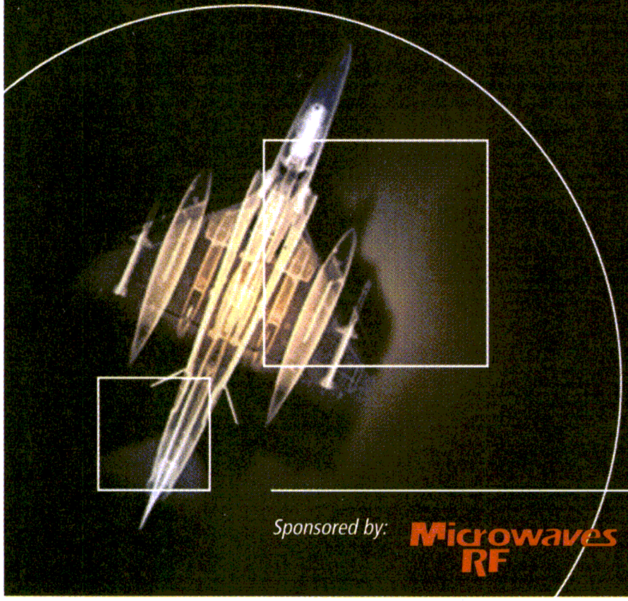

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CIRCLE NO. 258

Bluetooth gains another application—sensors

Already growing in potential by leaps and bounds for computer and consumer applications, the Bluetooth wireless protocol can also serve in the industrial instrumentation world, according to Michael Horton of Crossbow Technology (San Jose, CA). A Bluetooth network can be used to enable communications between remote sensors and data-acquisition (DAQ) electronics, eliminating the expensive, time-consuming, and failure-prone hardwiring of sensors. The author claims that Bluetooth can replace RS-232, parallel, the universal serial bus (USB), and other standard cables with a single standard wireless link. Bluetooth, combined with an established sensor ID standard such as IEEE 1451, permits the creation of a reliable, low-cost architecture with a broad range of applications. These include building monitoring and maintenance, medical and consumer product monitoring, DAQ and control, as well as many others. The architecture supports a wide range of sensor types through a smart interface. See "Standard Wireless Protocols Create New Sensor Networks," *NASA Tech Briefs*, September 2000, Vol. 24, No. 9, p. 30.

Efficient RF antenna system based on traps

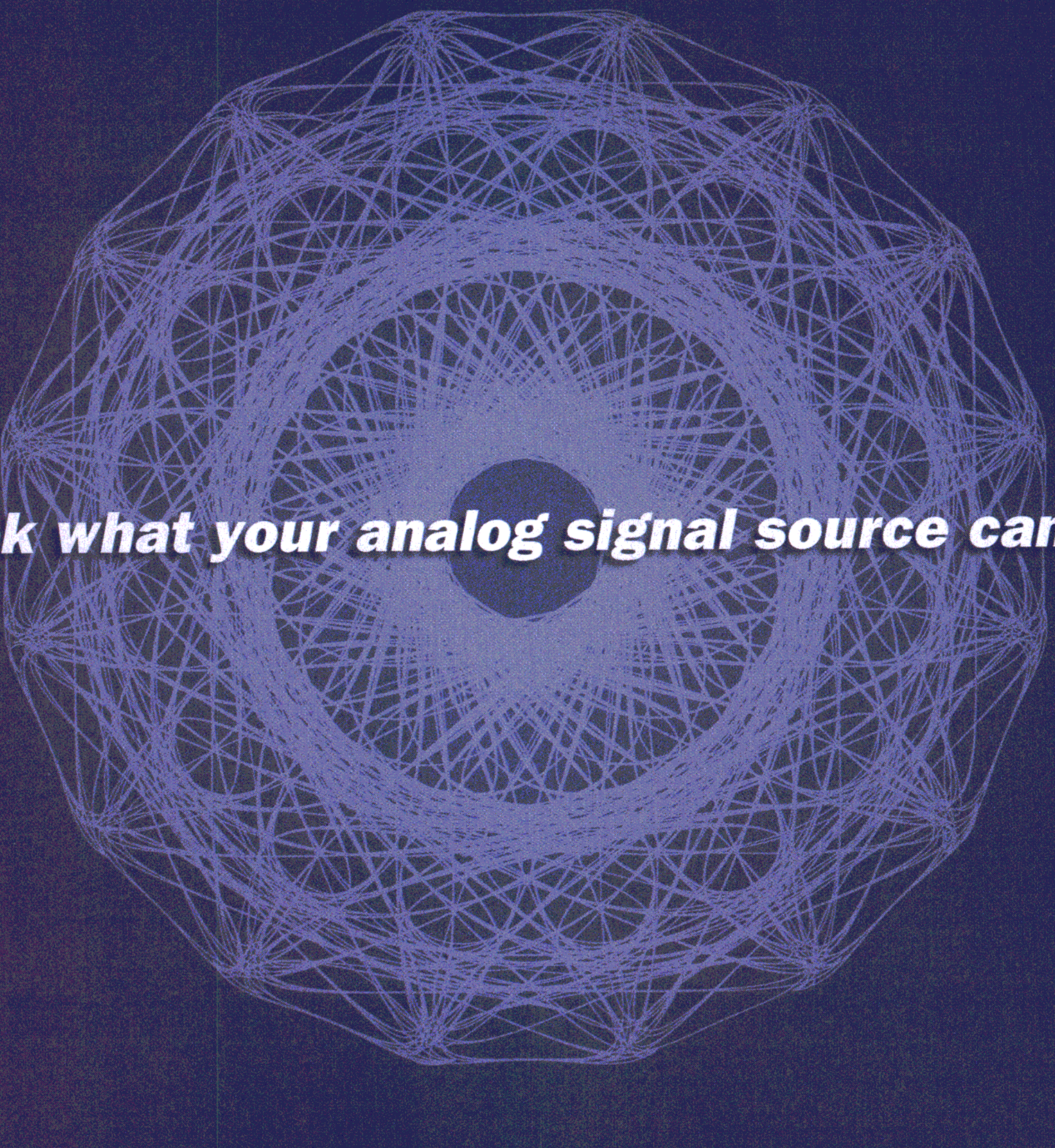
Antenna systems using traps for operating at different frequencies are well-known in amateur radio, but cannot meet the bandwidth, efficiency, directivity, response time, and other requirements demanded by modern high-frequency communications. Traditional systems are single-wire types with traps (inductor-capacitor circuits), but Daniel D. Reuster and Kevin J. Cybert of ARINC, Inc. (Annapolis, MD) propose the use of traps in a multiple wire-element arrangement using different physical antennas for adjacent subbands. The traps divide the physical antenna into varying radiating lengths, which provides continuous coverage over the entire frequency band of interest. The use of subbands reduces impedance-matching network requirements and helps to more efficiently radiate energy over the entire frequency band of interest. The difference between the author's design and the traditional design is that the former is a true broadband antenna while the latter is simply a multiband antenna. See "A High-Efficiency Broadband HF Wire-Antenna System," *IEEE Antennas and Propagation Magazine*, August 2000, Vol. 42, No. 4, p. 53.

Special oven design enables low-frequency measurements

Low-frequency-noise measurements are important in characterizing the behavior of materials and devices used in integrated circuits (ICs), but the measurements must be made at a tightly controlled temperature. A special high-stability, temperature-controlled oven designed especially for characterizing electromigration in the metal interconnects of ICs has been built by Carmine Ciofi, Ivan Ciofi, Stefano Di Pascoli, and Bruno Neri of the Dipartmento di Ingegneria dell' Informazione, University of Pisa (Pisa, Italy). Most low-frequency-noise measurements are in the extremely low frequency range of 10 MHz to 1 Hz and the background noise must be also extremely low; for example, 3 $\text{nV}/\sqrt{\text{Hz}}$ at 1 Hz. The concept of the oven design is to calculate the maximum value of the power spectral density of the temperature fluctuations that do not cause a significant increase in the background noise. The authors have developed a unique two-stage oven together with a microcontroller board that implements all operations for warm-up and closed-loop control under steady-state conditions. See "Temperature Controlled Oven for Low Noise Measurement Systems," *IEEE Transactions on Instrumentation and Measurement*, June 2000, Vol. 49, No. 3, p. 546.

Neural networks reduce complex-design time

Neural network modeling is becoming an important design-automation technique for reducing the central-processing-unit (CPU) time necessary to model intensive electromagnetic (EM) simulations in a classical optimization procedure. One type of neural network, the multilayer perceptron, was investigated by A. Gati, G. Alquie, and V. Fouad Hanna of the Laboratoire Instruments et Systèmes, Université Pierre et Marie Curie (Paris, France), and M.F. Wong of CNET, France Telécom (Issy-Moulineaux, France) and applied to a coplanar-waveguide (CPW) T-junction. CPW technology has proven its superiority in microwave hybrid and monolithic integrated circuits (ICs) and, while a full-wave three-dimensional (3D) EM simulation is very accurate, it is time-consuming (CPU time). The validity of neural networks is demonstrated by the geometrical and frequency parameterization of a complex CPW component such as a lowpass filter and a uniplanar antenna using a CPW feeding line. See "Neural Network Modeling and Parameterization Applied to Coplanar Waveguide Components," *International Journal of RF and Microwave and Computer-Aided Engineering*, September 2000, Vol. 10, No. 5, p. 296.



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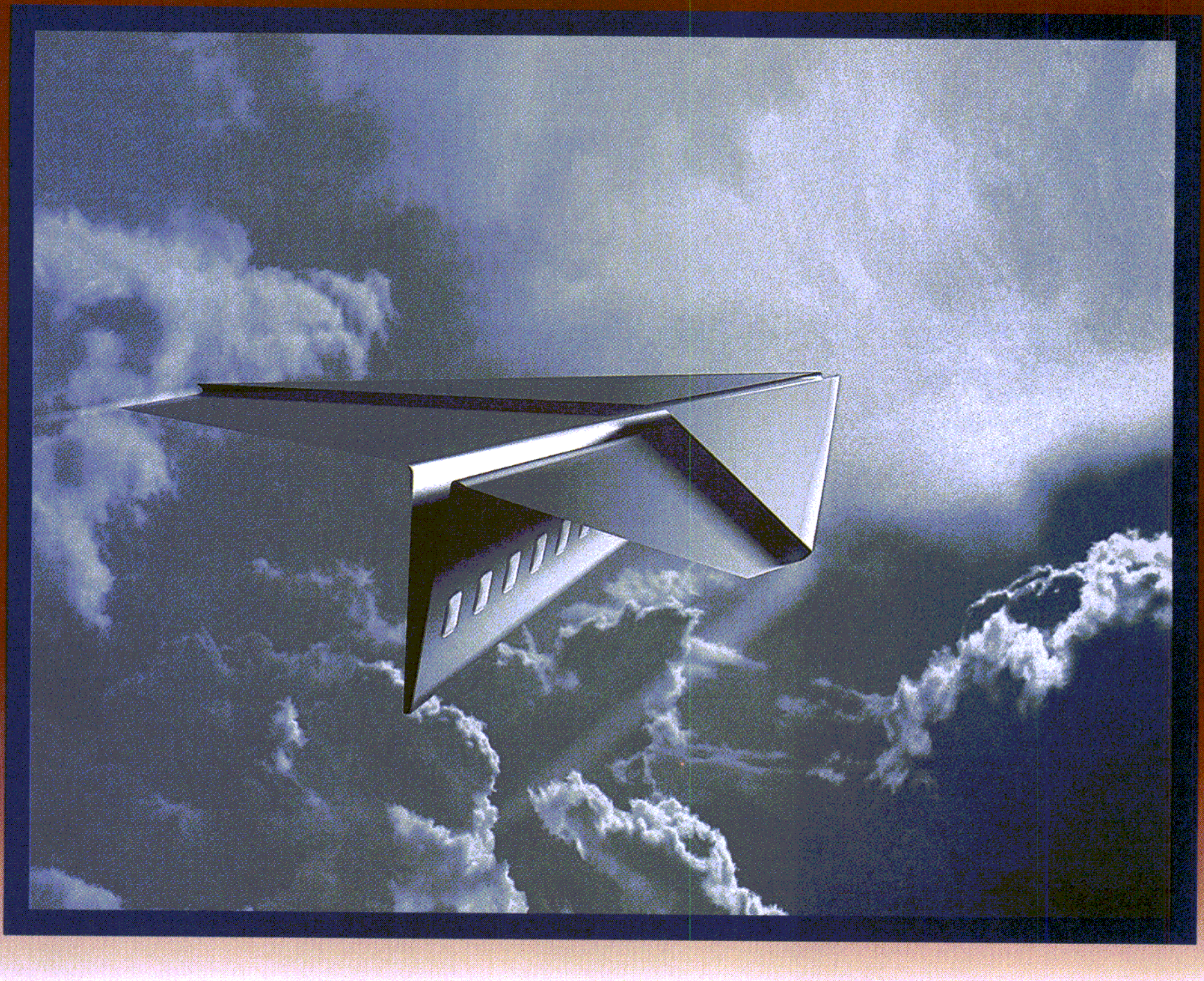

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NGA-286	0.1-6.0	4.0	50.0	15.5	15.2	32.0	120
NGA-386	0.1-5.0	4.0	35.0	20.8	14.5	25.8	144
NGA-486	0.1-6.0	5.0	80.0	14.8	18.3	39.5	118
NGA-586	0.1-6.0	5.0	80.0	19.9	18.9	39.6	121
NGA-686	0.1-6.0	5.9	80.0	11.8	19.5	37.5	121

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CIRCLE NO. 250

S-Band Amplifier Modeled For Wireless Data

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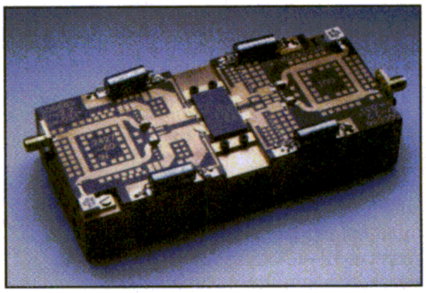
Shansong Song and Raymond Basset
California Eastern Laboratories, Inc.,
4590 Patrick Henry Dr., Santa Clara,
CA 95054-1817.

MULTICHANNEL-MULTIPOINT-DISTRIBUTION-SYSTEM (MMDS) and wireless-data/Internet technologies are the latest transport media for television broadcasting and the Internet. Since these technologies require the use of reliable linear power amplifiers (PAs), a compact amplifier was designed based on device modeling and characterization with a two-tone signal for optimum third-order intermodulation distortion (IMD3) at a defined output power. The amplifier configuration was also an important factor in obtaining good linearity, stability, and external matching, along with low circuit loss.

The development of this amplifier includes the modeling and the characterization of the "twin" gallium-arsenide metal-semiconductor-field-effect-transistor (GaAs MESFET) device, NES2427P-60 used in this design.[1] This article also presents the design methodology of the external input- and output-matching circuits of the GaAs device and a two-arm branch 90-deg. coupler. The performance of this compact amplifier is presented and discussed.



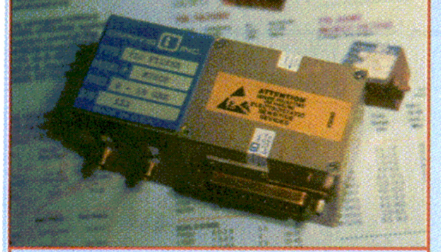
1. A complete balanced, linear-amplifier circuit based around the NES2427P-60 "twin" GaAs MESFET, the device incorporates a splitter and combiner in the package shown.

The goal was to design a compact linear PA using the 60-W Class A-B NES2427P-60 from NEC/CEL that delivers a power of +40 dBm for each tone with an IMD3 of less than −40 dBc over the 2.5-to-2.7- GHz, S-band instantaneous bandwidth. The requirement for the typical gain at 1-dB compression (P1dB) was 11 dB with a gain flatness of better than ±0.5 dB. The target for the typical output power at P1dB was +48 dBm.

The amplifier-circuit optimization for linearity and the accurate prediction of the amplifier's performance require an available device model and characterization for a two-tone signal. The device was modeled and characterized with a two-tone signal at an output-power level of +40 dBm

Table 1: Half-device impedances versus frequency

Frequency GHz	R_{IN} (ohms)	X_{IN} (ohms)	R_{OUT} (ohms)	X_{OUT} (ohms)
2.4	4.9	−9.1	17.2	16.4
2.5	4.0	−4.2	21.9	12.3
2.6	8.0	1.7	18.4	10.1
2.7	6.7	−2.4	16.8	11.1



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6- , 8- and 12-STAGES			
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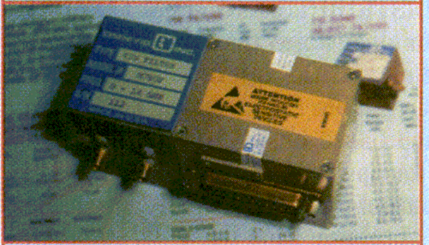
L106RX	1.0 – 2.0	1.5	10
C105RX	2.0 – 8.0	1.5	10
X106RX	8.0 – 12.4	1.5	20
Ku106RX	12.0 – 18.0	1.8	20
M102RX	4.0 – 12.4	1.5	8
M103RX	4.0 – 12.4	1.5	10
M104RX	4.0 – 18.0	1.5	10–60



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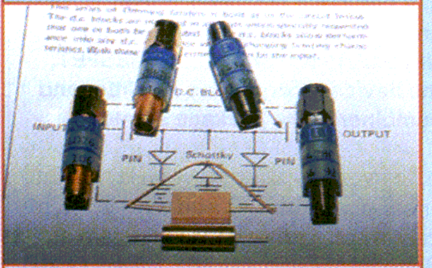

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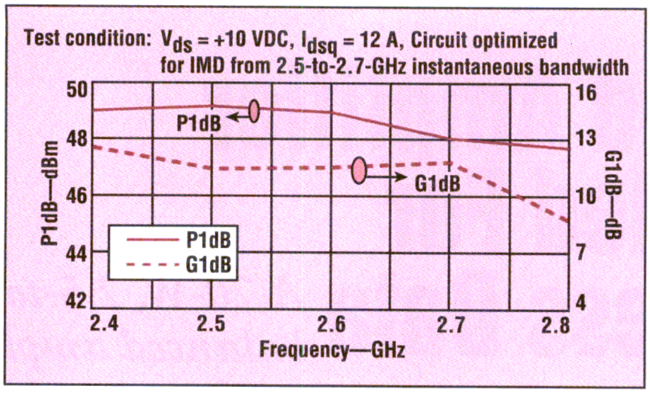


2. These curves show the P1dB and G1dB performance over the 2.4-to-2.8-GHz range of the NES2427P-60 amplifier at a V_{ds} of +10 VDC and an I_{dsq} of 12 A.

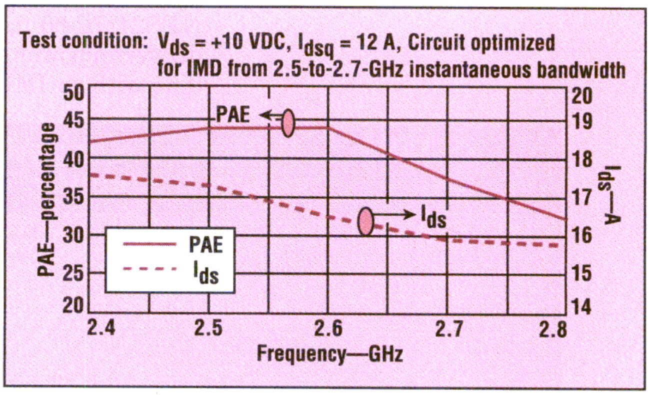
each tone in the 2.4-to-2.7-GHz band.

The design challenge was to match the device's optimum output impedance for IMD3 to a 50-Ω load over the 2.5-to-2.7-GHz bandwidth with an excellent return loss (better than 18 dB) and to minimize the loss of the output matching and combining circuit. The input circuit is not as critical as the output since it does not significantly affect the output power of the amplifier. Its matching and loss must be excellent only at the maximum frequency for gain consideration. Stability is always an issue with high-power amplifiers. The splitter and combiner have to be selected for high isolation (20 dB), low loss (0.25 dB), good balance, and ease of integration to the amplifier layout. The choice of the amplifier configuration is important if a good external match is needed to isolate the driver from the power stage.

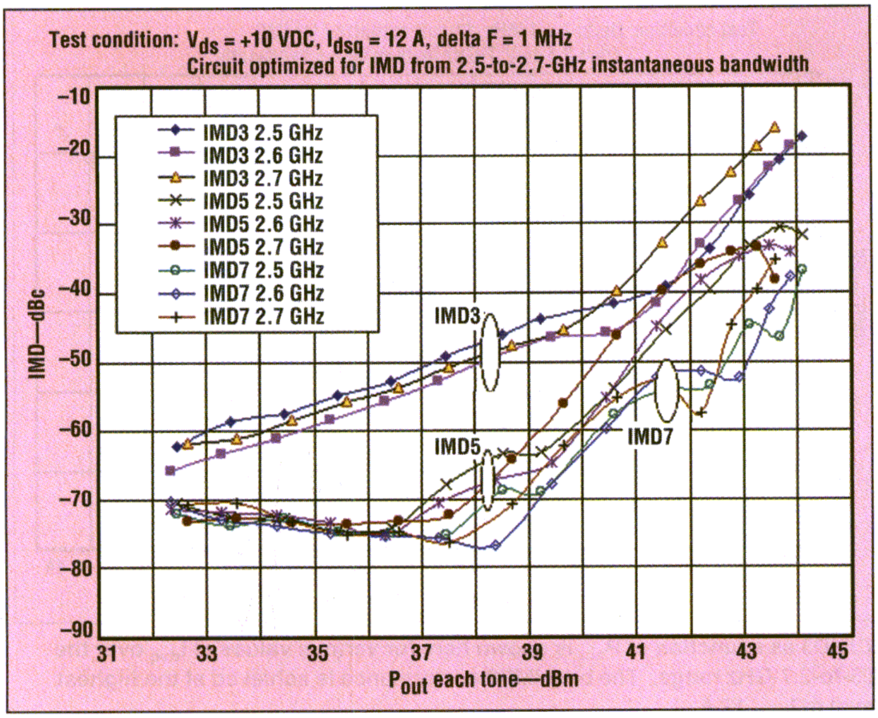
A MESFET MODEL

The device used in this amplifier (a Class A-B MESFET "twin" device) consists of two separated pairs of chips with their internal matching circuits mounted in the same package. The package has two gate connections and two drain connections without any internal connection between the two sides of the device. These devices are sometimes improperly called "push-pull" devices. This concept comes from lower-frequency applications—high frequency (HF), very-high frequency (VHF), and ultra-high frequency (UHF)—using silicon (Si) devices with internal connections between the two sides of the device to take advantage of the virtual ground created by the push-pull configuration.[2,3] The high-power GaAs devices at L-band and higher frequencies do not have these internal connections because their transverse dimensions are too large.[4] Therefore, these devices are not "push-pull." They are devices with two identical sides called (by CEL) "twin" devices. The configuration of the amplifier defines only how the devices operate. They can be combined the same way as single-ended parts in balanced configuration (90-deg. hybrid), in push-pull configuration (baluns[4-6]), and in phase combining (Wilkinson), etc.

To model these devices, only one side is considered. Simple narrowband external input- and output-matching circuits with biasing circuits were designed. At each frequency, the input circuit was tuned for maximum return loss and



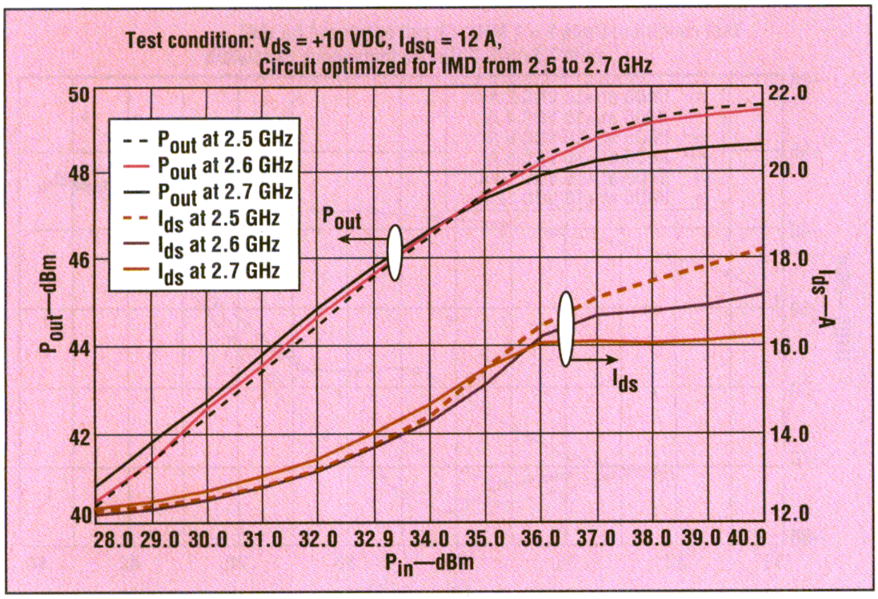
3. Another performance curve of the NES2427P-60, this one illustrates the amplifier's PAE and drain-to-source current (I_{ds}) at 1-dB gain compression.



4. The overall IMD performance of the NES2427P-60 amplifier is provided in these sets of curves for IMD3, IMD5, and IMD7 over the 2.5-to-2.7-GHz band. IMD3 remains below -40 dBc for output-power tones up to +41 dBm.

the output circuit was optimized for IMD3. The performance, the source, and the load impedances were measured and recorded. These impedances are relatively high and do not present any difficulty in measuring to good accuracy because the device is internally prematched.

Table 1 shows the half-device impedances versus frequency. Under these conditions and for half the device, the following performance was obtained in the 2.5-to-2.7-GHz bandwidth—P1dB = +45 dBm, IMD3 = −41 dBc to −43 dBc at +37 dBm each tone, and G1dB = 13.3 dB.



5. This set of curves shows the relationship between P_{out}, P_{in}, I_{ds}, and P_{in} for the amplifier. The measurements were made at three different frequencies for each set (2.5, 2.6, and 2.7 GHz).

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YOM1317	2.0 – 8.0	20	12
YOM818	8.0 – 18.0	20	12
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YM 1027	100 MHz	1 – 18	-40
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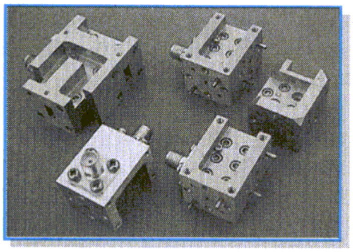

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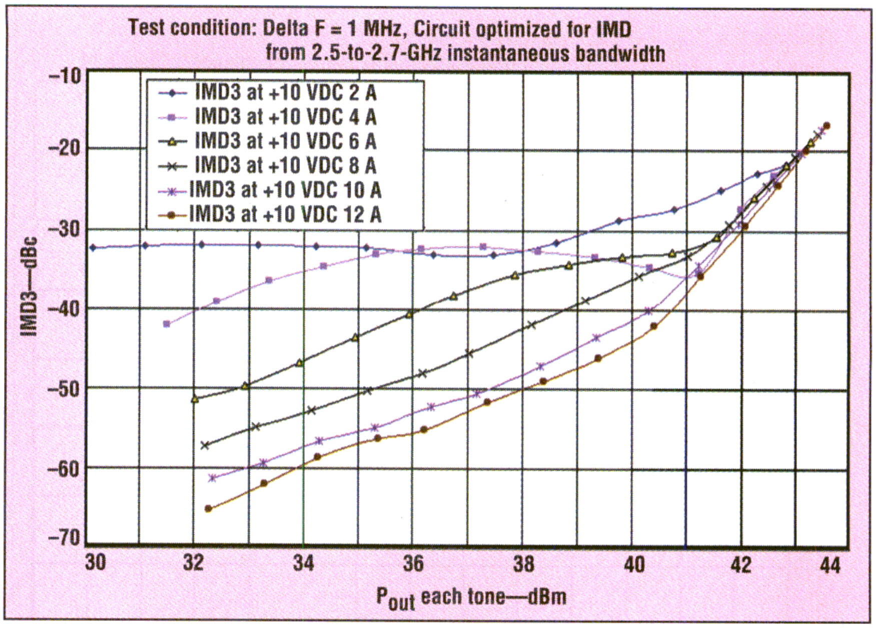


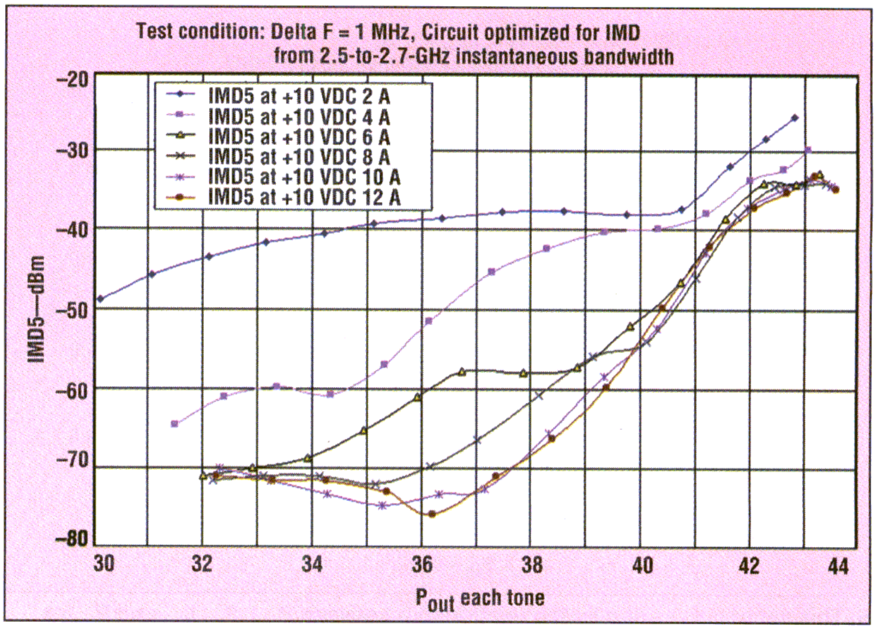
6. IMD3 as a function of P_{out} is shown here for various values of I_{dsq} over the 2.5-to-2.7-GHz range. The best IMD3 performance is achieved at the highest level of I_{dsq}, 12 A.

In addition to the device characterization with a two-tone signal, the device was characterized with one-tone signal at P1dB and its S-parameters were measured under small-signal conditions in the 1.5-to-3.5-GHz band. The S-parameters were used for gain and stability analysis under small-signal conditions.

Contrary to popular belief, for a bandwidth of less than one octave there are no advantages in using a push-pull amplifier in terms of bandwidth and linearity. The disadvantages are:

1. Low isolation between the two sides of the device if a classical balun is used (only 6 dB).
2. Poor external input and output match.

On the other hand, the balanced configuration has the same perfor-



7. These curves are similar to those of Fig. 6, except that IMD5 is plotted against power. And like Fig. 6, IMD performance improves for higher values of I_{dsq}.

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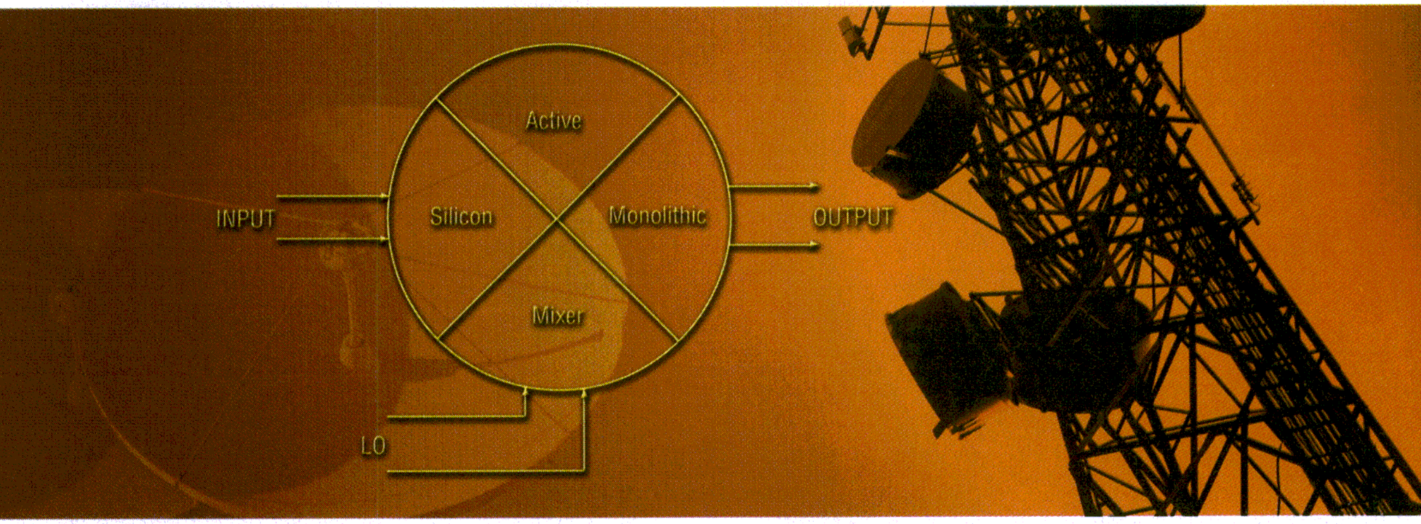

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Table 2: Comparison of simulated and measured impedances

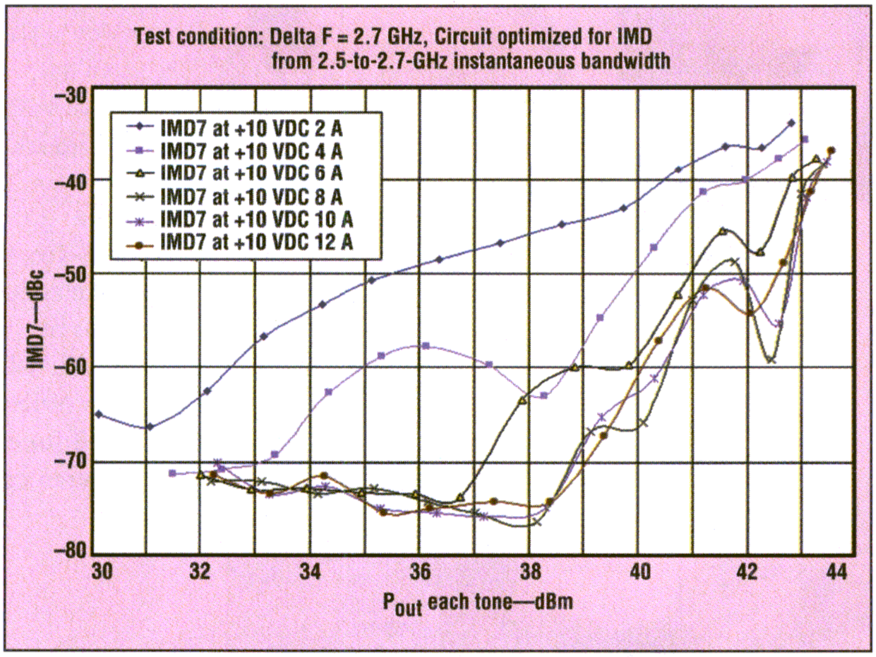
Frequency GHz	Source impedance of half NES2427P-60		Load impedance of half NES2427P-60	
	Simulated	Measured	Simulated	Measured
2.4	6.2 + j4.8	5.0 + j7.4	20.2 – j12.3	16.4 – j8.7
2.5	7.1 + j4.9	7.5 + j7.1	19.5 – j11.9	18.4 – j8.9
2.6	7.4 + j3.6	8.0 + j4.0	18.8 – j11.5	19.0 – j10.6
2.7	4.9 + j2.3	4.6 + j2.8	18.2 – j11.2	17.9 – j12.4

mance as the push-pull concerning bandwidth and linearity (bandwidth less than one octave) and has the following advantages:

- High isolation between the two sides of the device.
- Good external match.
- Easy-to-design printed 90-deg. splitters/combiners for narrowband applications that can be easily integrated with the layout of the amplifier.
- Good reliability—the failure of one device side does not result in total failure. Only the output power drops by 6 dB.

For balanced amplifiers, many kinds of couplers that can be used—Lange couplers, Wilkenson with 90-deg. phase jog on one port, two-arm branch 90-deg. hybrid, etc. In this project, a two-arm branch 90-deg. hybrid was selected due to its simple layout and ease in integrating with the amplifier layout. Since the input- and output-matching circuits realize the matching of the device to 50-Ω impedance, the hybrid does not perform any impedance transformation and all its port impedances are 50 Ω.

The design of the hybrid was performed with Hewlett-Packard (now Agilent) HP multipoint-distribution software (MDS). Also, HP MOMENTUM software was used to perform the electromagnetic (EM) simulation. The comparison of the simulated results between MDS and MOMENTUM software showed a slight difference in the coupler layout. The final EM simulation indicates that this 90-



8. The curves of IMD7 with respect to P_{out} follow the same pattern as those of Figs. 6 and 7, showing how IMD performance improves for higher values of drain current.

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deg. hybrid has more than 20 dB of isolation between its coupled ports and 20 dB of return loss for each port. Its balance is better than 0.1 dB in magnitude and 90 ±1 deg. in phase.

The input-matching circuit was optimized for flat gain over the 2.4-to-2.7-GHz bandwidth and to provide the best match to a 50-Ω impedance at the highest frequency. It consists of two sections using only transmission lines with all stubs being open. In order to cover the 2.4-to-2.7-GHz bandwidth, sections with low quality factor (Q) were selected for the design, especially the first one. A gain slope of −6.0 dB per octave for the device was assumed and the input-matching circuit was designed to compensate for this slope and to obtain an excellent match at 2.7 GHz. The result is a flat gain and a maximum gain over the desired bandwidth. From the device modeling and characterization data the predicted associated gain was G1dB = 13.0 dB and the gain flatness was 0.5 dB. The simulation also showed an input-return loss of more than 12 dB across the bandwidth for half the device. However, the full device return loss will be higher since a balanced configuration was selected and the two device sides are symmetric.

The design goal for the output matching network was to present the optimal load impedance for a two-tone signal at a defined output power with minimum loss. Since the NES2427P-60's device-output optimal impedance for a two-tone signal is not far from 50-Ω impedance, two sections of one-sixteenth wavelength Chebyshev impedance transformer circuit was selected. This circuit minimizes the dimensions and loss of the matching network. The simulation showed that the loss of this output circuit was less than 0.2 dB and the return loss was better than 19 dB across the bandwidth.

Figure 1 shows the complete balanced-amplifier layout. The two-arm branch 90-deg. hybrid, which uses the same substrate as the matching circuit, is integrated with the amplifier layout and does not require any additional connection. All the circuits are directly printed on the same $E_r = 2.2$,

(continued on p. 160)

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Design A Frequency Synthesizer For Mobile Communication Systems

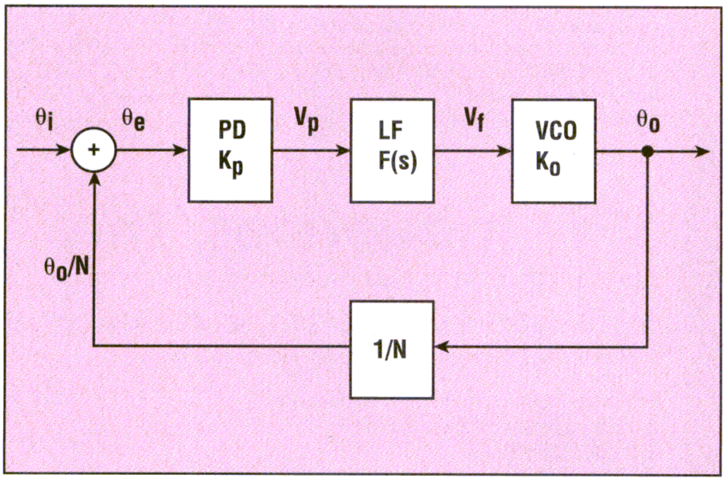
This approach uses a programmable synthesizer to drive a VCO, achieving high stability and low phase noise.

Jwo-Shiun Sun

Dept. of Electronic Engineering, National Taipei University of Technology, Taipei, Taiwan, ROC

PHASE-LOCKED loops (PLLs) are playing an increasingly important role in communications and tracking systems. In a mobile communication receiver, a PLL is used as part of the frequency synthesizer. It operates as a coherent detector by continuously correcting the voltage-controlled oscillator's (VCO's) output frequency using the phase error between feedback and input. There are three basic types of frequency synthesizer:[1-3] the direct frequency synthesizer, the direct digital frequency synthesizer, and the indirect frequency synthesizer. The indirect frequency synthesizer[1] has advantages over the other two types, including low power consumption, low phase noise, and high stability.

This article explores the design of an indirect frequency synthesizer that can be applied to DCS-1800 digital communication systems. The article discusses the design of the PLL, VCO, loop filter, divider, and methods of manipulating the required channel frequencies. Bandwidth, noise, and stability are considered in the frequency-synthesizer design. The authors use a Bode Plot of the loop stability to determine the optimal component values of the PLL loop filter.



1. This block diagram shows a linear model of the PLL circuit.

DESIGN AND THEORY

A series-tuned VCO is designed to suppress harmonics beyond the fundamental frequency.[4] A common-base transistor with additional feedback increases the VCO's negative-resistance region. The reflection coefficients Γ_{in} and Γ_{out} are expressed as:

$$\Gamma_{in} = S'_{11} + \frac{S'_{12}S'_{21}\Gamma_L}{1 - S_{22}'\Gamma_L} \qquad (1)$$

$$\Gamma_{out} = S'_{22} + \frac{S'_{12}S'_{21}\Gamma_r}{1 - S'_{11}\Gamma_r} \qquad (2a)$$

where:

$\Gamma\gamma$ and Γ_L = the reflection coefficients of a resonator and a matching network, respectively. In addition:

$$S'_{11}, S'_{21}, S'_{12}, S'_{22} \qquad (2b)$$

are the new S-parameters of the transistor with feedback. For a steady oscillation state:

$$Z_{out}(v,\omega) + Z_L(\omega) = 0$$
$$\Rightarrow R_{out}(v,\omega) + X_{out}(v,\omega) + R_L(\omega) + X_L(\omega) = 0 \qquad (3a)$$

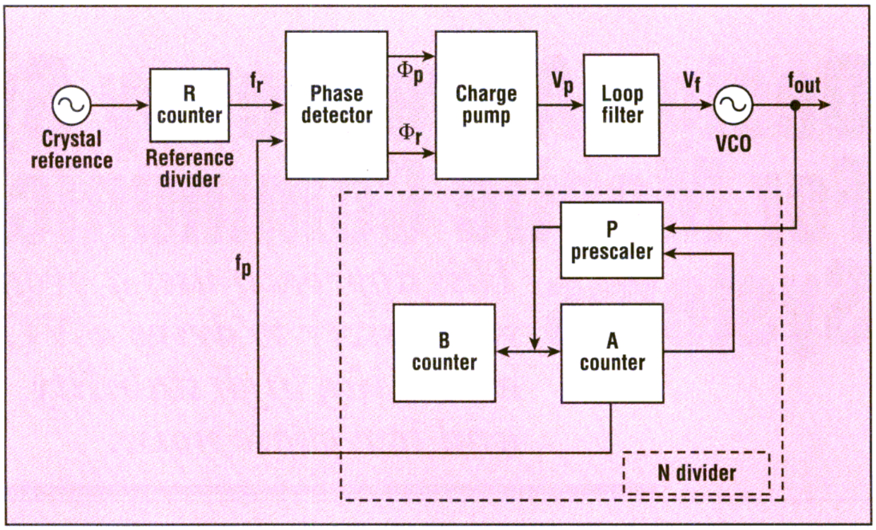
An oscillator's output power will increase until the negative resistance is equal to the load resistance. Generally, the output reactance X_{out} is a function of voltage (V) and angular frequency (ω). That explains the inequality of:

$$X_{out}(v,\omega) \neq X_{out}(v_o,\omega_o) \qquad (3b)$$

Kurokawa[5] indicates that oscillation will be stable as long as:

$$\frac{\partial X_L}{\partial \omega} > 0 \qquad (3c)$$

Figure 1 shows the PLL's linear model with feedback. The phase detector's output voltage V_p is expressed as:



2. This is a block diagram of the designed frequency synthesizer.

$$V_p = K_p\left(\theta_i - \frac{\theta_o}{N}\right) = K_p \cdot \theta_e \qquad (4)$$

where:
θ_i, θ_o, and θ_e = the input phase, output phase, and phase error, and
K_p = the phase-detector/charge-pump gain factor.

The designed VCO is assumed to be a linear device[6] whose output frequency deviates from its free-running frequency by a frequency variation

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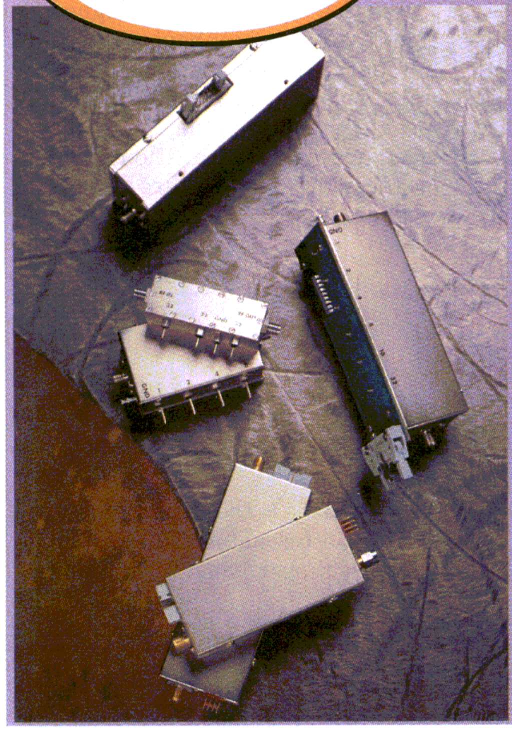

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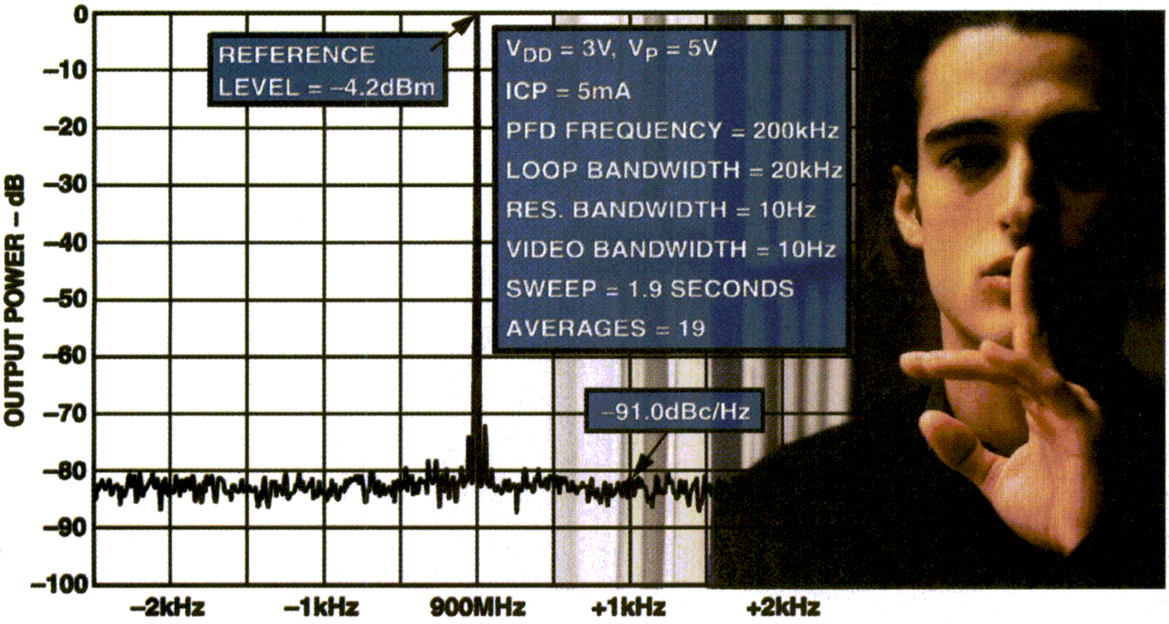

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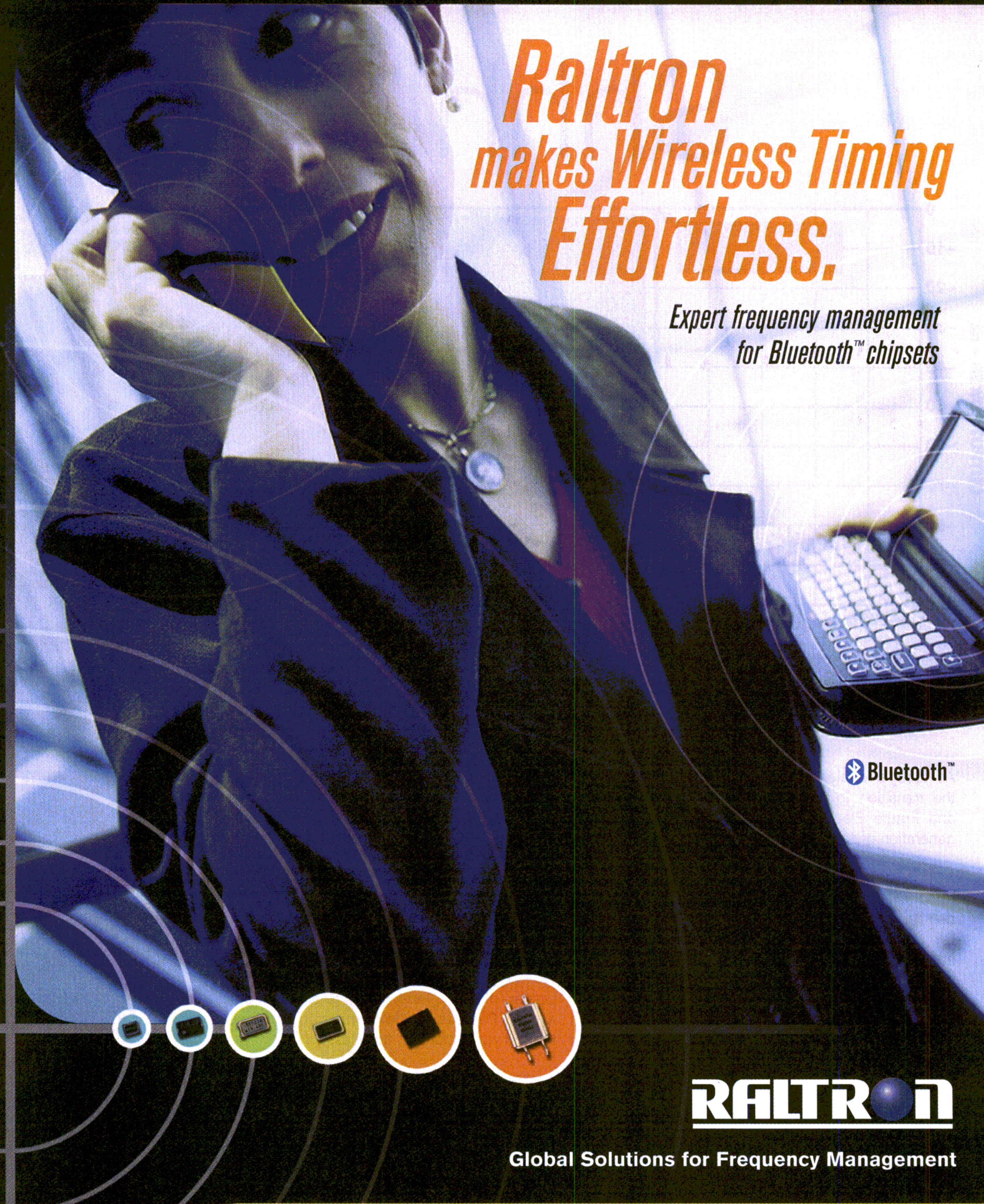

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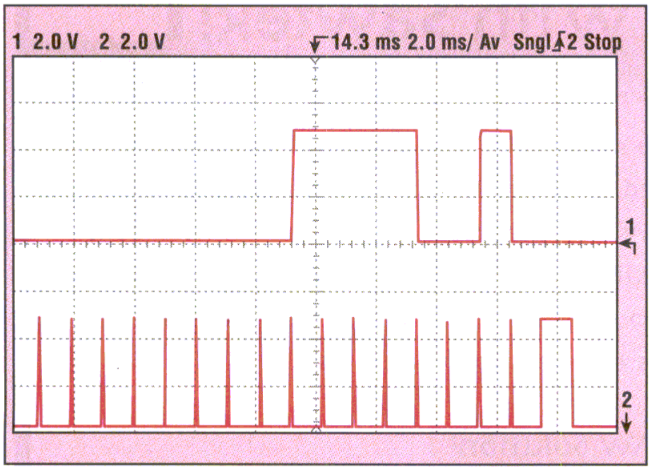
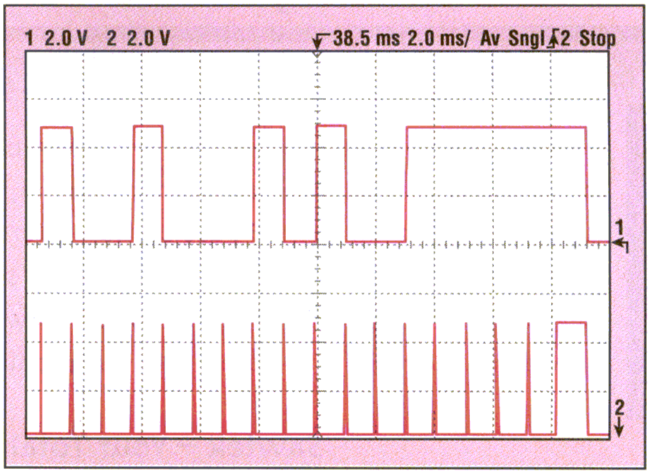


3. This oscillograph shows the data and clock of the R counter.



4. This oscillograph shows the value of the N counter and clock.

signified as:

$$\Delta\omega = \frac{\partial\theta_o}{\partial t} = K_o V_f(t) \qquad (5a)$$

where:

$V_f(t)$ = the loop filter's output voltage, and

K_o = the VCO's gain factor.

In the frequency domain, it is given as:

$$\theta_o(s) = K_o V_f(s)/s \qquad (5b)$$

and open-loop gain is expressed as:

$$T_o(s) = K_d F(s) K_o / Ns \qquad (6a)$$

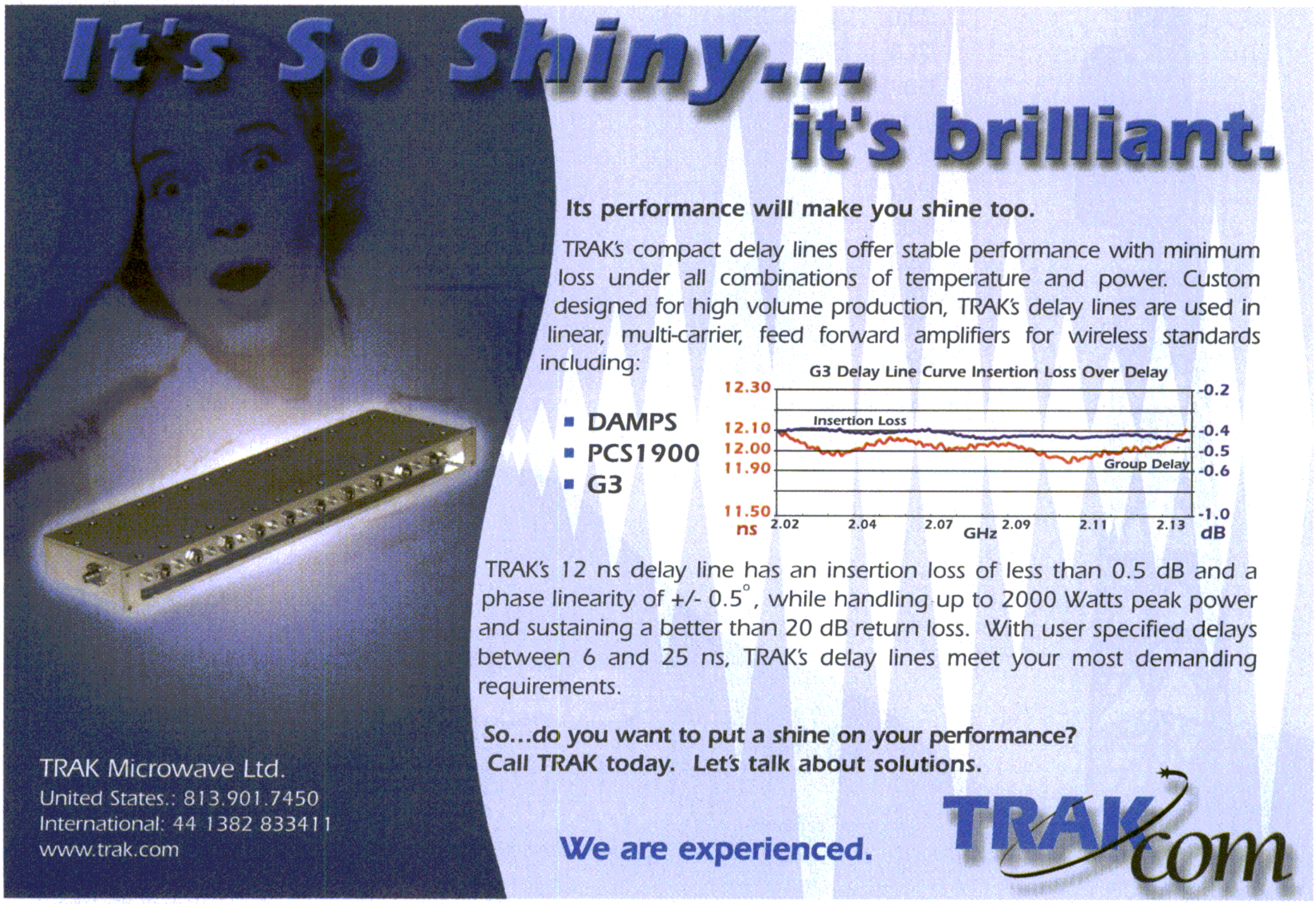
where:

F(s) = the transfer function of the loop filter.

The Bode-diagram method is applied to predict the loop stability by analyzing the open-loop gain $T_0(s)$. The frequency error:

$$f_e(t) = \frac{d}{dt}\theta_e(t) \qquad (6b)$$

is expected to be zero when it is in the steady state. The final value of the



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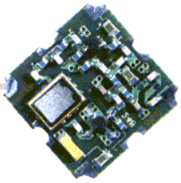

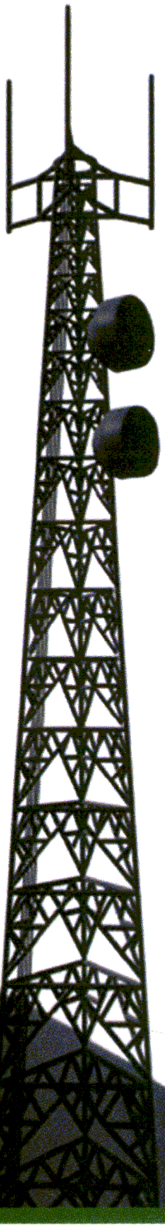

Frequency (MHz)	Passband (MHz)	Insertion Loss (dB)	Modes of Operation	Package	Part Number
69.99	1.26	22.0	SE, BAL	2.875 BT	854550-1
70.0	1.26	22.0	SE, BAL	2.875 BT	854747-1
70.0	1.23	22.0	SE	DIP V	855107
71.0	0.16	8.0	SE, BAL	15.3x6.5mm	855713
150.0	1.18	25.0	SE, BAL	19x6.5mm	854833-1
160.0	1.18	25.0	SE, BAL	19x6.5mm	855049
167.0	0.14	8.0	SE	20.1x9.8mm	854998
170.0	0.18	6.0	SE	13.3x6.5mm	855297
190.0	5.5	12.0	SE, BAL	13.3x6.5mm	855529
199.0	0.20	7.5	SE	19x6.5mm	855131
201.0	0.16	6.0	SE, BAL	15.3x6.5mm	855590
211.0	0.8	8.0	SE, BAL	13.3x6.5mm	855444
380.0	5.4	16.0	SE, BAL	13.3x6.5mm	855530

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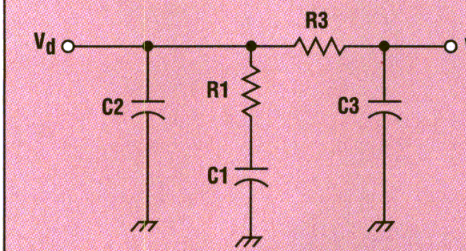
phase error θ_e is:

$$\lim_{t\to\infty}\left[\theta_e(t)\right] = \lim_{s\to 0}\left[s\theta_e(s)\right] = \lim_{s\to 0} s\cdot \left[\frac{\theta_i(s)}{1+K_pF(s)K_o/Ns}\right] = \frac{f_oN}{K_pK_oF(0)} \qquad (7a)$$

The phase error will be zero while

$$\lim_{s\to 0}\left[F(s)\right] \qquad (7b)$$

is an infinite value. This is achievable through the loop filter's pole at its original position. However, since that may compromise loop stability, a lead network is added in F(s).



$$Z(s) = \frac{1+k_1s}{sk(1+k_2s)}$$

$$F(s) = \frac{V_f}{V_d} = \frac{Z(s)\times(1/sC3)}{Z(s)+R3+(1/sC3)}$$

$$k = \frac{1}{C_1+C_2} \qquad k_2 = \frac{C_1\times C_2\times R_1}{C_1+C_2}$$

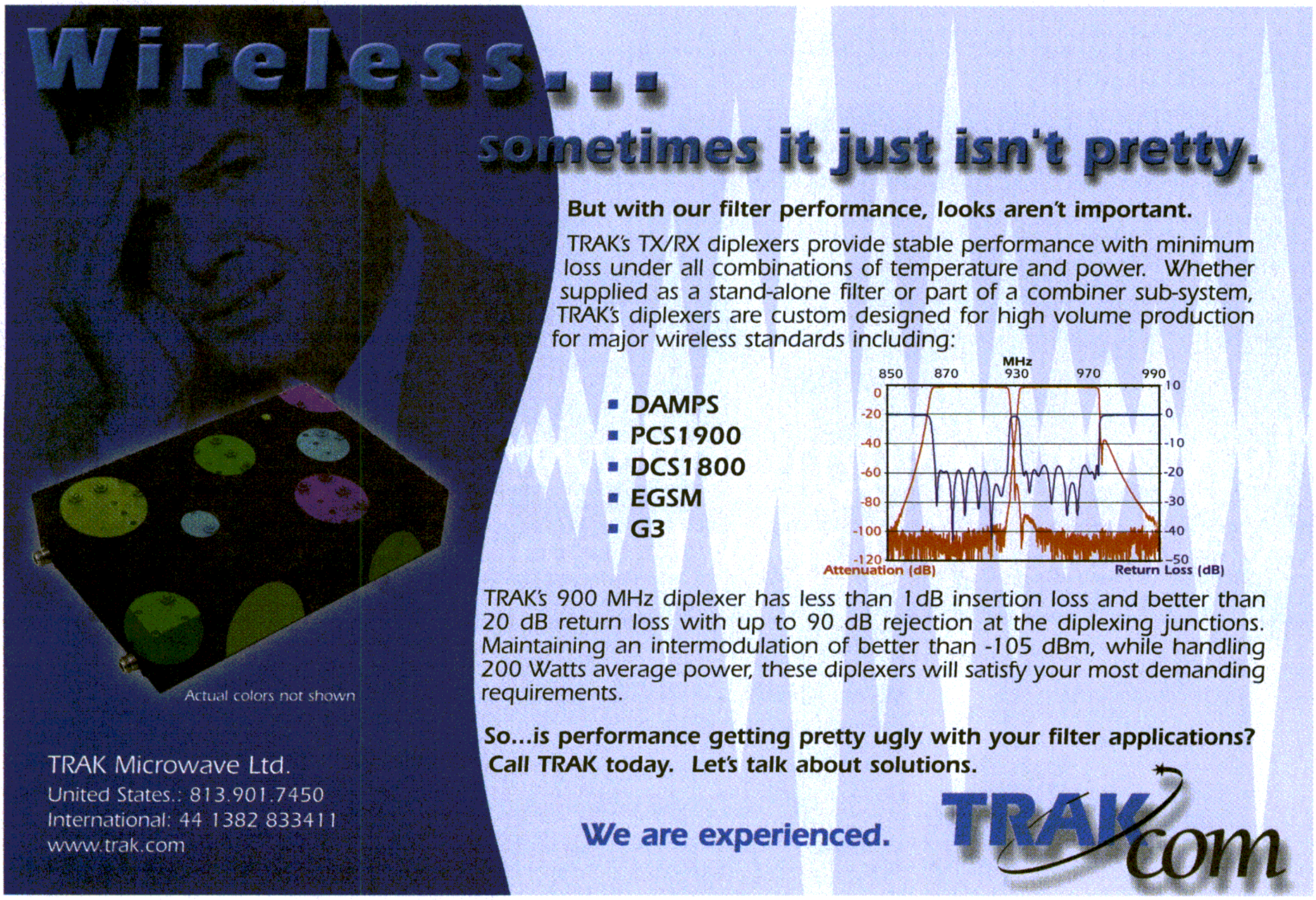
$$k_1 = C_1\times R_1$$

5. The schematic and transfer function of the designed loop filter are shown here.

DISCUSSIONS

A bipolar-junction transistor (BJT) is often employed in the common-base configuration as a wideband oscillator.[7] In the design described herein, additional feedback is applied to a common-base VCO to increase its negative resistance, and a series-tuning method is used to suppress the harmonics.[4] Since a two-port active-device circuit tends to be capacitive in nature,[4] it is hard to achieve wideband tuning if a parallel tunable varactor is used. Therefore, the series-tuned varactor is employed to reduce the equivalent-capacitance ratio and increase the VCO's output frequency.

An oscillator is triggered by thermal noise, and the small-signal S-parameters of the active device determine the initial oscillation condition. Accurate analysis of an oscillator design can only be accomplished by using the large-signal S-parameters.



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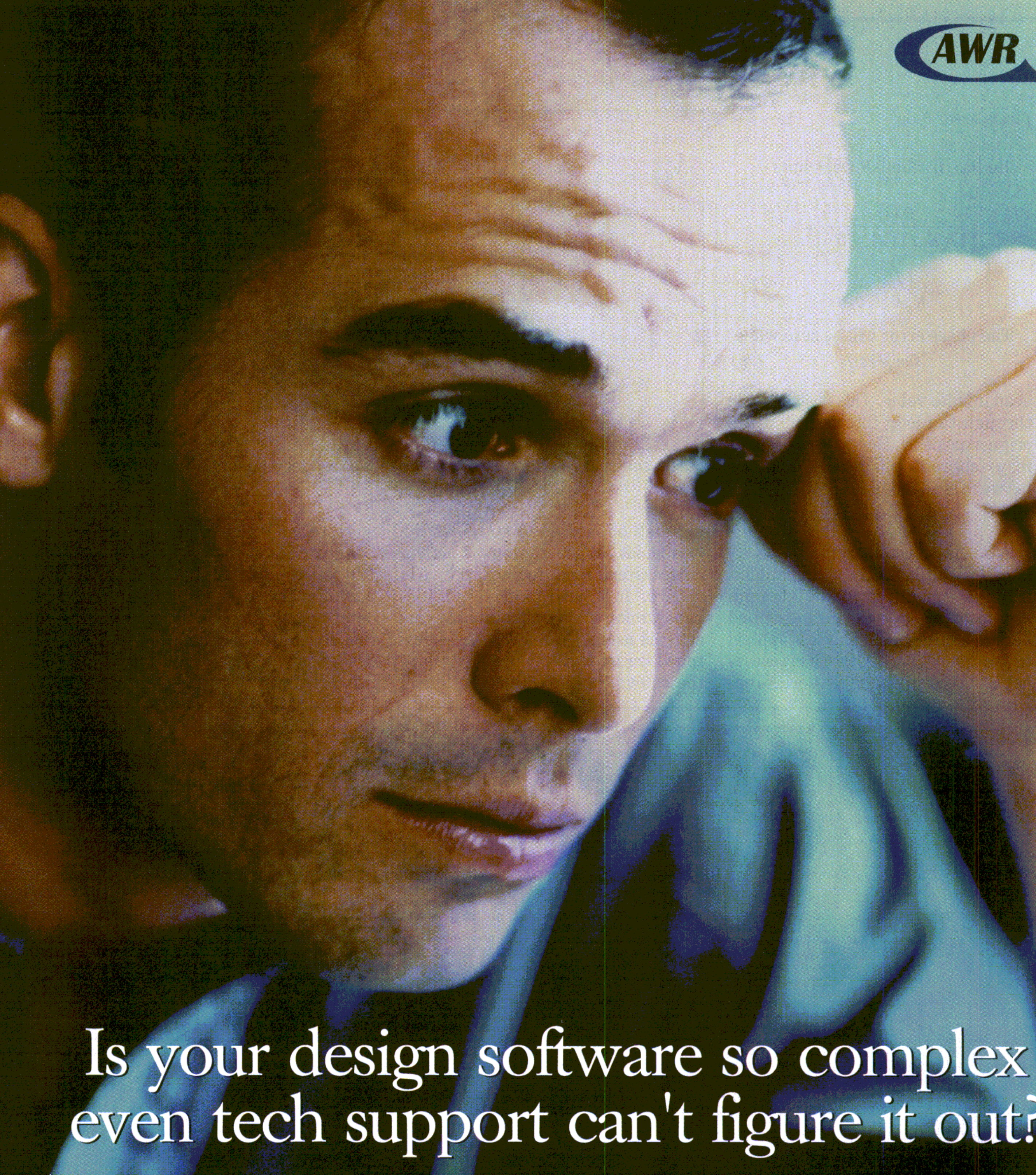
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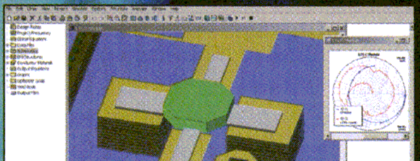
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But the large-signal S-parameters of an active device are difficult to measure. However, literature on this subject[8] indicates that the large-signal S-parameters only change a few percent from their small-signal parameters. For the design described in this article, a modeling program known as AWR Microwave Office is used to simulate the circuit. At frequencies from 1.8 to 2.0 GHz, the circuit provides a tuning linearity of 35 MHz/V and a phase noise of better than –105 dBc/Hz (at 100-kHz off-carrier frequency).

A National Semiconductor model LMX2330 programmable frequency synthesizer provides the tuning voltages for the designed VCO, and a clock stream is programmed to control the synthesizer in the PLL design. The PLL serial-control data are controlled by three inputs—data, LE, and clock. The serial-data input is designed to control the 15 b of the R counter and 18 b of the N counter (which includes 7 b from the A counter and 11 b from the B counter) for regulating the PLL's channel frequency (Fig. 2). When the loop is locked, the output frequency is equal to N times of the reference frequency (f_r). For example, in a radio designed to work in the DCS-1800 digital communication system, a 12-MHz crystal oscillator is used with a reference divider (R counter) of 60 for the 200-kHz channel spacing. Figure 3 shows the R-counter data and clock signal, and Fig. 4 shows the N-counter data and clock signal. Once the R-counter enable reads high, the data are transferred into the latch.

THE LARGE-SIGNAL S-PARAMETERS OF AN ACTIVE DEVICE ARE DIFFICULT TO MEASURE. HOWEVER, LITERATURE INDICATES THAT THE LARGE-SIGNAL S-PARAMETERS ONLY CHANGE A FEW PERCENT FROM THEIR SMALL-SIGNAL PARAMETERS.

For total 22 b of 010110 00000000 11110010

15-b R counter = 60 (00000000 111100)

For total 22 b of 010001 00100010 10011111

N-counter consists of 7 b from the A counter and 11 b from the B counter

A counter = 39 (0 100111)

B counter = 145 (0001 0010001)

The output frequency f_{out} is equal to M (= 64 B + A) × the reference frequency (200 kHz), the product of which is 1863.8 MHz.

The charge-pump circuit in Fig. 2 consists of an inverter and two complementary transistors. Its purpose is to convert the logic states of the comparator into analog signals to drive the VCO. The loop filter must

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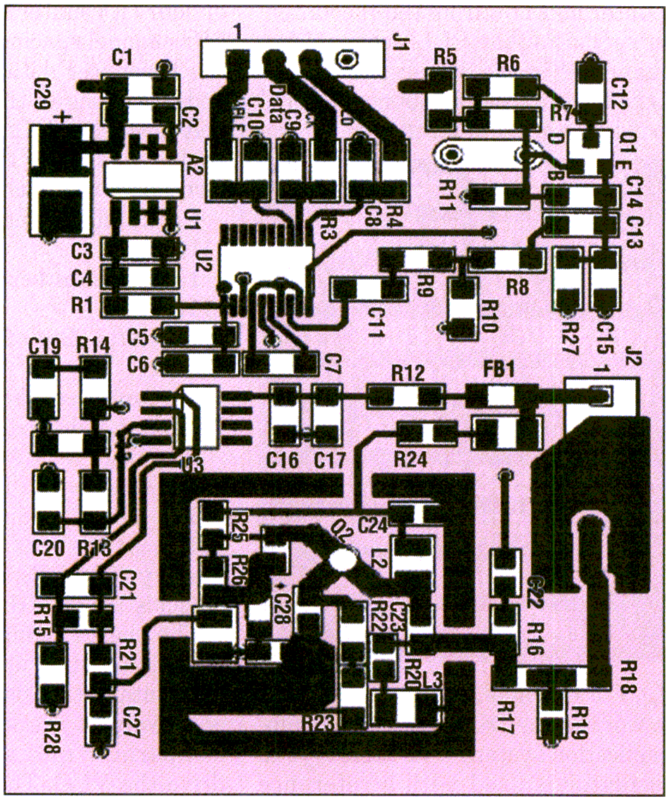

6. This is the PCB layout of the designed frequency synthesizer.

strike a balance between reducing the reference spurs and maintaining the lock time required by the system. An effective method to reduce the PLL's noise is to decrease its bandwidth.[2] The narrower that the loop filter's bandwidth is, the lower the reference spurs are, but the longer the lock time is. Since the sidebands' spurious noise affects the adjacent channels, the unwanted spurs can be suppressed by narrowing the loop filter's bandwidth. The Bode-diagram method is applied to judge the loop stability. Figure 5 shows the designed loop filter with its transfer function F(s). It provides a pole that increases the attenuation slope to –12 dB/octave, and a –6-dB/octave slope where it crosses the 0-dB gain level. To minimize the fundamental-frequency leakage in a synthesizer design,[9] the filter circuitry typically incorporates an additional R3 and C3 (Fig. 6), which provide a pole that increases the attenuation slope to –18 dB/octave.

Phase and gain margins of an open-loop PLL with the optimal loop-filter solution

Frequency (MHz)	Pushing figure (MHz/V)	Phase margin (deg.)	Gain margin (dB)
1800	30	31.2133	8.7183
	40	26.1369	6.2195
	50	20.0474	4.2813
1850	30	31.5465	8.9563
	40	26.7512	6.4575
	50	20.9122	4.5193
1900	30	31.8480	9.1879
	40	27.3294	6.6891
	50	21.6899	4.7509
1950	30	32.1060	9.4135
	40	27.8514	6.9148
	50	22.4565	4.9765

For example, four different kinds of oscillator frequencies and three different kinds of pushing figures in the table are selected. Of course, the optimal component values for the loop filter are concerned with stability, noise, and bandwidth considerations. One can select the component values of C1, C2, R2, C3, and R3 randomly in the 12 possible conditions. Afterward, the component values are put into the open-loop transfer function $T_0(s)$ to see if the phase margin is smaller than 180 deg. while the gain margin

$$\lim_{s\to 0}[F(s)] \qquad (7b)$$

equals zero. If so, then the determined values will be appropriate. The values of C1, C2, R2, C3, and R3 of the loop filter are simulated by more than a thousand times by controlling the bandwidth in the transferred function. From the 12 possible conditions, one can obtain many different values that satisfy the stability requirements. Generally, the optimal component values can be chosen with the phase margin outermost from π. The solutions C1 = 10 pF, C2 = 1 μF, C3 = 100 pF, R1 = 3.8 kΩ, and R2 = 15 kΩ, shown in the table, generate different phase and gain margins to satisfy the loop-stability limitations.

Figure 6 shows the layout of the designed synthesizer. It exhibits better phase-noise performance than the designed VCO, and it can lock signals at the desired DCS-1800 channels. The VCO exhibits a tuning linearity of 35 MHz/V and phase noise under –105 dBc/Hz (at 100-kHz off-carrier frequency) from 1.8 to 2.0 GHz.●●

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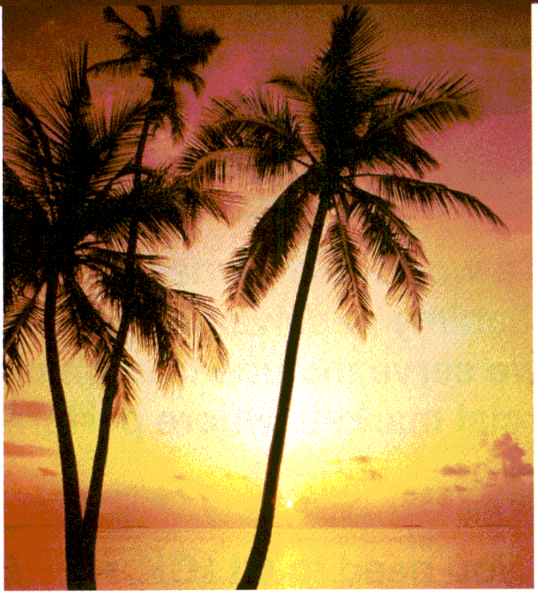

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MSA-2643	3.6	15.9	+21.9	27
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Defining And Testing Dynamic ADC Parameters

Part 1 of this two-part series defines some of the key dynamic parameters of ADCs, including signal-to-noise ratio and total harmonic distortion.

Dynamic ADC Testing, Part 1

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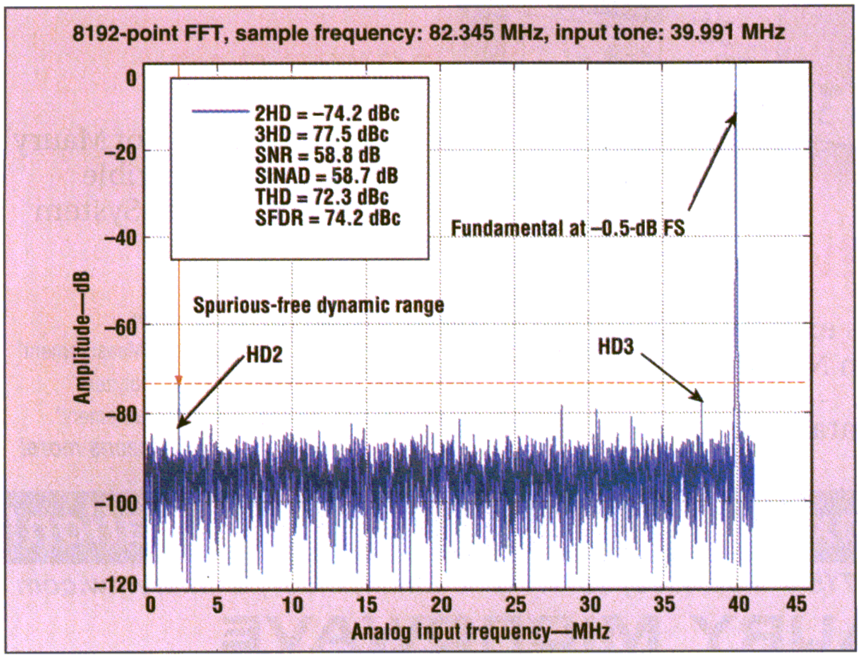
DYNAMIC specifications for high-speed analog-to-digital converters (ADCs) are very important in high-speed applications such as digital communications, ultrasound imaging, instrumentation, and intermediate-frequency (IF) digitization. This first of two articles provides definitions and mathematical foundations for each of the key dynamic ADC parameters, and explains how these dynamic parameters correlate with ADC performance. The key specifications include signal-to-noise ratio (SNR), signal to noise and distortion (SINAD), effective number of bits (ENOB), total harmonic distortion (THD), spurious-free dynamic range (SFDR), two-tone intermodulation distortion (TTIMD), multi-tone intermodulation distortion (MTIMD), and VSWR.

This short article series will conclude next month with insights into the practical aspects of dynamic performance testing. Note that some specifications allow more than one approach for measurement and even for definition. Thus, the test techniques of Part 2 represent one approach, but are not mandatory. Any of the methods described can be extended or altered as necessary to suit the application at hand.

When testing high-speed ADCs, one emulates the operation of an instrument used to quantify linearity in analog circuits—the spectrum analyzer. For this instrument and for the test procedure, dynamic specifications are usually expressed in the frequency domain, using the Fast Fourier transform (FFT). In both cases, the data output represents the magnitude of this FFT. As an example (Fig. 1), consider the FFT plot for an 80-MSamples/s, 10-b ADC designed and optimized for ultrasound imaging and the digitization of baseband/IF signals. These FFT plots contain impressive amounts of information, and they can be quickly generated. But to make use of an FFT, one must understand how its parameters are defined.



1. This 8192-point FFT plot was generated for the MAX1448.

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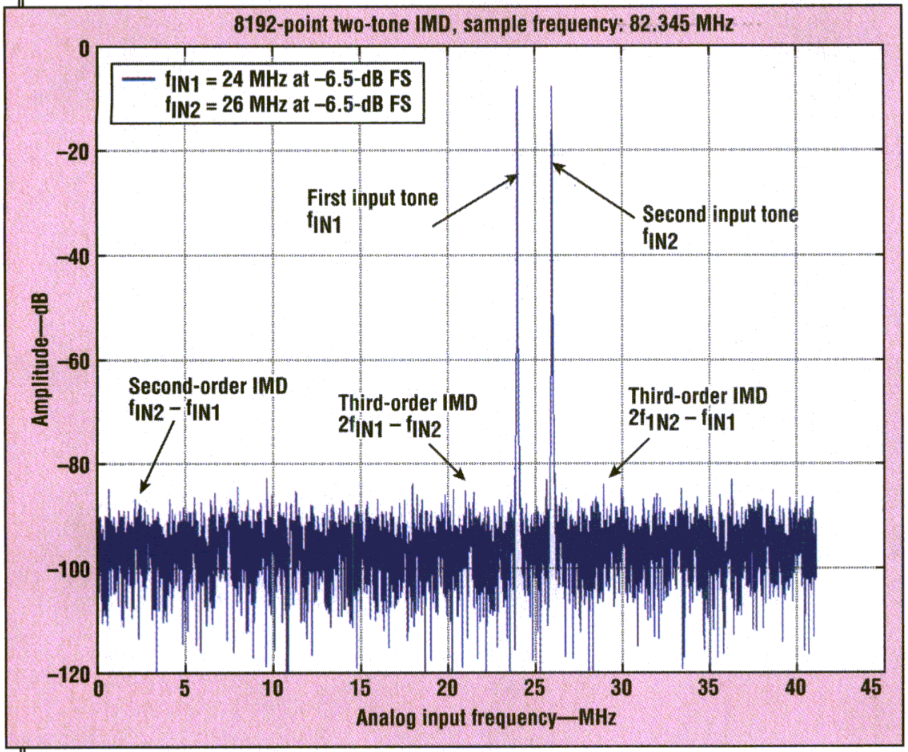

2. The two-tone intermodulation distortion for the MAX1448 was plotted with a sample frequency (f_{SAMPLE}) of 82.345 MHz.

structed from digital samples, the SNR is the ratio of a root-mean-square (RMS) full-scale analog input to its RMS quantization error, $A_{QUANTIZATION}[RMS] = A_{LSB}/(12)^{0.5} = A_{REF}/(2^N(12)^{0.5}$. The RMS value of a sine wave is one-half its peak-to-peak value divided by √2, and quantization error is the difference between an analog waveform and its digitally reconstructed replica, which is characterized by a staircase-shaped transfer curve. The difference function resembles a sawtooth wave that oscillates once per sample between the least-significant-bit (LSB) levels of +0.5 LSB and −0.5 LSB.

The difference function's RMS value is its peak value (0.5 LSB) divided by √3. For an ideal N-bit converter, the SNR is defined as:

$$SNR = 2^N \times (\sqrt{3}/\sqrt{2})$$
$$= 1.225 \times 2^N$$

Most of the dynamic specifications are expressed as a ratio of relative measurements rather than absolute units. Thus, the SNR for an ideal ADC, driven by a full-scale sinusoidal input with AC power equal to $A_{REF}/(2\sqrt{2})$ [in decibels], is:

$$SNR_{dB} = 20\log_{10}(A_{IN}[RMS]/A_{QUANTIZATION}[RMS])$$
$$SNR_{dB} = 20\log_{10}(A_{REF}[2\times\sqrt{2}]/A_{REF}/[2^N\times\sqrt{12}])$$
$$SNR_{dB} = 6.02N + 1.763. \quad (1)$$

SNR is diminished by many noise sources in addition to quantization noise (see the sidebar "Decoding noise"). A data converter's resolution and quantization level help to establish its noise floor. The actual SNR for a sinusoidal input signal can therefore be described as:

$$SNR_{dB} = 20\log_{10}(A_{SIGNAL}[RMS]/A_{TOTAL\_NOISE}[RMS]), \quad (2)$$

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$A_{SIGNAL}[RMS]$ = the RMS amplitude for the analog input signal, and

$A_{TOTAL\ NOISE}[RMS]$ = the RMS sum of all noise sources (thermal noise, quantization noise, etc.) that limit the converter's dynamic performance. Applying this definition to a 10-b ADC, such as the MAX1448 from Maxim Integrated Products (Sunnyvale, CA), yields a typical SNR value of 58.4 dB at the 40-MHz Nyquist frequency (f_{SAMPLE} = 80 MSamples/s). This SNR represents 94 percent of the ≈62-dB SNR exhibited by an ideal 10-b ADC.

For an ADC driven by a sinusoidal input with an amplitude equal to the ADC's full-scale input, the maximum theoretical SNR is:

$$SNR_{dB} = 6.02N + 1.763 + 10\log_{10} (f_{SAMPLE} / 2)(f_{MAX}), \quad (3)$$

where:

f_{MAX} = the maximum bandwidth of the input tone, and

f_{SAMPLE} = the converter's sampling frequency.

From this equation, note that SNR increases as the sampling frequency increases beyond the Nyquist rate of ($2f_{MAX}$). Known as processing gain, this effect is caused by spreading of the quantization noise power (which is fixed and independent of bandwidth) as the sampling frequency increases. This "oversampling" helps to minimize the effect of noise, which falls into the Nyquist bandwidth of DC to f_{MAX}.

For sinusoidal input signals, SINAD is defined as the ratio of RMS signal to RMS noise (including the first N harmonics of THD—usually the second- through fifth-order harmonics). For a particular sampling rate and input frequency, SINAD provides the ratio (in decibels) of the analog input signal to the noise plus distortion. SINAD describes the quality of an ADC's dynamic range, expressed as the ratio of the maximum amplitude output signal to the smallest increment of output signal that the converter can produce. Mathematically, SINAD is described as:

$$SINAD_{dB} = 20\log_{10} (A_{SIGNAL}[RMS] / A_{NOISE+HD}[RMS]), \quad (4)$$

where:

$A_{SIGNAL}[RMS]$ = the RMS output signal level, and

$A_{NOISE+HD}[RMS]$ = the RMS sum of all spectral components below the Nyquist frequency, excluding DC.

The quality of SINAD also depends on the amplitude and frequency of a sinusoidal input tone.

For actual (versus ideal) ADCs, a specification often used in place of the SNR or SINAD is ENOB, which is a global indication of ADC accuracy at a specific input frequency and sampling rate. It is calculated from the converter's digital data record as N − $\log_2$ of the ratio of measured and ideal RMS error:

$$ENOB = N - \log_2 (A_{MEASURED\_ERROR}[RMS] / (A_{IDEAL\_ERROR}[RMS]), \quad (5)$$

where:

N = the number of digitized bits,

$A_{MEASURED\ ERROR}[RMS]$ = the averaged noise, and

$A_{IDEAL\ ERROR}[RMS]$ = the quantization noise error, expressed as $q/(12)^{0.5} = A_{FS}/(2^N(12)^{0.5}$. Note that A_{FS} is the converter's full-scale input range as determined by the voltage reference, A_{REF}:

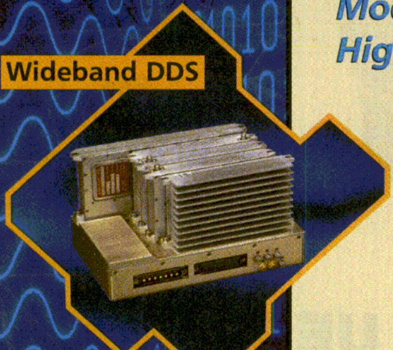
$$ENOB = \log_2 (A_{FS} / A_{MEASURED\_ERROR}[RMS]\sqrt{12} \quad (6)$$

or

$$ENOB = \log_2 (A_{REF} / A_{MEASURED\_ERROR}[RMS]\sqrt{12} \quad (7)$$

ENOB generally depends on the amplitude and frequency of the applied sinusoidal input tone, and both must be specified for this particular test. This method compares the RMS noise produced by the ADC under test to the RMS quantization noise of an ideal ADC with the same resolution in bits. If an actual 10-b ADC with a sine-wave input of a particular frequency and amplitude has an ENOB = 9 b, then it produces the same RMS noise level for that input as an ideal 9-b ADC would.

Directly related to SINAD, ENOB



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is frequently expressed as ENOB = (SINAD − 1.763)/6.02. The error of an ideal ADC consists solely of noise. For actual converters, however, the measured error includes quantization noise along with aberrations such as missing output codes, AC-to-DC nonlinearity, and aperture uncertainty (jitter). Noise on the reference and power-supply lines also degrades the ENOB.

Dynamic errors and integral nonlinearities contribute to harmonic distortion whenever an ADC samples a periodic signal. For pure sine-wave inputs, the output harmonic-distortion components are found at spectral values whose non-aliased frequencies are integer multiples of the applied sinusoidal input tone. The amplitudes of the non-aliased frequencies, which depend on the amplitude and frequency of the applied input sine wave, are generally provided as a decibel ratio with respect to the amplitude of the applied sine-wave input. Their frequencies are usually expressed as a multiple of the frequency of the applied sinusoidal input signal.

THD is the RMS sum of all harmonics in the output signal's FFT spectrum. In communications, high-speed instrumentation, and RF/IF applications, THD is often a more important figure of merit for ADCs than are the DC nonlinearity specifications that describe the converter's static performance. THD is given by:

$$THD_{dBc} = 20log_{10} (\{A_{HD\_2[RMS]}{}^2 + A_{HD\_3[RMS]}{}^2 + \ldots + A_{HD\_N[RMS]}{}^2\}^{0.5} / A[f_{IN}]_{RMS}), \quad (8)$$

where:

$A[F_{IN}]_{RMS}$ = the RMS fundamental amplitude, and

$A_{HD\_2[RMS]}$ through $A_{HD\_N[RMS]}$ = the RMS amplitudes of the second- through Nth-order harmonics.

The choice of harmonic components included in a set is usually a trade-off between the desire to include all harmonics with a significant portion of the harmonic-distortion energy, and the exclusion of Discrete Fourier Transform (DFT) frequency bins, whose energy con-

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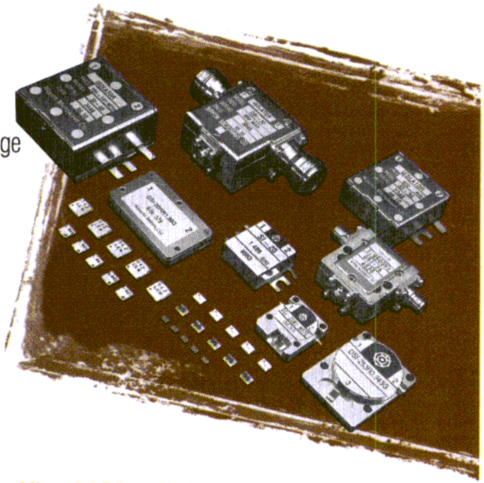

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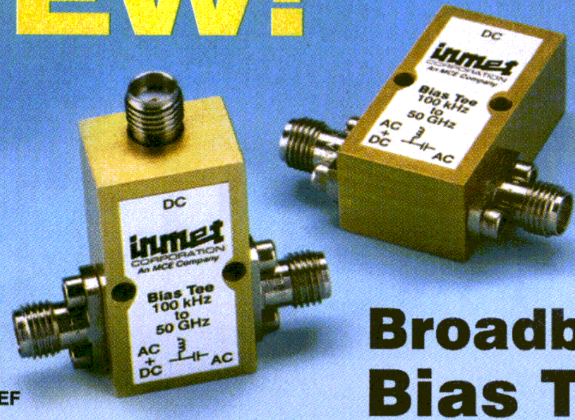

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tent is mainly dominated by random noise (see the sidebar "DFTs or FFTs?").

Unless otherwise specified (refer to the manufacturer's specification in the data sheet), THD normally consists of the lowest four to nine harmonics (the second through tenth harmonics, inclusive) of the sinusoidal analog input tone. Note that manufacturers may specify their THD values either in decibels (dB) or decibels with reference to the level of the carrier frequency or fundamental (dBc). Both units are commonly used, and THD is defined with respect to the analog input tone.

The term SFDR is usually applied to cases where the harmonic distortion and spurious signals are regarded as undesirable content in the output spectrum of a sampled pure sinusoidal input tone. SFDR indicates the usable dynamic range of an ADC, beyond which a spectral analysis poses special detection and threshold problems. Though similar to THD, SFDR addresses the converter's in-band harmonic characteristics.

SFDR is the ratio of RMS amplitude of the fundamental tone (the maximum signal component) to the RMS value of the largest distortion component in a specified frequency range. In well-designed systems, this spurious signal should be a harmonic of the fundamental. SFDR is important because noise and harmonics restrict a data converter's dynamic range. In an IF bandpass converter, for example, spurious signals may be interpreted as adjacent-channel information.

In other applications, signals of interest, such as low-level radar signals, cannot be distinguished from the harmonic content. To help determine the SFDR value, a spectrum analyzer with an integrated digital-to-analog converter (DAC) for reconstruction is recommended. The usual procedure is to apply a near full-scale input signal (the preferred input-tone amplitude is −0.5 dB to −1 dB FS), measure the response, and then acquire and measure the amplitude of the largest spurious component. SFDR is the ratio of the first to the second measurement. SFDR can also

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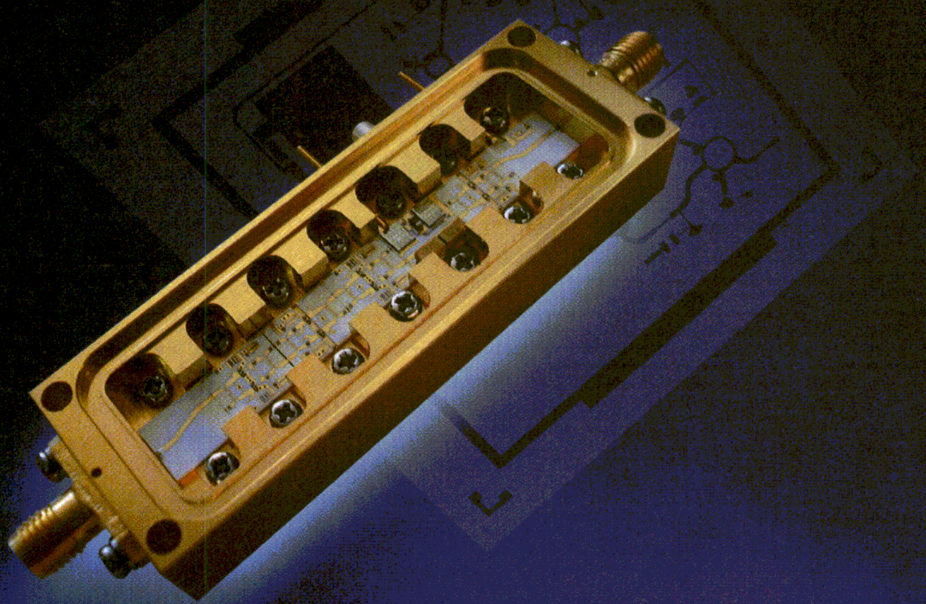

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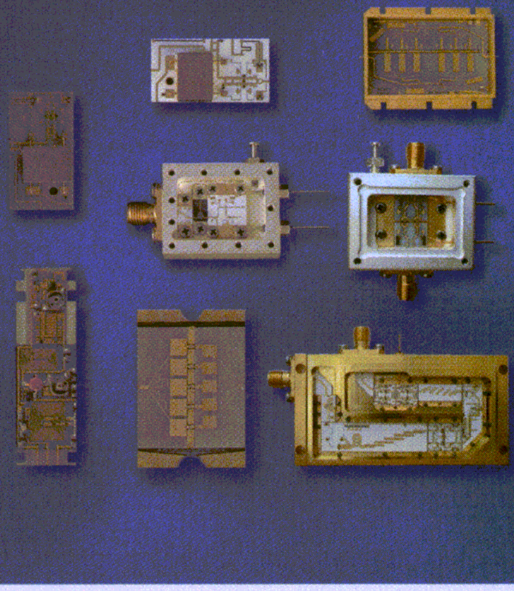

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be determined by inspecting the FFT spectrum (plot) of an ADC under test.

For spectrally pure sine-wave inputs, SFDR is the ratio of the amplitude of the averaged DFT value at the fundamental frequency ($A[f_{IN}]$) to the amplitude of the averaged DFT value of the largest-amplitude harmonic ($A_{HD\_MAX}[RMS]$) or spurious signal component ($A_{SPUR\_MAX}[RMS]$) observed over the entire Nyquist band:

$$SFDR_{dBc} = 20log_{10} (|A[f_{IN}]_{RMS}| / |A_{HD\_MAX}[RMS]|), \text{ or} \quad (9)$$

or

$$SFDR_{dBc} = 20log_{10} (|A[f_{IN}]_{RMS}| / |A_{SPUR\_MAX}[RMS]|). \quad (10)$$

In general, SFDR is a function of the amplitude ($A[f_{IN}]$) and frequency (f_{IN}) of the analog input tone, and, in some cases, even the sampling frequency (f_{SAMPLE}) of the converter under test. When testing an ADC for its SFDR, therefore, the sampling frequency, as well as the input frequency and amplitude, should be specified.

TTIMD is generally caused by modulation, and it can occur when an ADC samples a signal composed of two (or multiple) sine-wave signals. IMD spectral components can occur at the sum (f_{IMF_SUM}) and difference ($f_{IMF\_DIFFERENCE}$) frequencies for all possible integer multiples of the fundamental (input frequency tone) or signal-group frequencies.

When performing TTIMD testing, the input test frequencies—f_{IN1} and f_{IN2}($f_2>f_{IN1}$)—are set to values that are odd numbers of the DFT bins, and away from the Nyquist frequencies ($f_{SAMPLE/2}$). These settings guarantee that the difference between the two input tones is always an even number of DFT bins. The resulting spectrum is the averaged amplitude spectrum, $A[f_{IMF}]_{RMS}$. The IMD amplitudes for a two-tone input signal are found at the specified sum and difference frequencies:

$$f_{IMF\_SUM} = |mf_{IN1} + nf_{IN2}| \text{ and}$$
$$f_{IMF\_DIFF} = |mf_{IN1} - nf_{IN2}|, \quad (11)$$

where:

m and n = positive integers.

The condition that m and n are greater than zero creates the second-order ($f_{IN1} + f_{IN2}$ and $f_{IN1} - f_{IN2}$) and third-order ($2f_{IN1} + f_{IN2}$, $2f_{IN1} - f_{IN2}$, $f_{IN1} + 2f_{IN2}$ and $f_{IN2} - 2f_{IN2}$, $3f_{IN1}$ and $3f_{IN2}$) IM products.

Since test parameters are generally application specific, no particular guidelines are necessary (or available) to specify the frequencies and signal amplitudes used for IM tests. The size of $| f_{IN2} - f_{IN1} |$ depends entirely on the application and the information desired. Note that small differences in the two input tones cause the IM frequencies to be clustered around the harmonic-distortion components of f_{IN1} and f_{IN2}.

(continued on p. 162)

DFTs OR FFTs?

Testing high-speed analog-to-digital converters (ADCs) for their dynamic performance often requires a frequency transform of the captured data record, using Discrete Fourier transform (DFT) or Fast Fourier transform (FFT) analysis. An FFT produces the same results as the DFT, but minimizes the computation requirements by taking advantage of computational symmetries and redundancies within a DFT analysis. By speeding up the computation, this approach enables a spectral analysis in virtual real time.

Provided that a periodic input signal is sampled frequently enough (i.e., $\geqslant 2f_{MAX}$, where f_{MAX} is the maximum bandwidth of the sinusoidal-input test tone, not the bandwidth of the data converter to be tested), the DFT equation pair is defined as:

$$x[n] = 1/N \sum_{k=0}^{N-1} X[k] \times e^{-j(2\pi kn)} \quad (1)$$

$$X[k] = \sum_{n=0}^{N-1} x[n] \times e^{j(2\pi kn/N)} \quad (2)$$

The acquired data record usually contains sinusoidal input signals, harmonics, intermodulation (IM) products, and other spurious signals that must be analyzed to properly characterize an ADC. Assuming that all input signals are periodic, the DFT of a data record not containing an integral number of cycles of all sinusoidal input signals will contain spectral components at frequencies other than those corresponding to the chosen input tones. Also known as spectral leakage, these components should be avoided because they mask spurious performance of the ADC itself. For a precise characterization, spectral leakage must be kept at a minimum by choosing the proper input tones (with respect to f_{SAMPLE}), and by the use of low-noise, high-precision signal sources.

To avoid spectral leakage completely, the method of coherent sampling is recommended. Coherent sampling requires that the input- and clock-frequency generators are phase locked, and that the input frequency be chosen based on the following relationship:

$f_{IN}/f_{SAMPLE} = N_{WINDOW}/N_{RECORD}$,
where:
f_{IN} = the desired input frequency,
f_{SAMPLE} = the clock frequency of the data converter under test,
N_{WINDOW} = the number of cycles in the data window (to make all samples unique, choose odd or prime numbers), and
N_{RECORD} = the data record length (for an 8192-point FFT, the data record is 8192 points long).

Since the ratio of f_{IN} and f_{SAMPLE} is an integer value, the signal and clock sources must have adequate frequency tuning resolution to prevent spectral leakage.

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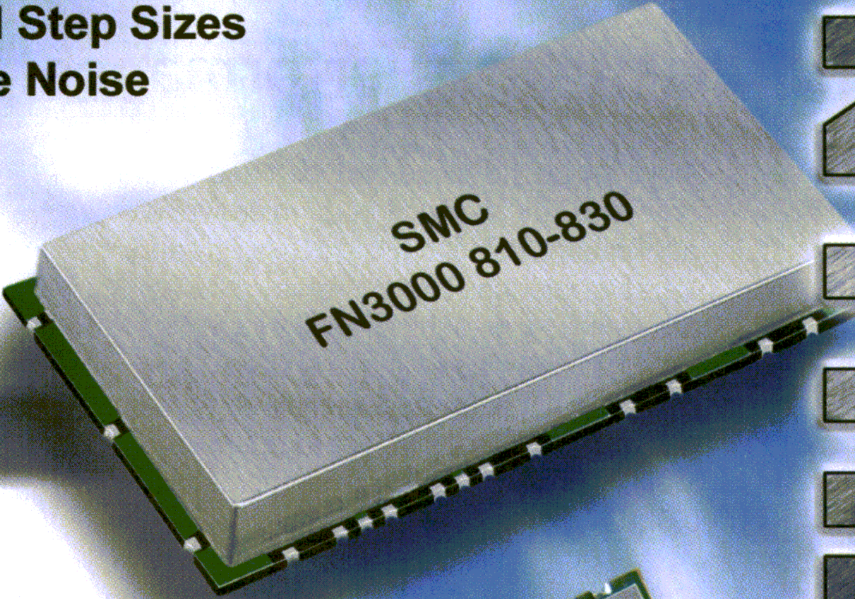

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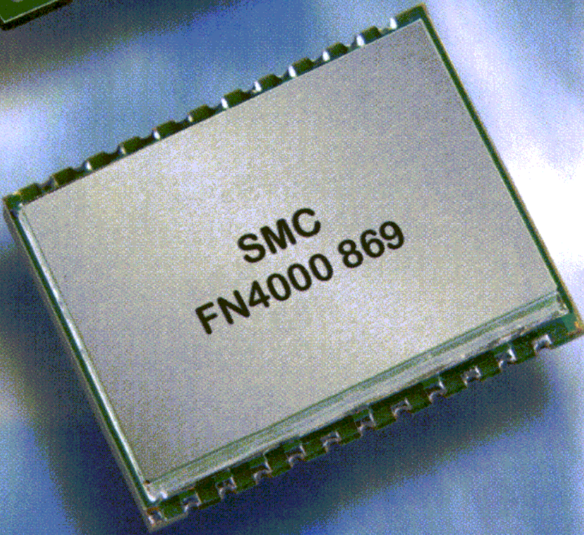

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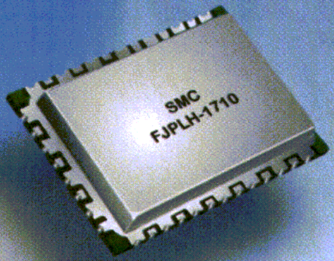

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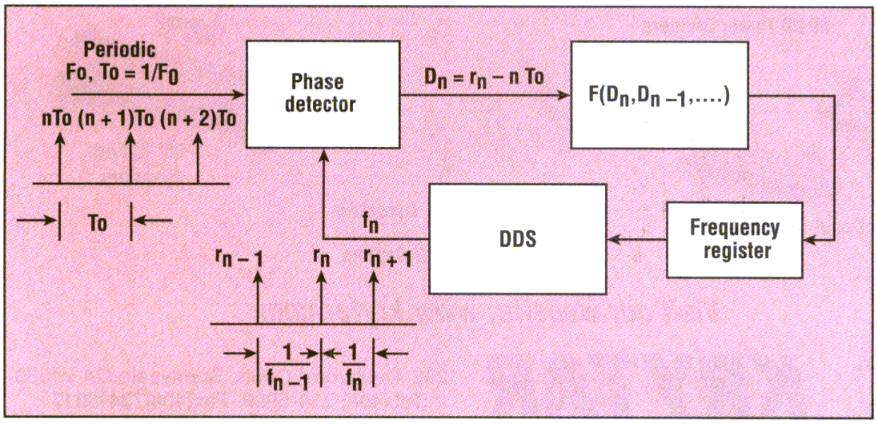
PHASE-LOCKED loops (PLLs) used in frequency synthesizers are based generally on a voltage-controlled oscillator (VCO). The frequency range of a usable VCO is limited to an octave. The direct-digital-synthesizer (DDS) approach is increasingly popular due to its wider frequency range, high resolution, and because its clock frequency can reach up to 200 MHz. But can a PLL with DDS be locked to a baseband signal even when the signal frequency is a not stable one?

The basic idea of this loop is similar to a popular synthesizer (Fig. 1). The phase detector is a counter, which runs continually and can be initialized by the periodic signal nTo or by the DDS-produced pulses r_n. If it is initialized by nTo and stopped by r_n, the count value displays a positive sign. Conversely, if the counter is initialized by r_n and stopped by nTo, it displays a negative sign. The filter $F(D_n, D_{n-1})$ is second order, that is $F = AD_n + B D_{n-1}$. The frequency of the DDS is determined by the value of a frequency register. For convenience, the convergence of the loop is tested by assuming the circuit has high resolution (i.e., a high clock frequency). The value of time D_n and frequency f_n are analog values.



1. A DDS-based frequency synthesizer contains a phase detector, filter, and a frequency register that sets the operating frequency.

LOOP CONVERGENCE

A pair of difference equations is used to describe the phase-locked process, while differential equations are used in a VCO synthesizer. The basic difference equations are:

$$r_{n+1} = r_n + 1/f_n \qquad (1)$$

$$f_{n+1} = f_n + A\,[r_{n+1} - (n+1)\,To] + B\,[r_n - n\,To] \qquad (2)$$

Finding values of A and B that can guarantee the convergence of this loop is not easy. Since a closed solution cannot be found, trial-and-error solutions must be used. In Eq. 2, f_n is on the order of 1/To. So, A and B must be $(1/T)^{\alpha}$ where α is a power. Extensive trial-and-error testing yields the following effective values:

$$A = 1/To^{1.7} \quad B = -A + 12/To \quad (3)$$

The test range is:
$1/1000/s \geq To \geq 1/20 \times 10^{-6}$. That means the bit rate of the input-signal ranges from 1000 b/s up to 20 Mb/s.

While this is not the optimum solution, it is one attempt at a universal solution across a wide range. If only a narrow-range solution is desired, there could be better A and B values.

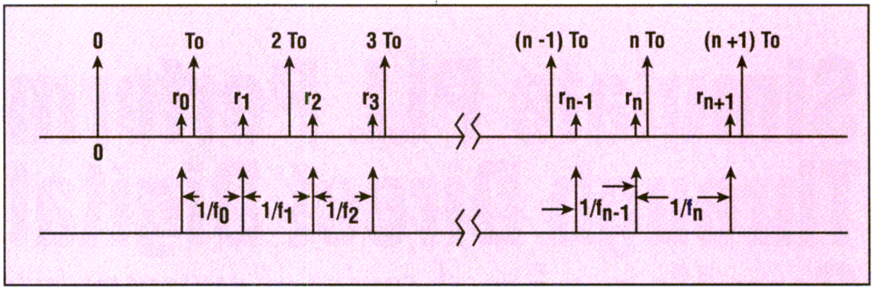
In Eq. 2, the frequency adjustment term, $A[r_{n+1} - (n+1)To] + B[r_n - nTo]$, could be rewritten as $AD_{n+1} + BD_n = A(D_{n+1} - D_n) + (A+B) \times D_n$.

According to Eq. 3, A is a large number, while A+B is a relatively small number. D_n is the phase difference.

All the simulations are under these assumptions:

- To is constant.
- The origin of the time begins with $0 \times To$.
- The first DDS pulse time r_0 is 0.9To.

Naturally, r_0 is random, $r_0 = 0.9To$ is the worst case for convergence. The first two pulses, $0 \times To$ and r_0, do not lead to frequency adjustment, because the second-order filtering needs two successive D_n. In the meantime, assume the filtering calculation is fast, and the result appears instantly. In the real world, it takes time to accomplish the arithmetical operations. The time delay does not change the convergence attribution, but it prolongs the convergence time.



2. The relationship between the pulse sequences that are input to the phase detector (Fig. 1) will determine the phase and frequency characteristics of the synthesizer.

In general, the nominal-input bit rate is known. In the first case, assume the initial frequency f_0 of the DDS is set to approximately 1/To.

Case 1. $f_0 \approx 1/To$.

Example 1: $f_0 = 20.1 \times 10^6$Hz, $r_0 = 0.9To$ and $To = 1/20 \times 10^{-6}$s (bit rate of input signal is 20 Mb/sec). The sequence pairs are shown in Fig. 2.

The results are shown in Fig. 3a and b. The y-axis of Fig. 3a is the phase difference:

$r_n/To - n$, which represents the relative position of r_n and D_n To in the units of To. If the difference holds at a small constant amount, the loop is locked. The y-axis of Fig. 3b is the frequency difference, or the DDS frequency minus input-signal frequency; $f_n - 1/To$. Both x-axes represent the sequential number, n, of

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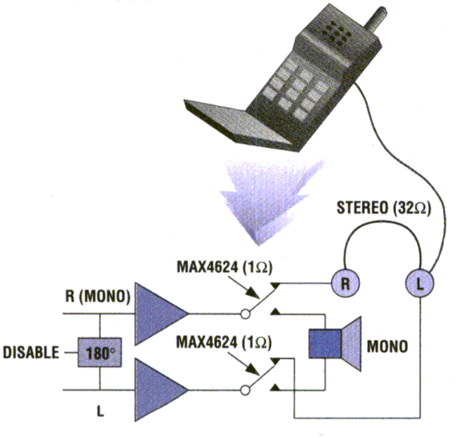

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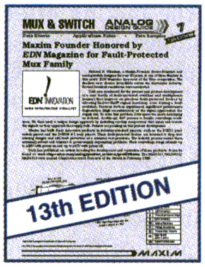

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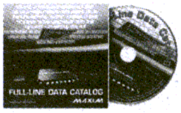

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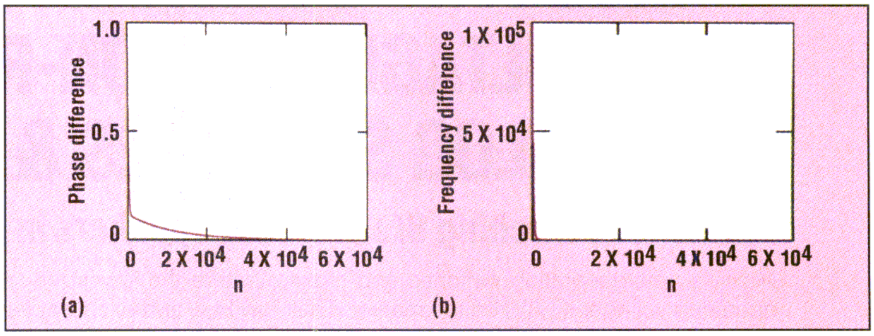
iterations.

After 50,000 iterations, which takes 2.5 ms, the phase difference is 0.103 percent of To, and the frequency difference between 1/To and the DDS is 1.941 Hz, or 0.097 PPM.

Case 2. Mixing Effect (a more detailed consideration of Case 1).

In Fig. 2, the r_n and nTo pair is the pulse time that begins and stops the counter. After the nTo pulse, the frequency is adjusted and results in a new frequency, f_{n+1}. Before the nTo pulse, the frequency is still f_n. In Fig. 2, the arrow designated as $1/f_{n+1}$ is really a combination of $1/f_n$ and $1/f_{n+1}$. To study this effect, use a different version of Eq. 1, and solve the two difference-equation system.



3. The phase (a) and frequency (b) convergence times (number of iterations) determine how fast the loop comes into lock.

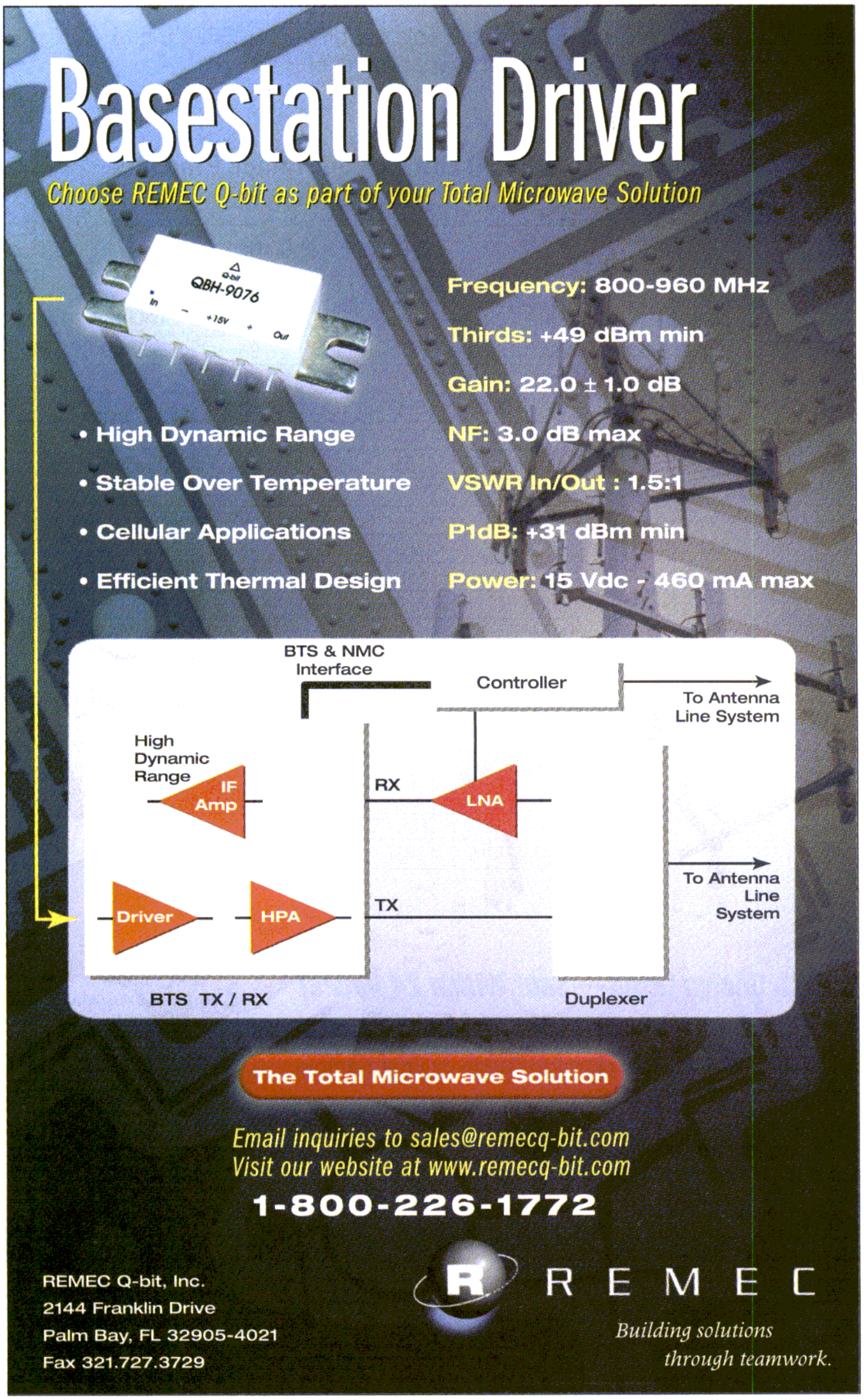
Example 2: $f_0 = 2.05 \times 10^6$ Hz, $r_0 = 0.9$To, and To = 0.5×10^{-6}s, a (bit rate of input signal of 2 Mb/s).

The result is shown in Figs. 4a and b. The y-axis of Fig. 4a is still the same as in Fig. 3a, the phase difference between r_n and nTo. One curve is the result of not considering the mixing effects (Eq. 1). The second is the result of considering the mixer effect. These two curves—the frequency difference—almost coincide. Fig. 4b is the DDS frequency difference, f_n, between both. It shows a difference of less than 70 Hz at the beginning. After 200 iterations, there is no need to consider the mixing effect because f_n is stable.

The convergence characteristics are as follows:

- Not consider the mixing effect: After 15,000 iterations, the elapsed time is 7.5 ms, the phase difference is –0.0813 percent of To, the frequency difference between 1/To and DDS is 0.7879 Hz, or –0.39 PPM.
- Considering the mixing effect: After 15,000 iterations, the elapsed time is 7.5 ms, the phase difference is –0.0813 percent of To, the frequency difference between 1/To and DDS is 0.7881 Hz, or – 0.39 PPM.

Case 3. If the initial frequency f_0 is larger than 1/To, the r_n pulses will successively hit the counter before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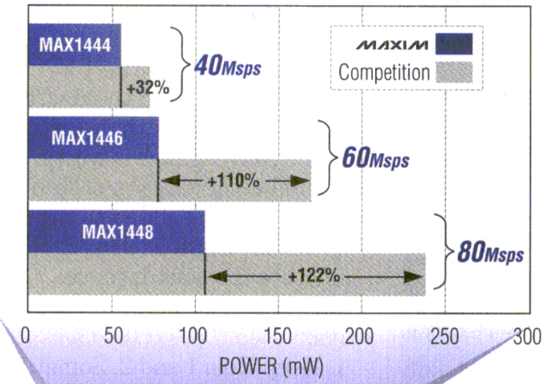

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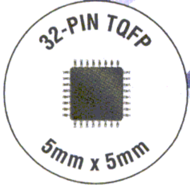

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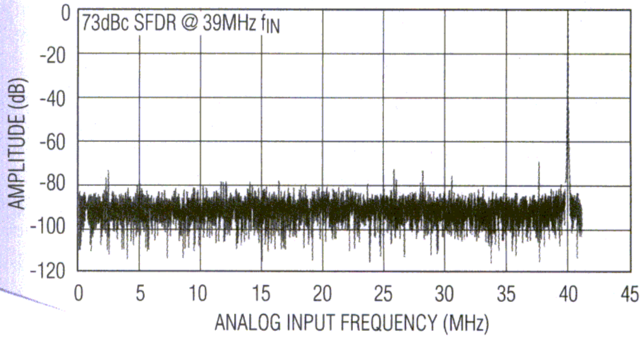

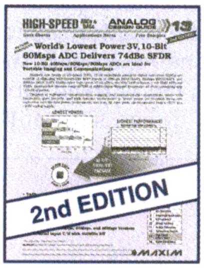
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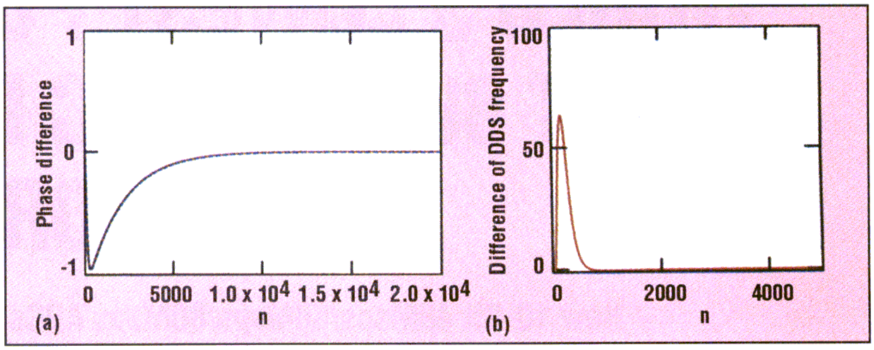


4. Mixing effects produce the phase (a) and frequency (b) curves shown here. The frequency difference between this curve and the one in Fig. 3b are virtually indistinguishable.

the nTo pulse ends the counting. Fig. 5 shows there are r_2 and r_3 pulses before the 2To pulse.

In general, r_n, r_{n+1},... r_{n+k-1} is smaller than (before) nTo, but r_{n+k}, where $k \geq 1$, is not smaller than nTo. So the counter begins counting with r_n (it is active) and skips r_{n+1},...r_{n+k-1} pulses (they are inactive). Then r_{n+k} is renamed as the new r_{n+1}. The new r_{n+1} is larger than the old r_{n+1}, as defined by Eq. 1. So the decrease of f_{n+1} in Eq. 2 is less than the old one (i.e., the convergence is worse.) But if the number of skips is stored as ns, and one –nsTo is added to the output D_{n+1} on the right of Eq. 2, this will accelerate the convergence.

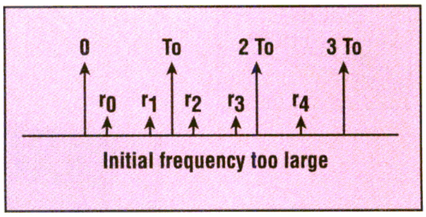
This case can be studied in three ways:

1. Keep Eqs. 1 and 2. Sometimes in the convergence process, the phase difference will be larger than To, and is not realizable.

2. Skip the successive r_{n+1},...pulses, and do not make compensation to D_{n+1}.

3. Skip the successive r_{n+1},...pulses and store the number of skipped pluses and compensate D_n+1.

Example 3: To = $1/10^5$; (that is, 100 Kb/s) with the initial DDS frequency f_0 = 14000 Hz and r_0 = 0.9To. The results are shown in Figs. 6a and b. The y-axis in Fig. 6a is also the phase difference. There are three curves. Keeping Eqs. 1 and 2, the resultant curve is shown with a solid line. If it swings to –9 (To), the counter cannot work. The dotted line represents the second way—skips, but with no compensation. It is a dense sawtooth



5. This pulse sequence is input to the counter when the initial frequency f_0 is larger than 1/To.

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10.9-12.8	11.8-14.0	DC-2.0	+10	5.5	41	42	MC64SMD-3
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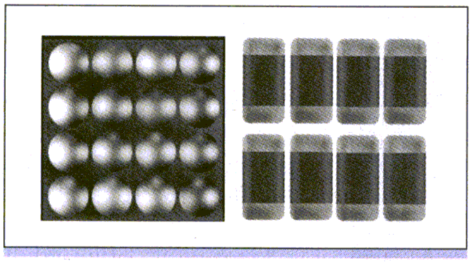

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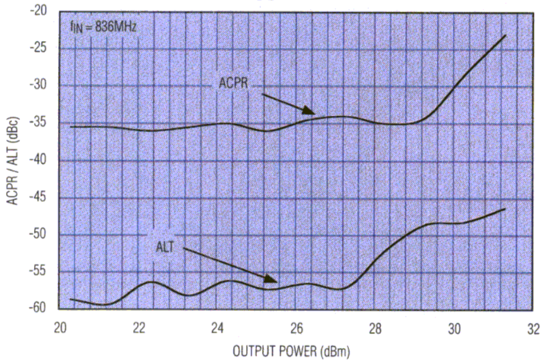

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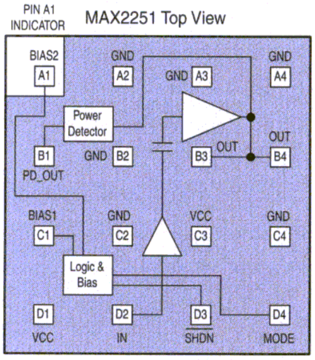

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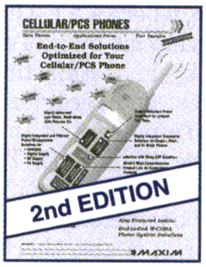

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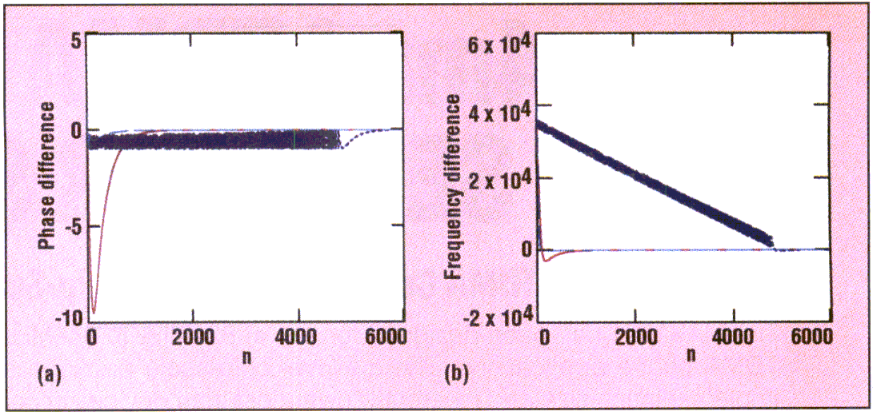


6. The fastest convergence of phase (a) and the most accurate frequency (b) is achieved when successive r_{n+1} pulses are skipped and compensation occurs.

curve. The phase difference is within To, but it converges after 5000 iterations. The least-important curve is the dashed curve that presents the third way—skips and compensation. It converges quickly, even better than the first way.

The convergence characteristics are as follows:

- First way (solid line):

iteration n = 3000, elapsed time = 30 ms, phase difference = –0.00298 percent of To, and the DDS frequency difference from 1/To is –0.0013Hz, or –1.3 PPM.

- Second way (dotted line):

iteration n = 6000, elapsed time = 60 ms, phase difference = 0.71 percent of To, and the DDS frequency difference from 1/To is –3.14 Hz, or –31.4 PPM.

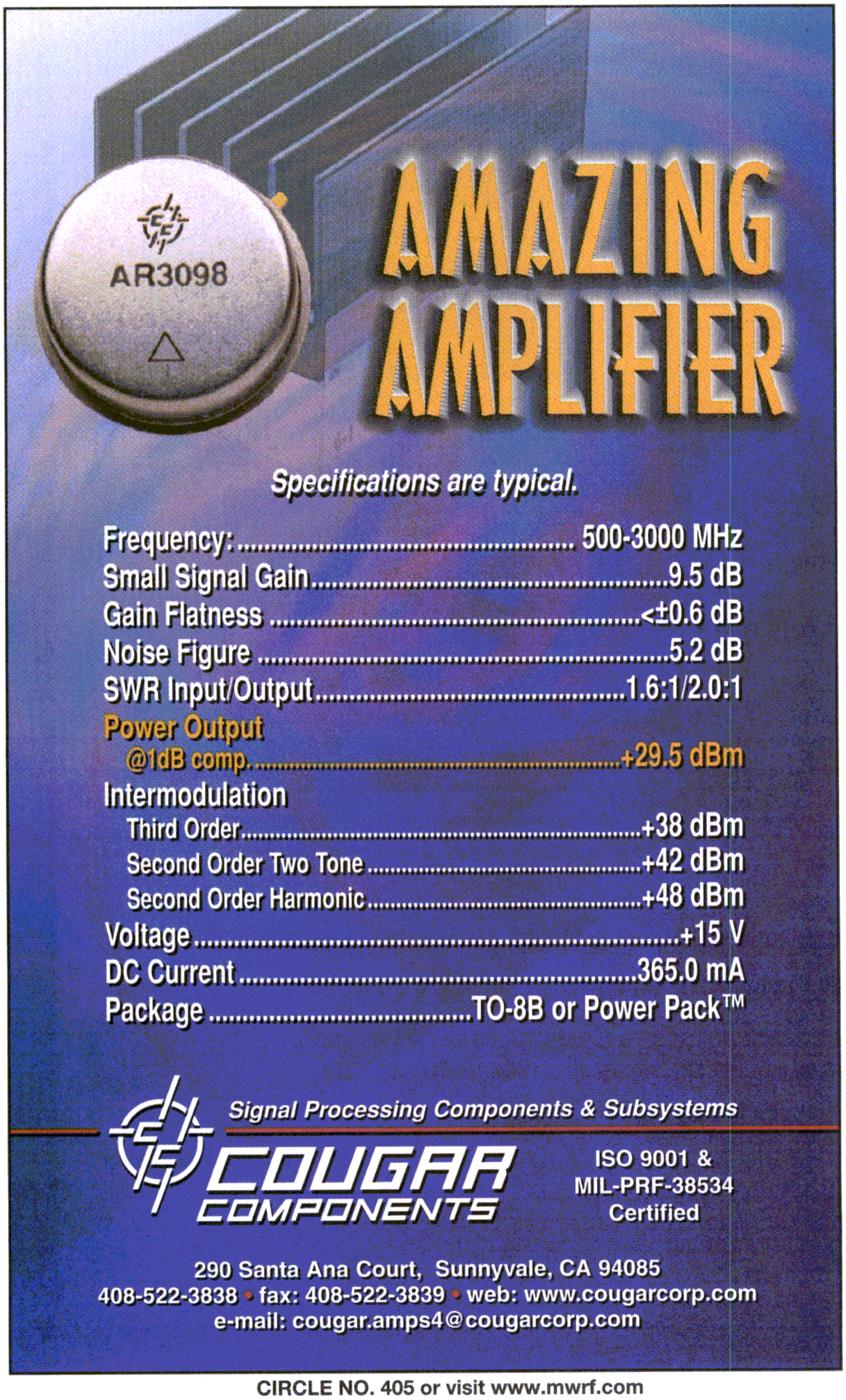
- Third way (dashed line):

iteration n = 3000, elapsed time =30 ms, phase difference = –0.0002 percent of To, and the DDS frequency difference from 1/To is –0.00083 Hz; that is, –0.0083 PPM.

Thus, the third way is best—skips and compensation.

Case 4. If the initial frequency f_0 is smaller than 1/To, the nTo pulses will successively hit the counter before the r_n pulses end the counting. Fig. 7 shows the To and 2To pulses before the r_1 pulse.

In the general situation, r_{n+1} is larger than (n+2)To,...(n+k)To, but less than (n+k+1)To, where k ≥ 1. The counter is successively hit by k (To) pulses, then ends by r_{n+1}. Skip the first k–1 (To) pulses, and recognize only the last (n+k)To pulse. Let the new (n+1)To pulse be the old (n+k)To pulse. That means the t-axis is artificially shortened or extracted by (k-1)To. The value of r_{n+1} is also shortened by (k–1)To. This means the positive adjustment of f_{n+1} in Eq. 2 is lessened, and the convergence is worse. Also, adding (k – 1)To to the output D_{n+1} on the right side of Eq. 2 makes this compensation an accelerated convergence.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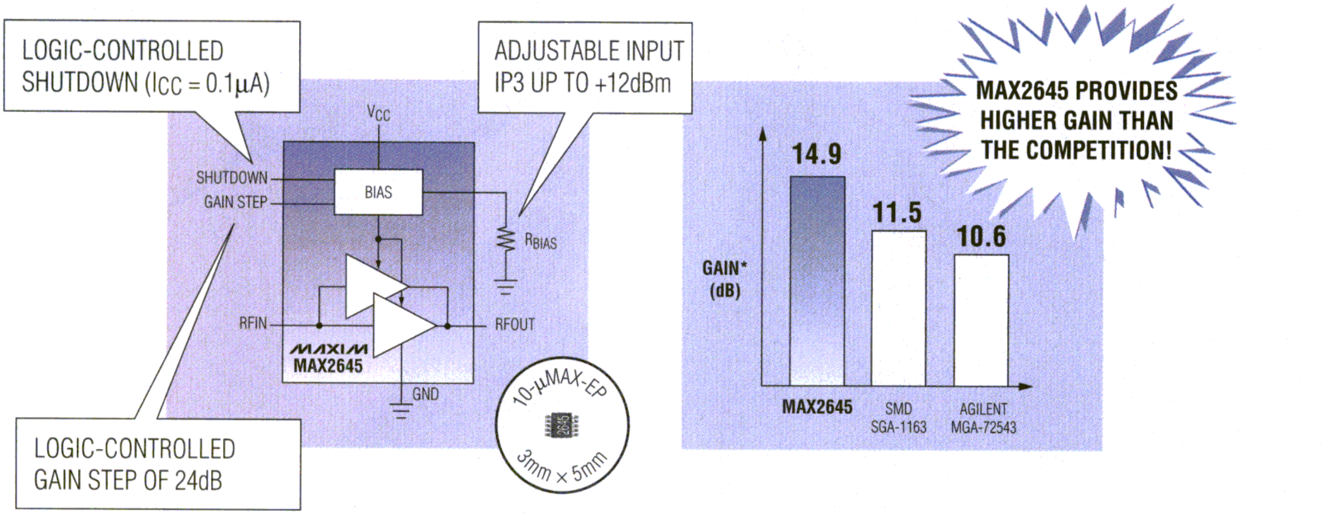

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11.5
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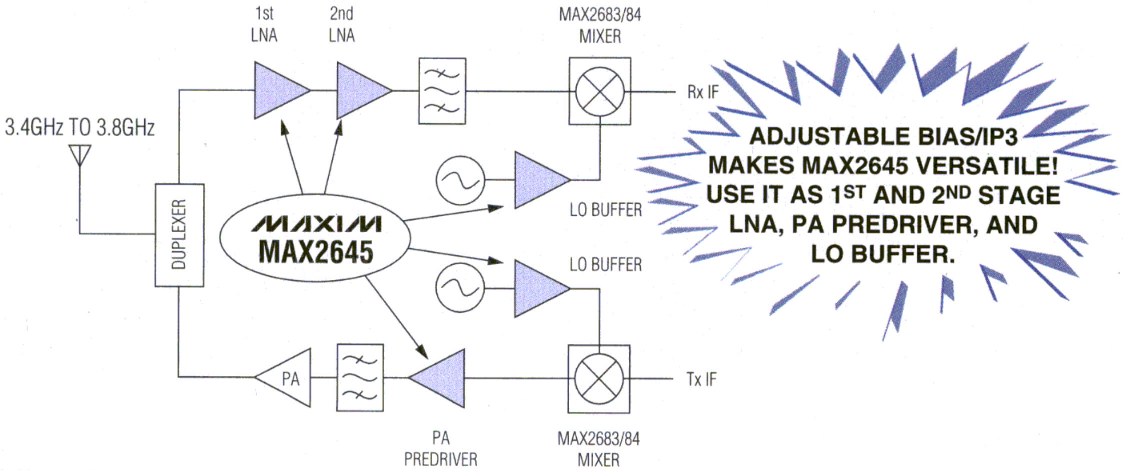
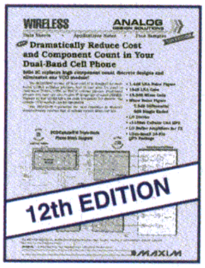


*f = 3.55GHz, I_{CC} ≈ 10mA, V_{CC} ≈ 3V



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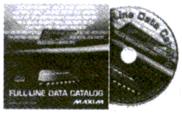

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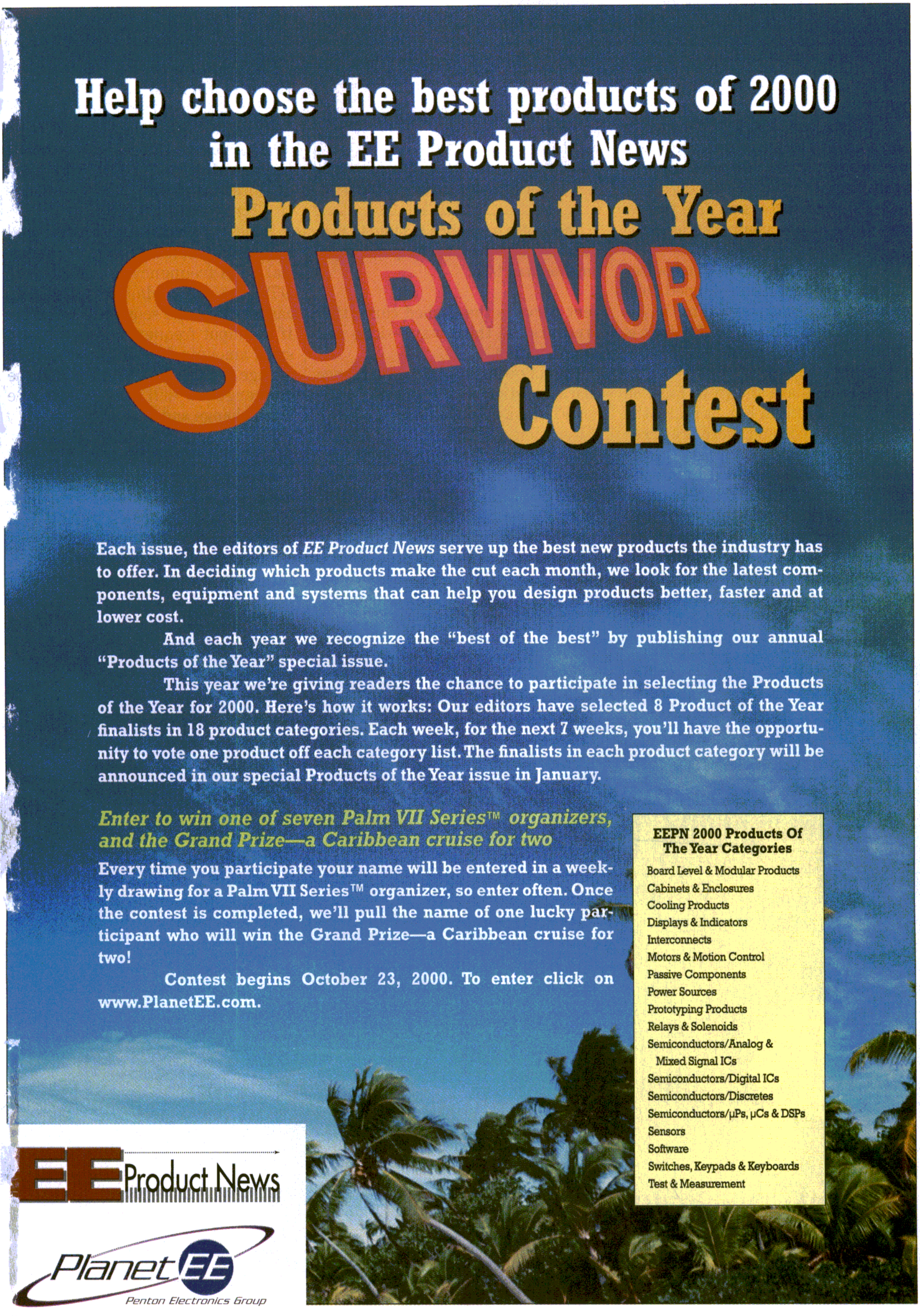

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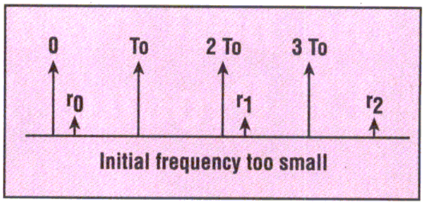
This case can be studied in three ways:

1. Keep Eqs. 1 and 2, because in some cases, the phase difference will be larger than To and is not realizable.
2. Skip the successive (To) pulses, and do not make compensation in D_{n+1}.
3. Skip the successive (To) pulses and save the number of skipped pluses and make compensation to D_{n+1}.

Example 4: To = 1/1000 s (that is 1bt/s), initial DDS frequency f_0 = 600 Hz, and r_0 = 0.9To.

The results are shown in Figs. 8a and b. The y-axis in Fig. 8a is also the phase difference and there are three curves. Keeping Eqs. 1 and 2, the resultant curve is shown by a solid line. If it swings to 3 (To), the counter will not work. The dotted line represents the second way; skipping but no compensation. It is a dense curve at the beginning and the phase difference is within To, but converges after 200 iterations. The quickest convergent curve is the dashed curve that presents the third way—skips and compensation.



7. This pulse sequence is input to the counter when the intial frequency f_0 is smaller than 1/To.

The following are the convergence characteristics:

- First way (solid line):
  iteration n = 250, elapsed time = 0.25 s, phase difference = 0.00015 percent of To, and the DDS frequency difference from 1/To is -1.735×10^{-4} Hz or –0.1735 PPM.
- Second way (dotted line):
  iteration n = 250, elapsed time = 0.25 s, phase difference = 0.0007 percent of To, and the DDS frequency difference from 1/To is -1.533×10^{-3} Hz, or –1.633 PPM.
- Third way (dashed line):
  iteration n = 250, elapsed time =0.25 s, phase difference = –0.00004 percent of To, and the DDS frequency difference from 1/To is -1.325×10^{-5} Hz, or –0.01325 PPM.

Thus, the third way, skipping and compensation, is best. But if the initial DDS frequency is not close to 1/To, the skip and/or compensation cannot make the sequence converge. If the input frequency 1/To drifts in small amounts, it is not necessary to change the value of A and B, and the convergence will catch up to the small change and lock.

THE DIGITAL CASE

In a digital system, time measurement is quantified and the smallest time step is the clock period 1/fc where fc is the clock frequency.

The signed counter (phase detector) measures the number of clock pluses as positive or negative integers. The integer is a truncation of

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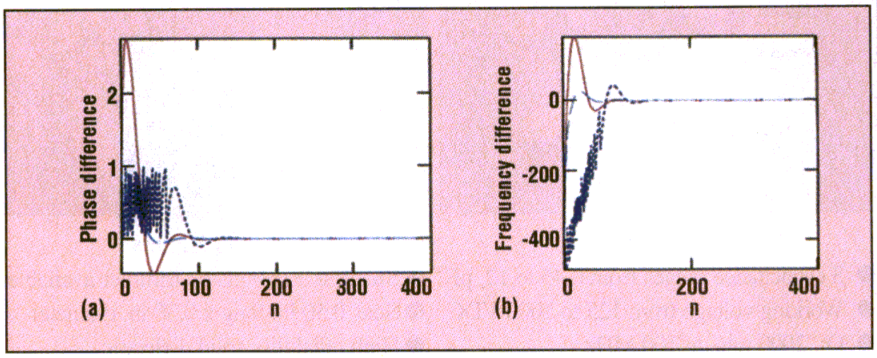


8. These phase (a) and frequency (b) curves present similar results as those of Fig. 6; namely, that phase and frequency converge fastest under a skip-and-compensation routine.

$(r_n$ –nTo)fc. When compared with Eq. 2, there is a factor fc times $(r_n$ – nTo). So, the new coefficients are a = A/fc and b = B/fc. The truncation of –2.8 is –2, an error of 0.8. This is the quantified error. If $(r_n$ – nTo) is smaller than 1/fc, the counter reads 0. Therefore, there is no contribution to the frequency adjustment. The maximum phase error is 1/(Tofc) in units of To. The higher clock frequency has less error. For example, if fc is 60 MHz, the input bit rate is 20 Mb/s. The phase error may reach 1/3 maximum, in units of To.

Now defining d_n = trunc(r_n - n To)fc, the contribution to frequency adjustment is $d_n(a+d_{n-1})b$. Let b = –a + delta, where delta ≥ 0, d_{n+1} a + d_n b = (d $_{n+1}$ –d_n a + delta × d_n. Comparing delta with a, delta is a small number. Nevertheless, even though $d_{n+1} - d_n$ is a small number, delta × d_n still makes little contribution to the frequency adjustment. Suppose the DDS frequency is stable after many iterations, but the phase difference still slowly changes. Finally, the trunc(d_n –nTo)fc emerges as a quantified jump, +1, so the DDS frequency has a jump. Therefore, convergence is not very stable; it always has ripples.

The frequency adjustment is (d_{n+1})a + d_nb. But when the frequency register increases by 1, the DDS output frequency increases by fc/2^{32} where 32 is the length of internal DDS register. The number in the register is an integer and the value put in the register must be rounded. Then Eq. 2 has its digital counterpart.

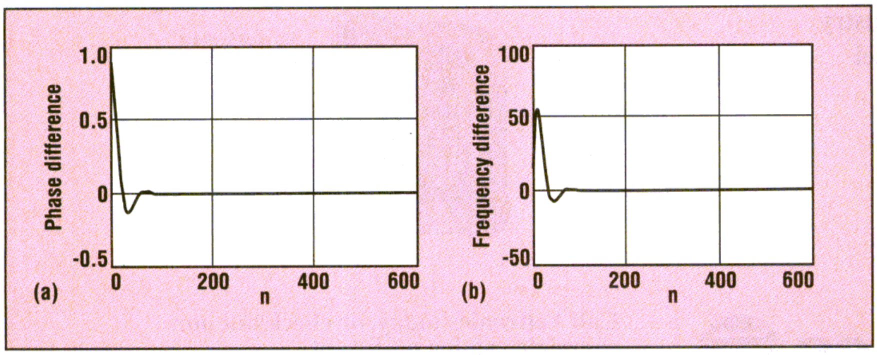
$$f_{n+1} = f_n + (fc / 2^{32})\ round$$
$$[(d_{n+1}\, a + d_n\, b)\, 2^{32} / fc] \qquad (4)$$

Formerly, A and B were only a function of To, but now they are also a function of fc. Change Eq. 4 to:

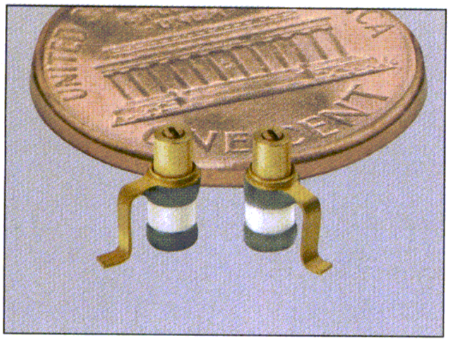
$$f_{n+1} = f_n + (fc / 2^{32})\ round$$
$$[(d_{n+1} - d_n)\, a' + d_n\, (a' + b')] \qquad (5)$$

where:

$$a' = 2^{32} / [To^{1.7}\, fc^2] \qquad (6)$$



9. The phase (a) and frequency (b) convergence characteristics of a DDS with a small number of iterations and f_0 of 1 kHz illustrate how fast both reach their end values.



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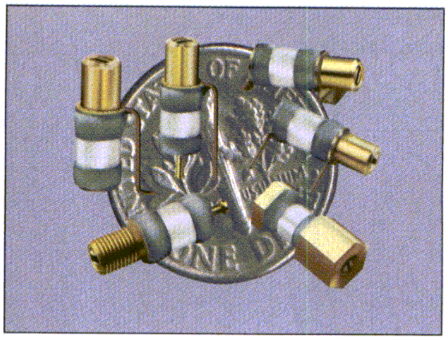

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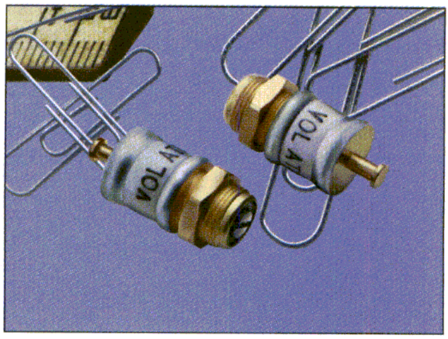

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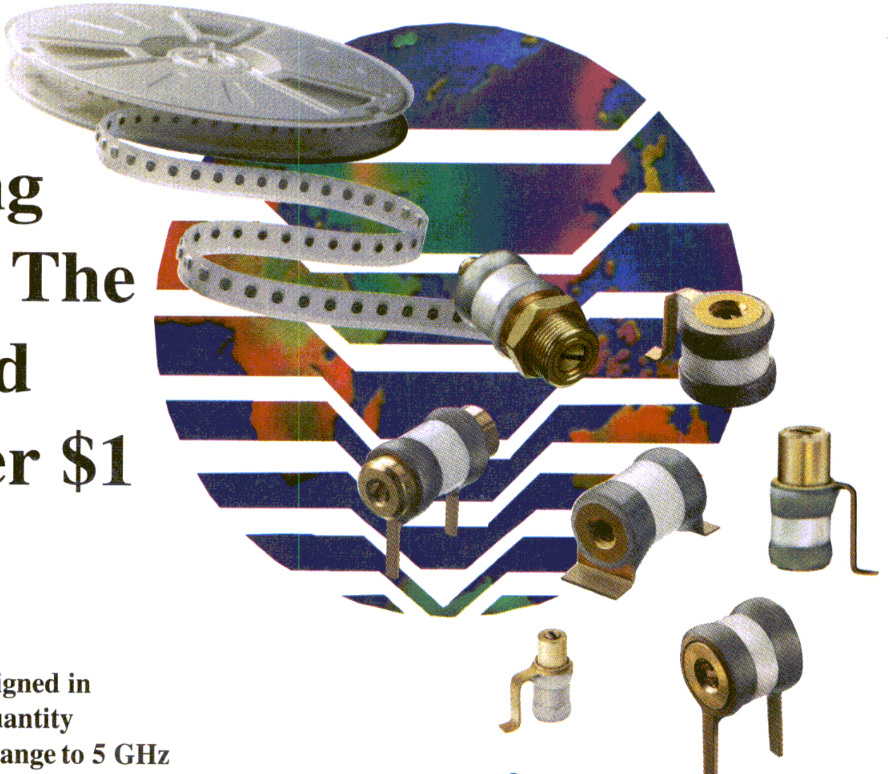

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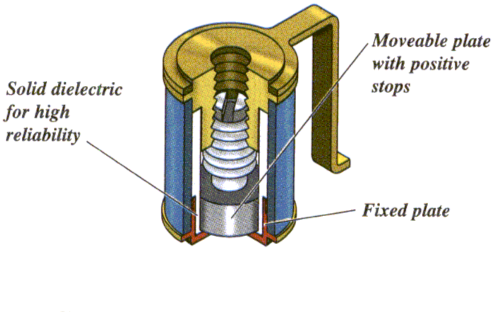
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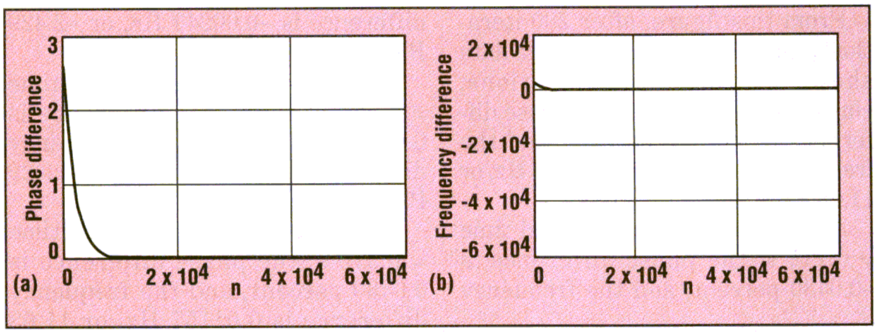


10. The phase (a) and frequency (b) curves of a DDS with a large number of iterations and f_0 of 1.5 MHz are similar to those of Fig. 9.

$$a'+b' = 12\,2^{32}/(To\,fc^{2}) \qquad (7)$$

where:

d_n = the counter output integer.

The calculation of Eqs. 7 and 8 is difficult. It is easy to store a lookup table of a' and a'+b' in memory. There are several ways to create the table. If the nominal bit rate is not known (1/To), it is possible to count the number of clocks in a period To. The number is N0 = floor(Tofc). Suppose the system clock frequency fc is 60 MHz at a bit rate between 15 and 20 Mb/s, N=3. The range-tolerance error is 1.33:1. Counting the number of clocks in the interval of 10To, and dividing the number by 10, N1 = 0.1 × floor(10T0fc). N1 has one more decimal of precision than N0.

Now suppose the lookup table is in the decimal system. The input variable is N = floor(Tofc). If 1/To is larger than 6 Mb/s, N ≥ 10, it is better to count the clock number in 10To and obtain one more digit of precision. The enlisted numbers of N are 1×10^n, 1.5×10^n, 2×10^n, 3×10^n,... 9×10^n,

where:

n = 0, 1, 2, 3, 4. The bit-rate of the input periodic signal is from 1 kb/s to 20 Mb/s. One can calculate one table for a' and a'+b', by using Eqs. 8 and 9.

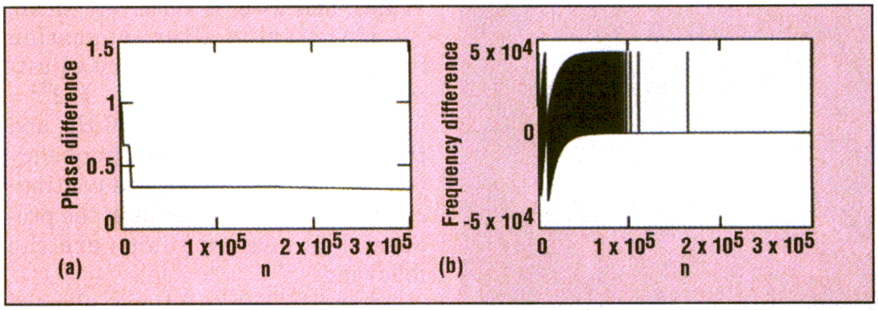
$$a' = 2^{32}/[N^{1.7}\,(fc^{0.3})] \qquad (8)$$

$$a'+b' = 12\,(2^{32})/[N\,fc) \qquad (9)$$

The table has 44 rows, with one N column as an input, a' and a' + b'columns as an output. When one N is not listed, go to the nearest number. The maximum deviation is smaller than 25 percent. For example, if 1/To is 9.6 kb/s, N= 6.25 × 10^3, take the input as 6 × 10^3 in the table.

The following examples use this lookup table, but different lookup tables are possible. Eqs. 7 and 8 could be used to build a table with input variable 1/To. It is also possible to create a binary-base input variable table.

Based on the previous "analog" simulation, if the initial frequency deviates far from the correct value, the counter will produce successive To pulses or r_n pulses. In this case, the counter skips successive pulses and compensates at the d_{n+1} in Eq. 6, and convergence will be kept and occur even faster.



11. In these curves using lookup tables, only a single iteration can cause a huge change in the frequency convergence of the DDS. The lookup table values are responsible for this change.

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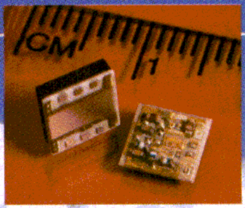

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The following examples use Eqs. 1 and 6 without skips and compensation, and attention is paid to the convergence characteristics.

Example 5: 1/To = 990 b/s, fc = 60 MHz, r_0 = 0.9To, f_0 = 1000Hz. The look-up table considers N as 6 × 10^4, shifted 1/To to 1 kb/s. From the table, a' = 0.15 and a'+ b' = 0.0143. The simulation result is shown in Fig. 9.

From this figure, after 200 iterations, there is convergence. The data shows that after 214 iterations, elapsed time = 0.216 s, the phase difference is –0.0418 percent, and the frequency difference is 0.01157 Hz, or 11.68 PPM.

After 215 iterations, elapsed time = 0.217 s, the phase difference is –0.0430 percent, and the frequency difference is –0.00240 Hz, or –2.429 PPM.

After 568 iterations, elapsed time = 0.574 s, the phase difference is 0.0427 percent, and the frequency difference is –0.00240 Hz, or –2.429 PPM.

After 569 iterations, elapsed time = 0.575 s, the phase difference is 0.0430 percent, and the frequency difference is 0.01157 Hz, or 11.68 PPM.

There is a negative frequency jump after 214 iterations, and there is a positive frequency jump after 568 iterations. The jump value is 2^{32}/fc = 0.01397 Hz. That is the quantified error. Since the phase difference is not zero, d_n always changes slightly when the accumulated change contributes a 1-b variation to the frequency register, and a jump occurs. So, in the convergent state it is not absolutely stable because there is a small jitter. This is the normal behavior of a digital system.

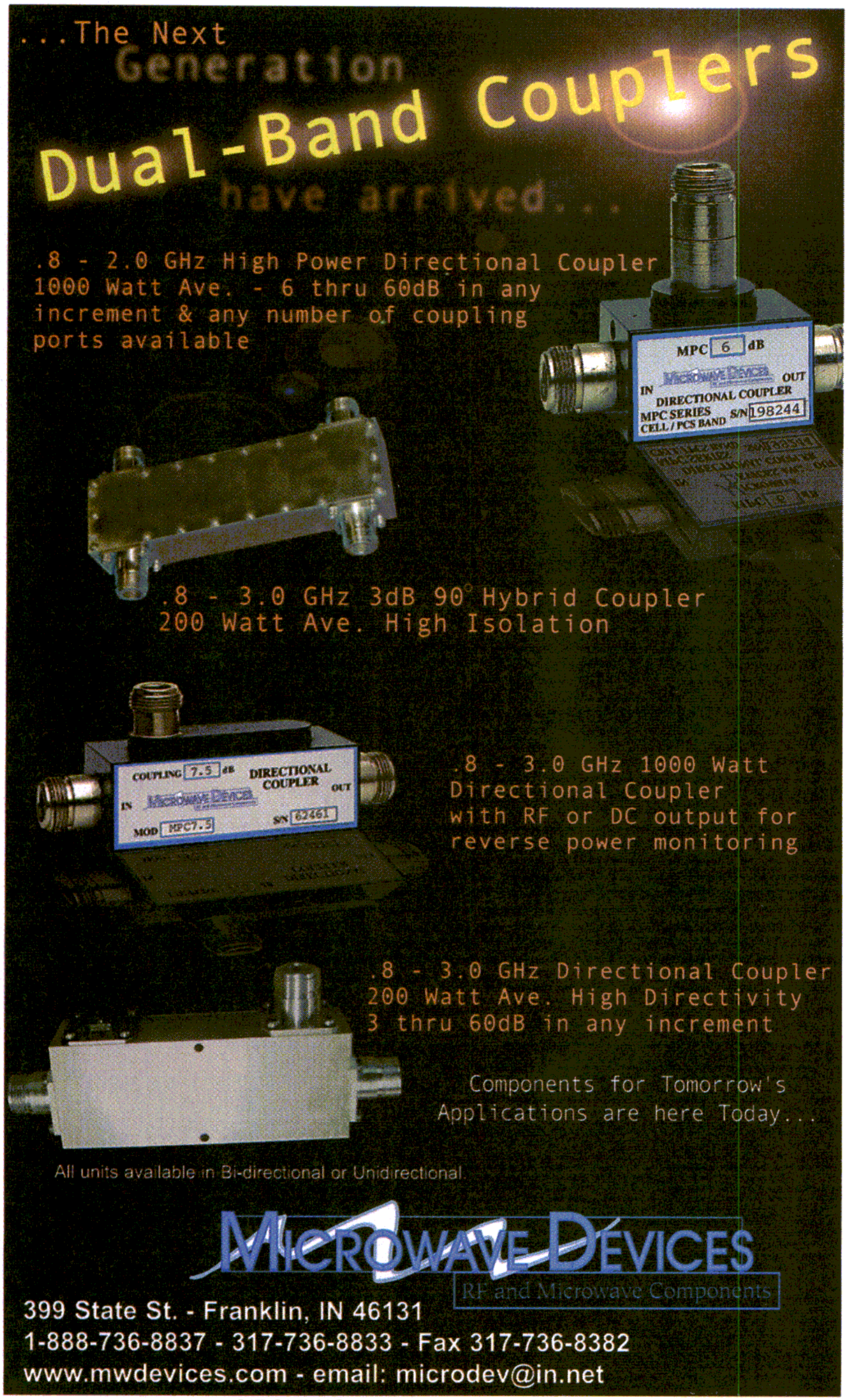
Example 6: 1/To = 1.54 Mb/s, fc = 60 MHz, r_0 = 0.9To, and f_0 = 1.5 MHz. From the lookup table, N is 40. From the table, al = 3.767 × 10^4 and a'+ b' = 21.47. The simulation result is shown in Fig. 10.

After 30,000 iterations, there is convergence. Actually, convergence occurs in 25,271 iterations. After 25,271 iterations, elapsed time = 0.0164 s, the phase difference is 2.5667 percent, and the frequency difference is 526 Hz, or 342 PPM.

After 25,272 iterations, elapsed time = 0.0164 s, the phase difference is 2.5325 percent, the frequency difference is 0.16 Hz, or 0.1 PPM.

One iteration can change the value of DDS frequency several hundred hertz, because a' and b' are very big values, and a'+b' is relatively small. So, if the value after truncation increases by 1, the frequency adjustment is approximately a' × fc/2^{32} = 526Hz, or b' × fc/2^{32} = –526 Hz, and this quickly pulls back the frequency to convergence. There are continuous DDS frequency jumps in the process. The nearest jumps are the following:

After 515,369 iterations, elapsed time = 0.3347 s, the phase difference is –2.5667 percent, and the frequency difference is 0.16 Hz, or 0.1 PPM.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Between 25,272 and 515,369 iterations, the frequency difference does not change.

After 515,370 iterations, elapsed time = 0.3347 s, the phase difference is –2.5667percent, and the frequency difference is –526 Hz, or –342 PPM.

After 515,371 iterations, elapsed time = 0.3347 s, the phase difference is –2.5325 percent, and the frequency difference is –0.133 Hz, or –0.086 PPM.

These three successive iterations change the DDS frequency approximately ±626 Hz, but the phase difference changes little. At high bit-rate, a' and b' are large, and the phase difference is within ±1/(Tofc) of the unit of To. Now Tofc =1.54/60 = 2.5667 percent, which coincides with the previously listed data. But this time jitter is unavoidable.

Example 7: bit-rate is 20 Mb/s, (To = $1/[20 \times 10^6]$), f_0 =19.99 MHz, r_0 = 0.9To. In the lookup table, N is 3, and al = 3.079×10^6, a'+ b' = 286.3. The simulation result is shown in Fig. 11.

The data are as follows:

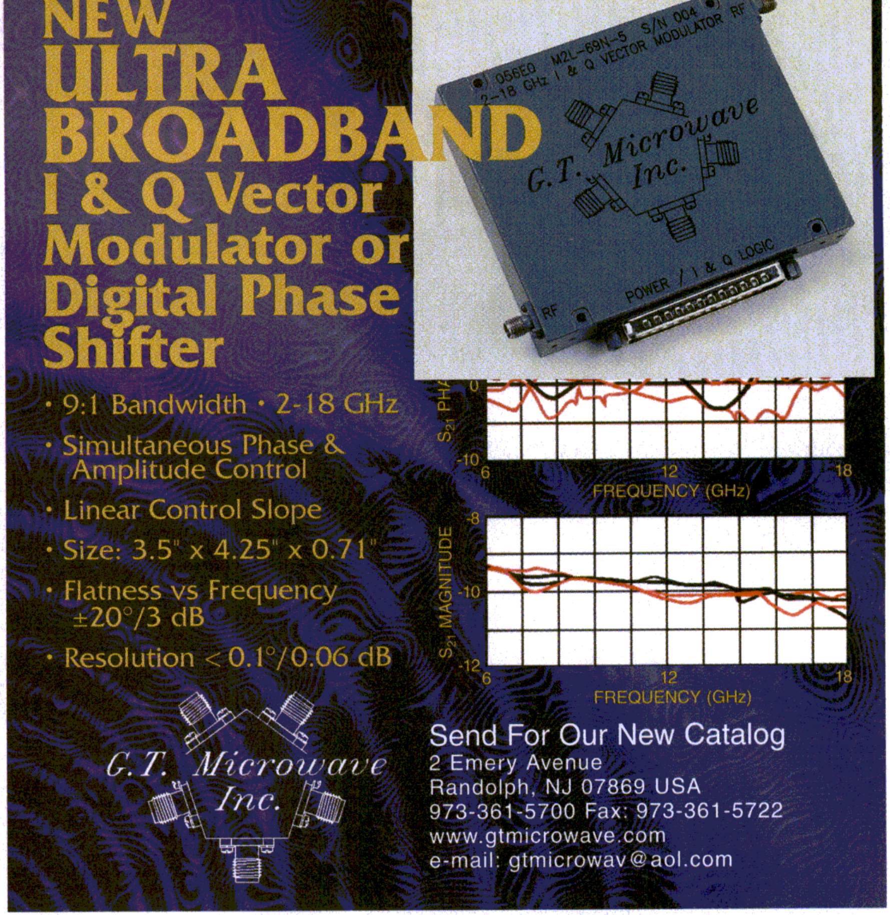
After 164,892 iterations, elapsed time = 0.0082 s, the phase difference is 33.33 percent of To, and the frequency difference is 4.3×10^3 Hz, or 2151 PPM.

After 164,893 iterations, elapsed time = 0.0082 s, the phase difference is 33.12 percent, and the frequency difference is 3.22 Hz, or 0.16 PPM.

In this case, the frequency adjustment is approximately $al \times fc/2^{32} = 4.3 \times 10^4$ Hz, or $b' \times fc/2^{32} = -4.3 \times 10^4$ Hz, and this high value quickly induces convergence. There are jumps of DDS frequency continuously. Due to the limitations of the computer, the next jump point cannot be found. From the 164,893th iteration to 10^6 iterations, the DDS frequency does not change, but the phase difference is changing.

After 1 million iterations, elapsed time = 0.05s, the phase difference is 19.7 percent, and the frequency difference is still 3.22 Hz, or 0.16 PPM.

Tofc = 20/60 = 33.33 percent. After millions of iterations, the phase difference will reach –33.33 percent of To. It will swing continuously between ±33.33 percent. When it reaches the limits, the DDS frequency will experience a jump of ±43 kHz. ••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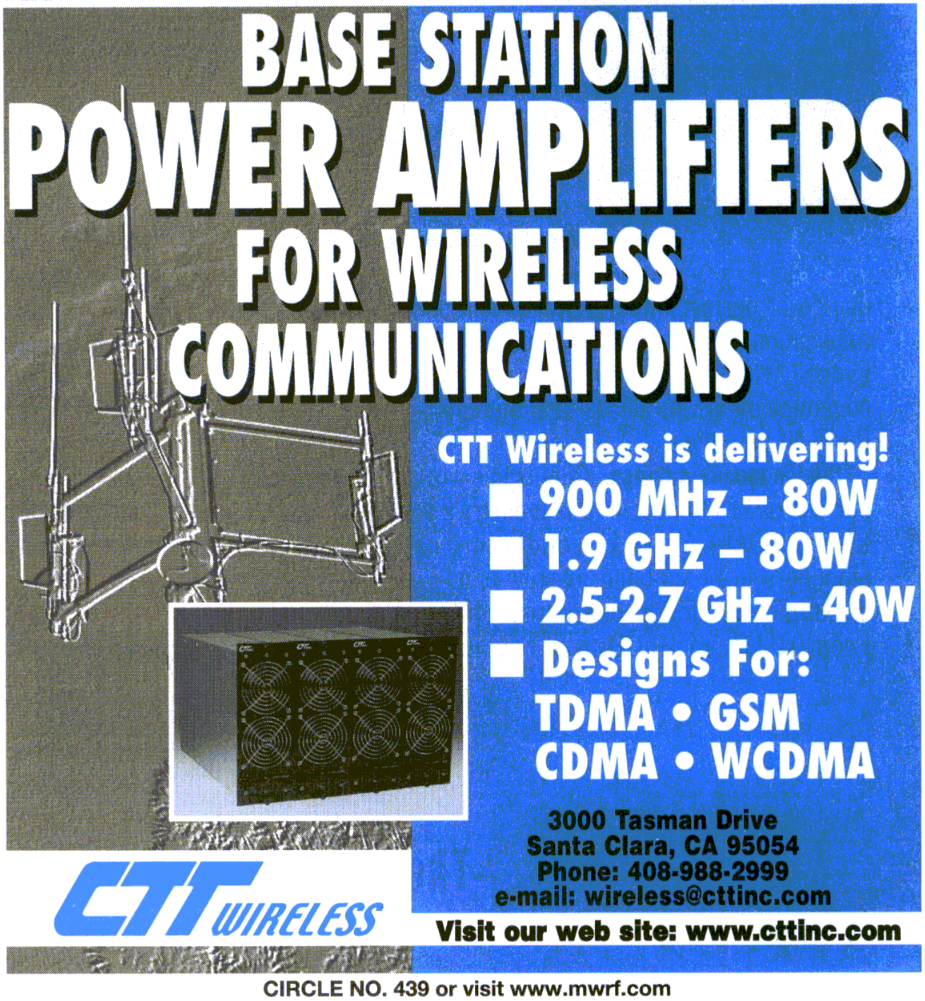
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CIRCLE NO. 303

High-Linearity RF Amplifier Design

Peter B. Kenington

High Linearity RF Amplifier Design offers a comprehensive treatment of RF amplifiers, focusing on the issues and techniques associated with maximizing linearity in their design.

Chapter 1 begins with an overview of distortion and linearity requirements. It discusses power efficiency, the effect of nonlinearity on a wideband-code-division-multiple-access (WCDMA) system, and the requirement for linearity in adaptive-antenna systems.

Chapter 2 focuses on the theory behind distortion in amplifiers. It discusses amplitude distortion, two-tone and multi-tone tests, harmonic and intermodulation distortion (IMD), nonlinearity measures for multi-tone and modulated signals, crest-factor reduction, phase distortion, white-noise testing, spurious signals, and cross modulation.

Chapter 3 discusses RF power-amplifier (PA) design. It covers the various classes of amplifiers and their variations, and discusses various biasing methods. Chapter 4 introduces feedback and discusses various feedback methods to achieve linearization. Included in the discussion are techniques such as series and shunt feedback, modulation feedback, and polar- and cartesian-loop transmitters (Txs).

Chapter 5 introduces feedforward systems and discusses their application to linearization. The chapter covers topics such as multiple feedforward loops, gain and phase matching, error-amplifier design, power efficiency, power loss and efficiency, linear distortion correction, and temperature drift and aging. It then identifies the requirements of the major components of feedforward systems.

Chapter 6 covers predistortion techniques. It introduces RF and intermediate-frequency (IF) predistortion theory and discusses issues such as gain- and phase-matching characteristics, and implementations, such as single-diode, field-effect-transistor (FET)-based, anti-parallel diode-based, and digital predistorters. It also covers baseband and adaptive-baseband predistortion, as well as postdistortion linearization.

Chapter 7 considers linear Txs that employ signal processing which radically alters the original signal. Major topics include envelope elimination and restoration (EE&R), linear amplification using nonlinear components, the vector-locked loop, and linear amplification employing sampling techniques (LIST).

Chapter 8 discusses methods of boosting the efficiency of Class A, AB, and B amplifiers. It covers the Doherty technique, adaptive biasing, envelope tracking, Class H amplification, and dual-bias control. (2000, 531 pp., hardcover, $109.00) **Artech House, 685 Canton St., Norwood, MA 02062; (781) 769-9750, FAX: (781) 769-6334, Internet: http://www.artechhouse.com.**

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Neural Networks For RF And Microwave Design

Q.J. Zhang and K.C. Gupta

One of the latest, yet least understood, high-frequency design techniques is revealed in *Neural Networks For RF And Microwave Design*, a text intended for RF and microwave designers, researchers, and graduate students. Artificial neural networks (ANNs) have been used in a number of engineering disciplines, but only recently in RF/microwave design. An ANN is an information-processing system akin to the human brain that learns from observation and generalizes by abstraction. It is an unconventional alternative to modeling difficult RF/microwave design problems with computer-aided-design (CAD) software. The authors claim that while research papers on ANNs have begun to appear in technical literature, theirs is the first book to explain the technology from the RF/microwave engineer's perspective. Accordingly, the book begins with an explanation of why neural networks are important for high- frequency design. Chapter 2 is an overview of modeling and optimization with an emphasis on the two major aspects of any design—design-in-the large and design-in-the small.

Chapter 3 describes the various neural network structures (architectures) that are candidates for RF/microwave designs. The most common type is a feedforward configuration known as multilayer perceptrons (MLP). The key to understanding and employing ANNs begins in chapter 4 where the training of neural networks is described. Training involves operations such as data generation, range, distribution of samples in the input-parameter space, data scaling, and others. The actual training process is one of optimization, and various optimization algorithms are described.

Chapter 5 includes a general discussion of modeling procedures. It offers several examples of the techniques described in earlier chapters for the development of CAD models of RF/microwave components. The examples are divided into three groups. Group one is microstrip transmission lines including vias and vertical interconnects in multilayer circuits. Group two deals with components used in coplanar-waveguide (CPW) circuit design, and group 3 discusses passive components, such as spiral inductors, microstrip patch antennas, and waveguide filter components.

An important application of ANN—high-speed integrated-circuit (IC) interconnects used in digital circuits—is the subject of chapter 6. This is a detailed description of modeling and signal-integrity analysis for a variety of interconnect and transmission-line configurations. Chapter 7 presents the use of ANNs for modeling active components, another critical application of the technology. Direct modeling of a device's external behavior, such as DC, small-signal, and nonlinear models; large-signal simulations; and indirect modeling through known equivalent circuit models of a device are covered.

Chapter 8 describes the use of neural network models for analysis and optimization of RF/microwave design. Chapter 9 addresses the topic of embedding knowledge in ANN models and the use of prior knowledge in reducing the training time of ANN structures. Two methods used in the training process are described. Entering the world of ANN can be somewhat intimidating because the introduction to topics such as modeling, optimization, and neural network structures is highly mathematical. However, the authors are mindful of their original claim for the book (i.e., making it a useful tool for practicing engineers and researchers). Thus, they try to illustrate many of the principles with RF/microwave components that are familiar to working designers. Included with the book is a compact-disc read-only memory (CD-ROM) containing the introductory version of NeuroModeler, a program to aid designers in developing neural network models of active and passive high-frequency components. It operates on Windows NT4.0 and Windows 95/98. (2000, 369 pp., hardcover, $93.00, ISBN: 1-58053-100-8). **Artech House Publishers, 685 Canton St., Norwood, MA 02062; (800) 225-9977, FAX: (781) 769-6334, e-mail: artech@artech-house.com, Internet: http://www.artechhouse.com.**

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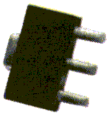

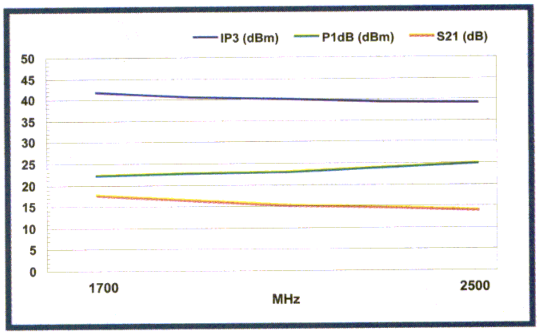

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CIRCLE NO. 228

Devise A GMSK Modulator Using The Power Of DSP

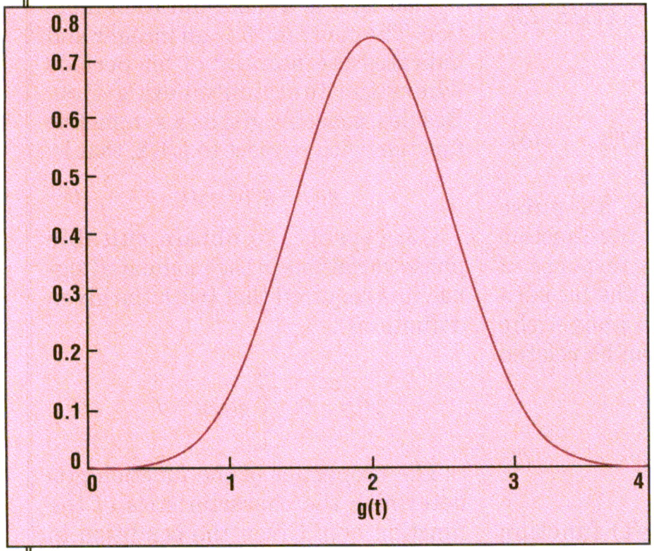
A low-power, low-parts-count modulator for GSM communications can be designed using digital-signal-processing (DSP) principles.

Dr. César Benavente
Dpto. Ingeniería de Circuitos y Sistemas, EUIT Telecomunicación, Universidad Politécnica de Madrid, Ctra. Valencia km. 7, 28031 Madrid, Spain; +34 913367830, FAX: +34 913367832, e-mail: cbpeces@euitt.upm.es.

DIGITAL modulator design is evolving toward systems that have lower power consumption and simpler architectures. But digital modulation usually includes complex operations that must be performed by systems with high-order processing capabilities. Advances in digital signal processing (DSP) have produced integrated-circuit (IC) devices with the necessary computation properties, high speed, low power, and interfacing capability. These devices can perform numerous tasks while reducing system-component count and power dissipation. As a result, DSP ICs occupy an important position in telecommunications system design. Each year, DSP developers announce new devices with improved capabilities—memory, speed, ports, tools, etc.—and lower power consumption.

Another important characteristic desirable in modern communications systems is flexible equipment. Flexibility can be defined as the ability to handle several communications standards, protocols, access techniques, or other communication characteristics. Using programmable DSP ICs is one way to obtain flexibility. A programmable device can be reprogrammed for the performance characteristics necessary to meet a specific communications requirement. For example, the modulation type can be changed by changing the values of data stored in the DSP's memory or by reprogramming the DSP itself.

One of the popular digital-modulation schemes is Gaussian minimum-shift keying (GMSK), the foundations of which will be developed in this article. This modulation is a particular case of continuous-phase modulation (CPM). These types of modulation are bandwidth efficient and are employed in several applications, such as Global System for Mobile Communications (GSM) and Digital European Cordless Telephone (DECT) standards. Some IC developers have built specific DSP-based codecs to perform the GSM standard. The price of these devices is competitive due to their extensive use in the high-volume communications market.

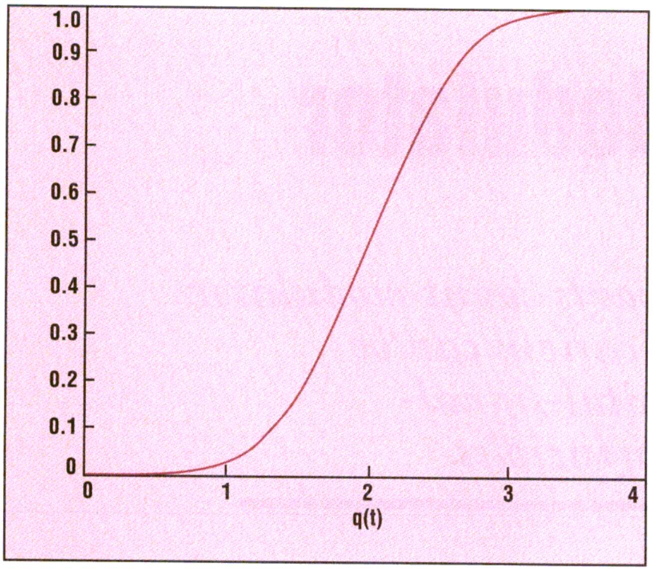


1. The sharpener pulse [g(t)] required in GMSK modulation accounts for the phase variations in the modulated signal.

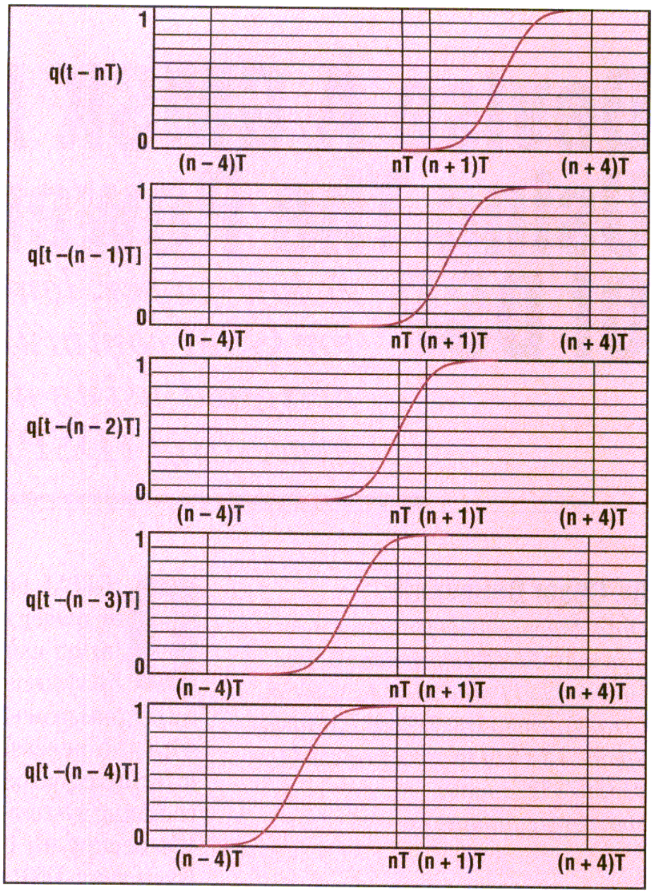
CPM THEORY

CPM is a type of digital modulation where the envelope of the signal remains constant. This property provides the modulation-bandwidth efficiency. That is, for fixed bit rates, the bandwidth of CPM is shorter than other modulation types such as quadrature phase-shift keying (QPSK). This is because there are no phase discontinuities in CPM.

A CPM signal can be expressed by



2. The sharpener pulse (Fig. 1) determines the characteristics of the q(t) function shown here, which is responsible for the phase change between a transmitted bit and the one following it.



3. Phase changes [q(t – kT)] during the transmission period of a bit are shown over the interval nT ≤ t ≤ (n + 1)T, L = 4.

its general equation:

$$x(t) = A\cos(\omega_o t + \varphi(t)) \qquad (1)$$

where:

A = the maximum amplitude the signal can take,

ω_0 = the carrier frequency in radians, and

$\phi(t)$ = the signal phase that depends on the time. $\phi(t)$ also depends on the data to be transmitted, and is given by:

$$\varphi(t) = \pi h \sum_{k=-\infty}^{n} a_k q(t - kT)$$

$$nT \le t \le (n+1)T \qquad (2)$$

where:

q(t) = the function that sharpens the phase-term $\phi(t)$ variation with time. Equation 1 could include an initial phase-term θ so that the modulated signal may be expressed as:

$$x(t) = A\cos(\omega_o t + \varphi(t) + \theta) \qquad (3)$$

From a signal-analysis point of view, it is advantageous to manage the analytic signal corresponding to Eq. 1 so that the modulated signal can be expressed by:

$$s(t) = A\exp[\omega_o t + \varphi(t) + \theta] \qquad (4)$$

In this expression, it is simple to separate the terms corresponding to the carrier frequency, the initial phase, and the phase term varying with time (the modulation term). The constant h in Eq. 2 is the modulation index that determines the phase change between one transmitted bit and the next transmitted bit.

The function $\phi(t)$ may be obtained as follows:

$$\varphi(t) = \pi h \int_{-\infty}^{t} \sum_{k=-\infty}^{n} a_k g(\tau - kT)d\tau \qquad (5)$$

where:

$$\phi(t) = \sum_{k=-\infty}^{n} a_k g(t - kT)) \qquad (6)$$

corresponds to the bit-pulse sequence sharpened by the function g(t), that is the impulse response of the sharpener filter. T is the bit period. Following the step above, the functions g(t) and q(t) can be related by the expression:

$$q(t) = \int_{-\infty}^{t} g(\tau)d\tau \qquad (7)$$

Depending on the g(t)-function characteristics, CPM has different properties. Consider the general case and that the function g(t) has the following property:

$$g(t) = \begin{cases} \neq 0 & 0 \le t \le LT \\ 0 & resto \end{cases} \qquad (8)$$

That is, it is a finite-duration function. The constant L is an integer and indicates the number of bit periods where the function meets g(t)≠0. Besides, consider g(t) as a symmetric function with respect to LT/2, that is:

$$g(t) = g(LT - t) \qquad (9)$$

Conversely, to obtain suitable phase variations in the modulated signal, it is required that (with the given definitions):

$$g(0) = 0 \qquad \int_{-\infty}^{\infty} g(t)dt = 1 \qquad (10)$$

The properties of the function g(t) determine the characteristics of the function q(t). Given the condition in Eq. 10:

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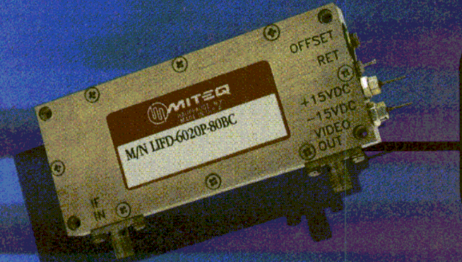

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LIFD-7030P-80BC	70	-80 to 0	±0.5	30	25
LIFD-16040-80BC	160	-80 to 0	±1.0	30	25
LIFD-300100-70BC	300	-70 to 0	±1.0	20	15

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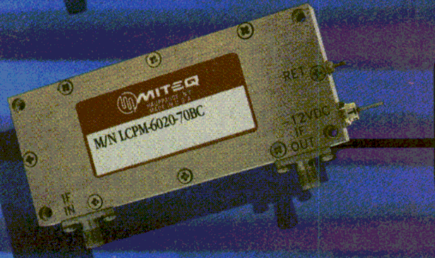

MODEL NUMBER	CENTER FREQUENCY (MHz)	DYNAMIC RANGE (dB, Min.)	OUTPUT POWER (dBm, Min.)	POWER VARIATION (dB, Max.)	PHASE VARIATION (Max.)
LCPM-3010-70BC	30	-70 to 0	10	±0.5	±3°
LCPM-6020-70BC	60	-70 to 0	10	±0.5	±3°
LCPM-7030-70AC	70	-65 to 5	10	±0.5	±5°
LCPM-16040-70BC	160	-65 to 5	10	±1.0	±3°

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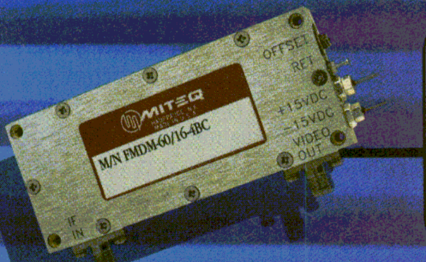

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FMDM-60/16-4BC	60	16	250	±3	90
FMDM-70/36-10AC	70	36	50	±2	50
FMDM-160/35-15BC	160	35	100	±2	30
FMDM-160/50-15AC	160	50	40	±2	25
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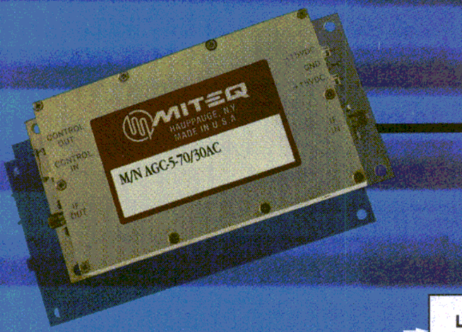
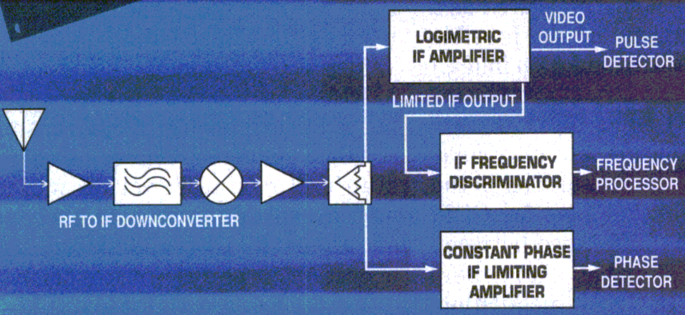


MODEL NUMBER	CENTER FREQUENCY (MHz)	BANDWIDTH (-3 dB) (MHz, Min.)	DYNAMIC RANGE (dBm, Min.)	OUTPUT POWER (dBm, Min.)	POWER VARIATION (dB, Max.)
AGC-7-10.7/4AC	10.7	4	-70 to 0	10	±0.5
AGC-7-21.4/10AC	21.4	10	-70 to 0	10	±0.5
AGC-5-70/30AC	70	30	-50 to 0	-4	±0.5
AGC-7-160/30AC	160	30	-70 to 0	8	±1.5
AGC-7-300/400AC	300	400	-65 to 0	3	±1.0



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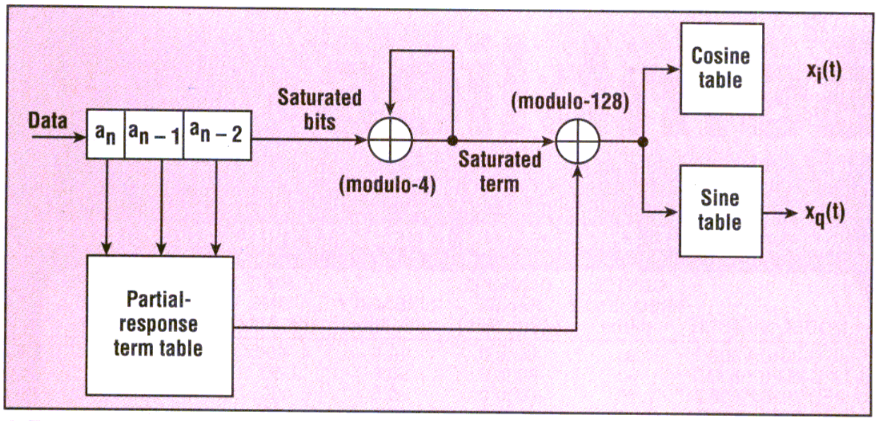

4. This DSP implementation of a look-up table for sine and cosine values is the preferred way to generate phase-change information.

$$q(0) = 0$$
$$q(LT) = 1$$
$$(q(t) = 1t \geq LT) \qquad (11)$$

Besides:

$$1 - q(t) = q(LT) - q(t) = q(LT - t) \qquad (12$$

The function q(t) is that represented in Fig. 2.

Thus, solving Eqs. 4 and 6, the value that the phase term $\phi(t)$ takes depends on all the bits that have been transmitted where the function q(t) does not drop to zero. Then, the value of ϕt) over a certain time interval nT≤t ≤t≤(n+1)T cannot be known only from the bit transmitted during that interval, but from what was transmitted before, and that information provides the signal phase before transmitting the actual bit.

Consider that L = 4 and it is desired to know how the phase $\phi(t)$ varies over the time interval nT≤t≤(n+1)T, i.e., during the transmission interval of the bit a_n. In this case, the term $\phi(t)$ is composed of [and, more concretely, the functions q(t-kT)] which are represented in Fig. 3).

In Fig. 3, it can be concluded that the only functions that vary in the interval nT≤t≤(n+1)T are:

$$q(t - nT)$$
$$q(t - (n-1)T)$$
$$q(t - (n-2)T)$$
$$q(t - (n-3)T) \qquad (13)$$

[that in a general way corresponds to the function q(t–(n–L+1)T)]. The rest of the functions of the form q(t–(n–k)T) con k≥4 (k≥L in the generalization) take the unity value from the instant t = nT, providing a constant contribution from that instant.

In this way, the function $\phi(t)$ may be expressed as:

$$\varphi(t) = \pi h \sum_{k=n-L+1}^{n} a_k q(t - kT) + \varphi_n$$
$$nT \leq t \leq (n+1)T \qquad (14)$$

where:

$$\varphi_n = \pi h \sum_{k=-\infty}^{n-L} a_k \qquad (15)$$

Finally, the modulated signal is given by the equation:

$$s(t) = exp(j(\omega_o t + \theta + \varphi(t)) = exp \left\{ j\left(\omega_o t + \theta + \pi h \sum_{k=-\infty}^{n} a_k q(t - kT) \right) \right\}$$
$$= exp \left\{ j \left(\begin{array}{l} \omega_o t + \theta + \pi h \displaystyle\sum_{k=n-L+1}^{n} a_k q(t - kT) + \\ \pi h \displaystyle\sum_{k=-\infty}^{n-L} a_k \end{array} \right) \right\}$$
$$nT \leq t \leq (n+1)T \qquad (16)$$

[It must be taken into account that the phase term $\phi(t)$ is measured as a modulo-2π function and the possible restrictions of the modulation index.]

GMSK AND DSP

GMSK modulation is a particular case of CPM. The formulation is the same as the general case for CPM modulation. The difference between particular cases is the sharpener-pulse function. In GMSK modulation, the sharpener pulse is a Gaussian function as represented in Fig. 1. Then, the integral of the Gaussian pulse results in the function represented in Fig. 2, showing how the phase varies with time.

A parameter that determines the GMSK characteristics is the B × T product, where B is the modulated signal bandwidth and T is the bit period. When developing a system based on GMSK modulation, these parameters must be determined to fix the signal characteristics.

It must be taken into account that a Gaussian function is an unlimited function in time. That is, it takes nonzero values from minus infinity to infinity. These functions cannot be managed in practical realizations and the function must be truncated. As indicated in the previous section, the function g(t) is truncated to a duration of L bits and the function q(t) saturates at a finite time instant.

In the present case, the parameters are as follows: BT = 3, modulation index h = 0.5 and L = 3.

The power of a DSP can be exploited in different ways. The functions that define the modulation can be implemented directly. This is accomplished by saving the values of the function q(t) into memory and performing operations that include sine and cosine calculations. But this is an inefficient implementation.

A simpler way is to use lookup tables. Since the in-phase and quadrature components of the signal are sine and cosine values of the phase function, discrete values of the sine and cosine for arguments in the interval 0 to 2π can be stored in the DSP memory. All that remains is to obtain the memory address of sine and cosine functions for each argument (Fig. 4).

In Fig. 4, consider the case for L = 3. Values of the sine and cosine functions are discrete—that is, the argument can only take a finite number of values from 0 to 2π. The values of the functions have addresses corresponding to the arguments of the transmitted data sequence. The values of the sine and cosine function will be addressed depending on the data sequence to be transmitted. The operation of the system is as follows.

Two terms are well separated—a

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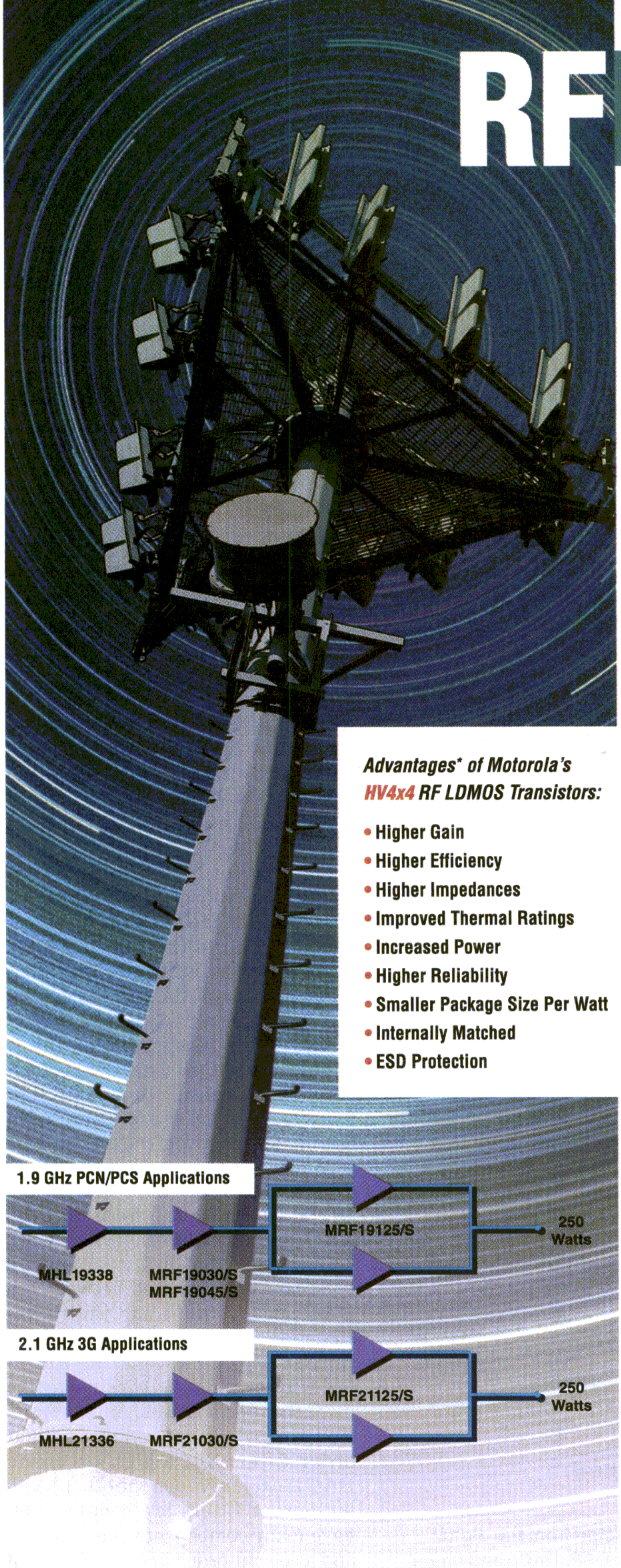

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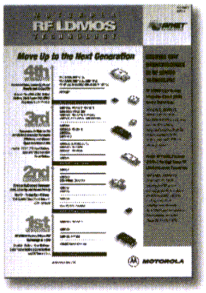

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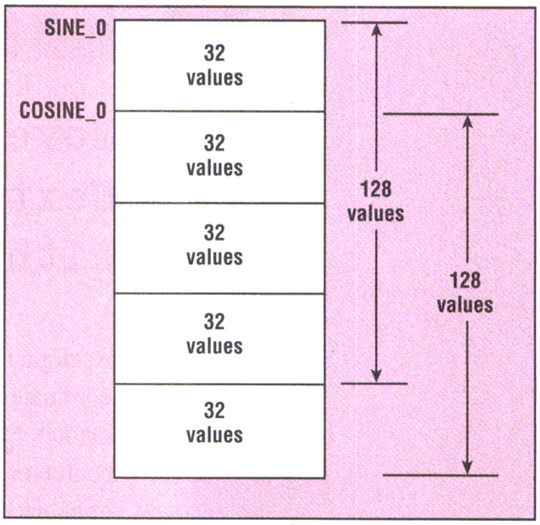
partial response term and a saturated term. The saturated term is composed of the phase caused by all the bits corresponding to values of the q(t) function that reach the saturated value in the analysis interval. The phase change due to saturated data for the modulation parameters considered in this development is π/2. Then the only possible values by which the saturated term can contribute to the signal phase are 0, π/2, π, and 3π/2. This is because the saturated bit adder is modulo-4—that is, the binary accumulation can take the values 0, 1, 2, or 3 corresponding to the phases previously listed.



5. Sine and cosine values are stored in memory to take advantage of the fact that the two functions are the same, only π/2 radians out of phase.

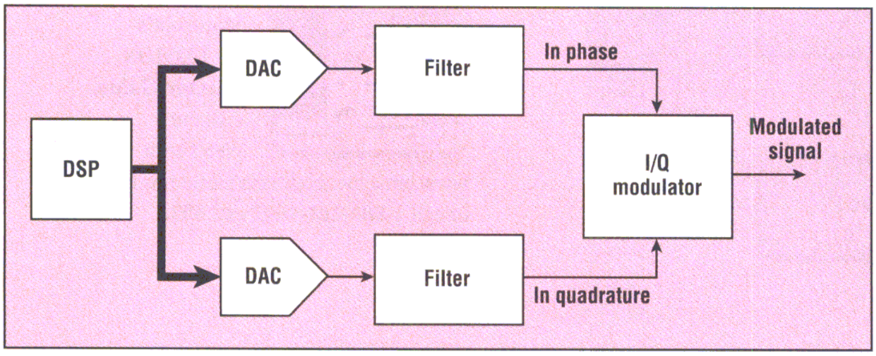
Due to the choice of L = 3, only the last three bits form part of the partial response term. As the q(t) function is well known and only eight different bit combinations are possible, these values can be prestored into memory. Instead of storing these values, it is useful to store the values that will address the corresponding phase change. This involves an additional phase change over that due to the saturated term.

The saturated term and the partial response term are added in a modulo-128 adder. This is because only 128 discrete values of the argument have been considered. This covers the phase range from 0 to 2π.

The addition of the address generated by the saturated term and that generated by the partial response term will produce the addressing of the corresponding values of the cosine and sine functions, that is, the in-phase and quadrature components.

The organization of the cosine and sine values stored in memory is shown in Fig. 5. To avoid repeated values of the cosine and sine functions, the π/2 radian's phase relationship between the two functions is used. The memory is organized as 32 memory positions for each quadrant. Beginning with the sine function as the cosine function is delayed π/2 radians, the values of this function will be allocated 32 memory positions below. Since the cosine and sine are periodic with period 2-π radians the values of both functions are organized as circular buffers with a range of 128.

The modulated signal is generated with four samples per bit period. This means that during a bit period, the DSP must generate four samples each for the sine and cosine functions. Before performing this operation, the partial response and saturated terms must be generated. During a bit period the saturated term remains constant and only the partial response term varies. This is the term that is responsible for the generation of the four samples per bit period. The partial response term table shown in Fig. 4 is organized so that the first address to generate the partial response term during a bit period is incremented in successive sampling instants at one position, addressing the next value. Eight memory blocks corresponding to the eight different values can be generated with three bits for the partial response term. Each block contains the four values that will generate the four samples per bit period. Digital samples obtained from the cosine and sine tables are outputted from the DSP to digital-to-analog converters (DACs) to obtain the analog in-phase (I) and quadrature (Q) components which, after filtering, feed the I-Q modulator, as shown in Fig. 6.



6. The I and Q signals for GMSK modulation are generated from the digital samples provided by the sine and cosine tables in the DSP. DACs and filters are needed to convert the digital signals to analog form.

The precision of the DACs depends on the system-quality requirements. The quality of the I and the Q components applied to the I-Q modulator depends on the DAC precision, that is, the number of bits with which samples are represented and the performance of the reconstruction filter. The quality of the signal can be identified with harmonics and noise suppression. Harmonics are generated in the digital-to-analog conversion process and have to be suppressed. Noise is generated by the finite precision of DACs. The higher the precision of the DAC, the lower the conversion noise.

Once fixed, the quality parameters must be a compromised between the DAC specifications and the reconstruction filter requirements. If the choice is a low-precision DAC, a high-order filter must be constructed to meet the specifications. A high-precision DAC does not require a high-order filter. But a high-precision DAC is expensive. On the other hand, high-order filters are complex and have many components. An 8-b DAC and a second-order lowpass filter can be employed with good results. ••

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

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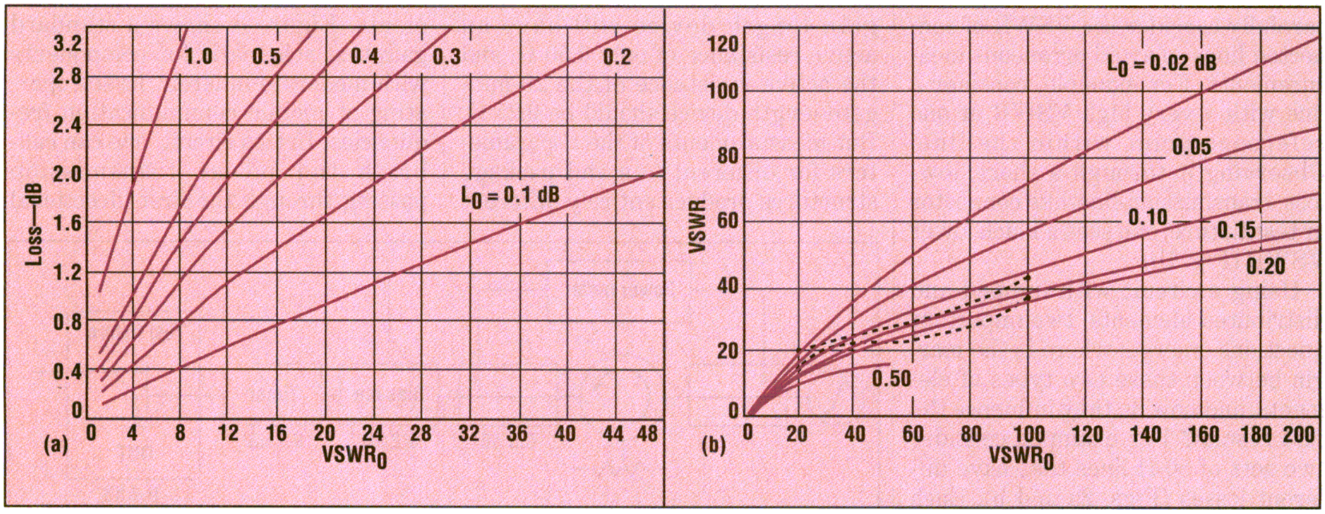
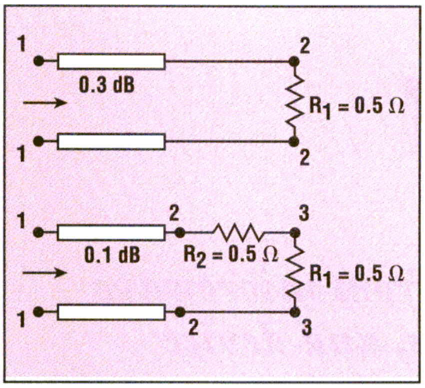
GAIN measurements of high-power transistors are strongly affected by losses at microwave frequencies, where the real portion of the matching impedance is very small. Microwave tuners play a crucial role in gain measurements, even for the very low losses attributed to matching lines. Moreover, the results of these measurements depend on the character of the losses (lumped or distributed) inside the transistors. This article presents an analysis of a gain-test circuit, including losses in the matching lines. It presents test results for different power metal-semiconductor field-effect transistors (MESFETs) to validate the analysis. The results are applicable to metal-oxide-semiconductor field-effect transistors (MOSFETs), high-electron-mobility transistors (HEMTs), heterojunction bipolar transistors (HBTs), and other types of microwave power transistors. The results can also help engineers choose the right test circuit to achieve the best parameter values. This can be especially important for high-volume production lines.

Coaxial cables, microstrip lines, and other transmission lines all have inherent losses. But what is the impact of these losses on the parameters measured, and how does one correctly choose the best line? The losses in homogeneous transmission lines between a well-matched line generator and a reflective load are defined as:[1,2]



1. This graph shows the total losses in a line containing standing waves (a) while the other graph shows the resulting VSWR through the line (b).



2. These two circuits have the same line VSWR but different useful load resistance.

$$L = 10\log\frac{1-|\Gamma_L|^2 e^{-4\beta l}}{e^{-2\beta l}\left(1-|\Gamma_L|^2\right)}, [dB] \qquad (1a)$$

where:

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

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

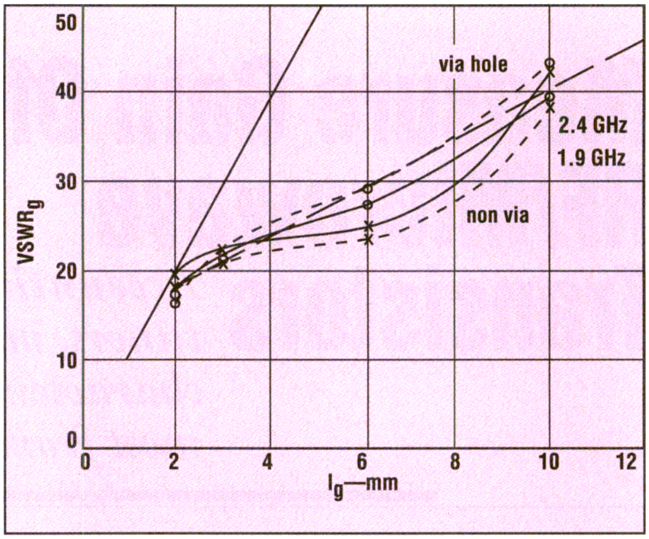
Ω_L = the load's reflection coefficient.

The value $L_0 = e^{\beta l}$ represents the generalized loss for a line with traveling-wave conditions, and Eq. 1a shows the contribution of standing waves to the total losses. Figure 1a represents total calculated losses, which depend on the VSWR:

$$VSWR_0 = \frac{1+|\Gamma_L|}{1-|\Gamma_L|} \qquad (1b)$$

It shows that the total losses in a line increase significantly with increasing VSWR, especially for high initial-loss values. Thus, one must carefully consider the VSWR of connected lines to avoid erroneous measurements. For example, consider a line with a very high VSWR at one reference point within the line. Observing it through a short, low-loss, transmission-line-like connector or test fixture, one sees a much lower VSWR (Fig. 1b).

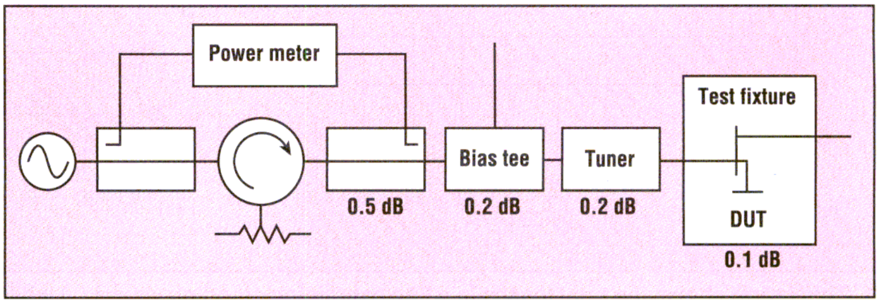
Using a circuit with lumped and distributed elements, two important questions emerge: where is the margin between these two types of elements, and how do they influence the test results? For example, consider two sets of 50-Ω lines that have different losses (Figs. 2a and b). Each set of lines has a lumped, 0.5-Ω resistor (R_1) connected as a load. In addition, the set of lines in Fig. 2b contains a series, 0.5-Ω resistor (R_2). Looking from ports 1-1, the VSWR in both cases is approximately 30:1. For the circuit in Fig. 2a, the total loss from ports 1-1 to the load (R_1) is L = 14.07 + 6.49 = 20.56 dB. For the circuit in Fig. 2b, L = 11.14 + 1.98 + 3.0 = 16.12 dB. The losses in the circuit in Fig. 2b are diminished because the elements are lumped. To distinguish between distributed losses and lumped losses, one must consider the physical characteristics of the components in each case.



3. This graph shows the FET's theoretical and measured VSWR attributed to the gate resistance.

TESTING TRANSISTORS

Consider a series of L- and C-band, GaAs power MESFETs produced by Hexawave, Inc.[3,4] The devices have gate lengths ranging from 1 to 20 mm, and they are composed of identical unit cells connected in parallel. The device with a 1-mm gate length has an active input resistance of approximately 5.0 Ω, so one can assume that its parameters will not be influenced by the matching-line losses during measurements. One can also assume that the 10-mm-gate-length device will have an active resistance of only 0.50 Ω, and the active resistance of the 20-mm-gate-length device should be 0.25 Ω. But measurements of the S-parameters for different lots and package structures produce entirely different results. Figure 3 shows the measured $VSWR_g$ versus l_g data for devices with 0.7- and 1.0-μm gate widths. $VSWR_g$ is attributed solely to the active gate resistance. The straight, solid line shows the theoretical values for a chip. Losses inside the package play a significant role with increasing gate length. If one represents these losses in lumped form, the probable decrease in a FET's power gain is presented in Fig. 5. If one considers the losses in distributed form, their contributions are represented by the dashed lines in Fig. 1b. Having some preliminary test data and knowing, in principle, what devices are measured and the expected results, the goal is to find out how to provide correct measurements. Figure 4 shows a standard gain-test schematic for components with moderate inherent losses operating at mobile communication frequencies. In this circuit, one can calibrate the power transmission passing through the first directional



4. This block diagram shows a standard gain-test setup.

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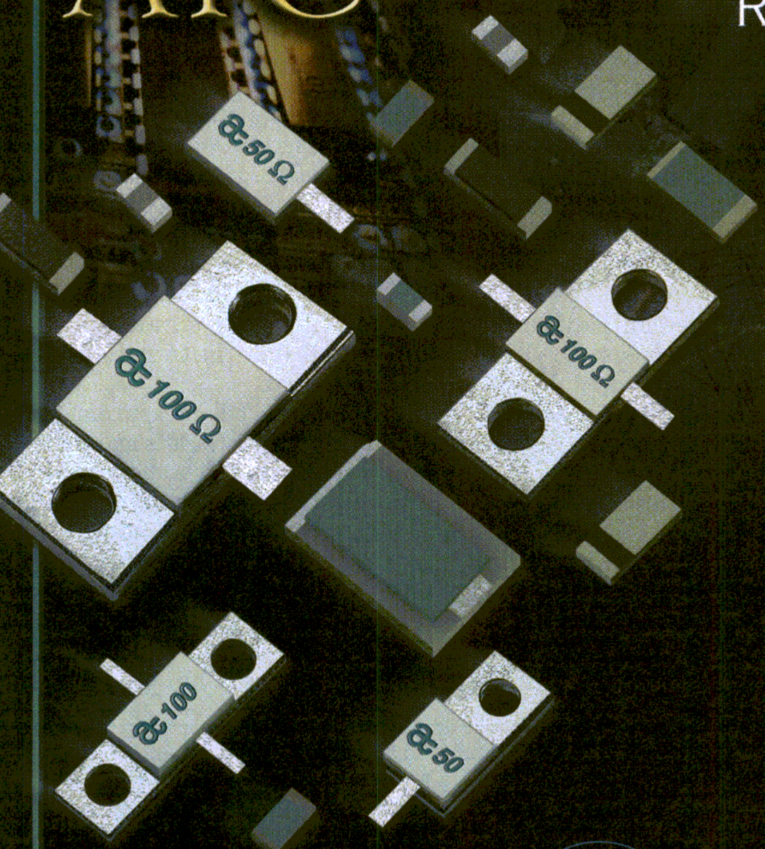

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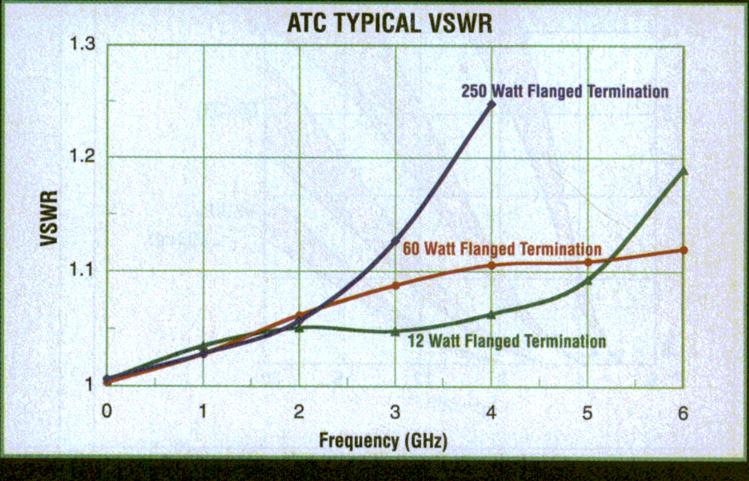

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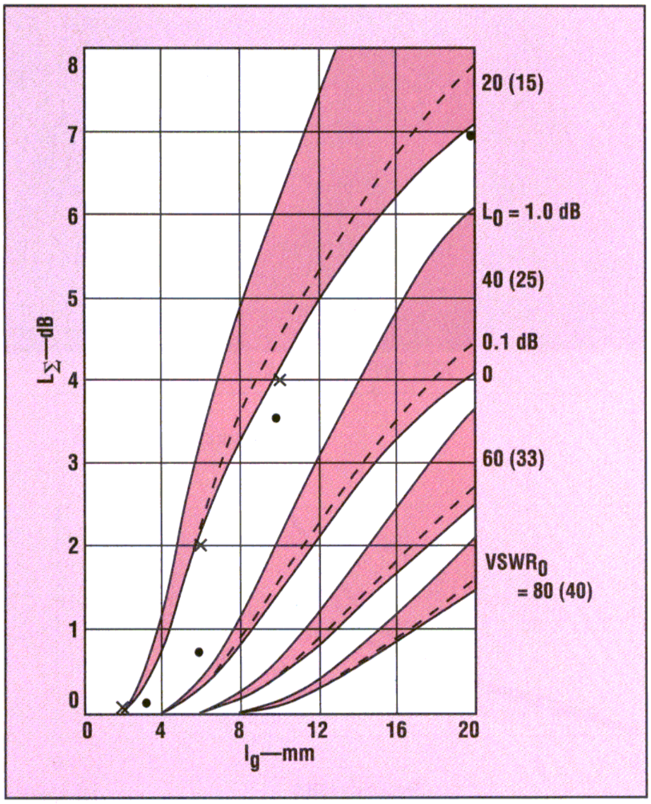
coupler to the transistor's gate lead using a 50-Ω load. But if improper matching causes standing waves in a line between the circulator and the device under test (DUT), the results are unknown. In fact, using a correctly calibrated vector analyzer instead of a power meter might make matters worse. The analyzer does not "know" what kind of losses are inside the transistor package.

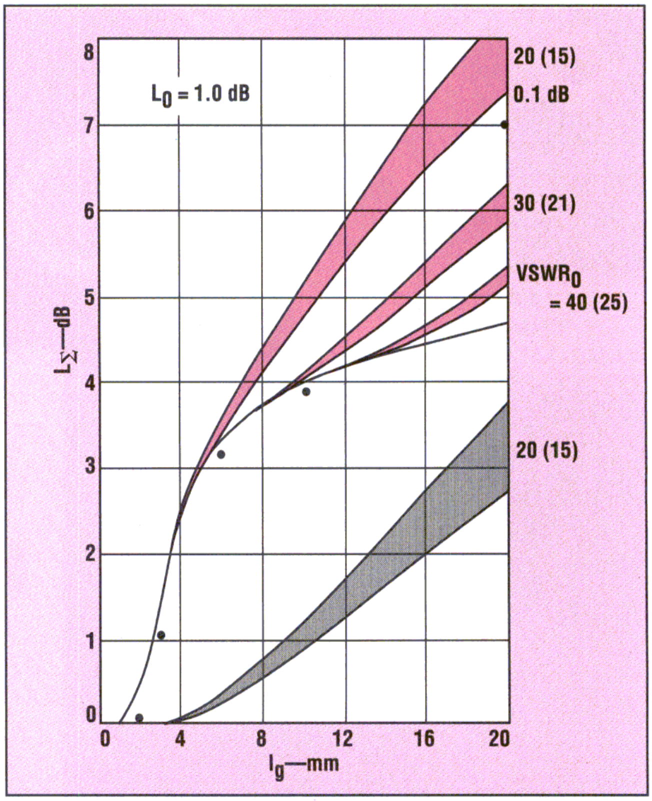
Figure 5 assumes the losses are distributed, and shows the total additional unknown losses in the line due to standing waves (between the circulator and the DUT) for the idealized transistors previously mentioned at different tuner VSWR conditions and at several initial line-loss values. The VSWR values are presented at the tuner-stub reference plane for the initial loss of a tuner having L_0 = 0.1 dB. This is more realistic, considering some additional losses when characterizing tuners. One can see that since the tuner's 50-Ω output impedance does not match the gate resistance—the result is a sharp growth of total losses with increasing gate length, especially for fairly low values of tuner VSWR. Furthermore, it can be seen that the initial losses contribute significantly to the total loss. If one does not take these circumstances into account, the high power gain of a DUT can be lost. The lower the initial loss value and the higher the achievable tuner VSWR are, the lower the loss is. Even if the reflected power is measured, the situation is analogous for the reflected wave. The further the directional coupler is located from the DUT, the lower the measured reflected power is. Moreover, its value depends on the phase distribution in the line with the standing wave. This yields contradictory requirements—to decrease the loss indefinitely, it is necessary to move the directional coupler as far as possible from the DUT and increase the initial-loss value.

Consider the real parameters measured in Fig. 3, and assume that all transistor losses are lumped. The useful resistance is defined by the straight solid line, and the total resistance is defined by the straight dashed line. In this case, the results presented in Fig. 6 are analogous to those in Fig. 5. If one disregards the location of the useful load, the result will be as shown by the dashed lines in Fig. 6. These results can neither explain the features of the real device, nor can they predict its behavior by changing some parameters of a circuit. Comparing Figs. 5 and 6, one can see quite different values and shapes of the curves. For Fig. 6, some saturation effect of the losses is observed with increasing gate length and tuner VSWR. The contribution of an initial line loss to the total loss is lower, owing to the lower VSWR value in the whole system. That, again, significantly reduces the indefinite reflected power measured.

Analyzing the RF measurement data of many transistor lots and comparing gain figures at a maximal P_{1dB} power, it was found that for via-hole chip structures, the total FET losses may be represented in distributed form. Figure 5 shows the average measurement results at 2.4 GHz for packaged and unpackaged (chip) transistors. Chip losses are



5. This graph shows the theoretical and experimental FET losses due to the distributed character of its initial losses.



6. This graph shows the theoretical and experimental FET losses due to the lumped character of its initial losses.

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higher due to additional test-fixture losses. For non-via chip structures with gate lengths to 10 mm, the total FET losses can be represented in lumped form (Fig. 6). But for longer gate widths, the distributed character of the losses begins to have more influence on the total loss value.

An experiment conducted on a 10-mm-gate-length, non-via-structure packaged FET driven at 1.9 GHz shows that extending the achievable VSWR value from 20 to 30 increases the gain value from 1.0 to 1.5 dB, and an increase in power-added efficiency (PAE) from 7 to 10 percent, over drain-bias voltages from +3.5 to +10 VDC. Increasing the VSWR to 40 provides a 0.5-dB increase in gain at only +3.5 VDC. These results are consistent with the physical properties of the device and they are comparable with the results in Fig. 6.

For all FET structures, one must choose tuners with the highest possible VSWR and the least initial whole-line losses, especially for via-hole structures and long gate widths. Otherwise, one can expect much lower gain and PAE values, which may be crucial for high-volume production lines and for characterizing newly developed devices. It should be noted that commercially available tuners do not satisfy the measurement accuracy required for devices with gate lengths of 10 mm or larger and for via-hole structures diminishing real gain figures. In these cases, FET-chip manufacturers should provide the internal matching of packaged devices and additional low-loss transformation characterization. Designers can get the most from an active device by taking its physical properties into account. Recently, Focus Microwaves introduced its new two-stub prematching tuner and low-loss test fixture that can extend the tuning range significantly.[5]

Within a FET's normal operating mode, standing waves at its output port are not a problem. Therefore, this article does not discuss power measurements. The improvement in power measurement achieved in ref. 5 can be attributed entirely to advanced matching—not to standing waves.

The questions discussed here can partially explain why many manufacturers—using approximately the same technology—provide different gain figures but practically equal power figures for their power transistors. Furthermore, using an evaluation board instead of a standard test circuit often results in increased gain values. ••

Acknowledgement

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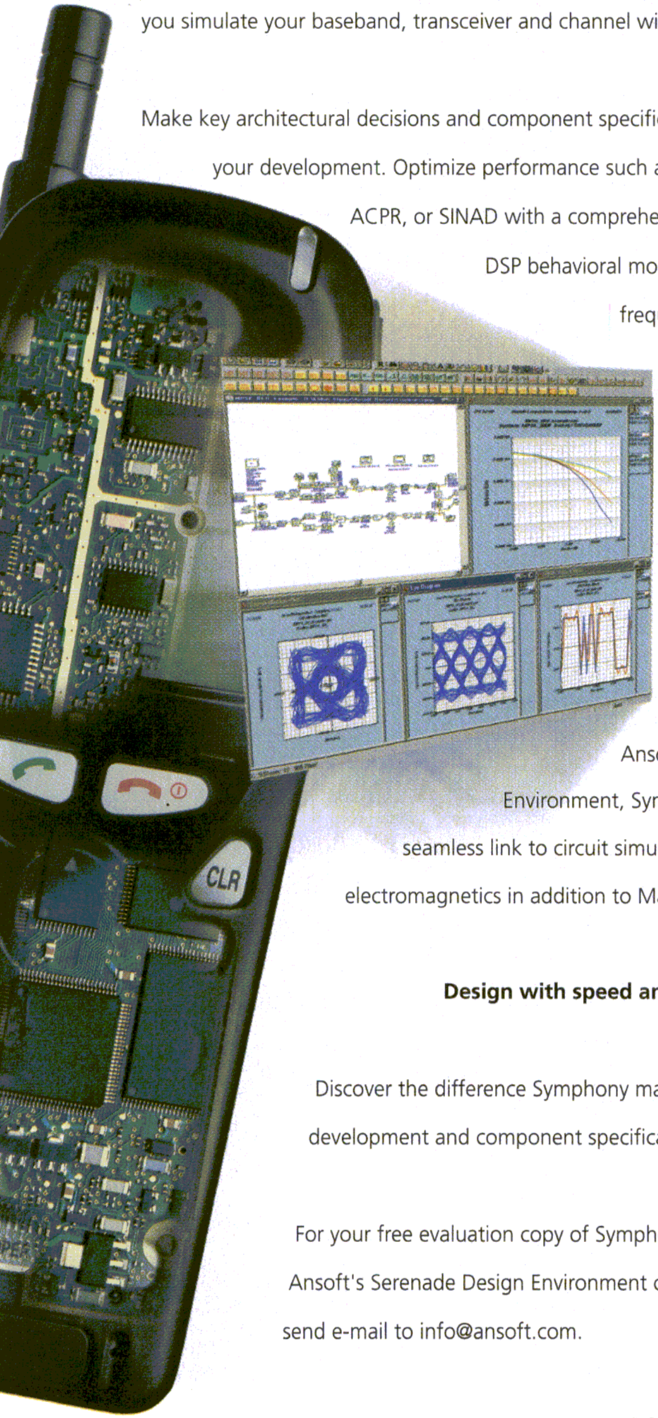

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Power Devices Stake Their Territories

Solid-state and vacuum-electronics devices may represent drastically different technologies, but they do compete for some amplifier applications.

Barry Manz
President
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DEVICES that are used to generate RF power—transistors and vacuum-electronics devices or tubes—are truly the odd couple of microwave technologies. They could not be more different technologically or in the power levels they attain, yet they exist together to achieve the same goal—to produce clean output-power levels for a variety of applications.

Manufacturers of solid-state amplifiers, the reigning champions at low power levels and frequencies in the lower microwave region, are faced with the fundamental challenge of the commercial wireless marketplace—reducing cost to lower levels every year. At the other end of the power-generation spectrum, vacuum-tube manufacturers strive to generate higher power levels at higher frequencies with greater efficiency for a wide array of customers in satellite communications, medical, scientific, radar, electronic warfare (EW), and even industrial markets.

In the microwave industry, which for most of its existence has focused on delivering heretofore unachievable levels of performance for aerospace and defense systems with cost as a secondary issue, designers of solid-state amplifiers for wireless base stations face an entirely new set of rules. Those rules are dictated by the increasingly complex modulation schemes used in digital communications systems and the need for their customers to deliver base stations at lower cost every year.

Unlike defense customers, which have traditionally placed a premium on performance, wireless infrastructure manufacturers take for granted that amplifier manufacturers will achieve the low distortion and clean power production dictated by international standards. Performance beyond this, unless truly exemplary and providing some tangible cost savings to the system builder and ultimately the service provider, is not something they are willing to pay for. "It's not 'give us the best you can,' it's 'make it as inexpensive as you can,'" says Murat Eron, director of engineering at Microwave Power Devices (Hauppauge, NY).

David Tai, president of CTT (Santa Clara, CA), tempers this argument slightly, but agrees in general. "Price is certainly the number one issue, because system prices drop 15 to 20 percent per year. But everyone uses more or less the same technology (to build wireless amplifiers), and systems companies tend to stay with the original vendor. So the key issues are to be part of the project from the beginning and to place a premium on servicing the customer and making sure your return rate is very low."

The trend toward increased wireless network capacity, more complex modulation schemes, and incorporation of data with voice is likely to keep solid-state amplifier designers busy for many years. The single-carrier-per-channel amplifiers of the past are rapidly being replaced with multicarrier designs that integrate the amplifier, power combiner, and a filter into a single enclosure. As an example, model G3S-1900-80 is a multicarrier modular power amplifier (PA) developed by Powerwave Technologies (Irvine, CA) which can provide up to 300-W power from 1930 to 1990 MHz in a four-module rack. Similarly, model LPA 890 is a multicarrier amplifier module from HC Electronics (Allendale, NJ) with 90-W output power from 869 to 894 MHz (see *Wireless Systems Design*, August 2000, p. 52). Up to four of the LPA 890 modules can be combined to yield 330-W output power.

In additon, Paradigm Wireless Communications (Irvine, CA) offers the MAF-S series of multicarrier PAs with feedforward linearization. Other suppliers of multicarrier PAs include Advanced Control Components (Eatontown, NJ) and Spectrian Corp. (Sunnyvale, CA).

These complex subsystems generally rely on laterally diffused metal-oxide-semiconductor field-effect transistors (LDMOS FETs) at frequencies to approximately 2.3 GHz, and gallium-arsenide (GaAs) FETs beyond. Bipolar transistors, which have powered wireless communications systems for decades, are slowly fading to the background as LDMOS technology advances in power output and overall performance.

Peak-to-average signal ratio is high in digitally modulated systems and as a result, amplifier designers must go to great lengths to accommodate the occasional high-power peak. "You're adding a lot of silicon just to capture peaks without clipping that occur only 1 percent of the time," says MPD's Eron. "The amplifiers are also biased Class AB, which means they generate a lot of heat, and when you put all of these devices together, mechanical and thermal considerations become extremely important."

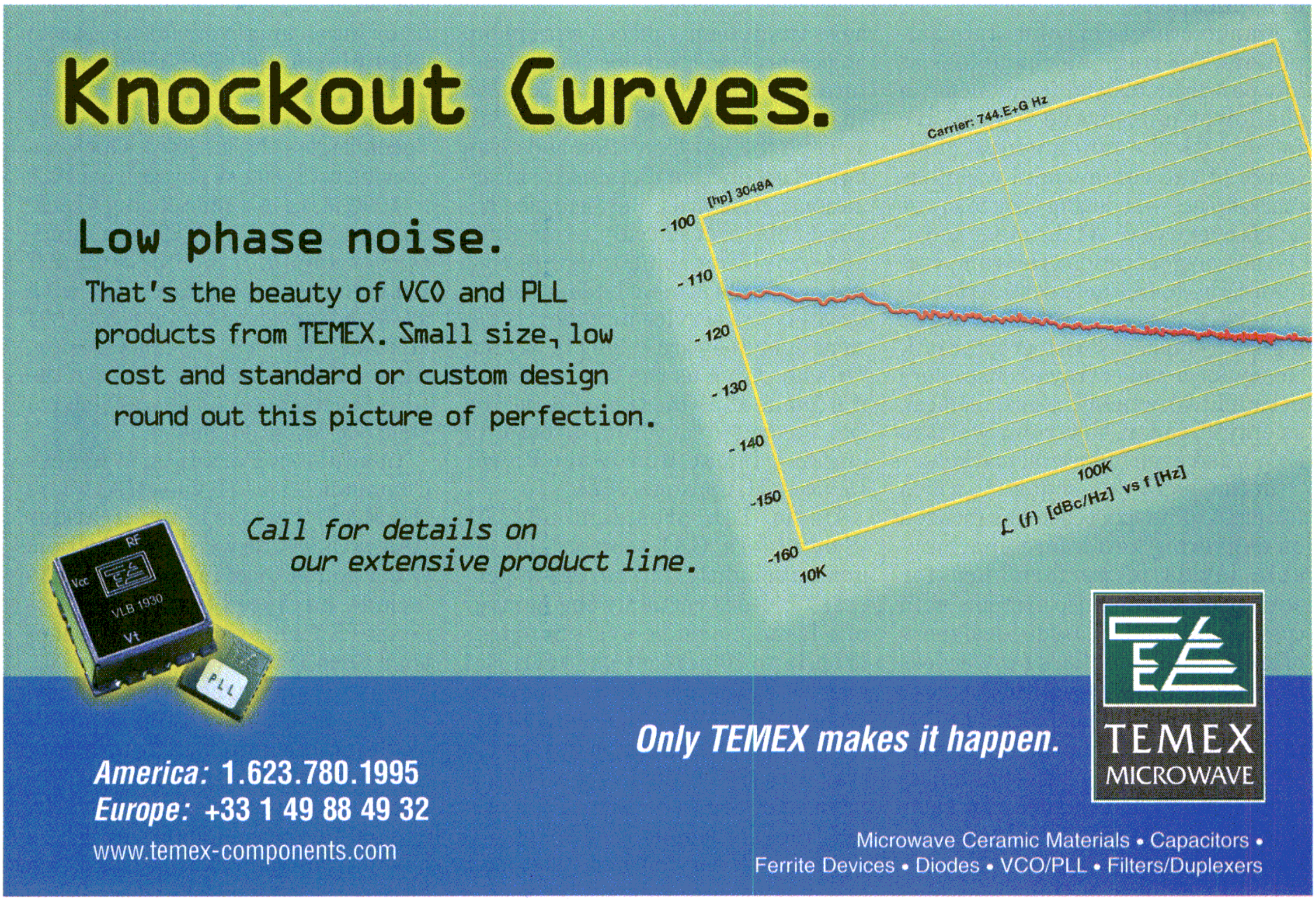
BEYOND CELLULAR

Although most of the attention (and money) for amplifier manufacturers comes from wireless communications applications, waiting in the wings are other applications that have enormous potential. First on the list is usually wireless links for satisfying the "last mile" problem of obtaining high-speed Internet access or telephone service to residential customers. In some countries where wired telephone service is spotty or nonexistent, wireless solutions provide compelling alternatives to stringing thousands of miles of copper (Cu) wire, many times over rugged terrain, in hostile climates. An offshoot of this concept, currently offered in several US cities by AT&T and Sprint, uses this technology to deliver relatively high-speed broadband Internet access to homes, in direct competition with digital subscriber lines (DSL) and cable modems.

The amplifiers powering the residential units can have low power output, while the communications hubs (one of which might serve all customers on a street, for example), require higher power levels. The frequencies chosen for this service are often unlicensed bands in the spectral region of approximately 2.3 GHz, with an accompanying band at 5.8 GHz. Amplifier designers employ GaAs FETs at the higher frequencies, since the performance of silicon (Si) devices fades at these frequencies and beyond.

Beyond this application lurks the first true commercial consumer application at millimeter wavelengths—local multipoint-distribution service (LMDS). This service has been heralded as the next competitor to optical fiber, since the bandwidth available is immense, and is able to support data rates an order of magnitude higher than available with Cu-based solutions.

However, LMDS continues to be slow in gaining acceptance, thanks to the combined effect of market, regulatory, and technological pressures. Creating a millimeter-wave transceiver and baseband-processing system for consumer applications at 28 to 30 GHz, at a reasonable price, is a challenge that will only be met when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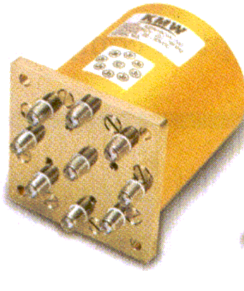

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volumes of LMDS systems reach an appreciable level. To get there, service providers will likely have to subsidize equipment cost initially in order to sign up the early adopters. Whether or not LMDS gains wide acceptance after that is a question that manufacturers, service providers, and investors ask. CTT has been delivering an amplifier for LMDS hubs that delivers approximately 4 W of pulsed power using monolithic microwave integrated circuits (MMICs). However, even CTT's Tai admits that LMDS activity has been slow.

In addition to communications applications, broadband solid-state amplifiers are also routinely used in measurement applications, such as for electromagnetic-compatibility (EMC) testing. Firms such as Amplifier Research (Souderton, PA), Instruments For Industry (Ronkonkoma, NY), Kalmus (Bothell, WA), and LCF Enterprises (Post Falls, ID) offer solid-state amplifiers that are capable of wideband coverage through 1 GHz with power levels exceeding 1000 W. Most of these companies also offer tube-based amplifiers for applications requiring power levels of greater than 2 kW.

Together, the opportunities for solid-state amplifier manufacturers are keeping production lines running at full tilt, and have generated concerns about parts shortages of resistors and capacitors that are holding back deliveries. The amplifiers required for coming 2.5G wireless systems and the third-generation (3G) systems to follow, will soon add to this backlog.

REAL POWER

The public perception of vacuum tubes is not helped when the nation's vice president uses a glass receiving tube as a prop in his pronouncement about what is wrong with the nation's aging air-traffic-control radars. Of course, Al Gore was actually holding a 12AU7 rectifier tube and was standing next to one of the first solid-state radar displays to include altitude and other information on the scan display. Other than that tube, the only other vacuum tube in the radar was the CRT itself.

Nevertheless, to people under the age of 40 or so, vacuum tubes are relics of an era gone by, an opinion supported by the fact that consumer products have not used them for decades. However, to anyone with a need to generate a high level of RF power at high frequencies, vacuum tubes are, as they have been for decades, the only technological choice.

As Robert Symons, technical director at the Litton Electron Devices Div. (San Carlos, CA) puts it, "Semiconductors are marvelous for low-power signal processing and logic functions, where tubes cannot compete at all. However, in the power business, where the size of the amplifier is determined by how much power you can get rid of with a particular type of coolant, what the operating temperature and efficiency of the thing is, tubes are still smaller, lighter, and more cost-effective." Symons has been designing high-power microwave tubes for more than 40 years, and invented the clustered-cavity klystron.

This is not to say that power transistors have not encroached on markets that were the exclusive domain of

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vacuum tubes. They are actually eroding some traditional tube markets at lower frequencies and power levels. "The AM radio market used to be tubes alone," says Bart Petrini, president of Communications and Power Industries (CPI) [Palo Alto, CA], "but now it's almost all solid-state to about 15 kW." CPI is the successor to the Electron Device Business of Varian Associates, and includes the Eimac division. CPI manufactures helix traveling-wave tubes (TWTs), klystrons, gyrotrons, coupled-cavity TWTs, and high-power transmitters (Txs) for satellite earth stations, medical, scientific, terrestrial communications, defense, and broadcast markets.

Solid-state Txs account for approximately 20 percent of the broadcast market (by number of customers). Powered almost exclusively by LDMOS FETs, these amplifiers deliver on the semiconductor's promise of high reliability, although at a higher initial cost. The potential 10× improvement in longevity of solid-state amplifiers continues to make them appealing choices in many markets. The level of encroachment depends largely on four factors—output power, efficiency, heat generation, and heat dissipation.

As broadcasters cautiously embrace high-definition television (HDTV) between now and 2006 (when all broadcasters must offer this technology), more tube-based PAs will be employed to serve this wideband, high-power application. Specifically, klystrons that deliver between 30 and 120 kW will be the power generator of choice. In this application, the tube is still king, and probably will remain enthroned for the foreseeable future.

One of the most talked-about new entertainment delivery services—satellite-radio broadcasting—is deploying repeaters throughout the US that use either transistor or klystron amplifiers, depending on the power output required. The satellite-to-car-radio digital-audio-broadcasting market consists of XM Satellite Radio championed by General Motors, and Sirius Satellite Radio (formerly CD Radio) championed by Ford and Daimler Chrysler. Each plans to offer 100 channels of compact-disc (CD)-quality programming beamed by high-power (TWT-based) satellite transponders direct to vehicles equipped with amplitude-modulation/frequency-modulation/signal-audio-tone (AM/FM/SAT) radios, at a monthly rate of $9.95. Manufacturers of transistor and tube amplifiers are shipping subsystems in large numbers for competitors to serve the repeaters required to fill in coverage gaps. These services will be operational in 2001, with Sirius expected to debut first.

Even in the wireless industry, where the primary technology drivers are smaller, lighter, and less expensive, base-station designers are eyeing 'mini-TWT' amplifiers as alternatives to solid-state designs. The appeal is the tube's inherent ability to handle the high peak-to-average power ratios of complex modulation schemes, and its small size when compared to conventional tubes. So while solid-state Txs rule the base-station market and will continue to do so, there may be room for TWTs in certain applications.

The defense community which, along with the broadcast industry, has always been an enormous consumer of vacuum tubes (there are more than 800 on each Navy Aegis cruiser), continues to buy them in large numbers for radar and EW systems. The market for "spares" alone is estimated to be approximately $100 million annually. The millimeter-wave region, which is out of the range of semiconductors at more than tiny power levels, is generating increasing revenues for tube manufacturers as well. Since, for example, every new radar or battlefield communications system operating at 77 GHz will later produce a need for corresponding EW systems at this frequency, the prospects are good for manufacturers of the tubes that power them.

Solid-state phased-array radars were big news in the microwave industry, since the many transmit/receive (T/R) modules they employ hold the promise of electronic-beam steering. The power dissipation required by solid-state radars with thousands of T/R modules, and the corresponding reliability questions, are controversial points within Department of Defense (DoD) budget discussions, and one that tube manufacturers have used as an argument for using their technology.

UNLIKELY PLACES

In a world that hardly recognizes their existence, tube manufacturers are continually looking for new markets where their particular strengths are well-suited. In most cases, the perception they face is the one fostered by Gore and others who use vacuum tubes as examples of "the early days of radio."

"I've been at CPI for 11 months," says Petrini. "My background is in tube manufacturing, but I was out of the manufacturing area for quite awhile. When I returned, I was amazed at the yields that can be achieved in tube manufacturing today. A lot has been learned from traveling-wave-tube development for the satellite-communications industry. Transponders in space have to last 15 years, and techniques employed in tubes for market have spilled over to other areas as well."

John Beighley, acting president of CPI's satellite-communications division, notes that vacuum-tube applications are more widespread than most people know. "In February, the FDA approved irradiation of meat (in response to public outcry over the e. coli virus and other outbreaks). Manufacturers are now rushing to develop systems that can irradiate millions of pounds of meat per year. Klystrons are one of the top choices for generating the power required in these systems. They need to generate lots of power, so that hamburger patties can be stacked six high on a conveyer and still be treated."

Beighley cites a host of other examples as well, from sterilizing medical waste, to cooking dog food, to generating plasmas in ion implanters for semiconductor processing, as well as curing laminated wood beams. "We're even working with a brewer that is interested in heating their beer with microwave energy," says Beighley, "since there is evidence to show that beer heated in this way is of better quality and has a longer

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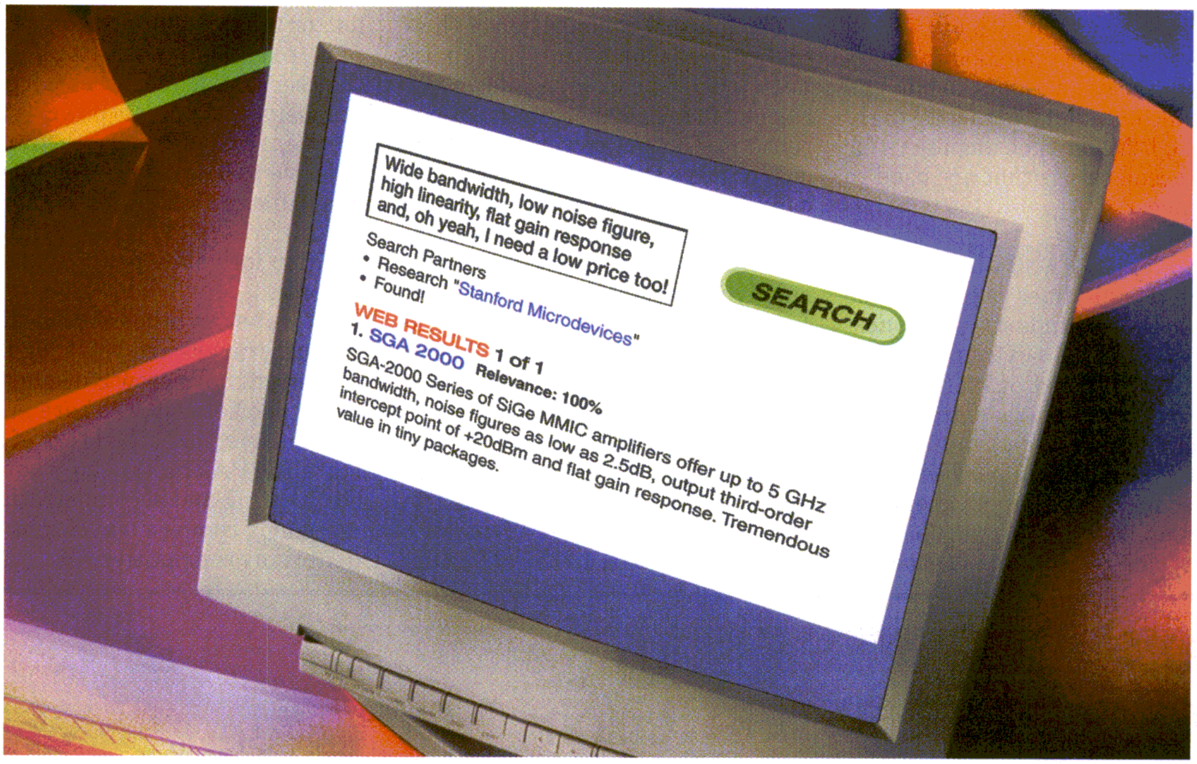
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Gain (dB)	10.5	15.0	17.4	19.6
TOIP (dBm)	20.0	20.0	20.0	20.0
P1dB (dBm)	7.0	7.0	7.0	7.0
N.F. (dB)	4.1	3.2	2.9	2.5
Supply Voltage (Vdc)	2.2	2.2	2.7	2.7
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shelf life than with conventional methods."

The travel industry is also a market, since huge cruise ships must store enough food to feed 1000 people for more than a week. To do this, the food is kept frozen, and rapidly defrosted when needed. Large industrial ovens, powered by 60-to-75-kW magnetrons, allow the food to be defrosted between meals. Also, the medical community is increasingly using high-power tube amplifiers for linear accelerators that are used to treat cancer.

NEW TUBES

Even though tubes reign supreme in the broadcast industry, the quest for greater efficiency remains constant because a major increase can save broadcasters tens of thousands of dollars in energy costs, effectively obsolescing competitive tubes.

Two tube developments exemplify the trend to extract more performance from technologies that have been in existence for decades. The first is the constant efficiency amplifier (CEA), developed by the Litton Electron Devices Div. (San Carlos, CA). The CEA is essentially an inductive output tube (IOT) with a multistage depressed collector (MSDC).

The unusual combination of these techniques (for which Litton has been granted patents), allows efficiency to reach 60 percent at 25-percent-rated output power with 70 percent or more at full output. The CEA can serve commercial broadcasting of NTSC or 8-VSB advanced-television (ATV) signals, reducing power-supply requirements by 50 percent. Litton believes that by simply employing this device, US TV broadcasters will collectively save approximately $25 million per year.

Another advance comes from CPI, the developer of the extended interaction klystron (EIK), which can deliver continuous-wave (CW) power of 1.5 kW at 30 GHz, 100 W at 95 GHz, 30 W at 140 GHz, and pulsed peak power of 3 kW from 30 to 95 GHz and 400 W at 140 GHz. The device has a 1-dB instantaneous bandwidth of up to 4 percent which, for example, has produced 100 W at 30 GHz with a bandwidth of 2.4 GHz.

Vacuum-electronics and semiconductor technologies are, for the most part, exclusive in the applications they serve. Vacuum tubes can generate megawatts of RF power at high frequencies, and there are no competitive technologies. For this reason, tube manufacturers can comfortably expect to remain in this position for years, perhaps decades, to come.

However, while vacuum-device manufacturers are working in the corners of the microwave industry to serve markets where there are few competitors, manufacturers of solid-state RF PAs fight a daily battle of cost cutting and accommodating increasingly complex waveforms in order to stay competitive. ●●

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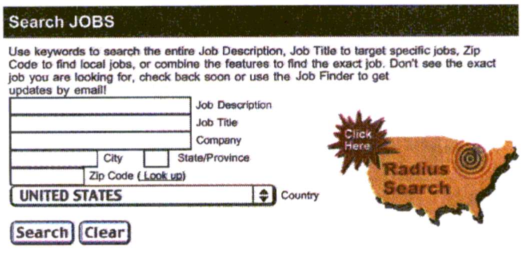

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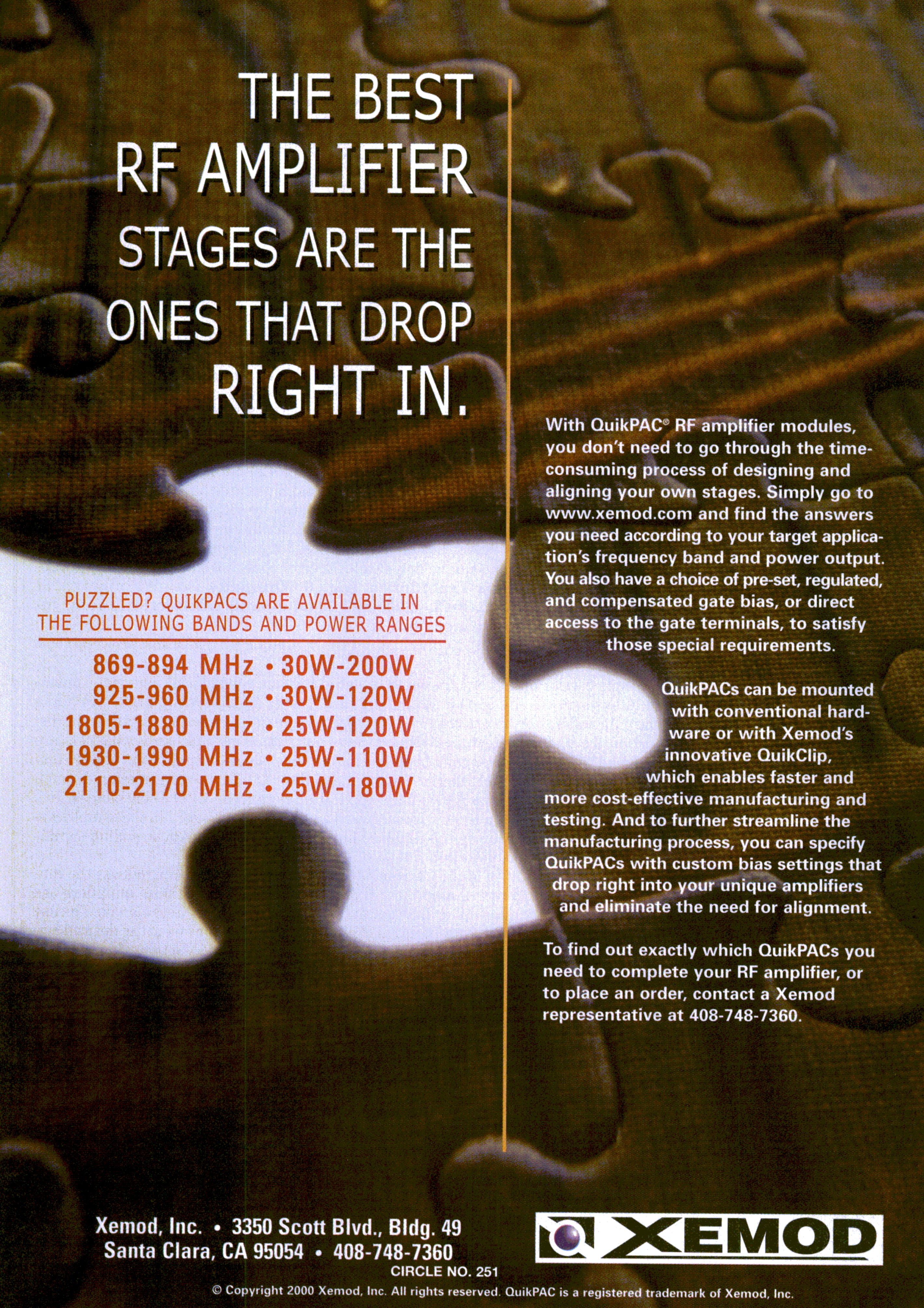

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Understand phase-noise needs of digital radios

Digital-modulation schemes such as quadrature amplitude modulation (QAM) allow relatively narrowband digital microwave radios to support high-speed data transmissions to 155 Mb/s. The design of such radios is often a careful trade-off between the performance of the demodulator and the levels of phase noise in the local oscillator (LO). Fortunately, an application note from Communications Techniques, Inc. (Whippany, NJ), "Source and Synthesizer Phase Noise Requirements for QAM Radio Applications," explains what levels of phase noise are needed for a particular radio-link budget and expected bit-error rate (BER).

Written by William Reuter, Senior Design Engineer with the company's Synthesizer Group, the application note includes a block diagram of a typical QAM radio link using dual conversion for the receiver (Rx) and transmitter (Tx). The LOs in the final upconversion and the first downconversion stages are the largest contributors to phase noise since they are at the highest operating frequencies. The note also offers a set of curves for BER as a function of signal-to-noise ratio (SNR), where a BER of 10^{-6} is used as the typical threshold for data transmitted on microwave links.

The comprehensive application note reviews residual SNR and its effects on the fade margin and the influence of the carrier-recovery phase-locked loop (PLL) on radio performance. It offers guidelines on making carrier-recovery PLL phase-noise-suppression measurements on a QAM system, and illustrates how to characterize the integrated LO phase noise. The note includes a step-by-step procedure for selecting and specifying phase-locked dielectric-resonator oscillators (PDROs) or frequency synthesizers for QAM systems based on a given set of performance specifications. The 16-page application note can be downloaded free of charge from the company's website. **Communications Techniques, Inc., 9 Whippany Rd., Whippany, NJ 07981; (973) 884-2580, FAX: (973) 887-6245, e-mail: sales@cti-inc.com, Internet: http://www.cti-inc.com.**

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Applying feedforward amplification techniques

Single-carrier amplifiers are being replaced in some cellular and personal-communications-services (PCS) base stations by multicarrier amplifiers capable of providing the gain and power boosts required by these complex signals without significant distortion. In such architectures, the low-level outputs of multiple radio channels are summed through a power combiner. The higher-level single output of the power combiner is then fed to a multicarrier combiner and filtered prior to transmission through the base-station antenna. A two-page application note from MPD Technologies, Inc. (Hauppauge, NY) explains how feedforward amplifiers can be used in these multicarrier applications to provide the required output-power levels without adding significant intermodulation distortion (IMD).

The note, "Benefits of Multi-Carrier Feed Forward Amplifiers," introduces the concept of feedforward power amplifiers (PAs) and describes how these amplifiers are characterized with a spectrum analyzer for IMD, using multitone test signals. The use of feedforward requires the separation of signal-distortion terms from the desired signal (which can be performed by canceling the carrier signal and then rotating the IMD spectrum 180 deg. in phase; an auxiliary error amplifier is used to produce a canceling signal. The canceling signal is then injected 180 deg. out of phase at the output of the PA in order to cancel the distortion terms.

The application note provides a basic block diagram for a feedforward amplifier system. It reviews the many benefits of using a feedforward amplifier for multicarrier base-station applications, including the ability to handle mixed-mode applications, such as Advanced Mobile Phone Service (AMPS) and code-division-multiple-access (CDMA) signals, lower output losses due to the removal of filters and power combiners, and smaller total equipment size due to the capability of a single amplifier to process multiple carriers, in contrast to one amplifier per carrier. Copies of the application note can be downloaded for free from the company's website. **MPD Technologies, Inc., subsidiary of Microwave Power Devices, Inc., 49 Wireless Blvd., Hauppauge, NY 11788; (516) 231-1400, FAX: (516) 434-1725, e-mail: sales@mpd.com, Internet: http://www.mpd.com.**

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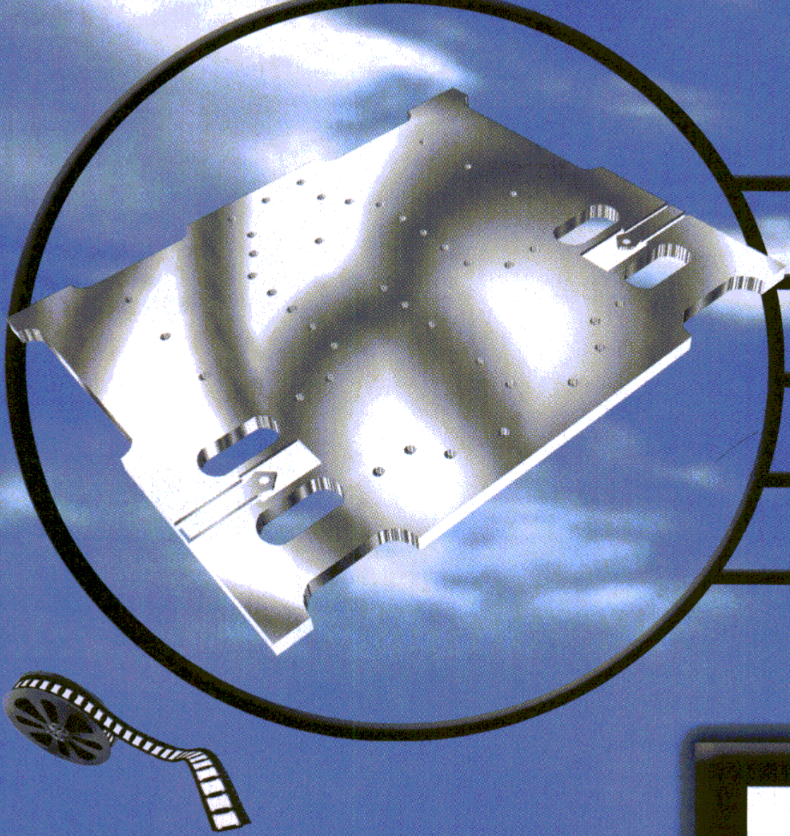

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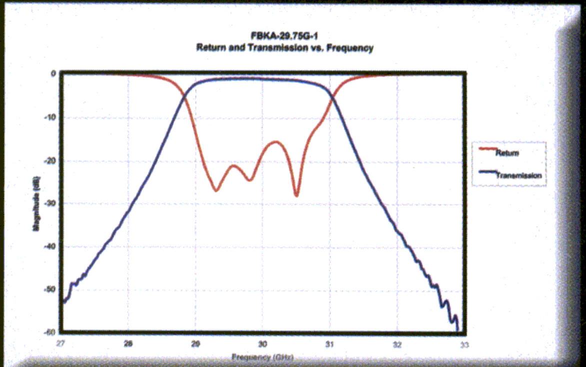

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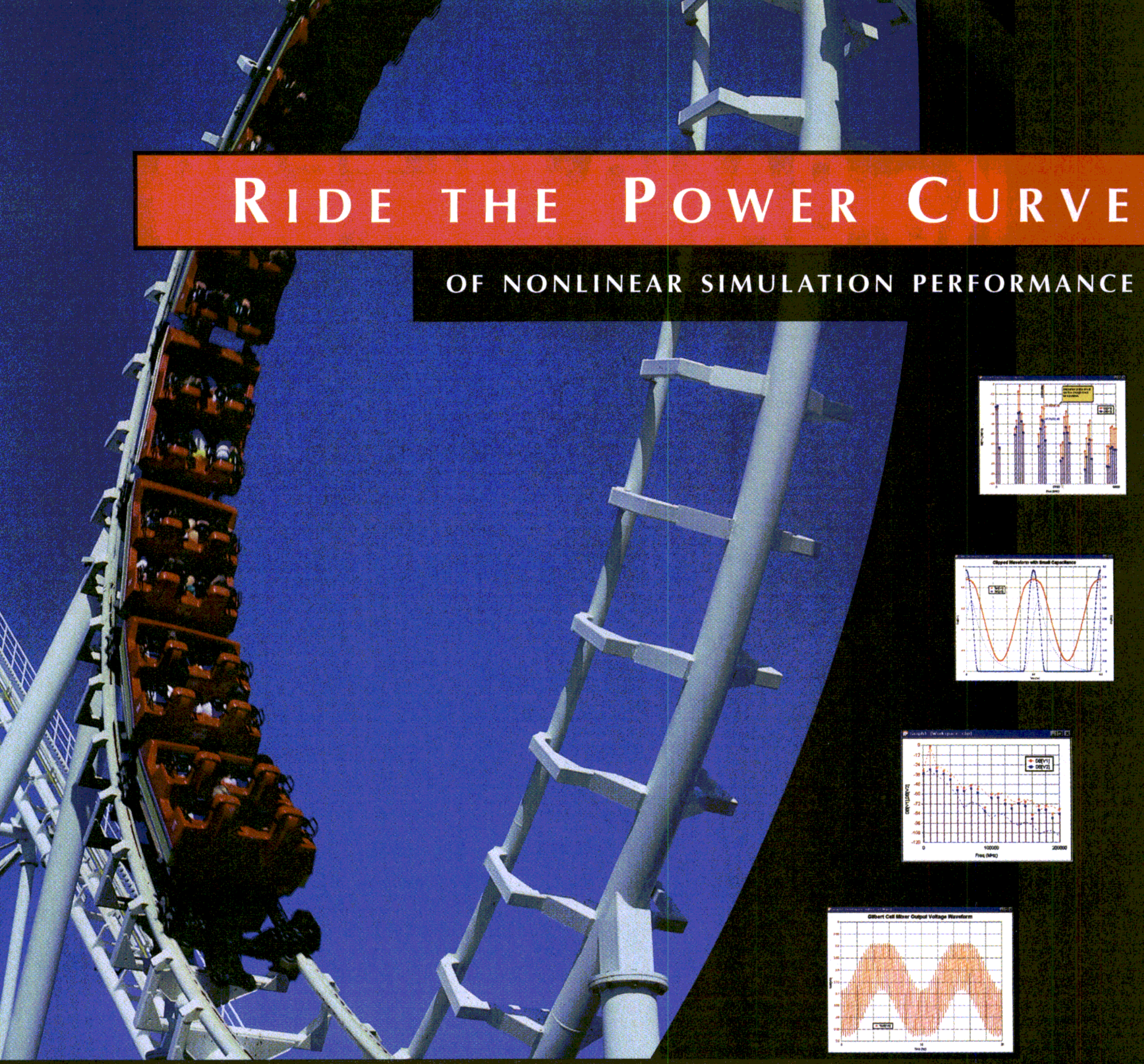

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Spectrum Analyzers Tune To New Levels Of Test Performance

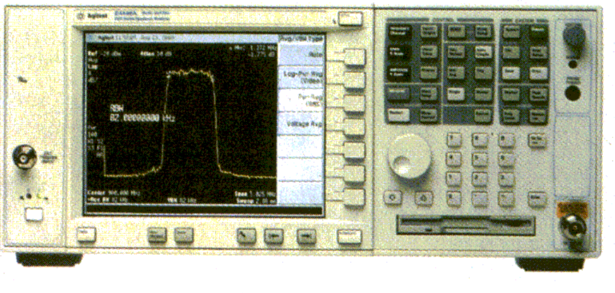
This line of performance spectrum analyzers makes the most of digital IF filtering and a dual-PLL scheme in order to increase measurement speed and accuracy at microwave frequencies.

JACK BROWNE
Publisher/Editor

SPECTRUM analysis is one of the fundamental test methods at higher frequencies. For many years, the HP 8566 spectrum analyzer from Hewlett-Packard Co. (Palo Alto, CA) was the reference standard for all other analyzers in terms of measurement power and accuracy. But as Hewlett-Packard's former instrumentation and medical lines have been transformed into Agilent Technologies, spectrum analysis will be transformed, with the introduction of the performance spectrum analyzer (PSA) series of instruments. The model E4440A is the first analyzer in the new line, with coverage from 3 Hz to 26.5 GHz. It embodies several innovations that enable it to set new standards for measurement speed, accuracy, and dynamic range at a competitive cost.

The E4440A (Fig. 1) carries an impressive list of credentials, including sweep speeds of 1 ms, sensitivity of −153 dBm, logarithmic linearity of 0.1 dB across a 100-dB dynamic range, one distortion-free dynamic range of 113 dB, and absolute amplitude accuracy of ±0.35 dB. Across a frequency range of 26.5 GHz, microwave engineers now truly have an open window to a wide range of continuous and intermittent signals.

With respect to the HP 8566A/B analyzers, the E4440A is an entirely new platform, built for modern applications. The complex intermediate-frequency (IF) circuitry found in the 8566A/B instruments has been reduced to a single integrated circuit (IC). And what required eight circuit boards in signal-processing electronics in the 8566 has been condensed to a single circuit board in the E4440A. The new analyzer, which is based on a reduced-instruction-set-computer (RISC) microcontroller, takes advantage of advances in microprocessor technology for enhanced processing speed and accuracy.



1. The 3-Hz-to-26.5-GHz model E4440A is the first member in a new line of performance spectrum analyzers (PSAs) based on extensive digital signal processing (DSP).

It also leverages digital-signal-processing (DSP) technology extensively. The E4440A, for example, features unprecedented 160 resolution-bandwidth filter settings, with filters adjustable from 1 Hz to 3 MHz in 10-percent steps, and fixed at 4, 5, 6, and 8 MHz. In the past, resolution-bandwidth filters to approximately 1 kHz were the standard. But with the E4440A spectrum analyzer, all of the resolution-bandwidth filters are digital, contributing to the absolute amplitude accuracy and measurement speed. By implementing the filters with DSP, it is possible to achieve near-ideal shape factors (the ratio of the 3-to-60-dB bandwidth). Traditional analog resolution-bandwidth filters

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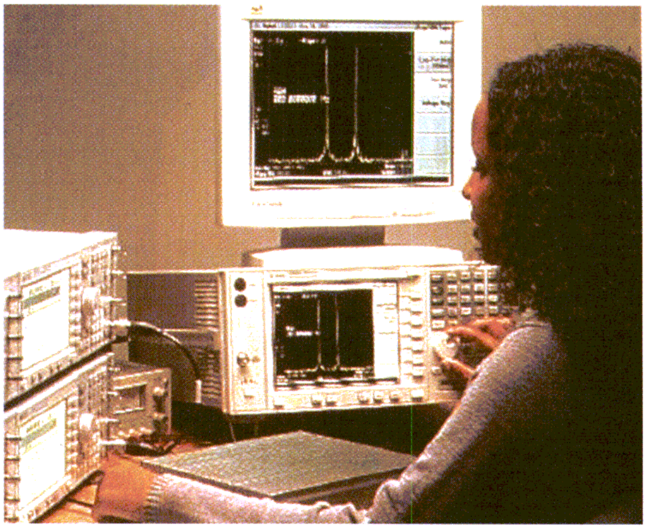
exhibit shape factors to 12:1. The digital filters in the E4440A boast nominal selectivity of better than 4.1:1. Due to this improved filter performance, operators can often use wider resolution-bandwidth filters with the E4440A than in conventional spectrum analyzers with analog filters, resulting in improved measurement speed (since narrower filters lead to longer sweep speeds).


2. With 160 resolution-bandwidth-filter settings, the E4440A spectrum analyzer offers unprecedented measurement flexibility along with the capability of optimizing measurement speed, dynamic range, and resolution.

The wealth of resolution-bandwidth filters—a total of 160 resolution-bandwidth settings in the E4440A—allows operators to fine-tune their measurements rather than settle for conventional filters in a 1-3-10 sequence. As a result, the resolution-bandwidth filters can be set up to optimize the measurement speed, dynamic range, and resolution, without sacrificing accuracy (Fig. 2).

When compared with analog filters, the E4440A's digital resolution-bandwidth filters are precise and predictable. The characteristics of the digital filters are stable with time and temperature, leading to improved amplitude accuracy when measuring noise or noise-like digitally modulated signals. The resolution-bandwidth switching uncertainty that plagues spectrum analyzers with analog filters is almost nonexistent in the E4440A, eliminating one of the error terms common to spectrum analysis prior to the E4440A. By minimizing resolution-bandwidth switching uncertainty (only ±0.05 dB for filters from 1 Hz to 2 MHz), E4440A operators are ensured of improved measurement repeatability and confidence.

Since the linear-phase characteristics of the E4440A's digital filters are precisely known, they can be swept and corrected at faster speeds than possible with analog resolution-bandwidth filters. In contrast, the sweep speed for analog filters is generally set to the square of the resolution bandwidth (in hertz) divided by 2 ($RBW^{0.5}/2$ Hz), which is a compromise between measurement speed and amplitude/frequency errors due to fast sweeping. The digital filters in the E4440A can sweep approximately twice as fast as analog filters, and with greater accuracy.

ABSOLUTE ACCURACY

One of the E4440A's most striking performance specifications (see table) is the absolute amplitude accuracy—±0.3 dB at 50 MHz. This is accuracy that rivals the performance of a stand-alone power meter, and applies to measurements on continuous-wave (CW) and modulated signals. As previously mentioned, the digital resolution-bandwidth filters contribute to this accuracy. The E4440A also employs a unique digital logarithmic amplifier in its front-end electronics, with a dynamic range exceeding 100 dB. Logarithmic amplification is performed digitally, so it is possible provide a log range that is limited only by instrument noise. Signals from −10 dBm to near the noise floor at −156 dBm can be measured without changing the IF gain or input attenuation.

Although the E4440A can examine signals up to +30 dBm when using its input attenuator, it also features an input mixer that can be directly driven with signals up to +5 dBm without attenuation, and without noticeable distortion of sideband information. Since the E4440A achieves a third-order-intercept-to-noise dynamic range of 73 dB, it is ideal for evaluat-

The E4440A spectrum analyzer at a glance

Frequency range	3 Hz to 26.5 GHz
Frequency spans	0 Hz, 100 Hz to 26.5 GHz
Frequency resolution	2 Hz
Frequency accuracy	±0.2 percent of span
Frequency counter (marker) resolution	0.001 Hz
Resolution bandwidths	1 Hz to 3 MHz (in 10-percent steps), 4, 5, 6, and 8 MHz
Filter selectivity (60 dB/3 dB)	<4.1:1
Video bandwidths	1 Hz to 3 MHz (in 10-percent steps), 4, 5, 6, and 8 MHz
Noise sidebands	
Offset 100 Hz	−88 dBc/Hz
Offset 10 kHz	−113 dBc/Hz
Offset 100 kHz	−118 dBc/Hz
Offset 1 MHz	−142 dBc/Hz
Offset 6 MHz	−145 dBc/Hz
Offset 10 MHz	−148 dBc/Hz
Amplitude measurement range (at 22 GHz)	−143 to +30 dBm
Reference-level range	−170 to +30 dBm in 1-dB steps
Input attenuator range	0 to 7 dB in 2-dB steps
Frequency response	
3 Hz to 3 GHz	<±0.35 dB
3.0 to 6.6 GHz	<±1.50 dB
6.6 to 26.5 GHz	<±2.00 dB
Absolute amplitude accuracy (at 50 MHz)	±0.3 dB

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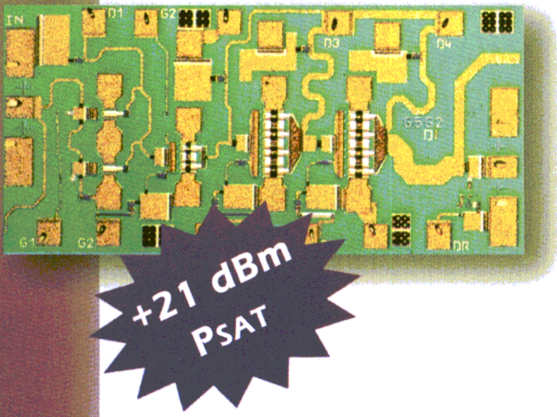

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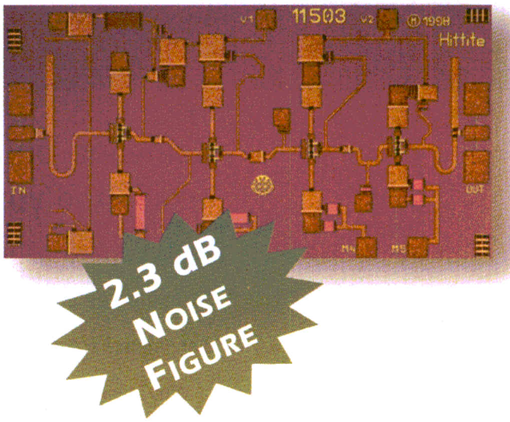

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HMC261	20 - 40	Driver AMP	13	7	+12
HMC262	15 - 24	LNA	25	2.0	+5
HMC281	18 - 32	LNA	22	2.5	+10
HMC263	24 - 36	LNA	25	2.3	+6
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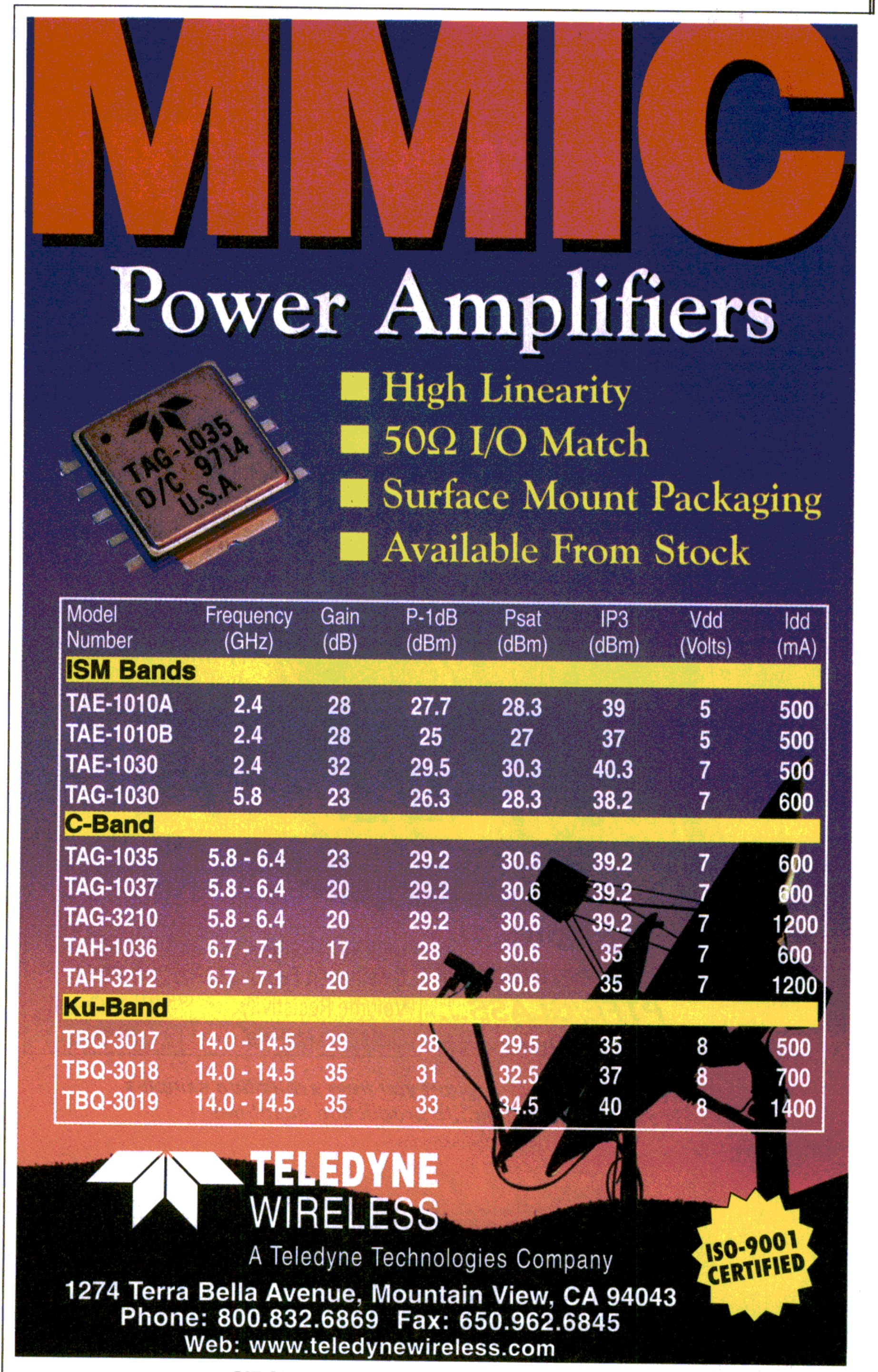
ing the adjacent-channel-power (ACP) characteristics of wideband-code-division-multiple-access (WCDMA) signals.

With its digital front end, the E4440A can offer a variety of useful functions, including multiple measurement modes—swept mode, Fast Fourier transform (FFT) mode, or a combination of the two. The swept-tuned mode is based on superheterodyne mixing of input signals with stable synthesized local-oscillator (LO) signals. Even the LO in the E4440A is unique, since it consists of two phase-locked loops (PLLs). One of the PLLs is optimized for close-in phase noise (at offsets closer than 30 kHz), while the other is optimized for further-out phase noise and faster tuning between center frequencies. So, within the swept mode, operators also have a choice of measurement modes. The phase noise for offset frequencies from 10 to 100 kHz is typically −113 to −118 dBc/Hz, but drops to −142-dBc/Hz offset 1 MHz from the carrier. The E4440A's phase noise can be optimized manually, or operators can choose an automatic mode that selects the phase-noise setting that optimizes measurement speed.

Sweep speeds can be set very slow—in the zero-scan mode, sweeps can range from 1 µs to 6000 s per sweep—for capturing intermittent signals. For frequency spans that are greater than 10 Hz, the sweep speeds range from 2 ms to 2000 s per sweep. Sweeps can be triggered in a variety of ways, and can be delayed by as much as 500 ms.

Fourier analysis is based on the mathematical analysis of digitized signals, and relies on the E4440A's high-speed 14-b analog-to-digital converter (ADC) working at 30 MSamples/s. The FFT measurement mode can provide great detail on signals within narrow bandwidths and with narrow resolution-bandwidth filters, but can be limited by the noise that is also sampled by the ADC. The E4440A allows operators to fine-tune the ADC through the use of dither. Dither, an additive signal that linearizes the ADC, can be set for best detection linearity. By using dither, the low-level logarithmic fidelity can be improved from approximately 0.50 to 0.07 dB. Without dither, the displayed average noise level (DANL) is approximately 1.5 dB lower than when using dither.

For operators favoring simplicity, the E4440A offers an autocoupling function where various measurement parameters, such as resolution bandwidth, video bandwidth, and sweep speed, can be linked. In addition, the choice of swept measurement or FFT can be automatically determined by the selection of frequency span and resolution bandwidth. An operator can even choose to divide a span into a specified number of FFTs. The dynamic range of the narrow FFTs can approach that of a swept measurement, while also providing the



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Model Number	Frequency (GHz)	Gain (dB)	P-1dB (dBm)	Psat (dBm)	IP3 (dBm)	Vdd (Volts)	Idd (mA)
ISM Bands							
TAE-1010A	2.4	28	27.7	28.3	39	5	500
TAE-1010B	2.4	28	25	27	37	5	500
TAE-1030	2.4	32	29.5	30.3	40.3	7	500
TAG-1030	5.8	23	26.3	28.3	38.2	7	600
C-Band							
TAG-1035	5.8 - 6.4	23	29.2	30.6	39.2	7	600
TAG-1037	5.8 - 6.4	20	29.2	30.6	39.2	7	600
TAG-3210	5.8 - 6.4	20	29.2	30.6	39.2	7	1200
TAH-1036	6.7 - 7.1	17	28	30.6	35	7	600
TAH-3212	6.7 - 7.1	20	28	30.6	35	7	1200
Ku-Band							
TBQ-3017	14.0 - 14.5	29	28	29.5	35	8	500
TBQ-3018	14.0 - 14.5	35	31	32.5	37	8	700
TBQ-3019	14.0 - 14.5	35	33	34.5	40	8	1400

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measurement-speed benefit of an FFT.

The E4440A offers a wide range of detectors, including normal, minimum, peak, sample, average, and root-mean-square (RMS) detectors. The sample detector, for example, is suitable for random noise measurements while the RMS detector is better suited for broadband amplitude measurements. But when in doubt, an operator can also allow the E4440A to automatically select the detection mode based on the measurement.

The instrument's range of averaging functions can also be selected manually or left to automatic control. Averaging processes include trace averaging, average detector, video-bandwidth filtering, and noise-marker averaging. Depending upon the needs of the measurement, the averaging functions can be performed on a power scale (when speed is required), on a logarithmic scale (for optimum CW measurements near the noise floor), and on a voltage scale (for analyzing pulsed signals).

The E4440A is also designed to simplify data transfers. It features local-area-network (LAN) connectivity and data transfer; the ability to capture screen and data to Microsoft Windows applications, such as Excel; and a host of plug-and-play drivers for Agilent VEE and other popular test automation software, such as LabView and LabWindows from National Instruments (Austin, TX). The analyzer includes general-purpose-interface-bus (GPIB), LAN, and RS-232 ports to simplify interconnections to networks and dedicated printers.

With its impressive accuracy, the E4440A allows operators to not only accelerate the pace of their measurements, but to work with closer tolerances and tighter guard bands on measurements. For example, filters that may have passed (or failed) testing in the past due to several decibels of uncertainty in the spectrum analyzer can now be evaluated and shipped (or rejected) with confidence.

The E4440A is the first member of the PSA line. It is available with an optional low-noise preamplifier (500 kHz to 3 GHz with 30-dB nominal gain) to improve the DANL from −153 dBm to −167 dBm. Future models will build upon the basic platform and will include demodulators for the Third Generation Partnership Program (3GPP) and EDGE signal analysis. The E4440A (and other members of the PSA line) are part of the company's test-and-measurement trade-up program, allowing owners of existing HP or Agilent spectrum analyzers to upgrade to an E4440A. P&A: $48,000 (E4440A); 30 days. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; (800) 452-4844 ext. 7176.**

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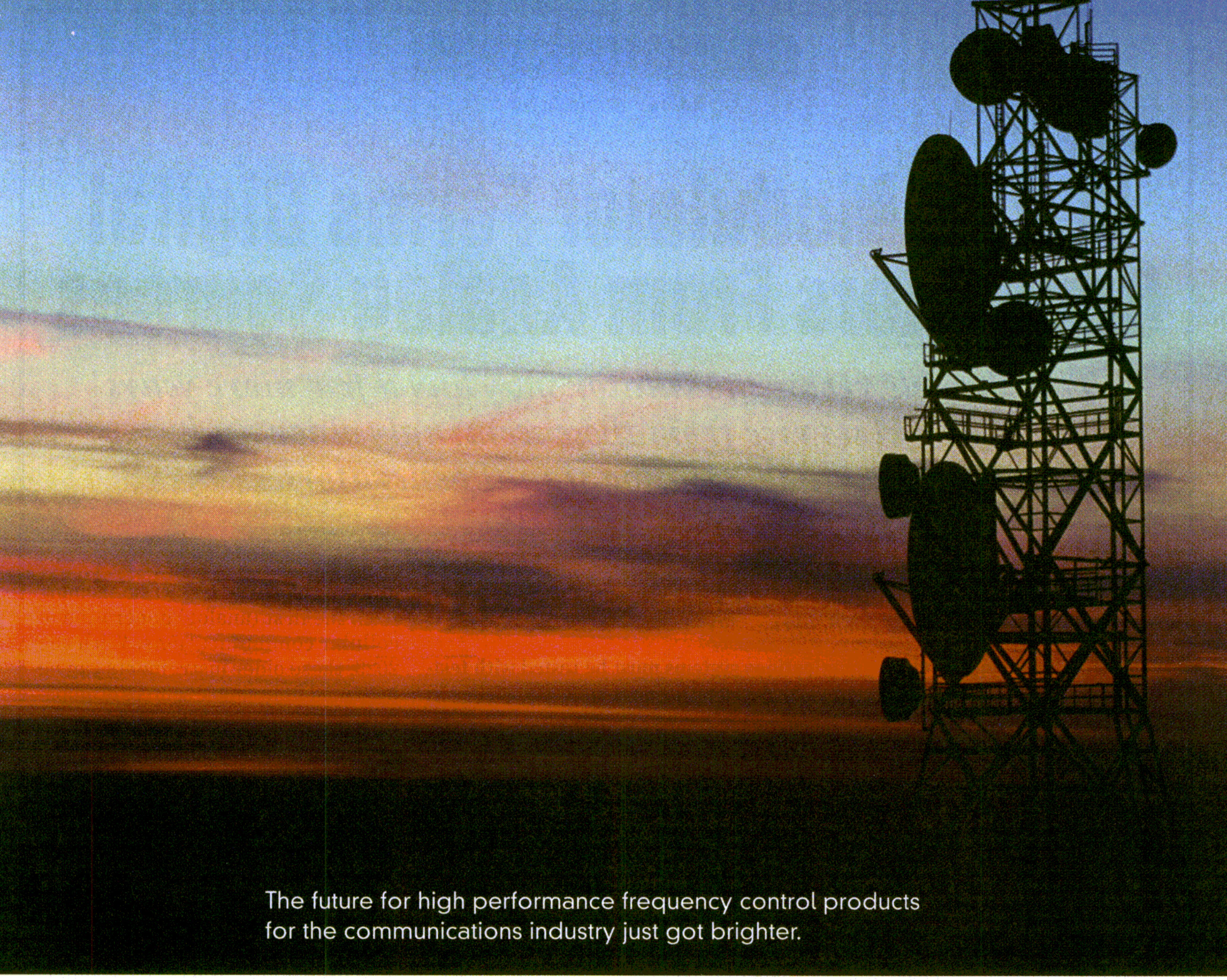
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CIRCLE NO. 200

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Vector Modulator Forms Digital Generators From Analog Sources

This flexible vector modulator may buy a few more years of useful service from those seemingly older and "outdated" analog signal generators.

JACK BROWNE

Publisher/Editor

DIGITALLY modulated communications systems must be tested with test-signal sources that closely emulate the signals found in these systems. In most cases, such signals are based on the use of complex phase modulation to represent different digital bits or states. Emulating these signals has generally required an expensive signal generator with built-in digital modulation. But with the model 2029 vector modulator from IFR (Wichita, KS), those older analog signal sources can now be converted into full-featured digitally modulated sources capable of emulating all of the latest digital communications systems. The vector modulator is suitable for use from 800 to 2510 MHz, and even the combination of the 2029 with an analog signal generator represents a fraction of the cost of a new digitally modulated signal generator.

The 2029 (Fig. 1) is designed primarily for production-line testing of digital cellular-telephone handsets, although it can be used with an analog signal generator for product and research-and-development (R&D) testing of any number of components and systems that fall within its frequency range. It serves as the modulation source for an analog signal generator. Rather than upgrade an analog source to include built-in digital modulation, the signal generator is used to drive the 2029 modulator, which delivers vector-modulated signals at its output port.

The vector modulator includes a powerful arbitrary waveform generator to create a wide range of complex digital modulation formats, including quadrature-amplitude-modulation (QAM) and quadrature-phase-shift-keying (QPSK) modulation. The arbitrary waveform generator can be programmed to create long, complex sequences of waveforms as might be used in modern communications formats, such as Bluetooth appliances, code-division-multiple-access (CDMA) systems, Digital European Cordless Telecommunications (DECT), and Global System for Mobile Communications (GSM) networks.

Most larger manufacturers of cellular handsets, and components and subsystems for those handsets, have banks of analog signal generators remaining from the first (analog) generation of cellular communications. With the 2029, these older instruments can be upgraded from simple sources of amplitude modulation (AM) and frequency modulation (FM) to more sophisticated sources of in-phase (I) and quadrature (Q) digital modulation.

The 2029 is easy to use. The output signal from an analog signal generator serves as the local-oscillator (LO) drive signal to the input port of the 2029 and its wideband I/Q modulator. The signal generator should provide a signal level of +7 dBm ±2 dB. The I/Q modulator is also fed by a pair of quadrature baseband I and Q signals produced by the arbitrary waveform generator (AWG) [Fig. 2]. The I/Q vector modulator features a bandwidth in excess of 10 MHz, enabling it to create multiple-signal formats and the latest wideband-CDMA (WCDMA) signal formats. The modulator is assisted by RF level-control circuitry which allows the 2029 to provide output power levels from −138 to 0 dBm with resolution of 0.01 dB and RF level accuracy of typically better than ±0.25 dB. The 2029 can generate output signals with crest factors (peak-to-average power) of better than 14 dB, making it suitable for multichannel IS-95 CDMA.

The 2029 and its companion analog signal generator can be fed programming commands and file instructions directly by general-purpose interface bus (GPIB) [the 2029 has no user interface of its own]. The vector

1. The model 2029 vector modulator can be used with any analog signal generator operating from 0.8 to 2.5 GHz to provide digitally modulated signals for communications system and component testing.

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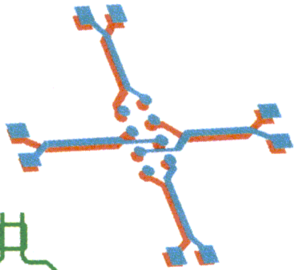

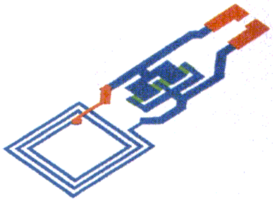

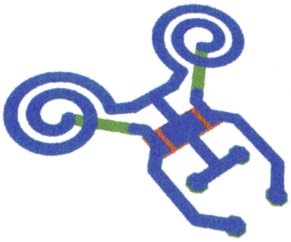

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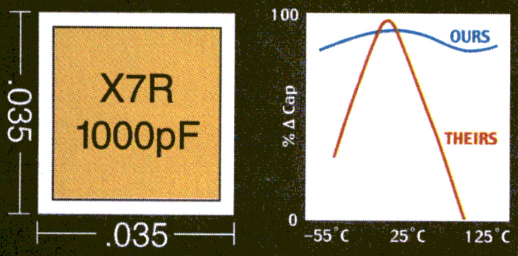

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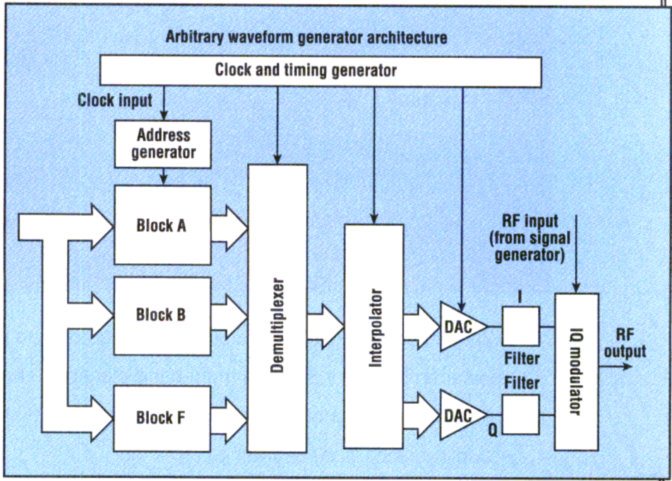

2. Files representing complex waveforms are processed through an interpolator and 14-b DACs to increase waveform precision and reduce noise.

modulator is shipped with software that produces a soft front panel on the display screen of a GPIB-compatible computer; the software requires Windows 95, 98, or NT. Files representing complex waveforms can also be stored in the 2029's internal memory in order to generate complex sequences of modulated waveforms. Waveform files can be created with standard software tools, such as MathCAD from MathSoft (Cambridge, MA). In addition, IFR offers files from a library of waveforms. The 2029 vector modulator is also equipped with an RS-232 port for importing firmware and software upgrades.

The 2029's built-in AWG is equipped with 14-b digital-to-analog converters (DACs) operating between 35 and 66 MSamples/s. The 14-b resolution helps to minimize spurious levels. The AWG works with an interpolator to increase the effective sample rate of the DACs, thereby exceeding the sample rates of the stored waveform files in the 2029. The interpolator can increase the sampling rate by factors of 1, 2, 3, 4, 6, 8, or 12.

The 2029 is supplied with an AWG packager and loader that can be accessed by the vector modulator's soft front panel. The packager configures source files into a format readable by the AWG, with additional header information attached. The loader transfers the files into the AWG through GPIB.

Combined with an analog signal generator, the 2029 vector modulator achieves a noise floor of −138 dBc/Hz when modulated with an AWG file having a value of +0.2 VDC root mean square (RMS) of full scale. The noise floor improves with increasing drive levels. The 2029 exhibits adjacent-channel power (ACP) of typically −70 dBm for an IS-95 pilot-channel file with a crest factor of 5.5 dB.

For those in need of digitally modulated test signals, who already have an analog signal generator and have considered the price of a stand-alone digital signal generator, the 2029 vector modulator may represent a practical alternative. **IFR, 10200 West York St., Wichita, KS 67215; (800) 835-2352, (316) 522-4981. FAX: (316) 524-2623, e-mail: infor@ifrsys.com, Internet: http://www. ifrsys.com.**

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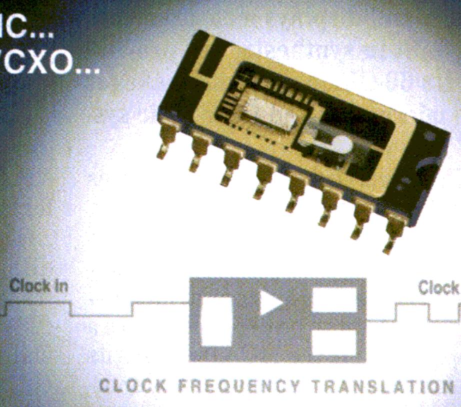

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CIRCLE NO. 239

Communications Analyzer Boasts Versatility And High Bandwidth

This analyzer offers a unique combination of versatility and high bandwidth for characterizing telecommunications and high-speed logic signals.

DON KELLER
Senior Editor

DUE to the complex nature of microwave signals, the instruments that are used to measure these signals tend to be highly specialized. Three of the most common types of measurement approaches are: viewing signals with an oscilloscope, measuring time and frequency response with a spectrum analyzer, and characterizing network performance with a vector network analyzer (VNA). LeCroy Corp. (Chestnut Ridge, NY) has introduced a communications analyzer that can perform all three of these measurements. The MCA1060 communications analyzer operates in the time and frequency domains, and displays measurements in rectangular and polar modes. It has a built-in synthesizer to provide stimuli to devices and circuits under test. Its 60-GHz bandwidth makes it suitable for testing virtually all of today's microwave and fiber-optic systems (see figure).

The MCA1060 can perform signal measurements in the time-domain, frequency-domain, and time-delay-reflectometry (TDR) modes. Display modes include linear, log magnitude, Smith, polar, group delay, and eye diagram. In the time domain, it measures time delay, rise and fall times, overshoot and undershoot, pulse width, amplitude, and eye diagrams. Frequency-domain measurements include spectrum analysis, log magnitude, group and phase delay, and fully error-corrected transmission and return loss. The analyzer's 60-GHz bandwidth allows it to characterize fiber-telecom products such as optical receivers (Rxs); modulators; ultra-high-speed logic blocks; and microwave and millimeter-wave components, such as mixers, filters, and amplifiers. A source synthesizer generates signals to stimulate devices under test (DUTs), while a receive synthesizer initiates sampling events.

The analyzer takes a novel approach to sampling. Instead of using a traditional triggering system, which requires an event to be detected before a sample is taken, the MCA1060 uses its internal receive synthesizer to generate narrow impulses which initiate sampling events. When a high-speed, repetitive signal is fed into either of the analyzer's two inputs, the synthesizer generates sequential impulses that trigger sequential samples. The impulse-repetition rate can be adjusted from 6.25 MHz to 1.6 GHz. This internal approach to triggering eliminates trigger-bandwidth limitations and trigger-jitter effects, and obviates clock regeneration. The samples accumulate to create a signal record, which can be displayed on the analyzer's 6.40-in. (16.26-cm), color, thin-film-transistor (TFT) screen or an external video-graphics-array (VGA) monitor and stored on the built-in floppy-disk drive for later analysis.

The analyzer has a noise floor of −80 dBm (−120 dBm with averaging), and its maximum measurement input (CW) is 0 dBm. DC drift is 2 mV/°C and system phase noise is <2 ps (without averaging). Frequency resolution is 0.0000014 Hz from 6.25 MHz to 1.6 MHz, and 0.0000525 Hz at 60 GHz. Channel-display modes include channel A or B, or the ratio, sum, difference, or X versus Y of both channels. The source synthesizer has three outputs: comb, emitter-coupled logic (ECL), and TTL. Other outputs include a D15 color VGA connector and a general-purpose interface bus. **LeCroy Corp., 700 Chestnut Ridge Rd., Chestnut Ridge, NY 10977-6499; (914) 425-2000, Internet: http://www.lecroy.com.**

CIRCLE NO. 53 or visit www.mwrf.com

The MCA1060 communications analyzer has a 6.40-in. (16.26-cm) color screen and a built-in floppy-disk drive.

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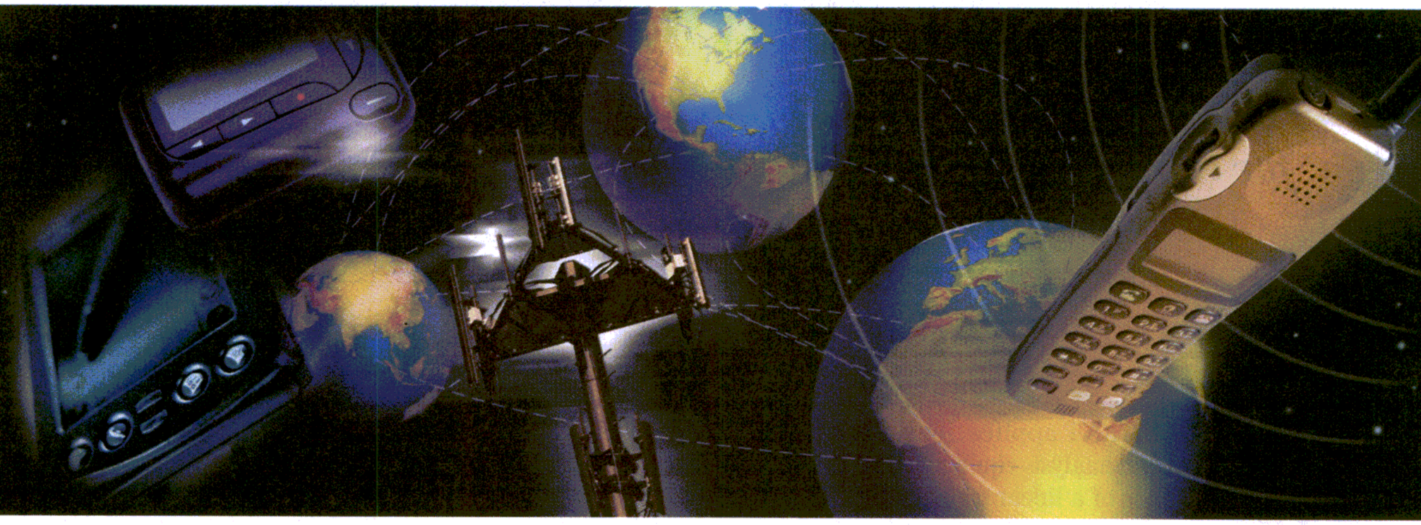

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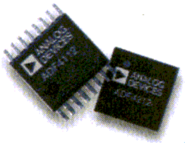

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These rugged high-gain, high-power amplifiers are suitable for communications component and system testing as well as in medical and automotive applications.

JACK BROWNE
Publisher/Editor

MEASUREMENT applications usually start by specifying signal sources and analysis equipment. But in assembling a test system, the importance of a linear measurement amplifier is often overlooked. The S series of linear amplifiers (see table) from Amplifier Research (Souderton, PA) succeeds where most amplifiers fail—they boost the level of test signals with minimal effect on distortion and noise. These are true Class A linear amplifiers, covering ranges from 0.8 to 4.2 GHz, and do not skimp on power, delivering 100 percent of their rated power levels (with choices from 1 to 200 W) into 50-Ω loads. The amplifiers are designed to faithfully reproduce amplitude modulation (AM), frequency modulation (FM), and pulse modulation, and exhibit high third-order intercept points (IP3s) for handling wide-dynamic-range signals.

The S series amplifiers have been available for some time, but are often an afterthought when assembling a measurement system. With the growing importance of high-power measurements, testing for linearity through intermodulation-distortion (IMD) measurements, and evaluating electromagnetic interference (EMI) and electromagnetic compatibility (EMC), the addition of amplifiers such as the S series to a test rig becomes nearly as important as the selection of a signal generator.

The S series amplifiers at a glance

Model	Frequency range (GHz)	Output power (W)	Maximum gain (dB)
1S1G4A	0.8 to 4.2	1	30
5S1G4A	0.8 to 4.2	5	37
10S1G4A	0.8 to 4.2	10	40
15S1G3	0.8 to 3.0	15	42
25S1G4A	0.8 to 4.2	25	44
30S1G3	0.8 to 3.0	30	45
50S1G4A	0.8 to 4.2	50	47
60S1G3	0.8 to 3.0	60	48
100S1G4	0.8 to 4.2	100	50
120S1G3	0.8 to 3.0	120	51
200S1G4	0.8 to 4.2	200	53

The S series of amplifiers currently includes 11 models, from the diminutive model 1S1G4A, with minimum rated output power of 1 W from 0.8 to 4.2 GHz, to the 400-lb. (280-kg) model 200S1G4, with 200-W minimum output power from 0.8 to 4.2 GHz. In between, are rack-mountable amplifiers ranging from 5-to-120-W output power, all specified at minimum rated output power, 1-dB compression, and 3-dB compression. The amplifiers output rated power levels when driving less-than-ideal loads without any foldback, and will not suffer damage or oscillation with any magnitude or phase of source or load impedance.

These are high-gain amplifiers, requiring only 1-mW input to achieve their rated output-power levels. The minimum gain for the model 200S1G4, for example, is 53 dB. Even for the smallest amplifier, model 1S1G4A, the minimum gain is 30 dB. Since maximum output-power levels may not always be needed, each of the amplifiers includes a minimum gain-adjustment range of 10 dB (in 4096 steps), except for the model 180S1G4, 120S1G4, and 200S1G4 which feature a 15-dB minimum gain-adjustment range.

The amplifiers provide output power that is almost ruler flat across their full frequency ranges, with typical output-power flatness of ±1.5 dB for all models except the 200S1G4, which delivers worst-case output-power flatness of ±2.5 dB. Dynamic-range performance is solid, with IP3s that range from a low of typically +39 dBm in the 1-W model 1S1G4A to typically better than +60 dBm in the higher-power models. The amplifiers boast spurious levels that are −73 dBc or better, and a phase linearity of typically ±1 deg./100 MHz of bandwidth.

Most of the amplifiers (except the two lowest-power units) are equipped with remote capability, through general-purpose-interface-bus (GPIB) and RS-232C connectors. The amplifiers are all equipped with type-N female input and output connectors or optional rear-panel connectors and internal fans for cooling. **Amplifier Research, 160 School House Rd., Souderton, PA 18964-9990; (215) 723-8181, FAX: (215) 723-5688, Internet: http://www.ar-amps.com.**

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	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

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CIRCLE NO. 268

Embedded Antenna Simplifies Integration In WLAN Devices

This tiny D-shaped antenna supports dual-polarization designs in 2.4-GHz WLAN terminals and portable devices and can be readily adapted to Bluetooth use.

JACK BROWNE
Publisher/Editor

NETWORK access is often critical in business and commerce. The growing acceptance of the IEEE's 802.11b standard for wireless local-area networks (WLANs) indicates the desire for mobility within the network by some network users. But in order to develop WLAN devices that are truly portable, the antenna must shrink along with the RF and supporting electronics. One possible solution for the antenna in portable WLAN devices is a new embedded antenna developed by RangeStar Wireless (Aptos, CA). The antenna is small enough to be handled by automatic assembly equipment but provides enough gain at 2.4 GHz to preserve the integrity of high-speed data streams through 11 Mb/s.

Intended for embedded applications, this miniature D-shaped antenna can be used in handheld terminals for WLANs and Bluetooth.

The miniature antenna (see figure) is suitable not only for WLAN applications but for in-building Internet access as well. With minor modifications, it can also be adapted to Bluetooth devices operating in the unlicensed industrial-scientific-medical (ISM) band at 2.4 GHz. The embedded antenna provides peak gain of 4 dBi from 2.40 to 2.45 GHz. It operates with 2.50:1 VSWR. The tiny D-shaped antenna measures 16 mm in diameter and 6.25 mm in height. It weighs only 1 g.

The design is small enough to permit the mounting of two antennas within a device in a diversity application. In these cases, the WLAN receiver switches constantly between the two antennas and selects the highest-level signal from either antenna. In that way, the receiver achieves an improvement in effective signal-to-noise ratio (SNR) and bit-error rate (BER) compared to the use of a single antenna.

The antenna is small enough to provide in-building wireless access wherever needed. According to John Harris, President and CEO of RangeStar Wireless, "Wireless Internet applications clearly represent the future for wireless, and RangeStar's newest embedded antenna will help improve the consumer's wireless Internet experience by improving the quality and durability of wireless modem applications."

The multipurpose embedded antenna embodies horizontal and vertical polarization characteristics. As a result, it can be installed and used in a variety of orientations, including flush-mounted or side-mounted on printed-circuit boards (PCBs), providing WLAN and Internet-device designers with a great deal of flexibility in their mechanical configurations without necessarily sacrificing performance under conditions of multipath interference.

The antenna's injection-molded plastic package is designed for use with automatic assembly equipment or continuous tape-and-reel handlers. The antenna can be handled and assembled on a PCB in similar fashion to surface-mount-technology (SMT) components.

The company, founded in 1995, manufactures a variety of other compact antennas for WLAN, cellular, and personal-communications-services (PCS) bands. A WLAN antenna with square footprint, for example, measuring 12.7 × 12.7 × 0.8 mm provides 1-dBi nominal gain from 2.40 to 2.45 GHz with less than 1.80:1 VSWR. It weighs less than 1 g, but handles up to 10-W CW power. More information on the WLAN antennas and other capabilities are detailed in the company's website. **RangeStar Wireless, 9565 Soquel Dr., Aptos, CA 95003-4153; (831) 661-4200, FAX: (831) 661-4201, Internet: http://www.rangestar.com.**

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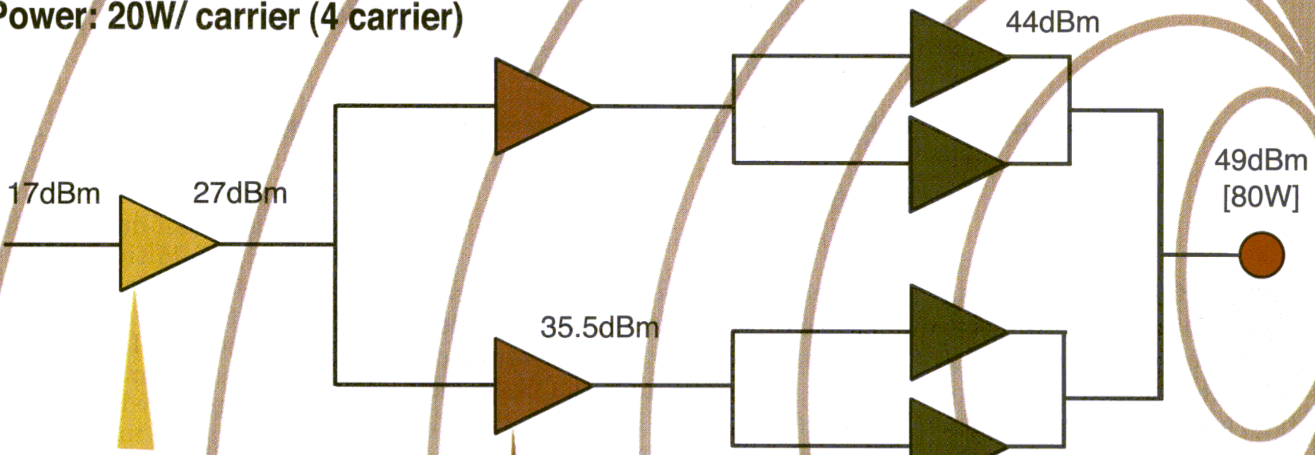
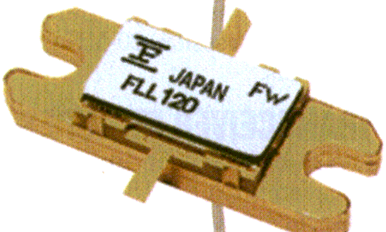
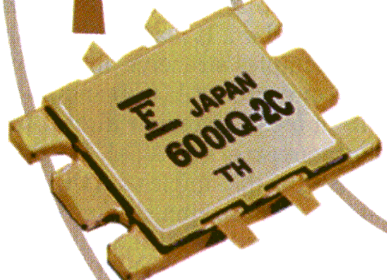




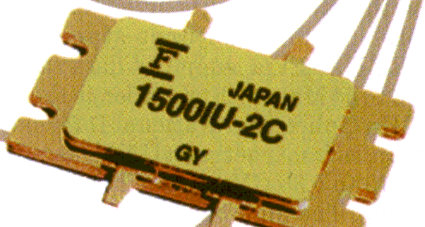
FLL120MK

FREQ. = 2.0 GHz (1)
P1dB = 40.0 dBm
G1dB = 10.0 dB
ηadd = 40%



FLL600IQ-2C

FREQ. = 2.17 GHz
Pout = 48.0 dBm
G1dB = 12.0 dB
ηadd = 51%



FLL1500IU-2C

FREQ. = 2.17 GHz
Pout = 51.8 dBm
G1dB = 12.0 dB
ηadd = 48%

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CIRCLE NO. 389

CMOS Yields High-Speed 5-GHz WLAN Chip Set

A pair of silicon CMOS ICs supports data rates to 54 Mb/s in the unlicensed 5-GHz UNII band.

JACK BROWNE
Publisher/Editor

SILICON (Si) just does not seem ready to run out of "gas." Several years ago, 2 GHz was a dividing line between applications for Si-device technologies and gallium arsenide (GaAs). With the introduction of the first integrated circuits (ICs) from Radiata (San Jose, CA), however, it is apparent that the dividing line will have to be raised. The chip set, which is fabricated entirely with Si complementary metal-oxide semiconductor (CMOS), is designed for 5-GHz wireless local-area networks (WLANs). The chips include a baseband modem and a radio transceiver.

WLANs are commonly centered in the 2.4-GHz band. After a long period of discussion, the industry has by and large adopted IEEE 802.11b as the standard for WLANs at 2.4 GHz, with data rates supported up to 11 Mb/s. But an adjunct standard, IEEE 802.11a, supports data rates up to 54 Mb/s with operation in the 5-GHz Unlicensed National Information Infrastructure (UNII) bands (5.15 to 5.35 GHz and 5.725 to 5.825 GHz). The high-data-rate standard uses a version of orthogonal frequency-division multiplexing (OFDM) specifically designed to enable large data capacity in indoor networks.

Radiata calls their pair of 5-GHz WLAN ICs the "wireless engine," since the two chips are meant to form early core building blocks for WLANs and other wireless communications systems. The two chips are the model R-M11a baseband modem and the model R-RF5 5-GHz radio transceiver.

The R-M11a is designed to implement all of the baseband functions of a 64-tone coded-OFDM (COFDM) system compliant with the IEEE 802.11a standard. It features dual 10-b, 80-MSamples/s receive analog-to-digital converters (ADCs) and 10-b, 80-MSamples/s transmit digital-to-analog converters (DACs). In addition, an 8-b, 5-MSamples/s ADC provides the digital signals required by the transceiver IC for the WLAN received-signal-strength-indicator (RSSI) function.

The 5-GHz CMOS modem includes a 4:1 interpolating finite-impulse-response (FIR) digital lowpass transmit filter, a 4:1 interpolating FIR digital lowpass receive filter, an IEEE 801.11a interleaver, a payload scrambler/descrambler, a transmit preprocessor for spectrum shaping and amplitude control, and a receive-channel equalizer.

The +2.5-VDC model R-M11a is fabricated with a 0.25-µm process. It has a single 80-MHz clock input port in addition to supporting data rates to 54 Mb/s. It is housed in a 208-pin chip-array-ball-grid-array (CABGA) package.

The companion model R-RF5 is a fully integrated half-duplex 5-GHz transceiver. In addition to IEEE 802.11a, it is suitable for HiperLan 2, European Telecommunications Standards Institute (ETSI) Broadband Radio Access Networks (BRAN), and MMAC high-speed WLAN applications. It connects to the R-M11a modem through analog, differential (for noise suppression) 20-MHz baseband signals.

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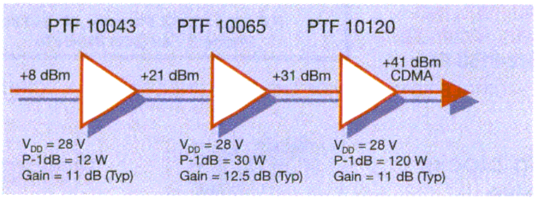

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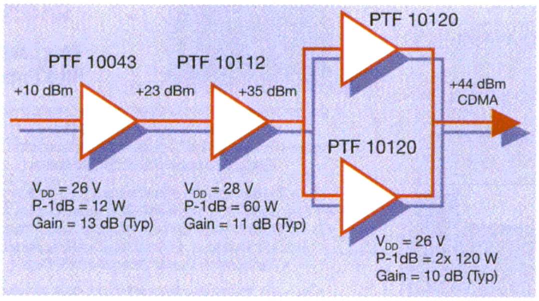

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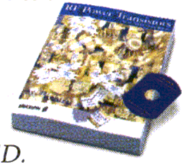

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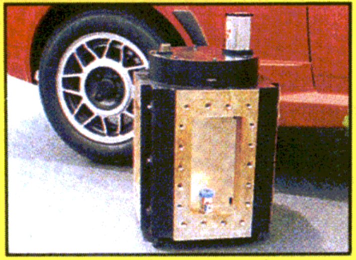

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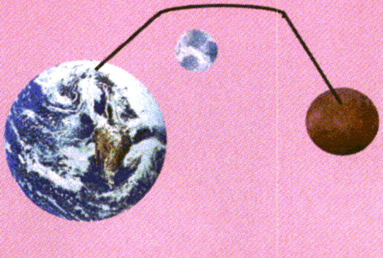

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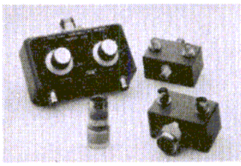

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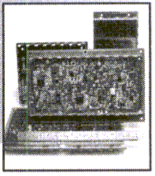

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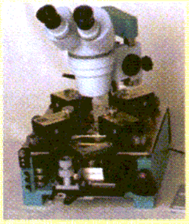

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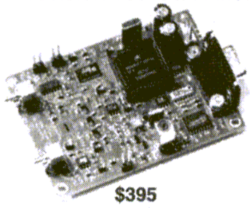

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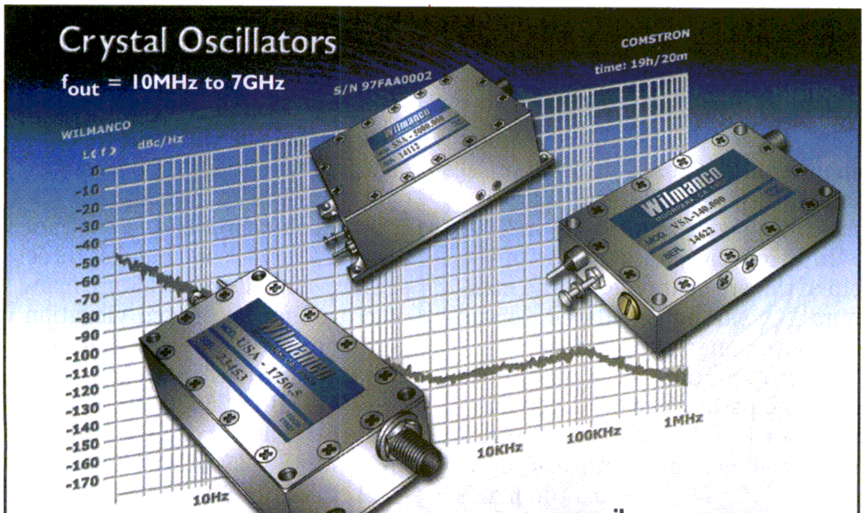

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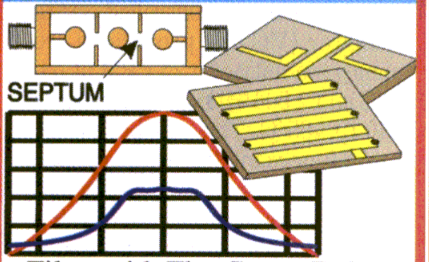

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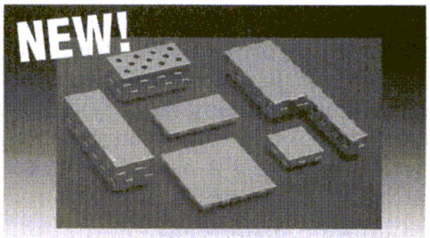

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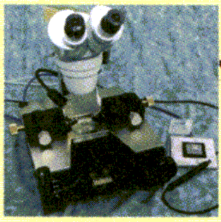

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(continued from p. 61)

31-mil-thick substrate. Standard 50-Ω loads are connected to the coupler-isolated ports. The amplifier is compact and easy to assemble. The total dimensions of the amplifier are 13.2 × 6.0 cm^2.

TEST RESULTS

The source and load impedances presented by the circuit to the device directly from simulation were measured versus frequency with a vector network analyzer (VNA). The experimental results showed good agreement between the measured and simulated impedances (Table 2). However, a slight tuning was performed with the use of a VNA on the input- and output-matching circuits to obtain the exact simulated impedance values.

The amplifier was tested in a 50-Ω system with fixed broadband tuning corresponding to an optimum IMD3 performance in CEL's high-power automated setup.[7] The device was biased at V_{ds} = +10 VDC and I_{dsq} = 12 A. Figure 2 shows the P1dB and G1dB performance versus frequency. Figure 3 provides power-added efficiency (PAE) and I_{ds} versus frequency at 1-dB gain compression. It shows that the amplifier exhibits a typical PAE of 40 percent from 2.4 to 2.7 GHz.

Figure 4 shows the IMD performance of the amplifier. These curves show that the amplifier has good IMD performance with an IMD3 lower than −42 dBc at +40-dBm output power, each tone across the 2.5-to-2.7-GHz MMDS bandwidth. P_{out} and I_{ds} versus P_{in} and frequency are shown in Fig. 5.

The amplifier-input return loss over the 2.5-to-2.7-GHz bandwidth was better than 15 dB, which is better than the expected return loss for a push-pull configuration. The amplifier's IMD performance was measured with fixed tuning at a constant, V_{ds} = +10 VDC, and at constant frequency of 2.7 GHz, versus P_{in} and I_{dsq}. Figures 6, 7, and 8 show respectively IMD3, IMD5, and IMD7. The curves show, as expected, that this amplifier exhibits the best IMD3 performance at any P_{out} level when biased at the highest I_{dsq}, 12 A. The maximum I_{dsq} is limited only at V_{ds} = +10 VDC by the maximum recommended channel temperature which is 150°C.

The mean time to failure (MTTF) of GaAs power devices is limited by their channel temperature. It is important for amplifier designers to be able to calculate this temperature for their applications from the device power dissipated, the case temperature, and its thermal resistance (R_{th}). This calculation is not as simple as it may appear since the R_{th} of GaAs devices is a strong function of the device-flange temperature and the channel temperature, or the power dissipated. The NES2427P-60 data sheet and CEL's application note[8] AN1032 give the information necessary to calculate Rth versus the flange and channel temperatures, or power dissipated.

The NES2427P-60 data sheet indicates R_{th} = 0.76°K/W maximum for T_f = 25°C, V_{ds} = +10 VDC, I_{ds} = 12 A and recommends a maximum channel temperature of 150°C. This temperature corresponds to a MTTF of 2.4 × 10^6 h. From these data and with the help of the application note, the maximum I_{dsq} can be calculated versus the device-flange temperature.

As an example, for a maximum flange temperature of 52°C, the maximum power dissipated corresponding to a channel temperature of 150°C is $P_{diss.}$ = 120 W. It means the maximum quiescent I_{dsq} for V_{ds} = +10 VDC and a flange temperature of 52°C is 12 A.

This current is relatively high, but it is lower than half of the device-saturated drain current (I_{dss}), which is 36 A typical. If the standard definition of class A for power devices is used, I_{dsq} = 12 A does not correspond to this device for a Class A operation but to a Class A-B one. ●●

References

1. NES2427P-60 data sheet, http://www.cel.com
2. L. Max, "Balanced Transistors: A New Option For RF Design," *MicroWaves*, June 1977.
3. L. Max, "Apply Wideband Techniques To Balanced Amplifiers," *MicroWaves*, July 1980.
4. G. Sarkissian, "An S-Band Push-Pull 60-Watt GaAs MESFET For MMDS Applications," *1997 IEEE MTT-S Digest*.
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6. D. Raicu, "Design of Planar, Single-Layer Microwave Baluns," *1998 IEEE MTT-S International Microwave Symposium Digest*.
7. R. Basset, "Automated Test Equipment For High Power Solid State Device Development And Characterization," *47th ARFTG Conference Digest*, June 1996.
8. Application note AN1032 "Microwave Power GaAs Device Thermal Resistance Basics," http://www.cel.com.

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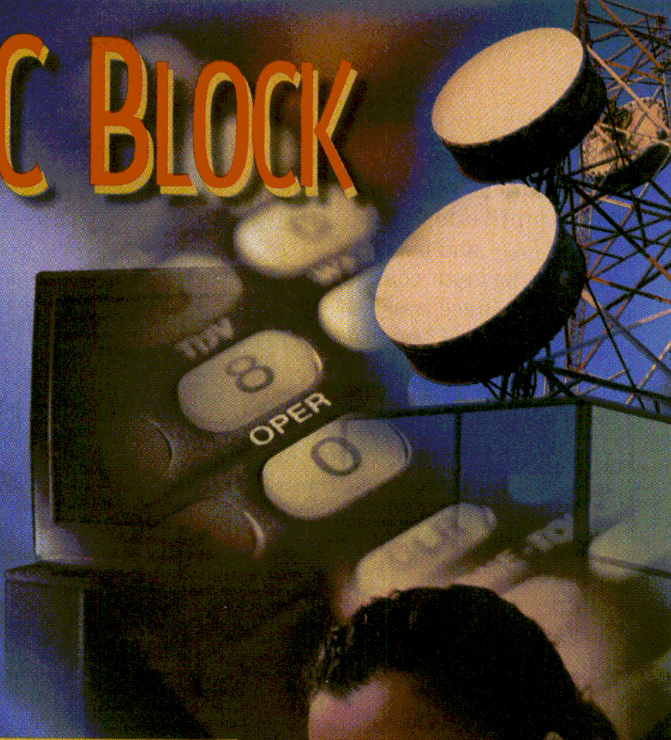

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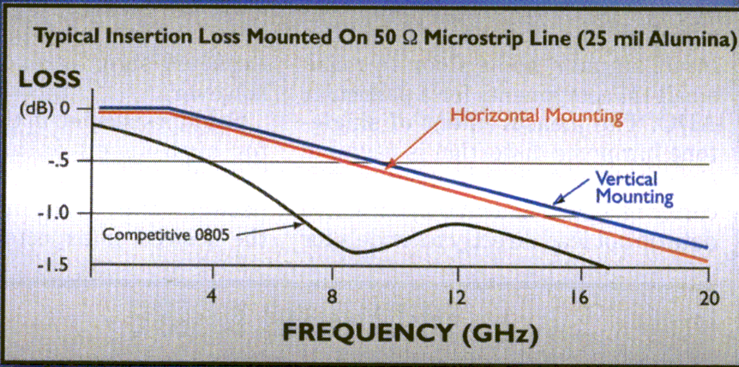

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(continued from p. 84)

TTIMD is generally a function of the amplitudes—A[f_{IN1}]$_{RMS}$ and A[f_{IN2}]$_{RMS}$ and frequencies (f_{IN1} and f_{IN2}) of the input components. Therefore, it is necessary to specify the input tones and amplitudes for which two-tone IMD measurements are performed. It is essential that the input test signal be virtually free of IM and harmonic distortion. For ADCs that have wide bandwidths and large dynamic ranges, this condition is increasingly difficult to achieve.

Two signal generators, containing output-leveling circuitry and linked through balanced or isolated outputs or any other coupling circuits, can easily generate IMD effects. In order to avoid IMD in the test signal, therefore, one should operate power splitters/combiners (used to combine or split two input tones) well within their linear range. Figure 2 depicts two-tone IMD with second- and third-order IMD products for a 10-b, 80-MSamples/s ADC. For best results, the two-tone envelope for this ADC was chosen to be −0.5-dB FS, and the amplitude for the two input tones was normalized to −6.5-dB FS.

Multi-tone IMD tests are often used in system design to determine limits for the signal dynamic range, useful frequency bands for different signal groups, and where to set the input signal's noise floor to mask small IM components for a particular ADC. The measurement of single-tone harmonic distortion is useful in obtaining general ideas about the linearity of a particular ADC, but these data do not lead directly to models for predicting useful measures of IM performance for independent input signal tones.

MEASURING NPR

Noise power ratio (NPR) defines the spectral power of contributed errors such as IMD and THD, in a small frequency band within the baseband of the composite input signal being analyzed.

For this test, one generates random noise whose spectrum is approximately uniform up to a predetermined cutoff frequency less than half the sampling frequency. Then, a notch filter removes a narrow band of frequencies from the noise. To improve the measurement, the notch depth is recommended to be at least 10 to 15 dB greater than the NPR value being measured. Compared to the overall noise bandwidth, the notch width should be narrow. With this notched noise applied to the ADC input, one computes the frequency spectrum of the resulting code sequence, and then calculates NPR as the ratio (in decibels) of the average power-spectral density inside the notched frequency band to that outside of the notched band.

TEST PROCEDURE

A typical test procedure features a computer-controlled DAC that generates a signal composed of a set of sine waves at DFT binary center frequencies. As the tone amplitudes are increased uniformly, beginning at the noise floor and continuing to the full-scale ADC level at which clipping begins, gaps between the tones serve as observation points to analyze any resulting IMD. These tests provide results similar to that of the noise-power-ratio (NPR) test (see the sidebar "Measuring NPR"). They support better simulation of the expected signal-group waveforms, however.

Seldom specified in the data sheets for high-speed data converters, VSWR is the ratio of mismatch between the actual impedance and the desired or expected impedance. It can be calculated by applying a test signal and measuring the reflection coefficient of the ADC input terminal. Calculated by the following, VSWR is directly related to the reflection coefficient, ρ, of a simple terminating impedance, Z_T:

$$VSWR = (1+|\rho|)/(1-|\rho|),\ where$$
$$\rho = (Z_T - Z_o)/(Z_T + Z_o). \quad (12)$$

where:

Z_T = the ADC input termination impedance, and

Z_O = the transmission line impedance (nominally 50 Ω).

To compensate for circuit inaccuracies in the measurement, it is recommended to use calibration standards if available (typically short, open, and 50 Ω).

In addition to the test setup information, Part 2 will provide samples of source code based on the MATLAB (from the MathWorks, Natick, MA) and LabWindows/CVI (from National Instruments, Austin, TX) software tools. The software will enable designers to analyze the dynamic performance of an ADC by capturing data records quickly and processing them efficiently. ●●

DECODING NOISE

The term "noise" is rather ambiguous if not qualified as to type. In general, it includes the effects of nonlinearities [such as integral nonlinearities (INLs) and differential nonlinearities (DNLs)], random and fixed-pattern effects, and sampling-time error. The total noise ($A_{TOTAL\ NOISE}$[RMS]) is any deviation of the output signal (converted to input units) from the input signal, excluding deviations caused by differential gain and phase errors, or DC-level shifts. Notable examples of these effects defined here as noise, include quantization error, harmonic and intermodulation distortion (IMD), and spurious distortion.

For Further Reading

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D. Johns and K. Martin, *Analog Integrated Circuit Design,* Wiley, New York, 1997.

R. van de Plasche, *Integrated Analog-to-Digital and Digital-to-Analog Converters,* Kluwer Academic Publishers, 1994.

Engineering Staff of Analog Devices, Inc., *Analog-Digital Conversion Handbook,* PTR Prentice-Hall, Upper Saddle River, NJ, 1986.

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CIRCLE NO. 387

Switch operates from DC to 2.5 GHz

The model SWN-218-2DR standard option DC205 HPR5W is a high-power (usable to 10-W) reflective single-pole, double-throw (SPDT) transmit/receive solid-state switch that operates from DC to 2.5 GHz (usable to 4 GHz). With an insertion loss of less than 0.75 dB at 40 MHz to less than 1 dB at 2.5 GHz, the isolation is >50 dB at 40 MHz and >20 dB at 2.5 GHz. Amplitude and phase are matched to within ±0.1 dB and ±1°, respectively, between port to port from 40 MHz to 2.5 GHz. VSWR is 1.5:1 typical, while delay on and off is less than 40 ns. Supply voltage is −5 VDC at 8 mA. **American Microwave Corp., 7311 G Grove Rd., Frederick, MD 21704; (301) 662-4700, FAX: (301) 662-4938, e-mail: amcpmi@aol.com, Internet: http://www.amwave.com.**

CIRCLE NO. 65 or visit www.mwrf.com

Amplifiers run from DC to 8 GHz

The NGA-x86 series of broadband monolithic-microwave-integrated-circuit (MMIC) amplifiers suits applications in the DC-to-8-GHz range. These devices were designed with indium-gallium-phosphorus/gallium-arsenide (InGaP/GaAs) process technology. The NGA-1 through NGA-6 offer small-signal gain ranging from 12 to 19 dB with output third-order intercept point (IP3) ranging from +27 to +39.5 dBm. **Stanford Microdevices, Inc., 522 Almanor Ave., Sunnyvale, CA 94086; (408) 616-5400, FAX: (408) 739-0970, Internet: http://www.standfordmicro.com.**

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Inductors reduce EMI

Available in current ratings up to 300 mA, the HK series of inductors is available from 1 to 120 nH (0402 case size), 1 to 220 nH (0603 case size), and 1.5 to 470 nH (0805 case size). The series features 100-percent silver (Ag) internal conductors, advanced dielectric ceramics, and unique fabrication techniques to achieve performance characteristics in small case sizes. Chief original-equipment-manufacturer (OEM) design benefits include surface mountability and small form factor. **Taiyo Yuden, Inc., 1930 Thoreau Dr., Suite 190, Schaumburg, IL 60173; (847) 925-0888, FAX: (847) 925-0899.**

CIRCLE NO. 67 or visit www.mwrf.com

Splitter covers 800 to 2200 MHz

The model DG-54FN unequal-way power splitter is designed to split a signal between two outputs in a 4:1 (80-percent/20-percent) ratio, covering all frequencies from 800 to 2200 MHz and providing DC continuity to both outputs with a VSWR of less than 1.25:1, passive intermodulation (PIM) is less than −140 dBc, and insertion loss is below 0.1 dB. The unit has no resistors to burn out, fewer solder joints, and reflects less power back to the transmitter (Tx). **Microlab/FXR, 10 Microlab Rd., Livingston, NJ 07039-1682; (973) 992-7700, (973) 992-0513, e-mail: sales@microlab.fxr.com, Internet: http://www.microlab.fxr.com.**

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Delay lines provide 3-dB bandwidth of 300 MHz

The TCD-B and TCD-M range of ultra-high-speed delay lines for wide-bandwidth applications provides a 3-dB bandwidth of 300 MHz. Amplitude roll of up to 100 MHz is less than 0.5 dB, and a return loss of greater than 20 dB is achieved, translating to minimal ringing with an applied rise time of 4 ns. Impedances of 50, 100, 120, or 150 Ω can be supplied, with delay times up to approximately 15 ns. The range is compatible with emitter-coupled-logic (ECL) high-speed logic circuits. **Faraday Technology Ltd., Croft Rd. Industrial Estate, Newcastle, Staffordshire, ST5 0QZ UK; +44 (0)1782 661501, FAX: +44 (0)1782 630101, e-mail: sales@faradaytech.co.uk, Internet: http://www.faradaytech.co.uk.**

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CATV amplifier targets 870-MHz systems

The ACA2401S7 is a +24-VDC cable-television (CATV) amplifier targeting 870-MHz CATV systems. The unit provides 4.5-dB maximum noise figure, 21.5-dB gain, and 18-dB minimum return loss. The device consists of two pairs of parallel amplifiers. A hybrid equivalent is created when one ACA2401S7 is cascaded between two transmission baluns. The amplifier is housed in a 16-pin small-outline integrated circuit (SOIC). **ANADIGICS, 35 Technology Dr., Warren, NJ 07059; (908) 668-5000, FAX: (908) 668-5132, Internet: http://www.anadigics.com.**

CIRCLE NO. 70 or visit www.mwrf.com

Telecom modulator spans −40 to +85°C

The VSC7991 is an OC-192 electroabsorption modulator (EAM)/laser-diode driver which spans −40 to +85°C. The VSC7991 operates at data rates to 10.7 Gb/s and +3 VDC peak-to-peak. Rise and fall times are less than 35 ps and the unit requires a single +6.5-VDC supply. The maximum current delivered to the load is 65 mA and the modulator dissipates less than 2 W even when supplying maximum modulation current to the load. **Vitesse Semiconductor Corp., 741 Calle Plano, Camarillo, CA 93012; (805) 388-3700, FAX: (805) 987-5896, Internet: http://www.vitesse.com.**

CIRCLE NO. 71 or visit www.mwrf.com

Switch ranges from 7.8 to 8.8 GHz

The model SFB-2DR-8F-AA870 is a dual-channel, high-isolation, low-loss switch-filter bank which operates from 7.8 to 8.8 GHz in two channels, with an isolation of 70 dB minimum. VSWR is typically 1.75:1 and channel rejection is 45 dB minimum at +450 MHz. The channel insertion loss is 4 dB maximum with an insertion-loss ripple of ≤0.5 dB peak-to-peak. With a switching speed of ≤1 μs, the RF power handling is +25 dBm maximum. Package size is 4.73 × 1.00 × 1.97 in. (12.01 × 2.54 × 5.00 cm). **Planar Filter Co., 7311-G Grove Rd., Frederick, MD 21704; (301) 662-7024, FAX: (301) 662-4938, e-mail: amcpmisales@aol.com, Internet: http://www.amwave.com.**

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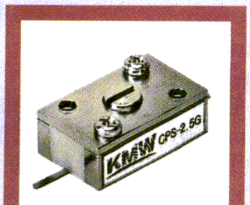
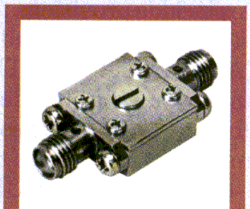
KPH90OSCL000

KPH90OSCL001

Standard Connectorized CPS

Product Code No.	A type : KPH90OSCL000 B type : KPH90OSCL001		
Frequency Range	~ 1GHz	1 ~ 2GHz	2 ~ 3GHz
Insertion Loss (Max.)	0.15dB	0.25dB	0.35dB
VSWR (Max.)	1.25:1	1.25:1	1.25:1
Incremental Phase Shift	90 degree min. @ 2GHz		
Electrical Delay	125 psec min.		
Nominal Impedance	50 ohm		
I/O Port Connector	SMA(F) / SMA(F)		
Average Power Handling	20W @ 2GHz		
Temperature Range	-30°C ~ +60°C		
Dimension (inch)	A type : 1.496*1.102*0.457 B type : 1.225*1.102*0.457		


KPH30OSCL000


KPH35OSCL000

Miniature CPS

Product Code No.	Drop-In type (KPH30OSCL000)			Connectorized type (KPH35OSCL000)		
Frequency Range	~ 1GHz	1 ~ 2GHz	2 ~ 2.5GHz	~ 1GHz	1 ~ 2GHz	2 ~ 3GHz
Insertion Loss (Max.)	0.15dB	0.25dB	0.35dB	0.15dB	0.25dB	0.35dB
VSWR (Max.)	1.3:1	1.3:1	1.3:1	1.25:1	1.25:1	1.25:1
Incremental Phase Shift	30 degree min. @ 2GHz			35 degree min. @ 2GHz		
Electrical Delay	41.7 psec min.			48.6 psec min.		
Nominal Impedance	50 ohm			50 ohm		
I/O Port Connector	Drop-In			SMA(F) / SMA(F)		
Average Power Handling	30W @ 2GHz			30W @ 2GHz		
Temperature Range	-30°C ~ +60°C			-30°C ~ +60°C		
Dimension (inch)	0.709*0.433*0.244			0.630*0.551*0.244		

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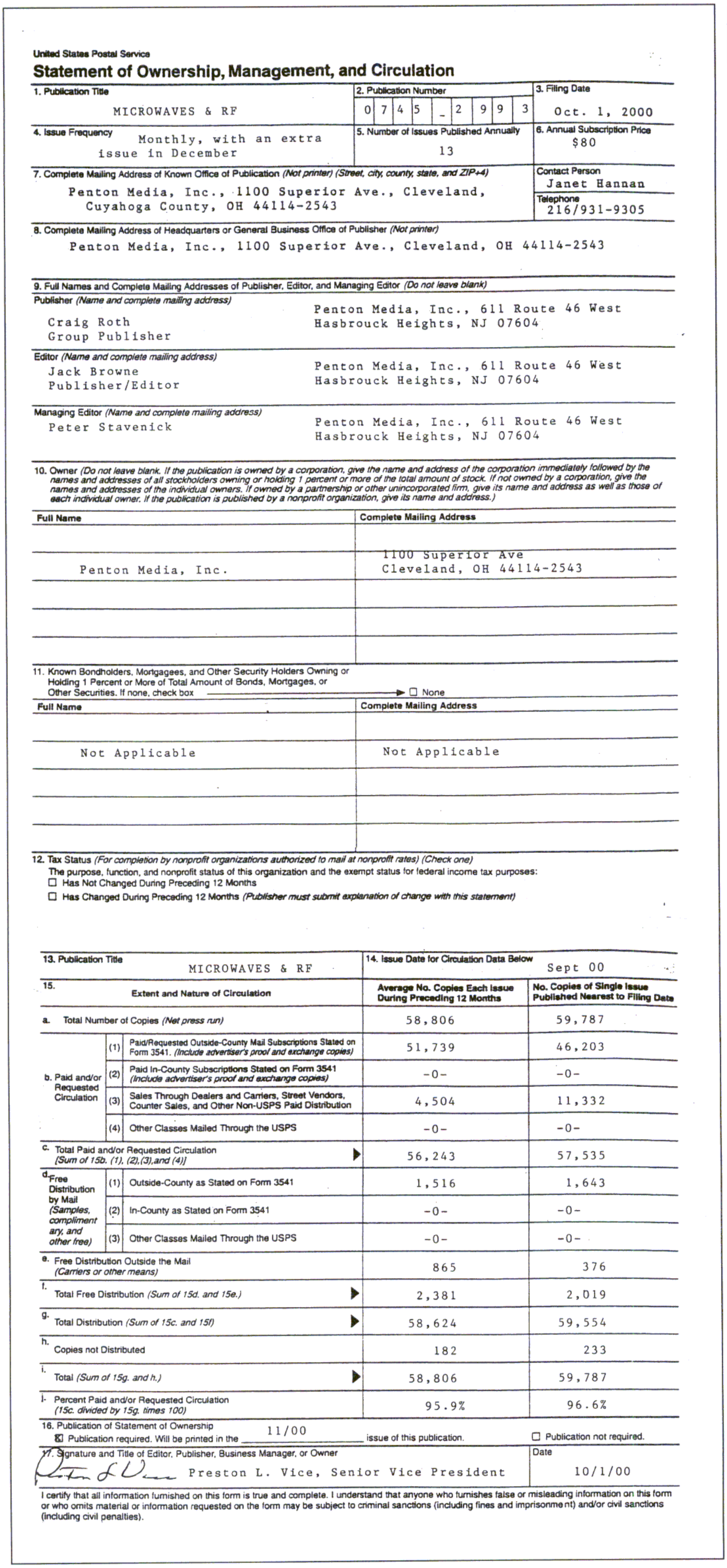

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The model DSC-R401ER is an amplified optical receiver (Rx) with typical −3-dB bandwidth of 100 kHz to 20 GHz. Applications include telecommunication system designers of OC-192/STM-64 systems using the return-to-zero (RZ) format with forward-error correction (FEC). **Discovery Semiconductors, Inc., 186 Princeton-Highstown Rd., Princeton Junction, NJ 08550; (609) 275-0011, FAX: (609) 275-4848, e-mail: sales@ chipsat.com, Internet: http://www.chipsat.com.**

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Filter operates from 2 to 8 GHz

Model MR1433DD is a multioctave band-reject filter that consists of eight stages covering the frequency range of 2 to 8 GHz. The rejection bandwidth ranges from 2 to 8 GHz. At 2 GHz, the 30-dB rejection bandwidth is 9 MHz, at 5 GHz the −50-dB rejection bandwidth is 30 MHz. The out-of-band insertion loss is typically 1.5 dB and the tuning sensitivity is typically 24 MHz/mA. **OMNIYIG, 3350 Scott Blvd., Bldg. 66, Santa Clara, CA 95054-3125; (408) 988-0843, FAX: (408) 727-1373, e-mail: Omniyig @ix.netcom.com, Internet: http:// www.Omniyig.com.**

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Tx modules suit telemetry

The model DR4000 series original-equipment-manufacturer (OEM) transmitter (Tx) modules for use in wireless data communications are designed for short-range wireless control and telemetry applications. The Tx modules include provisions for on-off-keyed (OOK) and amplitude-shift-keyed (ASK) modulation. The Tx module includes an amplifier-sequenced-hybrid (ASH) Tx plus configuration components in a ready-to-use printed-circuit-board (PCB) assembly. **RF Monolithics, Inc., 4347 Sigma Rd., Dallas, TX 75244; (800) 704-6079, (972) 448-3700, FAX: (972) 387-8148, Internet: http://www.rfm.com.**

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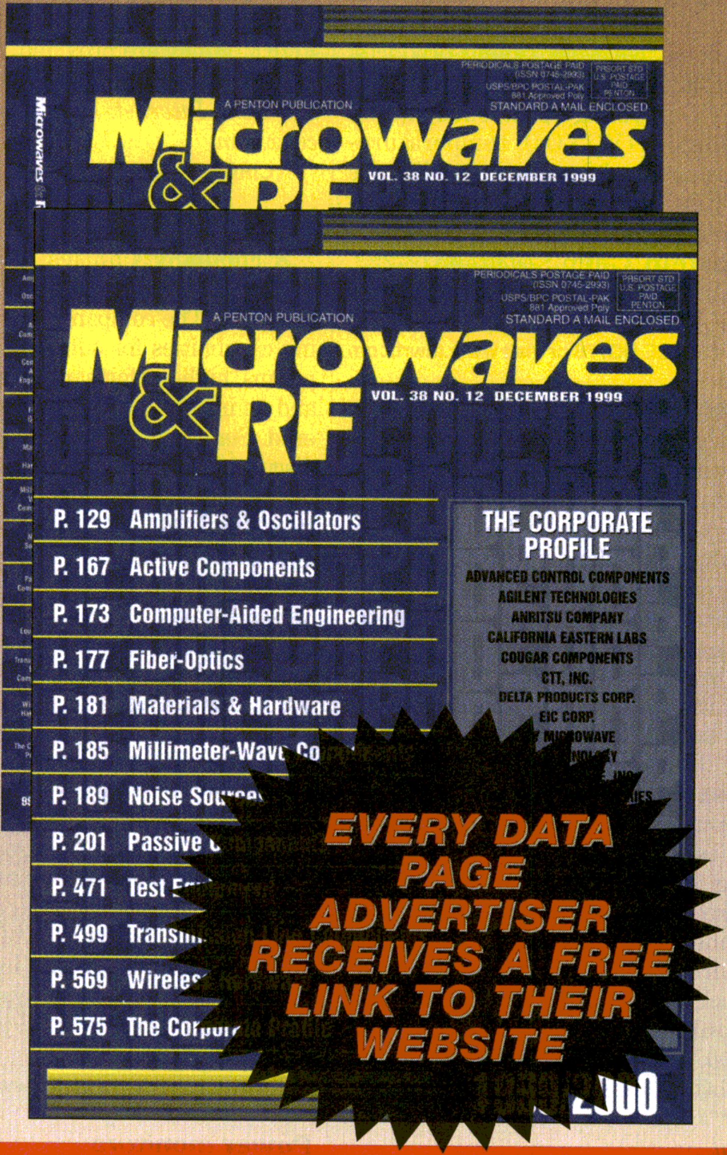

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Frequency control

Frequency-control product solutions are the subject of a six-page brochure. Network timing products, oven-controlled crystal oscillators (OCXOs), voltage-controlled crystal oscillators (VCXOs), temperature-compensated crystal oscillators (TCXOs), and voltage-controlled oscillators (VCOs) are presented. **Connor-Winfield Corp.;** *(630) 851-4722, FAX: (630) 851-5040, Internet: http://www.conwin.com.*

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Cable assemblies

High-performance flexible microwave coaxial-cable assemblies for military applications are discussed in a six-page brochure. Flexible polytetrafluoroethylene cable assemblies, microwave cable assemblies, and specialized interconnects are featured. **Kaman Aerospace Corp.;** *(719) 635-6954, FAX: (719) 634-8159, Internet: http://www.stablecable.com.*

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Broadband RF

Broadband RF products are the focus of a six-page foldout brochure. Readouts, RF-field probes, electromotive-force (EMF)-analysis software, and portable RF-survey systems are offered. Product specifications are included. **Holaday Industries, Inc.;** *(612) 934-4920, FAX: (612) 934-3604, e-mail: sales@holadayinc.com, Internet: http://www.holadayinc.com.*

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Microwave cables

A family of tin (Sn)-dipped, hand-formable microwave cables is highlighted in a six-page brochure. Specifications include insertion loss, shielding effectiveness, and power handling. Benefits of the 10 cables in the UTIFORM family are provided. **MICRO-COAX;** *(800) 223-2629, FAX: (610) 489-1103, Internet: http://www.micro-coax.com.*

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Precision optics

A company's optics-fabrication capabilities are described in a brochure. The company's manufacturing capabilities for substrates, lenses, prisms, high-performance laser optics, and an infrared (IR) detector are covered. Sensor and filter optics are also discussed. Specifications for each product category are included. **Meller Optics, Inc.;** *(401) 331-3717, FAX: (401) 331-0519, e-mail: sales@melleroptics.com, Internet: http://www.melleroptics.com.*

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Electronic test equipment is covered in a 50-page catalog. Oscilloscopes, software, generators, power supplies, probes, logic analyzers, data-acquisition (DAQ) units, signal generators, spectrum/network analyzers, meters, and monitors are specified. Calibrators, testers, AC power sources, DC electronic loads, and cables are offered. **TestEquity;** *(800) 758-3457, Internet: http://www.testequity.com.*

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Wireless components, networks, and instruments are overviewed in a 296-page catalog. Adapters, attenuators, VSWR/power monitors, circulators, couplers, custom assemblies, duplexers, electromechanical switches, and filters are offered. Isolators, local multipoint-distribution services (LMDS), oscillators, low-noise amplifiers (LNAs), monitors, power combiners/dividers, receiver (Rx) multicouplers, survey meters, and terminations are also presented. **Narda Microwave-West;** *(916) 351-4500, FAX: (916) 351-4550, e-mail: nardawest@L-3COM.com, Internet: http://www.nardamicrowave.com.*

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Power combiners

RF components are presented in a 14-page brochure. Adapters, cable assemblies, circulators, isolators, connectors, power combiners/dividers, and quadrature hybrids are covered. Specifications include frequency, isolation, phase balance, and VSWR. **RFcomps.com;** *(847) 926-9060, FAX: (847) 926-9061.*

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Semiconductors/transistors

Electronic components are offered in a 280-page catalog. Connectors, cable assemblies, integrated-circuit (IC) sockets, semiconductors, transistors, ICs, transistors, diodes, rectifiers, crystals, oscillators, inductors, and filters are provided. Capacitors, resistors, potentiometers, switches, relays, wire, cable, tools, solder, test equipment, batteries, fuses, transformers, power supplies, and liquid-crystal displays (LCDs) are specified. **Digi-Key Corp.;** *(800) 344-4539, (218) 681-6674, FAX: (218) 681-3380, Internet: http://www.digikey.com.*

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Filters and duplexers

Filter products and subassemblies ranging from 30 kHz to 40 GHz are featured in a 96-page catalog. Filters, tower-mounted amplifiers, directional couplers, power dividers, and delay-line assemblies are presented. Full-band and Global System for Mobile Communications (GSM) duplexers, a hybrid combiner, and a tower-top amplifier are specified. **FSY Microwave, Inc.;** *(410) 381-5700, FAX: (410) 381-0140, e-mail: sales@fsymicrowave.com, Internet: http://www.fsymicrowave.com.*

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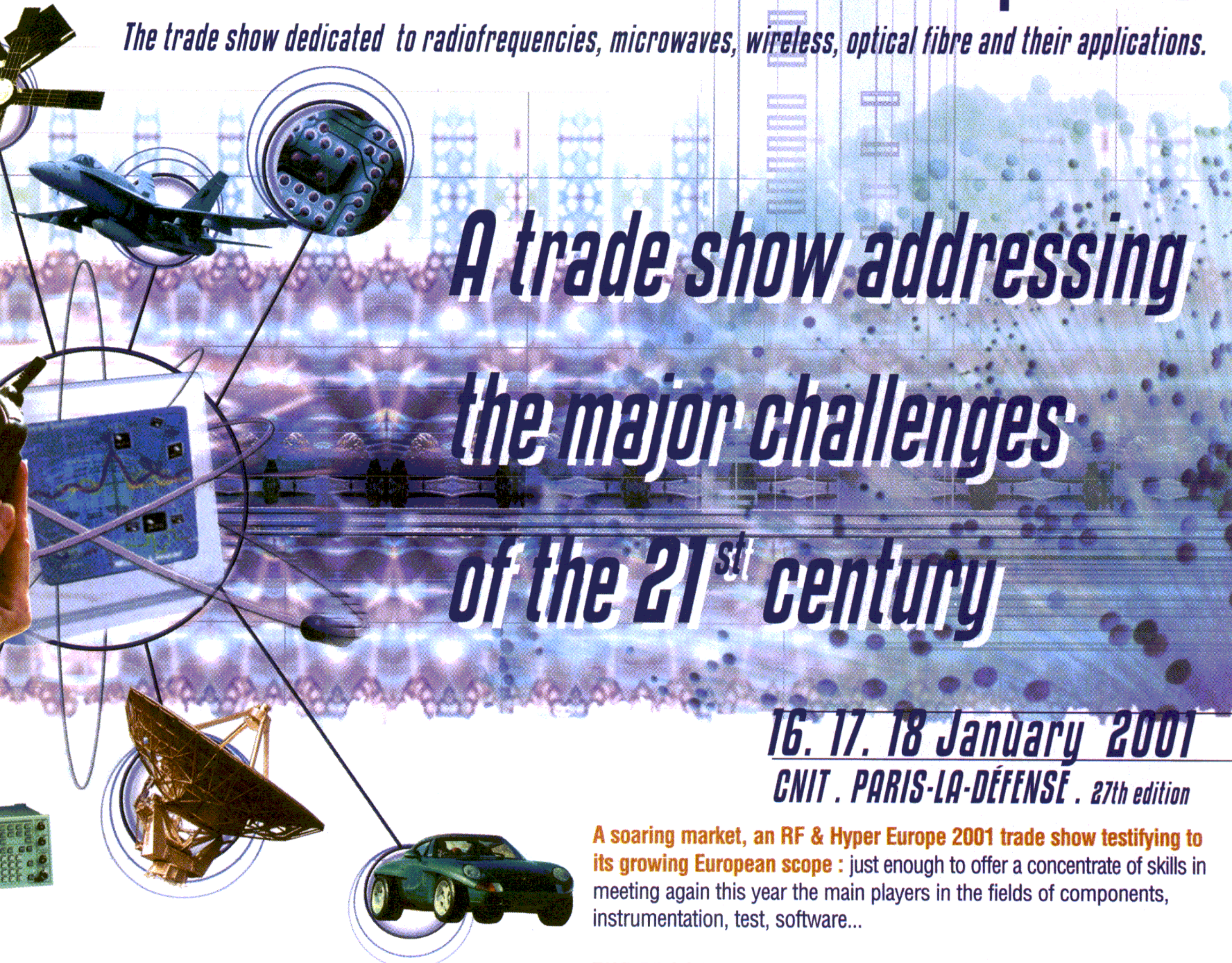

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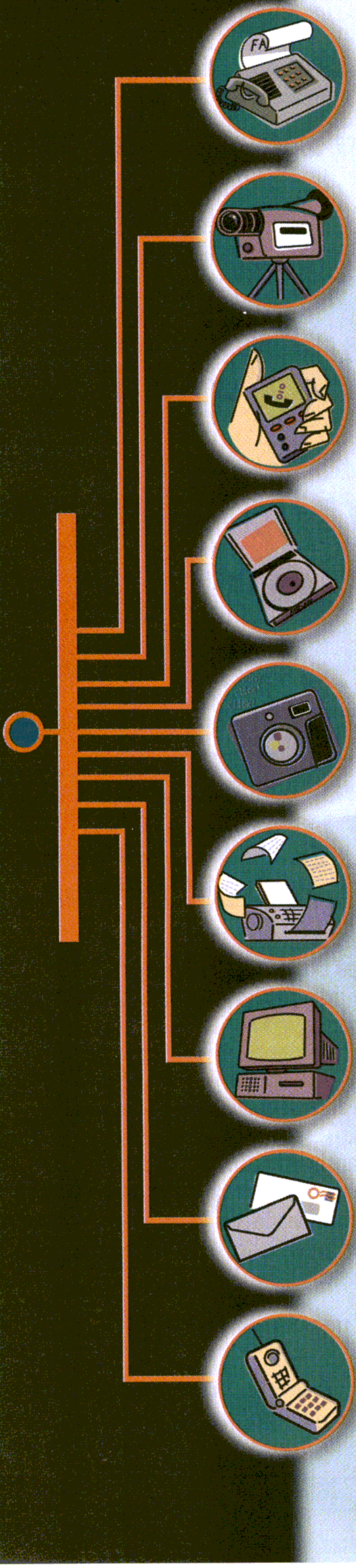

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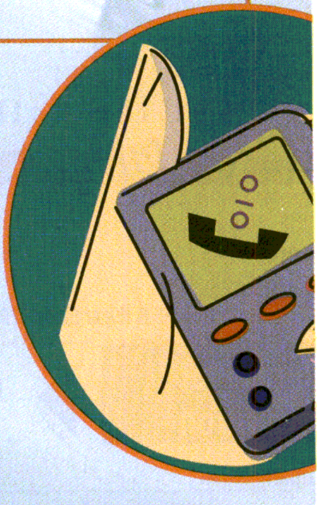

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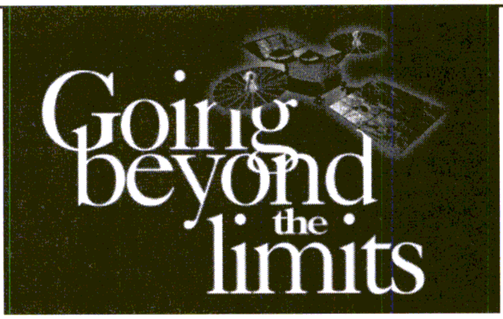

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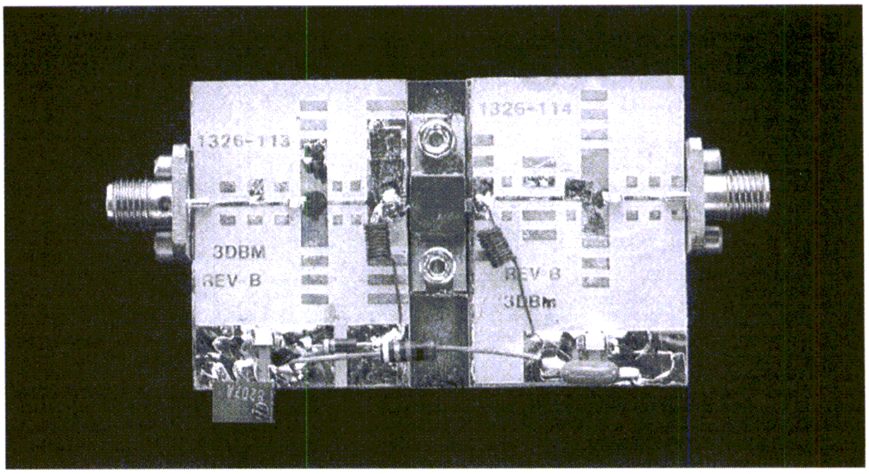

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Microwaves & RF December Editorial Preview

Issue Theme: Wireless Show Preview

News

Scheduled for February 12-16, 2001 at the San Jose Convention Center (San Jose, CA), the Wireless/Portable Symposium & Exhibition promises to be bigger than ever, with more technical presentations, more exhibitions, and more educational workshops than ever before. The December issue will feature summaries of the technical sessions, highlights of the workshops, and a sneak preview of the key products.

Design Features

December features a strong lineup of contributed technical articles, including a practical report from Rohde & Schwarz (Munich, Germany) on measuring the dynamic characteristics of cellular amplifiers with real-world modulated signals. Authors from China offer design notes on the construction of an integrated second-harmonic W-band Gunn oscillator, while an author from Taiwan provides insights into improving the precision of microwave measurements.

Product Technology

In December, a new approach to the design of RF and microwave mixers will be unveiled. Additional articles will detail high-power amplifiers for cellular and personal-communications-services (PCS) base stations, a line of compact integrated antennas that serves wireless data applications, and a wideband radio-channel simulator with maximum channel bandwidth of 35 MHz.

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Freq. Range (GHz)	Average Power (W)	Model No.	
		N Conn.	SMA Conn.
DC-18.0	1	N9412-*	9412-*
DC- 5.0	1	N4402-*	4401-*
DC- 4.0	5	N4405-*	4405-*
DC- 4.0	10	N4410-*	4410-*
DC- 4.0	25	N4425-*	4425-*
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*Value of attenuation

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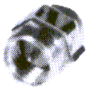

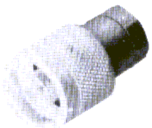

- DC-18 GHz model
- 50 W model
- SMA or Type N conn.

Freq. Range (GHz)	Average Power (W)	Model No.	
		N Conn.	SMA Conn.
DC-18.0	1		9512
DC-12.4	2	N9512	
DC-12.4	5	N9505	9505
DC-12.4	10	N9510	9510
DC- 8.0	25	N9525	9525
DC- 8.0	50	N9550	

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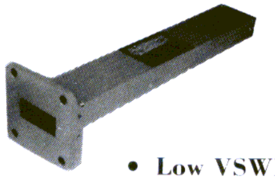

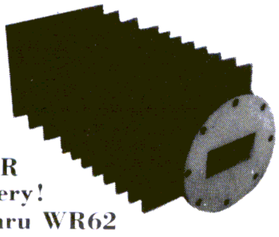

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Freq. Range (GHz)	Medium Power		High Power	
	Average (W)	Model No.	Average (W)	Model No.
2.60- 3.95	1200	284-925	4500	284-920
3.30- 4.90	1000	229-925	3000	229-920
3.95- 5.85	750	187-925	2000	187-920
4.90- 7.05	625	159-925	1500	159-920
5.85- 8.20	500	137-925	1000	137-920
7.05-10.0	425	112-925	600	112-920
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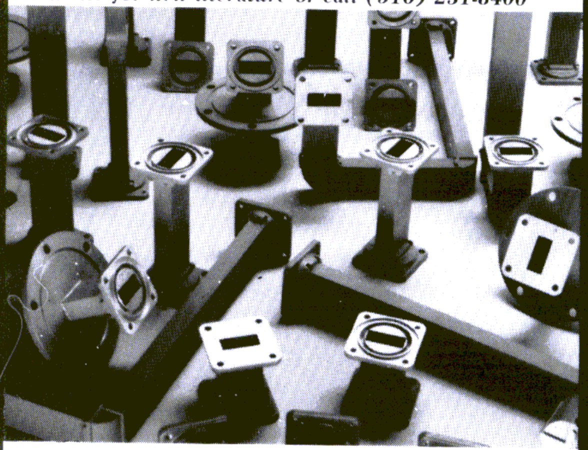

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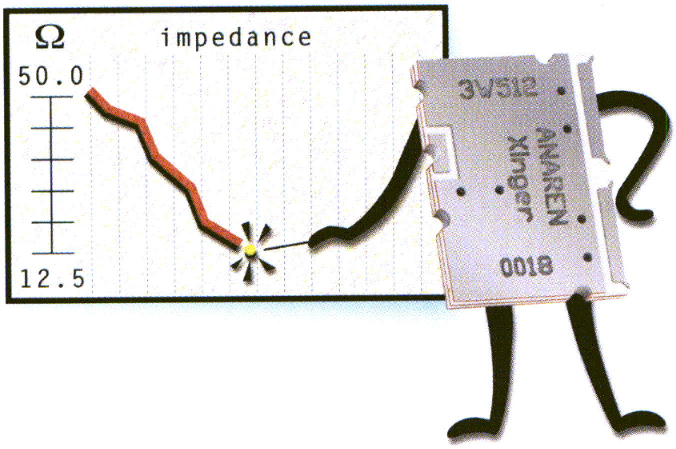

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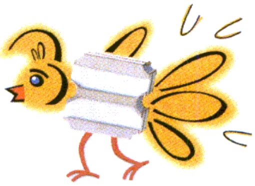
An infinitely smarter and safer way to cross the road.



How do you get to the other side? Cross any RF and DC line combination with the 0.20" x 0.20" surface mounted Xinger® Crossover.

- > NMT, GSM, UMTS, MMDS and HiperLAN applications
- > Frequency range of 0 to 6GHz
- > RF-DC and RF-RF components available
- > Insertion loss of less than 0.10dB; return loss of better than 20dB
- > RF-RF version maintains an isolation higher than 25dB
- > And our B-There™ service commitment and 100% On-Spec™ performance guarantee!

CIRCLE NO. 275

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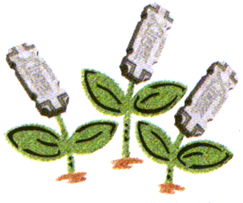
Look what just cropped up.



At only 0.56" x 0.20" x 0.075", the Xinger® 1.9GHz surface mount coupler is perfect for many applications out in the wireless field.

- > LNA and PA applications for DCS and PCS
- > Low 0.20dB max insertion loss and 2 degree phase balance
- > ± 0.20dB amplitude balance; 23dB minimum isolation; and a maximum VSWR of 1.17:1
- > A great price, in volume
- > And our B-There™ service commitment and 100% On-Spec™ performance guarantee!

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What'll we think of next?™