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Microwaves & RF

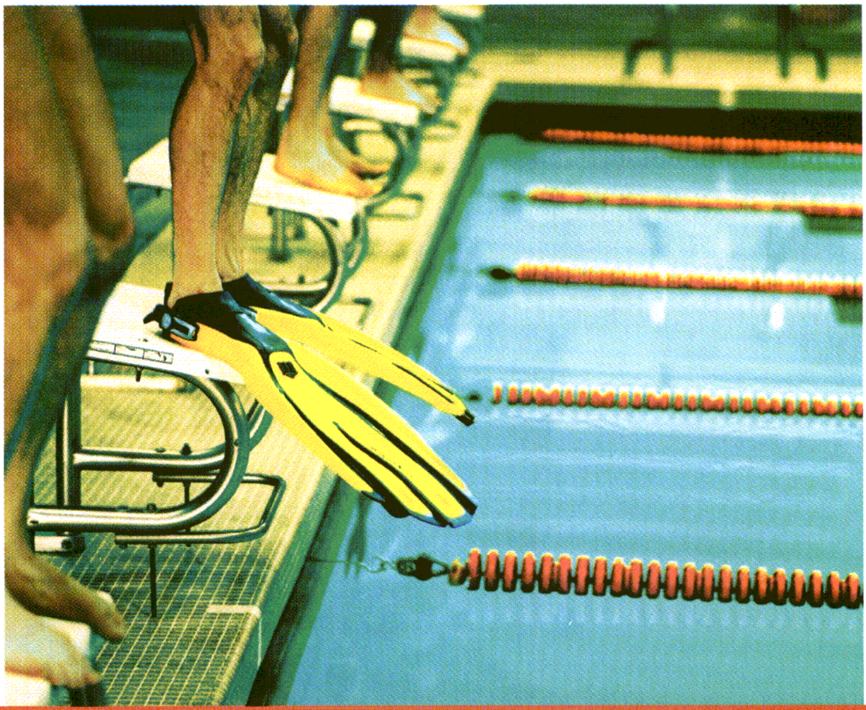
Wireless Applications Issue

NEWS
Marking Bluetooth and WLAN territories

DESIGN FEATURE
Analyzing VCOs and fractional-N synthesizers

PRODUCT TECHNOLOGY
Building a trimless IF oscillator

Digital upconverter IC tames complex modulation



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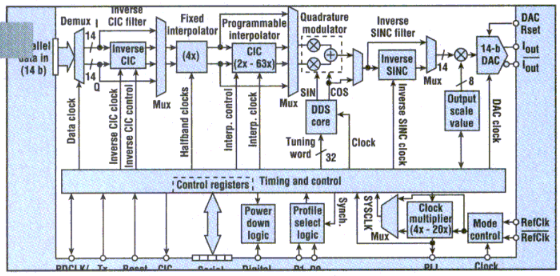
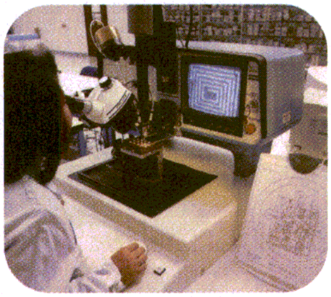


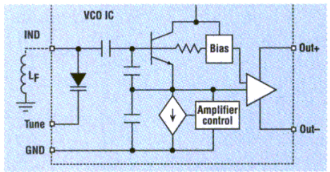
DEPARTMENTS

NEWS



DESIGN FEATURES

PRODUCT TECHNOLOGY



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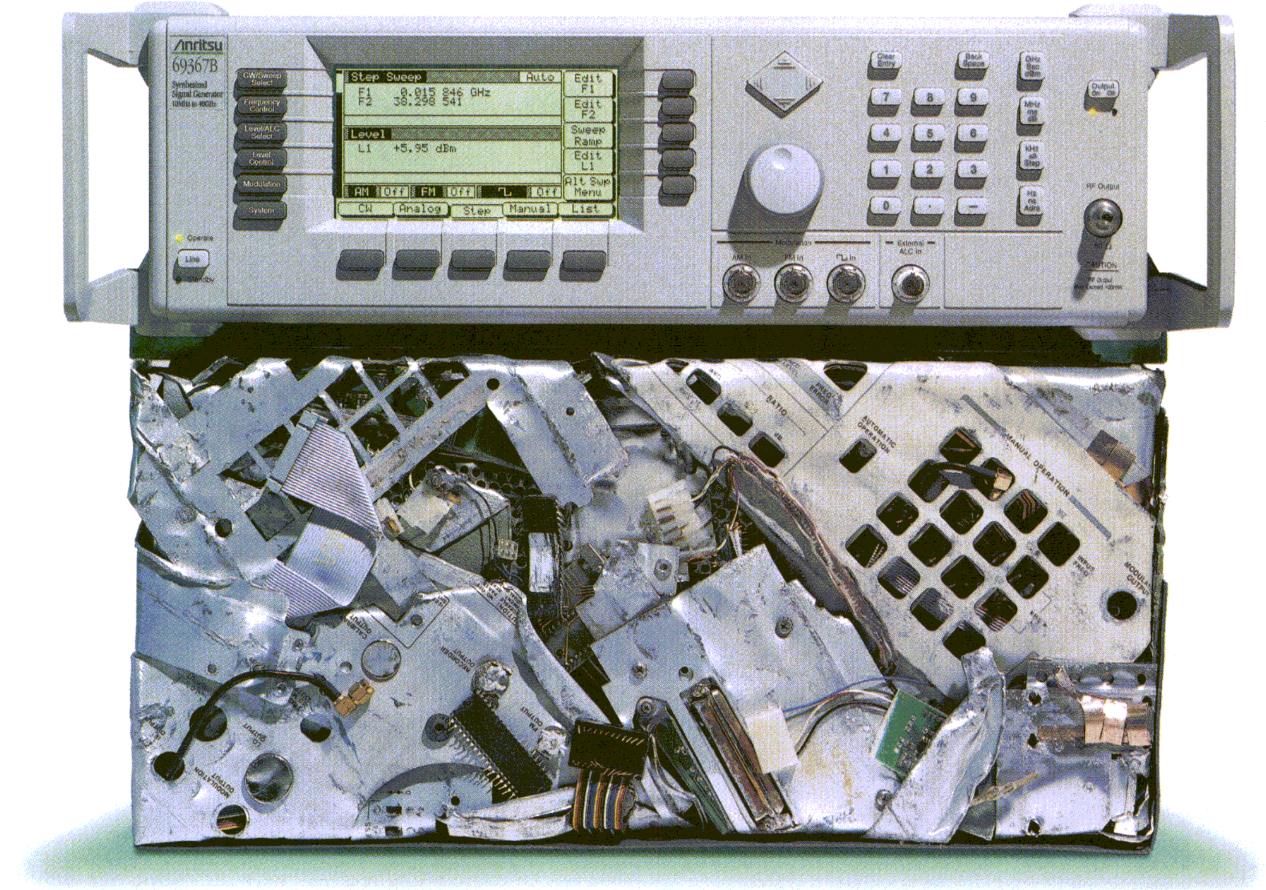
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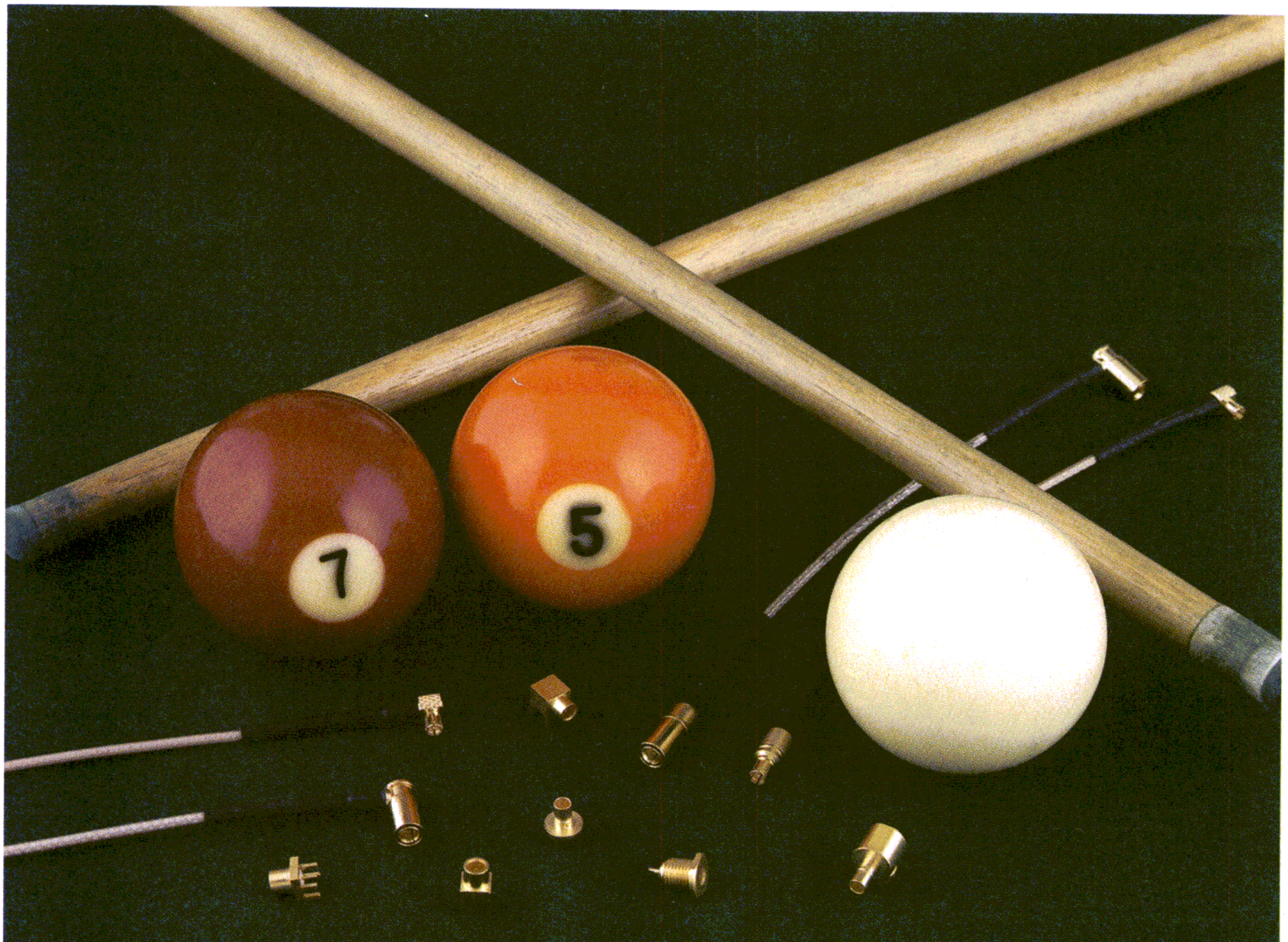

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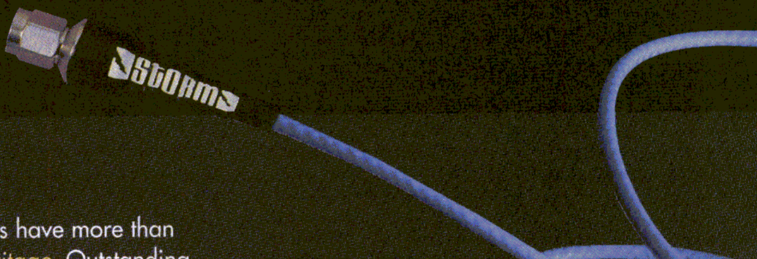

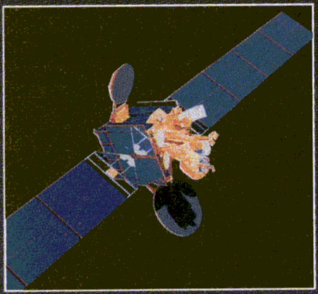

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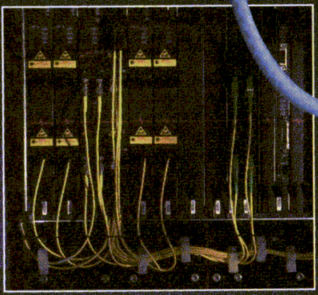

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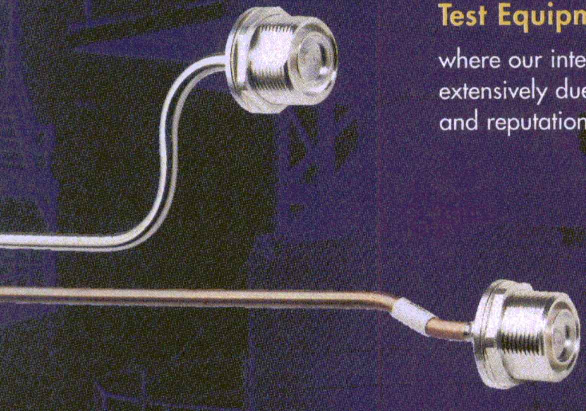

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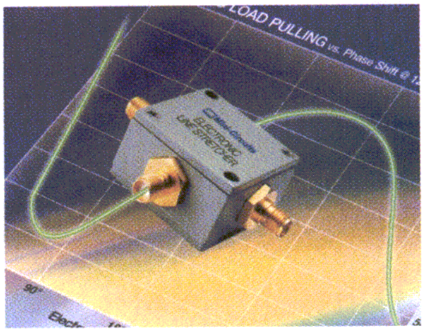
CIRCLE NO. 217

MISSING THE MARK

To the editor:

In an age when product branding is an important part of any marketing campaign, I was somewhat disappointed upon receiving my copy of the July issue of *Microwaves & RF*. On the front cover, which depicts one member of a family of novel electronic line stretchers from Mini-Circuits (Brooklyn, NY), the Mini-Circuits' logo did not appear on the top of the housing. The logo is representative of a company well-known for the quality of its many product lines, including amplifiers, filters, mixers, power dividers, and terminations, and a company associated with technological innovations, such as some of the industry's smallest and widest-dynamic-range mixers. I have included a photograph of the electronic line stretcher as it should have appeared (see figure).

The technological significance of the electronic line stretchers was also unnecessarily downplayed in the



article. It should have been noted in the initial paragraph that a patent is currently pending for the electronic line stretchers. The technology embodied within the devices is patent pending and is not only innovative, but also a practical advance for high-frequency engineers involved in measuring and modeling voltage-controlled oscillators (VCOs). Since these electronic line stretchers can cover wide frequency ranges from 110 to 950 MHz with the 10-to-12-dB return loss customarily exhibited by loads used when testing VCOs, and because they can provide the 360-deg. phase shifts needed to evaluate VCO frequency variations at center frequencies below 1 GHz, these devices represent significant improvements over mechanical line stretchers. And the fact that the phase of these electronic line stretchers can be adjusted electronically makes it a simple matter to incorporate the electronic line stretchers into automated-test-equipment (ATE) racks for VCO testing.

On behalf of Mini-Circuits, we were pleased with the chance to tell the story of these innovative electronic line stretchers in the July 2000 issue of *Microwaves & RF*, but disappointed that "patent pending" did not appear in references to the electronic line stretchers in the article and that the company logo and model number did not appear on the photograph on the front cover.

Bruce Marks
Mini-Circuits
Brooklyn, NY

TALENT

DESIGNERS

UNTIL YOU ARE BLUE IN THE TOOTH

Bluetooth is going to be big. All the signs point to it. The Bluetooth Special Interest Group (SIG) now numbers more than 1800 members. (A year ago, that number was just over 1000.) While many of these companies are small hardware and software developers, there are enough heavyweights on the list, including Ericsson, IBM, Lucent, Nokia, and Toshiba, to throw considerable marketing force behind the efforts to sell Bluetooth technology to the general public. Make no mistake—the "media blitz" for Bluetooth that started several years ago (in magazines such as this one) has only just begun.

Bluetooth is a marvelous example of technological reuse. Largely the brainchild of cellular-equipment giant Ericsson, Bluetooth essentially builds upon the foundation of Digital European Cordless Telecommunications (DECT) technology to create a personal wireless interconnectivity standard for low-to-moderate data rates [see p. 31 for a special news report on Bluetooth and wireless local-area-network (WLAN) markets and technologies].

Can marketing clout alone sell a product? In the case of Bluetooth, there are also many practical aspects to the standard. For example, it will allow users to establish miniature networks or "piconets" within their home or home office so that computers, peripherals, and other electronic devices can be connected by means of "invisible wires" (RF) and software protocols, rather than currently used cable assemblies.

Why is this so special? The answer can be found with just a quick glance at the various wire-and-connector approaches currently used with computers and their peripherals, such as parallel and serial connectors, universal-serial-bus (USB) connectors, and Firewire. In the world of digital audio, the proliferation of digital interconnect "standards," such as Toslink, AES/EBU, S/PDIF, and Firewire, has created a secondary market for interconnect devices that translate from one standard to another.

Bluetooth does away with the wires and the incompatibilities. Will the general public pay more for Bluetooth than for a cable connection? Probably not, which is the thinking of many wizened members of the Bluetooth SIG in stating that any complete Bluetooth original-equipment-manufacturer (OEM) solution, which includes the transceiver and baseband electronics, must be in the price range of $5 or less. A variety of low-cost integrated circuits (ICs) for various Bluetooth functions are already on the market, from such suppliers as Atmel Corp. (San Jose, CA), National Semiconductor (Santa Clara, CA), Maxim Integrated Products (Sunnyvale, CA), and Silicon Wave (San Diego, CA).

None of these ICs is the complete single-chip solution—integrating RF and baseband—that many SIG members have in mind. That solution is coming, and soon. When it does, prepare for the media blitz. It will be similar to the selling of cellular, and then some. We will be hearing the term "Bluetooth" until we are, well... ••

Jack Browne

Publisher/Editor

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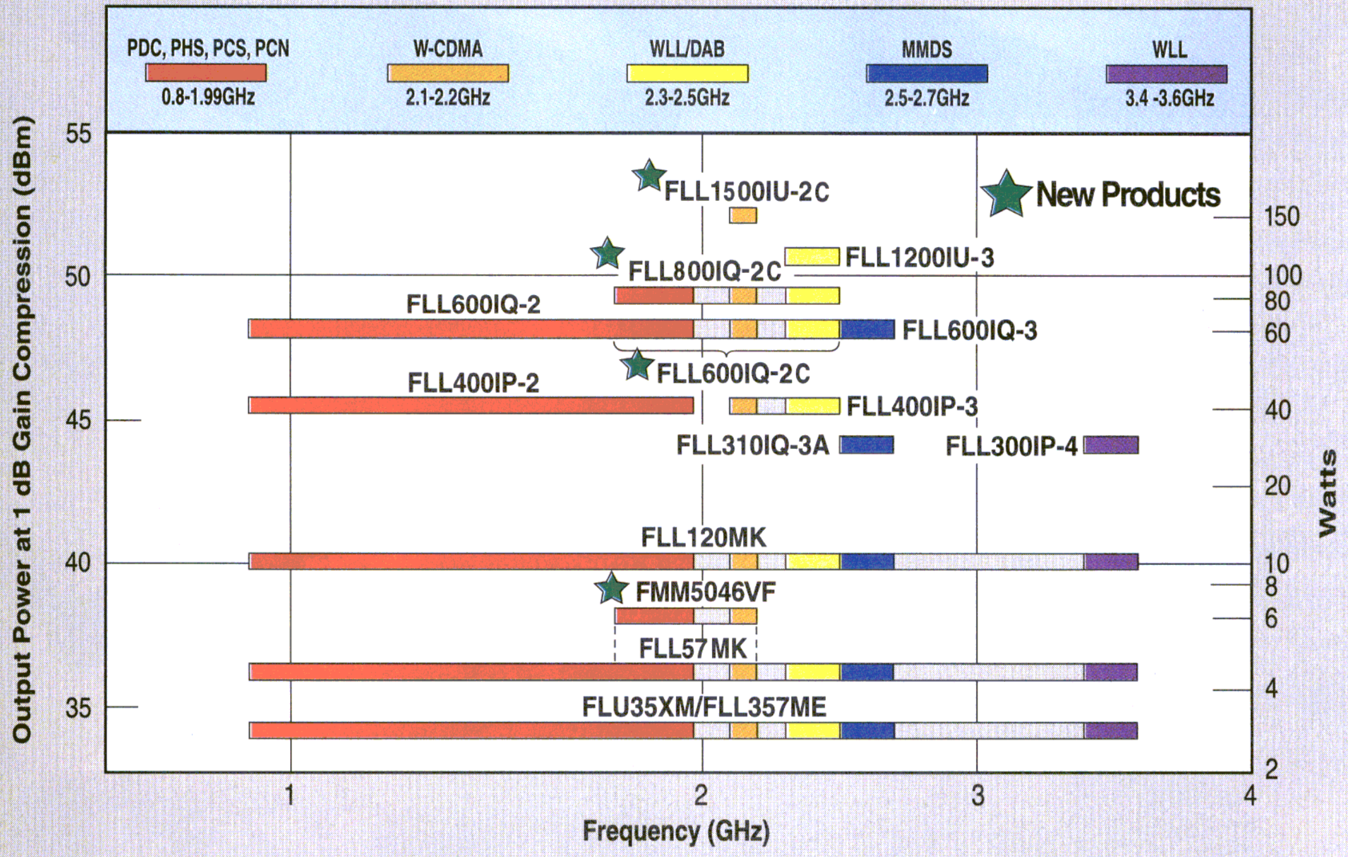

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CIRCLE NO. 238

French Prime Minister Opens State-Of-The-Art Front-End Plant

ROUSSET, FRANCE—The Prime Minister of France, Lionel Jospin, formally inaugurated Rousset8, the latest 8-in. (200-mm) STMicroelectronics wafer plant, which is located in Rousset (near Aix-en-Provence) in the south of France. The ceremony was hosted by Jean-Pierre Noblanc, chairman of the supervisory board, and Pasquale Pistorio, president and chief executive officer of STMicroelectronics.

Representing a planned total investment of approximately $1.4 billion, Rousset8 will have a capacity of 7000 8-in. wafers per week, which will make it (when completed) the largest 8-in. front-end facility in ST's worldwide manufacturing network, as well as the largest ST semiconductor manufacturing plant in France.

The Rousset fab uses state-of-the-art manufacturing techniques, including the innovative "mini-environment/isolation technology." This concept was developed to ensure that the increasingly stringent air-quality requirements of each future new technology generation can be met without escalating financial and environmental costs for air purification. In the mini-environment approach, each stage in the production process takes place in a localized environment, with wafers transported between these "mini environments" in hermetically sealed boxes. Wafer transfer and handling operations are performed by robots. In this way, the air in each mini environment can be economically maintained at a level that is approximately 1000 times purer than conventional cleanrooms, while simultaneously supporting a dramatic reduction in the frequency of air recirculation in the cleanroom.

"Mini-environment technology will be mandatory for the next generation of 12-in. fabs, but it has reached the stage where it offers financial and environmental benefits for 8-in. fabs," says Laurent Bosson, corporate vice president for central front-end manufacturing. Bosson also noted that in addition to providing higher air quality for a lower energy cost and fast manufacturing ramp-up, the elimination of wafer handling by humans improves yields by removing a possible source of contamination and misrouting.

Broadband Market Will Grow Exponentially

OYSTER BAY, NY—The broadband market will arrive in a grand fashion during the next five years as subscribers will grow six fold, according to a study from Allied Business Intelligence, Inc. (ABI). Subscribers will grow throughout the world, from 23 million in 2000 to 124 million in 2005 (see table), according to the findings in ABI's latest annual report detailing the broadband market, "Broadband Delivery in the Local Loop." As it has in almost all voice and data markets, wireless will play a significant role in the broadband space as well.

Subscribers to broadband services world market, 2000 to 2005

Year end	Subscribers (millions)
2000	22.8
2001	33.3
2002	48.8
2003	69.4
2004	93.6
2005	123.5
CAAG	40.1 percent

Source: Allied Business Intelligence, Inc.

The market will be led by asymmetric digital-subscriber line (ADSL) and cable modems, which are expected to account for 37 percent and 29 percent of the market in 2005, respectively. However, considerably greater subscriber numbers are possible if original-equipment manufacturers (OEMs) of wireless broadband equipment can achieve significant declines in equipment costs. "Wireless is also expected to account for a higher share of subscriber revenues, since many wireless services will target high-end users," says the author of the report, ABI senior analyst Andy Fuertes.

Meanwhile, integrated-services-digital-network (ISDN) BRI services are expected to slip from a leadership position in 2000 as new technologies increase their service footprint and decrease service fees and equipment costs. Nonetheless, a market for ISDN is expected to remain strong for several years to come, due to its wide availability and ability to offer switched services. ADSL and cable-modem services and technologies will continue to build upon already rapidly growing momentum.

Report Addresses Impact Of Packet Switching

AUSTIN, TX—Technology Futures, Inc., a telecom forecasting business, recently announced the publication of a report titled *Technology Forecasts for Local Exchange Switching Equipment.* The report was authored by Ray L. Hodges, senior consultant at TFI, and Lawrence K. Vanston, Ph.D., president of TFI. The research was sponsored by the Telecommunications Technology Forecasting Group (TTFG), a consortium of telephone companies comprised of Bell Atlantic, Bell Canada, BellSouth Telecommunications, Cincinnati Bell, GTE, Southwestern Bell, Sprint-LTD, and US WEST Communications.

The report addresses issues relating to the current trend of asynchronous-transfer-mode/Internet-protocol (ATM/IP) packet switching and its impact on the embedded digital circuit switches of incumbent local-exchange carriers (ILECs). It provides an update and comparisons to earlier forecasts conducted by Technology Futures, which were accurate in identifying the technology substitutions and predicting the pace of change in switching equipment. The report also includes a technology description and assessment of optical switching.

According to Hodges, "In the past, we have upgraded circuit switches to provide better services and features for voice traffic and to accommodate some low-speed data. In the future, the circuit switches must be replaced by packet switches, which are optimized for high-speed data. ATM/IP switches are optimized for data and, when data dominates, they will become cost-effective for all forms of traffic. Voice traffic will increasingly be disguised as data and transported over the packet network. The direction is clear, but the time of the transition is more difficult to determine and is the main subject of this report."

Precise Frequency Measurements Result From Research

GAITHERSBURG, MD—Researchers from the National Institute of Standards and Technology (NIST) and Bell Laboratories of Lucent Technologies recently teamed to produce a more precise method for measuring the frequency of visible and infrared (IR) light. The new technology one day may help to build more accurate atomic clocks, improve scientists' ability to identify molecules and elements by spectroscopy, and provide more reliable frequency standards for use by the telecommunications and related industries.

The technique uses a single laser to measure optical frequency instead of a cumbersome and expensive multiple-laser system. The measurements made by the NIST/Lucent system are more precise because they are compared to the well-defined frequency of a Cesium-133 atomic clock.

The researchers "locked" a special laser so that it generated a repeating series of ultrashort optical pulses. Each pulse is so short that it contains approximately three cycles of light. The new technique is based on controlling the phase of the pulses, effectively putting them in "lock step." Through a high-resolution spectroscope, the laser's output is seen as sharply defined lines, separated by the pulse-repetition frequency. The scientists refer to this spectrum as a "comb" because it has the appearance of a common pocket comb.

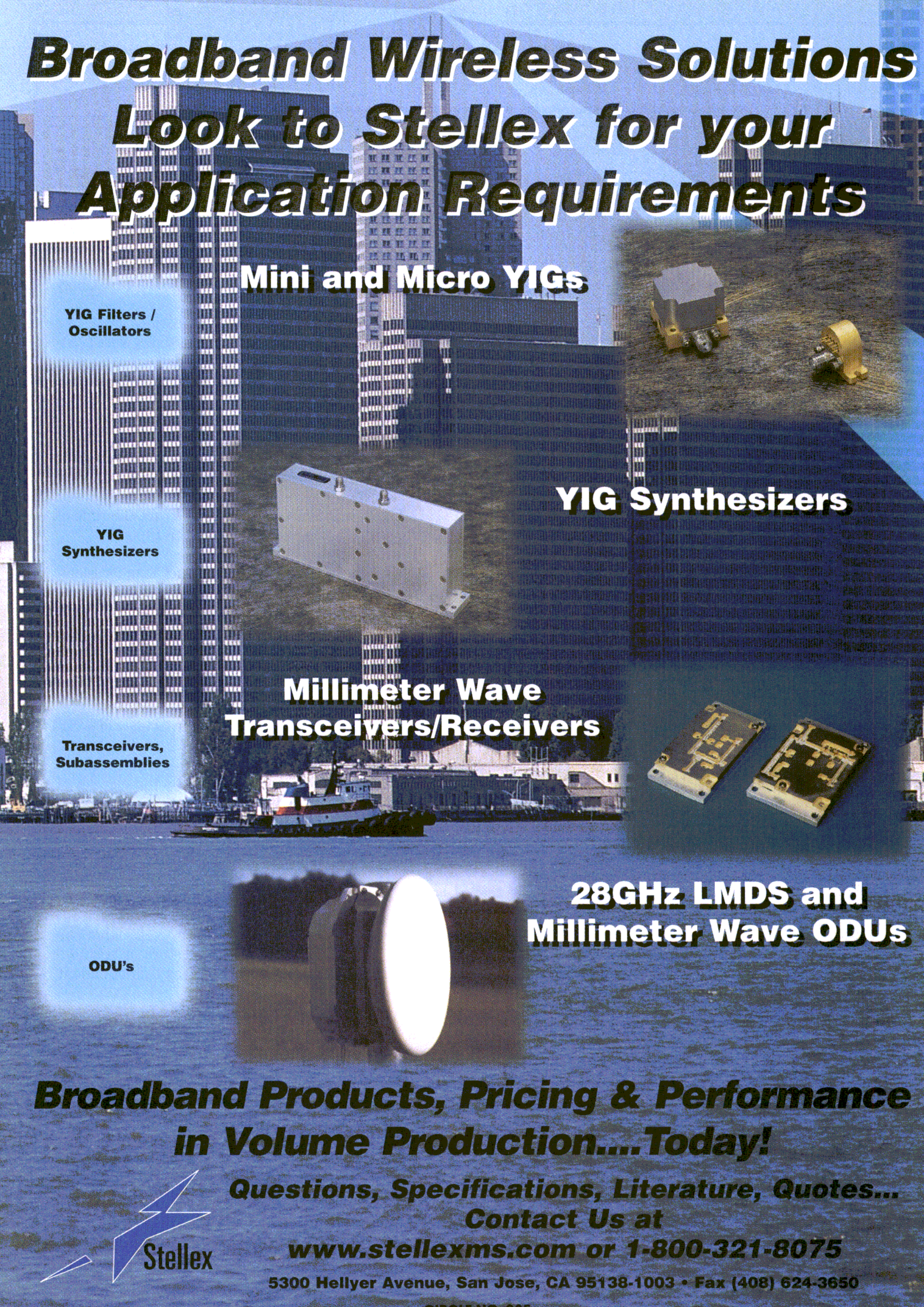
MSM Photodetector Array Is Developed

WARREN, NJ—ANADIGICS announced that it has developed a 12-channel metal-semiconductor-metal (MSM) photodetector array for use in 2x fibre channel, and Infiniband data-communications networks capable of data rates up to 3.125 Gb/s per channel. The devices, which integrate a 1×12 array of photodetectors on a single gallium-arsenide (GaAs) substrate, operate at 850 nm with a polarity-independent architecture.

"Parallel optical-communication systems are emerging as solutions for increased bandwidth applications," remarks Dr. Bami Bastani, president and CEO of ANADIGICS. "This highly integrated photodetector array, coupled with optical-link laser arrays, will provide the industry with scalable broadband fiber interconnects for networking and fiber-optic backplanes."

Dr. Bastani explains that transmit and receive arrays provide space-efficient, broadband-physical-layer building blocks for high-density fiber-optic parallel interconnects.

ANADIGICS' MSM technology integrates all 12 channels onto a single die, conforms to the 250-μm pitch fiber-ribbon cables, and displays superior crosstalk isolation while targeting responsivity in excess of 0.4 A/W. The scalable nature of the architecture enables the company to generate a family of products from 1×4 up to the 1×12 technology recently demonstrated.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Ink-Jet-Printed Full-Color LEP Display Announced

CAMBRIDGE, ENGLAND—Seiko-Epson and Cambridge Display Technology (CDT) have developed a 2.5-sq.-in. (6.35-sq.-cm) full-color display using CDT's light-emitting-polymer (LEP) technology, an innovative display science that paves the way for the creation of ultra-small, paper-thin consumer-electronic displays.

The prototype color display has been manufactured using CDT's red, green, and blue polymer materials and an industry-first ink-jet printing process developed for the project. The color display achieves color quality that is equal to current liquid-crystal-display (LCD) technology and is comparable to displays found in many portable-computer products.

The prototype color display measures 2.5 sq. in., has a resolution of 200 × 150 pixels, with 16 gray scaling at system level, and will be targeted at initial market-entry points for LEP display products such as mobile phones and personal digital assistants (PDAs). Beyond this, CDT and Seiko-Epson expect this technology to eventually penetrate all other display markets.

"The pre-production colorlight-emitting polymer display being shown by CDT and Seiko-Epson has a color density similar to current LCDs. The techniques being jointly developed by the two companies means that the manufacturing cost of an LEP display will be significantly less than the cost of producing conventional LCD or cathode-ray-tube displays," says Dr. Shimoda, general manager of basic research for Seiko-Epson.

Results Are Revealed For"GIGA-CHIPS"Project

STUTTGART, GERMANY—TEMIC Semiconductors and Alcatel recently announced the first results of the "GIGA-CHIPS" project within the European MEDEA program—devices for 40-Gb/s optical-transmission systems manufactured in silicon-germanium (SiGe) technology. TEMIC Semiconductors provides the SiGe technology on a foundry basis for the production of the new devices. The 50-GHz SiGe technology that TEMIC Semiconductors uses is the fastest Si-based technology in production. The first circuit developed by Alcatel is a new 2:1 multiplexer integrated circuit (IC) which combines cost-effectiveness and maximum performance.

The new multiplexer with its high-bit-rate capability responds to the market demand for increased transmission capacity by using already-existing fibers. The chip is driven by a standard supply voltage of +5 VDC and has a typical current consumption of 35 mA. It combines two 20-Gb/s data streams to one 40-Gb/s data stream to use 40 Gb/s as one signal on one line. This means that more information can be transmitted faster, without additional lines.

The successful realization of first circuits, such as the new 2:1 multiplexer, provides a promising outlook for further SiGe applications at 40 Gb/s. SiGe, the latest process for RF applications, is seen as the alternative for gallium arsenide (GaAs).

Kudos

Qualcomm recently announced a $25 million commitment for educational programs that will help with efforts nationwide to bridge the "Digital Divide." The company has designated the University of California at San Diego (UCSD), San Diego State University (SDSU), California State University San Marcos (CSUSM), and the Foundation for the Improvement of Mathematics and Science Education as recipients of this donation. The contribution is spread over a five-year period, beginning in fiscal year 2001...U.S. Wireless Corp. announced that it has received notice of allowance from the US Patent Office for four of its principal patents, covering key aspects of the company's multipath pattern-matching technology. An important area of intellectual property covered by the company's patent filings focuses on its unique strength in locating wireless calls in urban areas, where other solutions such as triangulation or Global Positioning System (GPS) suffer from signal blockage...ADC announced that its AAC-2™ asynchronous-transfer-mode (ATM)-access concentrator has been named a finalist by CMP Media, Inc. for Network Computings' 2000 Well-Connected Award in the category of ATM-access concentration. The product was chosen for its flexible, scalable architecture, multiple protocols, wide range of line speeds, large variety of customer interfaces, and inverse multiplexing over ATM (IMA) capabilities...Motorola has increased its gallium-arsenide (GaAs) capacity every year since 1997, and completed the process of tripling the wafer output of its Compound Semiconductor-1 (CS-1) wafer-fabrication facility in June, three months ahead of schedule. Motorola's GaAs semiconductor devices are currently fabricated in the CS-1 facility.

Model Number	Frequency Range (GHz)	Gain (Min.) (dB)	Gain Flatness (±dB)	Noise Figure (dB, Max.)	VSWR Input (Max.)	VSWR Output (Max.)	Output Power @ 1 dB Comp. (dBm, Min.)	Nom. DC Power (+15 V, mA)
MODERATE BAND AMPLIFIERS								
AFS2-00700080-05-10P-4	.7–.8	30	0.50	0.45	1.5:1	1.5:1	+10	90
AFS2-00800100-05-10P-4	.8–1	30	0.50	0.45	1.5:1	1.5:1	+10	90
AFS3-01200160-05-13P-6	1.2–1.6	40	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-01400170-05-13P-6	1.4–1.7	40	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-01500180-04-13P-6	1.5–1.8	40	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-01500250-06-13P-6	1.5–2.5	36	0.50	0.60	2.0:1	2.0:1	+13	150
AFS3-01700190-04-13P-6	1.7–1.9	36	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-01800220-05-13P-6	1.8–2.2	36	0.50	0.50	1.5:1	1.5:1	+13	150
AFS3-02200230-04-13P-6	2.2–2.3	36	0.50	0.40	1.5:1	1.5:1	+13	150
AFS3-02300270-05-13P-6	2.3–2.7	34	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-02700290-05-13P-6	2.7–2.9	32	0.50	0.50	1.5:1	1.5:1	+13	150
AFS3-02900310-05-13P-6	2.9–3.1	32	0.50	0.45	1.5:1	1.5:1	+13	150
AFS3-03100350-06-10P-4	3.1–3.5	29	0.50	0.6	1.5:1	1.5:1	+10	150
AFS4-03400420-06-13P-6	3.4–4.2	40	0.50	0.60	1.5:1	1.5:1	+13	225
AFS3-04400510-07-5P-4	4.4–5.1	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-04500480-07-5P-4	4.5–4.8	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05200600-07-5P-4	5.2–6	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05400590-07-5P-4	5.4–5.9	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-05800670-07-5P-4	5.8–6.7	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS3-07250775-06-5P-4	7.25–7.75	30	0.50	0.60	1.5:1	1.5:1	+5	100
AFS3-07900840-07-5P-4	7.9–8.4	30	0.50	0.70	1.5:1	1.5:1	+5	100
AFS4-08500960-08-5P-4	8.5–9.6	32	0.75	0.80	1.5:1	1.5:1	+5	125
AFS3-09001100-09-5P-4	9–11	26	0.50	0.90	1.5:1	1.5:1	+5	100
AFS4-09001100-09-5P-4	9–11	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS4-10951175-09-5P-4	10.95–11.75	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS4-11701220-09-5P-4	11.7–12.2	32	0.75	0.90	1.5:1	1.5:1	+5	125
AFS2-12201280-10-8P-4	12.2–12.8	14	0.75	1.00	1.5:1	1.5:1	+8	80
AFS4-12201280-10-12P-4	12.2–12.8	27	0.75	1.00	1.5:1	1.5:1	+12	200
AFS4-12701330-13-10P-4	12.7–13.3	27	0.75	1.30	1.5:1	1.5:1	+10	175
AFS4-13201400-14-10P-4	13.2–14	24	0.75	1.40	1.5:1	1.5:1	+10	175
AFS4-14001450-14-10P-4	14–14.5	24	0.75	1.40	1.5:1	1.5:1	+10	175
AFS4-20202120-20-8P-4	20.2–21.2	20	1.00	2.00	1.5:1	1.5:1	+8	175
AFS4-21202400-22-10P-4	21.2–24	18	1.00	2.2	2.0:1	2.0:1	+10	100
OCTAVE BAND AMPLIFIERS								
AFS3-00120025-09-10P-4	.12–.25	38	0.50	0.9	2.0:1	2.0:1	+10	175
AFS3-00250050-08-10P-4	.25–.5	38	0.50	0.8	2.0:1	2.0:1	+10	125
AFS3-00500100-05-10P-6	.5–1	38	0.75	0.5	2.0:1	2.0:1	+10	150
AFS3-01000200-05-10P-6	1–2	38	1.00	0.5	2.0:1	2.0:1	+10	150
AFS3-01200240-05-10P-6	1.2–2.4	34	1.00	0.5	2.0:1	2.0:1	+10	175
AFS3-02000400-06-10P-4	2–4	30	1.00	0.6	2.0:1	2.0:1	+10	125
AFS3-02600520-10-10P-4	2.6–5.2	28	1.00	1.0	2.0:1	2.0:1	+10	150
AFS3-04000800-07-10P-4	4–8	30	1.00	0.7	2.0:1	2.0:1	+10	125
AFS3-08001200-09-10P-4	8–12	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-08001600-15-8P-4	8–16	26	1.00	1.5	2.0:1	2.0:1	+8	80
AFS4-12002400-25-10P-4	12–24	20	2.00	2.5	2.0:1	2.0:1	+10	85
AFS4-12001800-18-10P-4	12–18	26	1.00	1.8	2.0:1	2.0:1	+10	125
AFS4-18002650-28-8P-4	18–26.5	18	1.75	2.8	2.5:1	2.2:1	+8	150
MULTIOCTAVE BAND AMPLIFIERS								
AFS1-00040200-12-10P-4	.04–2	15	1.50	1.2	2.5:1	2.0:1	+10	75
AFS3-00300140-08-10P-4	.3–1.4	33	1.00	0.8	2.0:1	2.0:1	+10	150
AFS2-00400350-12-10P-4	.4–3.5	22	1.50	1.2	2.0:1	2.0:1	+10	80
AFS3-00500200-08-15P-4	.5–2	38	1.00	0.8	2.0:1	2.0:1	+15	125
AFS3-01000400-09-10P-4	1–4	30	1.50	0.9	2.0:1	2.0:1	+10	125
AFS3-02000800-09-10P-4	2–8	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS4-02001800-23-10P-4	2–18	25	2.00	2.3	2.0:1	2.0:1	+10	175
AFS4-06001800-22-10P-4	6–18	24	2.00	2.2	2.0:1	2.0:1	+10	150
AFS4-08001800-22-10P-4	8–18	26	2.00	2.2	2.0:1	2.0:1	+10	150
ULTRA WIDEBAND AMPLIFIERS								
AFS3-00100100-09-10P-4	.1–1	38	1.00	0.9	2.0:1	2.0:1	+10	150
AFS3-00100200-10-15P-4	.1–2	38	1.00	1.0	2.0:1	2.0:1	+15	150
AFS3-00100300-11-10P-4	.1–3	32	1.00	1.1	2.0:1	2.0:1	+10	150
AFS3-00100400-13-10P-4	.1–4	28	1.00	1.3	2.0:1	2.0:1	+10	150
AFS3-00100600-13-10P-4	.1–6	28	1.25	1.3	2.0:1	2.0:1	+10	125
AFS3-00100800-14-10P-4	.1–8	25	1.50	1.4	2.0:1	2.0:1	+10	125
AFS4-00101200-22-10P-4	.1–12	28	1.50	2.2	2.0:1	2.0:1	+10	175
AFS4-00101400-23-10P-4	.1–14	24	2.00	2.3	2.5:1	2.5:1	+10	200
AFS4-00101800-25-10P-4	.1–18	25	2.00	2.5	2.5:1	2.5:1	+10	175
AFS4-00102000-30-10P-4	.1–20	20	2.50	3.0	2.5:1	2.5:1	+10	175
AFS4-00102650-40-8P-4	.1–26.5	18	2.50	4.0	2.5:1	2.5:1	+8	175

Note: Noise figure increases below 500 MHz in bands greater than 0.1-10 GHz.

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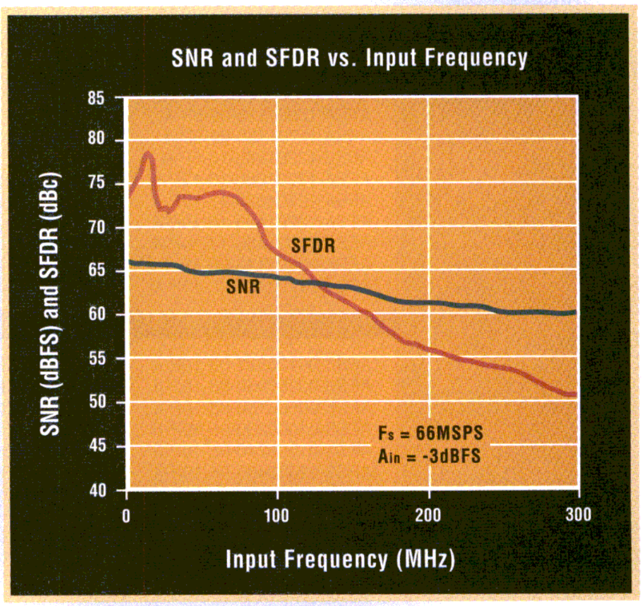

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CLC5957	12bit	70MSPS	67dBFS	74dBc	250MHz
CLC5958	14bit	52MSPS	71dBFS	90dBc	70MHz

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CIRCLE NO. 376

When Bluetooth-enabled hardware invades the turf of WLANs, will it be a clash of titans or will the two peacefully coexist?

Bluetooth Meets WLANs—Can They Live Together?

GENE HEFTMAN
Senior Editor

IF all the pundits, experts, and members of the Bluetooth Special Interest Group (SIG) are correct, it will be only a matter of time before the most heralded wireless communications technology since the cellular phone starts showing up in laptop computers, personal digital assistants (PDAs), digital cameras, and a host of other portable devices. These Bluetooth-enabled products will exchange information among themselves and fixed equipment over a short-range radio link that operates in the 2.4-GHz industrial- scientific-medical (ISM) band. This is the same frequency that wireless local-area networks (WLANs) using the well-established IEEE 802.11 standard have been transmitting on for the past several years. Since Bluetooth and WLANs share a similar technology—spread-spectrum communications—and will serve in identical data and voice applications in many instances, an obvious question is, can they both share the airwaves without interfering with one another? The answer is critical to both technologies since they are expected to proliferate tremendously in the next few years with a myriad of products incorporating both. In many applications, Bluetooth and 802.11 hardware will wind up in the same portable device and operate in the same location—office, factory, hospital, etc. This situation is known as the coexistence problem of the two technologies and it has attracted the attention of the Bluetooth and WLAN camps, IEEE standards groups, and the Federal Communications Commission (FCC).

No one associated with Bluetooth or WLANs doubts that there will be a coexistence problem. Likewise, no one doubts that the technical difficulties of two different spread-spectrum devices transmitting information at the same time in close proximity cannot be resolved. But the problem is complex and is being approached on a number of levels: the geometry or distance between devices, the applications (voice or data), packet size, packet length, operating rules and channel widths within the 2.4-GHz band, the characteristics of direct sequence, spread spectrum (DSSS) [used in most WLANs] and frequency hopping, spread spectrum (FHSS) [used in Bluetooth], and others. The IEEE has formed the 802.15 Coexistence Group to come up with guidelines for allowing the technologies to live together. Recent reports and simulations indicate that solutions to the potential interactions can be devised. Meanwhile, the FCC is in the process of hearing arguments for changing Part 15 of its rules (section 247 regarding the operation of spread spectrum devices) to widen the channel widths of the 2.400-to-2.483-GHz band to permit frequency-hopping devices to operate with a wider bandwidth. But organizations such as the IEEE 802 Standards Working Group oppose the proposed change and the issue is currently in the charge/counter-charge arena.

EXPERT OPINIONS

"There is sure to be a co-existence problem." says David McCartney, Executive Vice President of Rangestar Wireless, Inc. (Aptos, CA), "If you put a Bluetooth card and WLAN card in two adjacent slots of a notebook computer, they won't work." But he is optimistic about reducing the interference between the two technologies by separating them within the equipment. Simulations and tests run by the 802.15 Group show that by placing the devices 6 to 8 in. (15.24 to 20.32 cm) apart within the equipment will enable full operability.

In the view of David Lyon, Chairman and CEO of Silicon Wave (San Diego, CA), "Bluetooth will not be jammed all of the time. Even worst case, with devices closely co-located (Bluetooth and WLAN), Bluetooth should be able to get significant throughput." Lyon is optimistic about Bluetooth and WLANs sharing the same neighborhood. He says, "Bluetooth was designed to survive even with transmitters in the ISM band such as microwave ovens, the different flavors of 802.11, cordless phones, etc., so the protocol stack and physical-layer definition is built to survive the worst of the jammers."

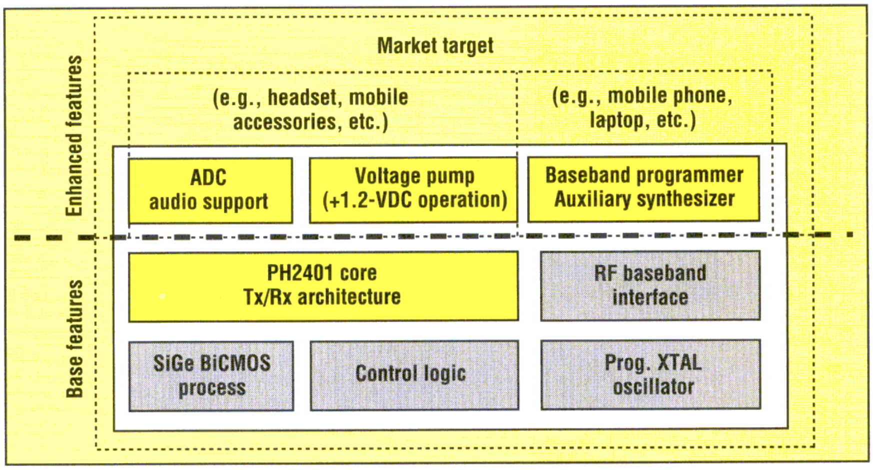
Another viewpoint comes from Paul Sofianos, Senior Staff Applications Engineer at Motorola's Semiconductor Power Sector (Tempe, AZ), who states, "One thing I haven't seen in the literature and will probably be critical is thinking in terms of Bluetooth and the applications. If you look at WLANs and the other potential interferers, it seems like the problems will not be as pronounced as people think they will be." For example, Sofianos points out that a heavy duty WLAN would be unlikely to have a Bluetooth voice application, and it is voice that would cause the major problem because one cannot retransmit data.

On the 802.11 side, Jim Zyren, Product Marketing Manager at Intersil Corp. (Irvine, CA) says of the possibility of interactions, "There will be some effect, but we don't think it will be catastrophic, but there could be some degradation to the abilities of both systems such as a marginal reduction in throughput. One system is frequency agile (Bluetooth) while the other is fixed frequency (WLAN) so inevitably there will be some collisions." The probability of collisions depends on the operating mode-voice or data downloads and the traffic density on 802.11. And another of the many factors that influence the occurrence of interference is the location of Bluetooth and WLAN devices with respect to one another.

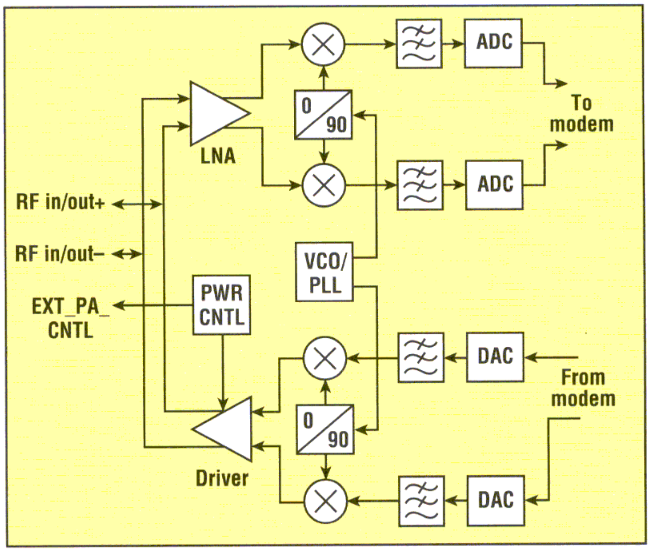
Collisions are a concern of Jim Wight, Senior Radio Architect for Conexant Systems Wireless Communications Division (Newport Beach, CA), who states, "Problems occur when you have two systems that use the same type of spreading. If you have different systems all doing frequency hopping, you can get into a very crowded signal space. It is more of a problem than if a hopper interacts with a direct sequence type. Two hoppers would collide with devastation. But direct sequence could have some impact on the bit-error rate (BER) of hoppers."

Many experts envision similar solutions to the co-existence problem. Intersil's Zyren and his associate Bruce Kraemer, also a Product Marketing Manager, believe that Bluetooth radios may be able to use intelligence to hop around and avoid other interferers in the band. Wight of Conexant Systems is on the same track when he says that you can use "traffic lights" (i.e., intelligence) to mitigate a collision. In other words, if the traffic light is "red" a new system is not allowed on the air, but it would take a high-level agreement between all of the system operators to do this. A theoretical possibility comes from Rangestar's McCartney who says, "Since we're both using the same 2.4-GHz band, why don't we share the front end. We could have a common RF front end and switch. If I'm using 802.11 in the transmit or receive mode, I shut off the Bluetooth and the two systems switch back and forth." Intersil's Kraemer believes, "There is a technical solution to the interference problem where you operate on non-overlapping frequency channels."

A recent report from Ericsson entitled, "Bluetooth voice and data performance in 802.11 DS LAN environment," examined the coexistence problem from the standpoint of a +20-dBm 802.11 direct-sequence system interfering with a 0-dBm Bluetooth link. The report does not cover how the Bluetooth system impacts on the WLAN. The setting is a "typical" office having a large number WLAN terminals but a small number of access points. It is assumed that in an office environment, both systems will operate simultaneously



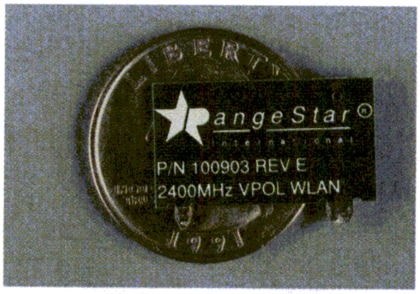
1. Single-chip radio transceivers for Bluetooth such as the PH2401 from Conexant Systems are designed to maximize battery life, have a small footprint, and be low in cost.



2. The architecture of Silicon Wave's SiW1501 radio-modem IC is a direct conversion type that converts the radio signal into digital bits and provides a digital interface. The synthesizer function is fully integrated and requires no external VCO.

and since they are in the same radio band, mutual interference is a concern.

The study differentiates between the impact on the Bluetooth voice link and data link. As might be expected, distance is a critical factor in the disturbance of a link. For example, under normal WLAN traffic conditions, Bluetooth voice users are not disturbed as long as the voice users stay within 2 m of the equipment that they are communicating with (disturbance occurs in less that 1 percent of cases). But if the Bluetooth voice user moves out to 10 m from the equipment, the probability of disturbance increases to 8 percent. In the data-link case, Bluetooth supports and experiences more degradation. At a distance of 10 m, a throughput reduction of more than 10 percent occurs with 24-percent probability. The authors claim that because Bluetooth and WLANs have limited frequency overlap, the throughput reduction can never exceed 22 percent.



3. A total Bluetooth solution from Rangestar Wireless includes the type of miniature embedded antenna shown here that meets a user's, size, cost, technical, regulatory, and volume-production requirements.

While the search for solutions to coexist go forward and interested parties make their case to change the FCC rules, Bluetooth is moving closer toward implementation in portable products. Earlier magazine article predictions called for 2000 to be a banner year for Bluetooth, but 2001 looks more like the breakthrough that SIG members are hoping for. IC suppliers, in particular, are jockeying to be first with Bluetooth-compliant chips on the market.

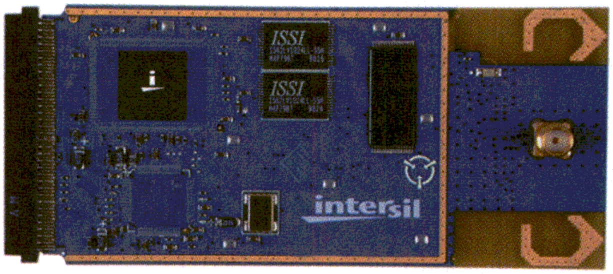
Single-chip low-power transceivers such as the PH2401 (Fig. 1) from Conexant Systems are springing up as IC suppliers rush to nail down orders for what appears to be an expanding market. The PH2401 is one of three radio transceivers optimized for 2.4-GHz operation that meets Bluetooth Specification v1.0. It incorporates a transmitter power amplifier (PA) having programmable output levels from –10 to 2 dBm. The modulation technique is Gaussian frequency-shift keying (GFSK). Channel bandwidth is 1 MHz with frequency deviation between 140 and 175 kHz. The device departs from the trend toward using mainstream semiconductor processes in Bluetooth products to reduce costs, simplify designs, and ensure an ample flow of products. The chip is fabricat-

ed as a silicon-germanium (SiGe) bipolar-complementary-metal-oxide-semiconductor (BiCMOS) application-specific IC (ASIC).

Cambridge Silicon Radio (Cambridge, United Kingdom) claims to have the first fully integrated 2.4-GHz radio, baseband, and microcontroller in a single IC. Known as the Bluetooth CMOS BlueCore™01, this is a pure plain-vanilla CMOS device. It is intended to drive Bluetooth costs down by integrating more functionality in a single IC. Built into the device are a 2.4-GHz radio, a baseband digital signal processor (DSP), a 16-b microcontroller and the necessary random access memory (RAM) to implement a Bluetooth solution.

Yet another semiconductor processing approach to Bluetooth is Silicon Wave's SiW1501 radio-modem IC, a BiCMOS device fabricated in silicon on insulator (SOI). It uses a direct-conversion architecture (Fig. 2) for the transceiver to directly convert an incoming analog signal into bits-direct in-phase/quadrature (I/Q) conversion (zero IF) with a DC output. Also on-chip are the GFSK modem and a synthesizer that requires no external VCO or resonator. The device is available on a stand-alone basis or with the SiW1601 link-controller IC and lower level Bluetooth protocol software.

An RF Bluetooth module based on thin-film technology is under development by Intarsia Corp. (Oxfordshire, United Kingdom) in collaboration with Ericsson Mobile Communications AB (Lund, Sweden). Integrating RF passive components in thin films on a glass substrate is said to offer a high-performance, miniature device suitable for mounting in handheld wireless devices such as mobile phones. The module to be built is a BiCMOS RF single-chip transceiver with numerous discrete passives mounted on a ceramic substrate. A key aspect of the design is Intarsia's PassPort™ design automation tool which can perform transmitter/receiver (Tx)/(Rx) balun matching, resonant tank-circuit design and the design of lumped-element components for the VCO loop filter. A full-function module has been fab-



4. PRISM 2.5, Intersil's latest version of its 2.4-GHz WLAN hardware set, reduces the chip count of its predecessor PRISM II from 5 to 4 devices. This WLAN adapter card is intended for laptops, PDAs, digital cameras, and other portable equipment.

ricated using Intarsia's direct-module-attach (DMA) module assembly technique, which eliminates the surface-mount assembly of more than 30 small discrete passive components onto the substrate.

The large influx of companies expected to manufacture Bluetooth-enabled products is likely to leave many with a difficult task of actually implementing the radio. According to Rangestar Wireless's David McCartney, many of the company's customers are not RF experts, but they are expert at what they do—making notebooks, PDAs, etc. Accordingly, they need a total solution provider for their Bluetooth product, from the software through the serial port. Although Rangestar's expertise is miniature Bluetooth antennas (Fig. 3), it offers a total solution package that includes the design and integration of the antenna, regulatory testing compliance, certification to the Bluetooth qualification tests, integration of radio and antenna, and software development. Some functions are performed by the company while other are performed through alliances with other companies.

While attention focuses on the data-handling capabilities of Bluetooth and WLANs, these systems will require voice capability as well. Mitel Semiconductor (Ottawa, Ontario, Canada) has come out with a baseband controller-the MT1020-that includes a full-duplex audio codec. The controller also supports data operations and includes an ARM7TDMI embedded microcontroller core, the company's Bluetooth baseband peripheral (BBP) clock, program memory, along with USB and UART host interfaces.

THE LAN SIDE

The other element in the coexistence-problem equation is WLAN, a technology that is growing at a rate of better than 30 percent a year and is expected to approach $3 billion by 2004. The IEEE 802.11 standard for WLANs is only three years old and already has outgrown its original 2-Mb/s data rate specified for the 2.4-GHz range. IEEE 802.11b is called the higher-rate extension for 5.5- and 11-Mb/s transfers at 2.4 GHz, while IEEE 802.11a is the 5-GHz extension that permits transfers at 6, 12, and 24 Mb/s. Although Bluetooth gets most of the attention these days, WLAN product manufacturers are making a few headlines of their own.

Almost from the inception of WLANs, Intersil (formerly Harris Semiconductor) has had one of the best-known chip sets on the market, the PRISM™. A new version was announced recently called PRISM 2.5, which marks a reduction in the number of chips—from 5 to 4—over the previous version, PRISM II (Fig. 4). The four ICs-3 RF devices and 1 digital device-can deliver 10BaseT Ethernet data rates of 11 Mb/s. The company has adopted a new manufacturing process called UHF-2, a 0.6-μm RF BiCMOS technology that integrates a 25-GHz f_T, 40-GHz f_{MAX} NPN transistor that can deliver higher performance, lower-power transistors. On-chip integration of passive components is another benefit of the new process.

Another 11-Mb/s WLAN card known as the ORiNOCO PC from Lucent Technologies (Research Triangle Park, NC) is aimed at high-speed wireless networking and Internet access for laptop and desktop computers. The card boasts a range of up to 1500 ft. and offers a choice of security levels from a 64-b key (silver level) to a 128-b key (gold level). Recently, IBM chose ORiNOCO technology as the wireless adapter in its ThinkPad™ notebook computers.

A complete access point for an 802.11b WLAN is offered by Symbol Technologies (Holtsville, NY) with its Spectrum24™ which consists of an access point with an integrated antenna, PC card for mobile users, and PCI card with an external antenna for desktop systems. The access point can serve as a bridge between a 10BaseT Ethernet hub and a variety of other equipment such as laptops and desktops, and handheld computers. The WLAN uses complementary code keying (CCK), which is a DSSS technology that uses particular codes to represent data bits. This enables receivers to filter out signals that do not use the same codes, such as interference or noise. ••

Value-Added Services

Value-added services, such as custom packaging, screening, and testing are essential ingredients for the successful growth of a long-time technical distributor.

Distribution Group Adds Value To Industry

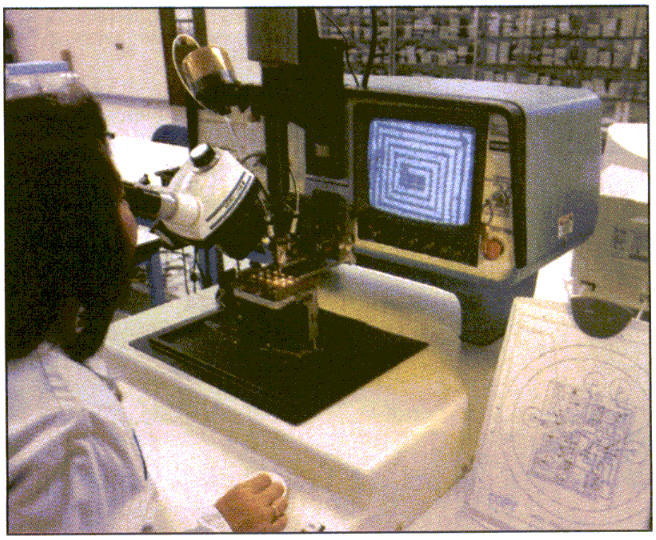
JACK BROWNE
Publisher/Editor

DISTRIBUTION of electronic products has never been trivial. Customers need and want their parts immediately and are rarely content to wait. But with the increased demands and reduced time to market of communications and other fast-growing electronics markets, the distribution business has become extremely complex. One possible solution is the approach taken by the Microwave Technical Solutions (MTS) division of Avnet Electronics Marketing (San Jose, CA), where technical solutions are part of nearly every product shipped.

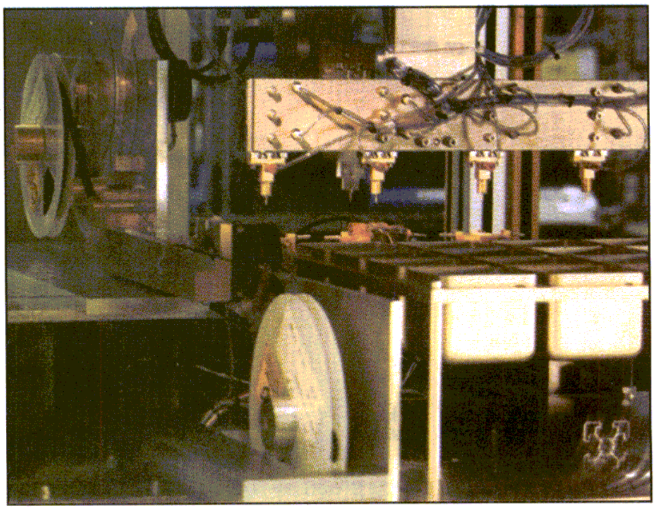
Veteran microwave engineers will remember Avnet's RF & Microwave Product Business Group by its former name, Penstock (see *Microwaves & RF*, June 1997, p. 121). Founded in 1971 by Bruce White, the firm broke new ground in distribution channels by offering not only standard products, but a variety of custom designs and specialized engineering services.

In 1994, Penstock became the RF and Microwave Product Business Group of Avnet Electronics Marketing, one of the largest global distributors of semiconductor products. Rather than playing a passive role in the success of Avnet, the group built upon the innovations fostered by its founder to become a multifaceted distributor/manufacturer/service organization with unique capabilities for commercial and military customers.

According to Avnet RF & Microwave Director Joel Levine, the division accounted for a little more than a \$100 million in annual sales only three years ago. Levine notes, "In North America alone, Avnet RF & Microwave division is now a \$300 million annual operation. In addition to being a distributor of electronic components and semiconductors for companies such as Agilent Technologies, Motoro-



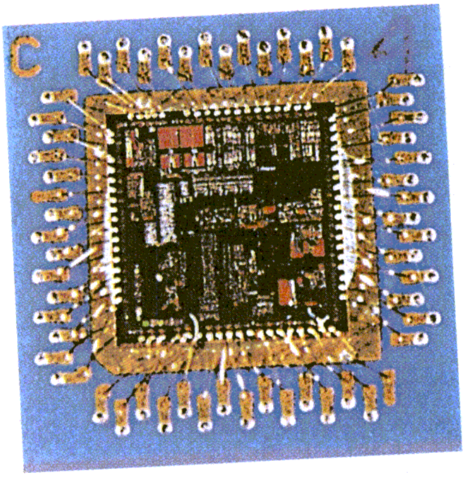
1. Avnet MTS offers numerous value-added services, including bond-pull testing. (Photograph courtesy of Avnet MTS, San Jose, CA.)



2. Avnet MTS offers a variety of specialized automatic test capabilities, including testing for output power, power, gain, and power-added efficiency. (Photograph courtesy of Avnet MTS, San Jose, CA.)

Application Profile

How to make Cell Phones Smaller and Lighter? BGA with Integrated Components using DuPont Green Tape™.



National Semiconductor is a leader in applying the LTCC advantages of high-density interconnect capability, ability to integrate passive components and functions, and low-loss performance. In a recent design, National chose to combine its advanced ICs for wireless communications with Green Tape™, DuPont's brand of LTCC tape dielectric material, to provide optimum performance in the smallest possible package.

Challenge: Decreased Size and Cost, Improved Performance for Wireless Devices

Portable wireless applications have quickly become the main driver for smaller, more cost-effective packaging and interconnects. For example, in the last few years, cell phones have evolved into lightweight, palm-size devices with a host of new functions. Their weight has decreased by a factor of 10, and the wholesale selling price by 75 percent.

OEM designers are now learning that integrating IC and package design to take advantage of the unique properties of Low Temperature Co-fired Ceramic (LTCC) technology can yield decreased size and improved performance in wireless devices.

Solution: Green Tape™ LTCC Allows for High I/O Counts in Chip Scale Package

National's newest chipsets use Green Tape™ packaging capabilities to provide a chip scale package that can accommodate the high I/O counts of a highly integrated RF analog front end using micro BGA (ball grid array) technology. The current package, only 9 x 9 mm, can provide 81 I/Os in a micro BGA array, plus topside pads for wire-bonding that interconnects to the BGA pads on the backside. The high number of I/Os allows for multiple grounds to improve RF performance, while the embedded multilayer structure contains 14 RF bypass capacitors constructed using a combination of high-K and low-K dielectrics.

The performance of the frequency synthesizer function can be enhanced through the use of an embedded VCO resonator that provides a high Q, and therefore lower phase noise, than that available using a VCO resonator located on the silicon.

This approach, co-designing the silicon and LTCC elements to achieve optimized size and performance, demonstrates the use of co-integration for wireless applications requiring smaller package size and higher performance at the lowest possible cost.

For more information, call DuPont at 1-800-284-3382, press 3, or visit the DuPont Microcircuit Materials website at http://www.dupont.com/mcm.

GREEN TAPE™ FODEL® BIROX® DIFFUSION PATTERNING™ LUXPRINT®

CIRCLE NO. 235

la, Philips Semiconductors, and M/A-COM, Avnet Microwave Technical Solutions (MTS) offers a variety of high-reliability and hybrid screening services as well as value-added materials and testing services.

Avnet MTS, for example, can perform a full range of space and military-level screening operations, in its MIL-STD-750 and MIL-STD-883 environmental laboratories. In support of high-reliability screening requirements, Avnet MTS can provide stabilization bakes, temperature cycling, constant acceleration testing, burn-in operations, high-temperature reverse-bias accelerated life testing, full electrical parameter testing, fine- and gross-leak seal testing, bond-pull and die-shear testing, lead and package integrity testing, as well as tape-and-reel packaging (Fig. 1).

The firm offers impressive value-added electrical test capabilities (Fig. 2). These capabilities include RF power testing and device selection at continuous wave (CW) and pulsed power levels for frequencies from 2 to 4000 MHz. Measurements can be made of standard parameters, such as output power, power gain, and power-added efficiency (PAE), as well as custom measurements, such as third-order intercept point (IP3).

"We currently have more than 90 RF electrical engineers globally, located in the field and in design centers throughout the world in Bulgaria, Singapore, New Zealand, and India to help us serve our customers with their specialized engineering needs," according to Levine.

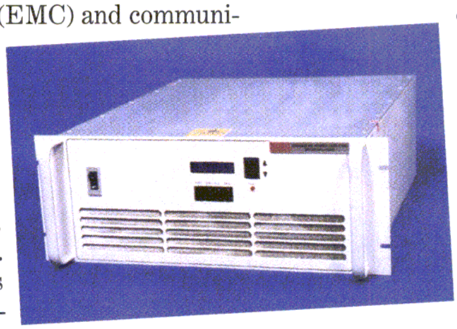
The company offers a variety of materials-related services, including tape-and-reel packaging, parts kitting, tape-and-reel marking, package lead trimming and forming, solder dipping, as well as color marking. Levine notes that if a manufacturer's production is slowed by symptoms such as excess preproduction non-value-added activities (including labeling and parts kitting), excessive tuning-in production, failures at final test, excessive rework on finished units, and excessive field returns, Avnet MTS can probably help through their value-added services.

Effective use of the Internet's World Wide Web has also made it easier for Avnet MTS to serve its customers. "The website is geared to help simplify our customers' searches for a solution," says Levine. "A customer can use our website to browse through the information, pick products, and download data sheets," he notes. "You can also use the website to perform a search through our product listings according to specific package types," he adds.

For more information on Avnet MTS, contact the company at the following address, or visit the company's website to explore the many materials, screening, and measurement services that are offered. Avnet Electronics Marketing, 6321 San Ignacio Ave., San Jose, CA 95119; (408) 360-4000, Internet: http://www.em.avnet.com. ●●

Power amp boosts MMDS

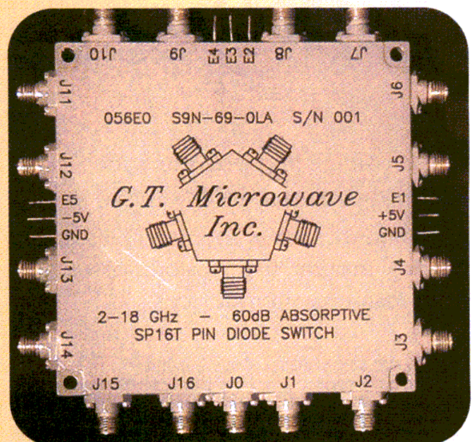
The model 4020 narrow-band, high-gain power amplifier (PA) boosts 2.5-to-2.7-GHz, multichannel-multipoint-distribution-system (MMDS) signals for electromagnetic compatibility (EMC) and communications applications. Applications include electromagnetic-interference (EMI)/RF-interference (RFI) susceptibility, high-power driver, satellite ground-station amplifier, and lab applications. At 0-dBm input, the amplifier achieves a saturated output power of 115 W. An auto-switching AC input allows the amplifier to be used transparently in 120- and 220-VAC environments. Available options allow the user to control the amplifier remotely through a custom controller or personal computer (PC). The custom, microprocessor-based-controller option includes a liquid-crystal display (LCD) and features functions such as automatic level control, VSWR protection, gain control, and forward monitoring. PC-communication-port options are IEEE-488, RS-232, and RS-422. **Ophir RF Corp., 5300 Beethoven St., Los Angeles, CA 90066; (310) 306-5556, FAX: (310) 577-9779, Internet: http://www.ophirrf.com.**



CIRCLE NO. 57 or visit www.mwrf.com

PIN-diode switch provides amplitude matching

The model S9N-0LA positive-intrinsic-negative (PIN)-diode, single-pole, 16-throw switch provides ±5-deg. phase matching and ±0.5-dB amplitude matching among its 16 output ports over a bandwidth of 2 to 18 GHz. The switch provides a minimum isolation of 60 dB, a maximum insertion loss of 6 dB, and a maximum VSWR of 2:1. It has a total switching speed of less than 2 μs and offers transistor-transistor-logic (TTL)-compatible control logic. The device switches RF at

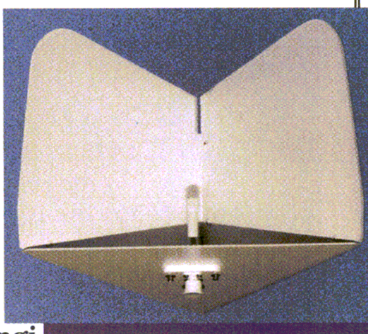


power levels up to +20 dBm and can tolerate RF power as high as 1 W CW. It draws +850/−200 mA from a ±5-VDC power supply and is housed in a package measuring 3.5 ×3.5× 0.88 in. (8.89 × 8.89 × 2.24 cm). **G.T. Microwave, Inc., 2 Emery Ave., Randolph, NJ 07869; (973) 361-5700, FAX: (973) 361-5722, Internet: http://www.gtmicrowave.com.**

CIRCLE NO. 58 or visit www.mwrf.com

Antenna serves ISM band

The model 0012-800 antenna operates in the industrial-scientific-medical (ISM) band from 902 to 928 MHz and is ideal for use in dense, urban, wireless-local-area-network (WLAN) links. The antenna can be horizontally or vertically polarized, and its trihedral-reflector design is said to optimize the front-to-back ratio. It has a minimum front-to-back ratio of 27 dB within its rear hemisphere, a nominal gain of 9 dBi, and a cross polarization of −20 dB. AZ pattern and EL pattern are 60-deg. beamwidth, and maximum VSWR is 2:1. **Seavey Engineering Associates, Inc., 28 Riverside Dr., Pembroke, MA 02359; (781) 829-4740, FAX: (781) 829-4590, Internet: http://www.seaveyantenna.com.**



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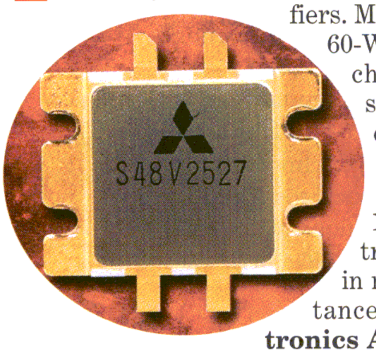
GaAs FETs target satellite amps

Four gallium-arsenide (GaAs) heterojunction field-effect transistors (HFETs) are said to boost output power, gain, and efficiency in base-station and satellite-communications amplifiers. Model MGFS48V2527 as shown is a 60-W, push-pull transistor for multichannel-multipoint-distribution-system (MMDS) transmitters operating from 2.5 to 2.7 GHz. Typical output power at P2dB is 48 dBm and typical power gain is 10 dB across the band. All four transistors are hermetically sealed in metal-ceramic, low-thermal-resistance packages. **Mitsubishi Electronics America, Inc., 1050 E. Arques Ave., Sunnyvale, CA 94086; (480) 730-5900, FAX: (480) 737-1129, Internet: http://www.mitsubishi.com.**



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WAP Needs More Zap To Grow

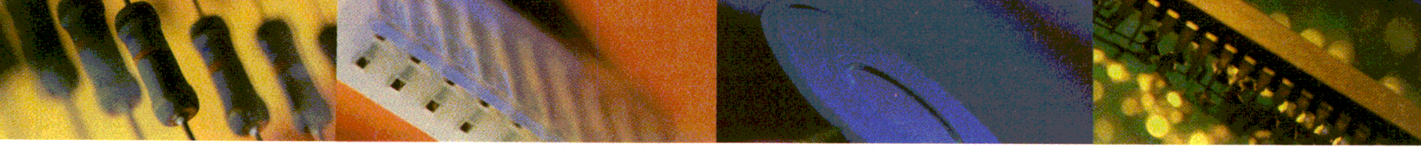
It is being heralded as the hot new marriage between the Internet and wireless technology, but so far, the wireless application protocol (WAP) has received a cool reception from users in Europe. WAP is a standard for allowing small terminals such as mobile phones, pagers, and personal digital assistants (PDAs) to receive Internet content and advanced data services. Most of WAP's current problems can be attributed to growing pains similar to those that users experienced in the early Internet days.

For one thing, according to a report in *The Wall Street Journal*, WAP-enabled phones have not been rolling off production lines in the quantities promised by major manufacturers such as Nokia, Ericsson, and Motorola. Part of the difficulty stems from factory decisions to extend testing time by up to three months to ensure that the phones work on existing networks. A second headache has been the failure of European telecommunications companies to provide a consistent and useful set of services.

Many phone users find that services are either not readily available or difficult to use. For example, a Swedish user bought a WAP phone to receive e-mails while in transit. This is easy with WAP simply by dialing into a portal and calling up the messages. But returning calls is more time-consuming and phoning the old-fashioned way is simpler and faster than with WAP.

Internet services are similarly eschewed according to mobile telecommunications provider Deutsche Telekom AB, whose records indicate that a typical WAP user accesses the net only once a week. And only slightly more than 1 percent of subscribers use WAP services at all. Many customers claim that the services are not always reachable and difficult to use. Yellow Page listings and other text-based information have been disappointing. Adding to user frustrations is the high price of a WAP phone—approximately $600. What is more, a handheld device is hardly the best appliance for downloading large amounts of information. The screen is small, has generally poor visual quality, memory is limited, and entering information from a keypad is tedious.

Does all this mean that WAP is destined for an early demise? More likely, the service's stumbling and fumbling is the common order of business when a new technology is rolled out. The current difficulties are putting a hitch in the plans of equipment manufacturers and service providers alike. But in the long run, the roadblocks will fall and the marriage of the Internet and wireless will be one that will turn out to have been made in telecommunications heaven. ••



Contracts

Giga-tronics, Inc.—Announced that its Instrument Division had received an order in excess of $2.3 million for its three-slot VXI microwave synthesizer.

EMS Technologies, Inc.—Announced, through its Space & Technology Group in Montreal, Canada, a $6 million ($9.5 million Canadian) contract with the Canadian Space Agency to provide the research and development (R&D) for a demonstration digital-mesh-connectivity system on Telesat's Anik F2 satellite.

Motorola Inc.'s Network Solutions Sector (NSS)—Signed a $123.5 million digital cellular-network expansion contract in China's Hunan Province with China's Hunan Mobile Communication Corp. (Hunan MCC).

Harris Corp.—Announced the signing of a $3.7 million contract to supply a cellular and fixed wireless system to Superior Wireless, a private telecommunications corporation located in Thunder Bay, Ontario, Canada. Superior Wireless will provide mobile and fixed wireless service to 50,000 people in 10 communities and along 1180 km of highway in Northwestern Ontario.

Fresh Starts

Volex Group, plc.—Has formed a new operating unit for RF products to be known as Volex RF Technologies. The unit will provide the engineering knowledge, customer support, and enhanced ability in manufacturing processes for RF products on a global basis for the rapidly growing telecommunications-market segment.

Vitesse Semiconductor Corp.—Announced that it has completed the acquisition of all equity interests of SiTera, Inc. on May 31 for $750 million in common stock.

Boldt Metronics International, Inc. (BMI)—Announced that it is expanding operations to Schaumburg, IL. The larger 161,000-sq.-ft. facility in Schaumburg will offer a greater capacity to meet its growing needs.

Agilent Technologies, Inc.—Announced a global license and support agreement with Motorola to supply software tools, training, and technical support for the design of Motorola's next-generation communications printed-circuit boards (PCBs), integrated circuits (ICs), and systems worldwide. Under the agreement, all Motorola designers worldwide will have broad access to Agilent's full line of electronic-design-automation (EDA) simulation tools.

Wireless Solutions—Became a member of the Wireless Ethernet Compatibility Alliance (WECA) in America. WECA works to ensure that all products and systems, which are based on Wireless LAN IEEE 802.11b (high rate), should be compatible with one another regardless of the manufacturer.

Cambridge Positioning Systems (CPS) and SignalSoft—Have signed a joint marketing agreement to promote mobile-location solutions to the Global Systems for Mobile Communications (GSM) industry. The alliance requires that both companies promote each other's services, and the benefits of mobile-location services (MLS) in general.

Hittite Microwave Corp.—Announced the appointment of Dytec, Inc. to represent Hittite in Ohio, Indiana, Kentucky, Minnesota, and Wisconsin.

Fujitsu Compound Semiconductor, Inc. (FCSI)—Named Arthur, Harris & Associates to sell Fujitsu's microwave, lightwave, and gallium-arsenide (GaAs) integrated-circuit (IC) data-communications products. The FCSI territory represented by Arthur, Harris & Associates includes the states of North Carolina, South Carolina, Georgia, Alabama, Mississippi, and Tennessee.

Qualcomm and NetZero, Inc.—Announced that Qualcomm has invested $144 million in NetZero. As a result of the transaction, Qualcomm now owns approximately 10 percent of NetZero's outstanding common stock.

The Society of Cable Telecommunications Engineers (SCTE)—Announced the formation of a standards-developing Cable Applications Platform (CAP) Subcommittee. CAP's mission is to explore the need for SCTE involvement in the development of standards for applications platform development through coordination with the International Telecommunications Union—Telecommunications (ITU-T), the Advanced Television Systems Committee (ATSC), and other related organizations.

The EMC/Aerospace Unit of TÜV Product Service, Inc.—Has opened an electromagnetic-compatibility (EMC) testing laboratory in Boston, MA offering East Coast avionics, aerospace, and defense manufacturers a local state-of-the-art testing facility, providing complete comprehensive testing services.

Gage Applied Sciences, Inc.—Has become a wholly owned subsidiary of Tektronix, Inc., and will operate under the name Gage Applied, Inc.

SiCon Video Corp.—Launched its "Helix" project to develop the world's most comprehensive V^{2}oIP™ for video- and voice-communications transport controller chips.

RangeStar Wireless—Announced the completion of its mezzanine round of funding in which it raised $25 million. The round included new investments from Dell Computer Corp., Intel Capital, and Summit Accelerator Fund. RangeStar also received continuing investments from existing shareholders led by ComVentures, GSM Capital, BJ Cassin, and RWI Group. RangeStar will employ the funds to continue development of its Invisible Antennaä technology, which enables enhanced connectivity for wireless devices including cellular/personal communications services (PCS), wireless local-area network (WLAN), Global Positioning Systems (GPS), and Bluetooth products.

Askey Computer Corp.—Has joined the Wireless Ethernet Compatibility Alliance (WECA). By joining WECA, an organization devoted to the testing of wireless local-area-network (WLAN) products, Askey will be able to submit its WLAN product, the WLC010, for WECA certification.

Trimble Navigation Ltd.—Has made a minority investment in Data on Air, Inc. of Orlando, FL, a privately held advanced wireless data-solutions company. Trimble also entered into a strategic marketing agreement with Data on Air. The agreement is intended to allow Trimble to enter into Internet-enabled fleet-management applications and become a full-service provider of wireless-location services for the mobile workplace.

Things are moving fast. Can you keep up with the demands of tomorrow's CDMA handset designs?

Now you can. Design faster and with more confidence with Agilent Technologies' flexible **CDMAdvantage** RF chipsets for your handset designs.

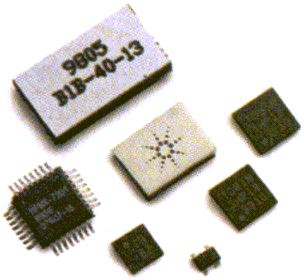
The **CDMAdvantage** complete product family of RF solutions is designed to work together in DBTM, DBDM and SBSM handsets. We supply extensive characterization data as well as complete receive and transmit chain demonstration boards. All to enable you to make better decisions faster– this is the **CDMAdvantage**.

Our transmit chain maximizes efficiency, from the adaptive current control on the upconverters to medium power optimization on the power module. The receive side has a switched LNA and an adjustable linearity downconverter to ensure compliance to even the most rigorous system specs.

Agilent RF semiconductors are already in millions of CDMA phones. And we are continuously adding more capacity, so you can move with the speed that the market demands.

At Agilent we listen to your needs to help you reach the right solution– fast.

World's smallest PCS duplexer,
Integrated DBTM Power Amplifier,
High output-level Upconverters,
Variable linearity Downconverters,
Miniature LNA with Bypass Switch



See how fast you can move with the flexiblility of the CDMAdvantage RF chipset. Visit our website to see product specifications and our latest technologies.

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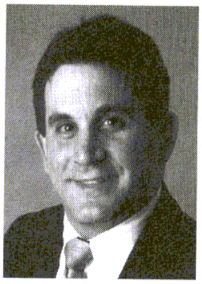
STMicroelectronics—Enrico Villa to corporate vice president, director of European region; formerly corporate vice president, region five. Also, Carmelo Papa to corporate vice president, region five; formerly director of product marketing and customer service for transistors and standard integrated circuits (ICs).

Current Technology, Inc.—Ronald L. Sutton to national sales director for the Joslyn™ branded AC surge-protection products; formerly regional manager for Teccor Electronics. Also, Michael Lucero to director of business development; formerly national account manager for Veeder-Root Corp.

Quad Systems Corp.—Cosmo Losco to vice president of operations; formerly senior director of operations.



LOSCO

WADORS

Wyle Electronics—Ken Wadors to president and CEO of RF Vision™; formerly senior vice president and director of RF/small-signal business for Avnet, Inc.

MCE Companies, Inc.—Stephen W. Shpock to chief operating officer; formerly vice president and general manager of Litton-Electron Devices' Williamsport, PA operation.

CTS Corp.—Roger E. Rickey to vice president and general manager of CTS' frequency products business; formerly vice president and general manager of the Transportation Electronics Division at Joseph Pollack Corp.

Varian, Inc.—Sergio Piras to corporate vice president; formerly general manager of Varian Vacuum Technologies' operation in Torino, Italy. Also, Wilson Rudd to corporate vice president; formerly general manager of the Varian Electronics Manufacturing business in Tempe, AZ and Poway, CA.

Advanced Thermal Solutions—Roger Alan Emigh, Ph.D. to director of marketing for the West Coast operations; formerly employed as technical marketing manager and research manager with Johnson Matthey Electronics.

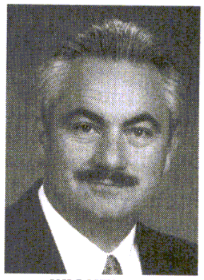
Technology Forecasters, Inc.—Beth Downey to vice president and chief financial officer; formerly vice president of administration. Also, Eric Miscoll, Ph.D. to chief operating officer; formerly senior consultant.

MVX.COM—Scott Schaefer to president and chief executive officer; formerly president of Interconnection Services at BellSouth Corp.

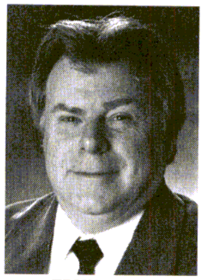
Scientific-Atlanta, Inc.—Sam Lim to president and general manager of the Media Networks business; formerly vice president of research and development (R&D) and product management for Media Networks.

Barry Industries, Inc.—James K. Wood to director of marketing; formerly in charge of national marketing and technical support for several Canadian and British communication-equipment manufacturers.

Dow-Key Microwave—David W. Wightman to president; formerly president of Sunbank.



WIGHTMAN



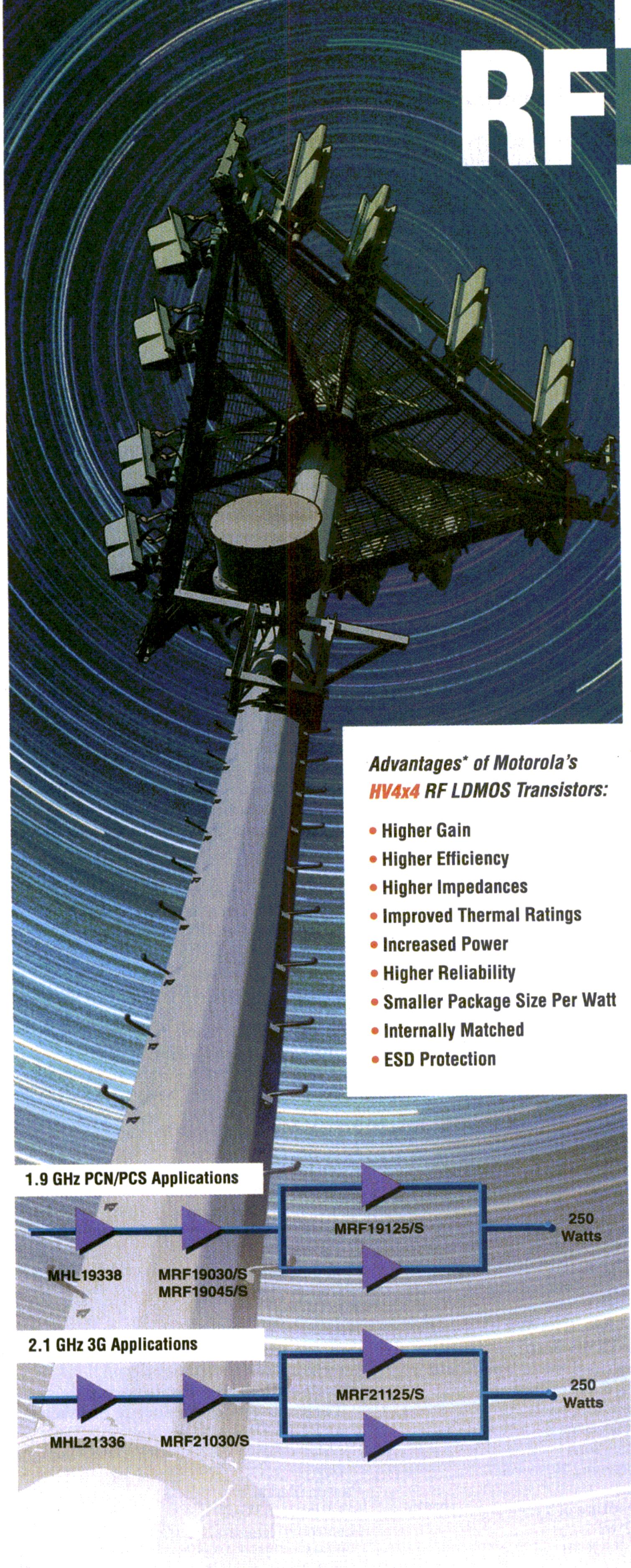
FRYKLUND

Alpha Industries—Dave Fryklund to general manager of RF gallium-arsenide (GaAs) control integrated-circuit (IC) products; formerly worked as director of commercial IC products.

SignalSoft Corp.—Steven D. Adams to vice president of operations; formerly technical business manager with InvisiX Corp. Also, Chris Mudock to vice president of quality assurance; formerly director of quality assurance. In addition, Doug Demmel to vice president of sales for the Americas; formerly director of North American sales.

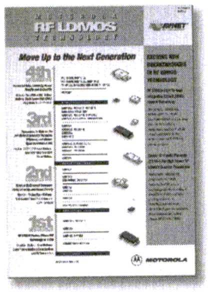
RF LDMOS

Move Up to the Next Generation with Motorola



As digital standards increasingly dominate the wireless communication market, Motorola's RF LDMOS continues to be the industry's technology of choice. And, Motorola continues to develop and improve its patented RF LDMOS design and has recently extended performance on new 4th generation RF LDMOS products. The newest product introductions incorporate the latest process improvements and the most advanced technology. With a dedicated business group for RF & Microwave products and services, local market resources for support, plus one of the largest RF inventories in the world, Avnet Electronics Marketing is uniquely positioned to be your source for RF LDMOS products from Motorola.

DigitalDNA™ from Motorola



For a copy of Motorola's RF LDMOS Selector Guide, visit our Web site at www.em.avnet.com/rfm/motldmos.html or call **1-800-261-9602 ext. 302.**

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* Compared to previous generations of Motorola RF LDMOS

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Short Courses

Asynchronous Transfer Mode Overview—Understanding ATM's Potential and Value
August 29-30 (Seattle, WA)
Telcordia Technologies (formerly Bellcore)
Learning Services
Registrar's Office, Room F212
6200 Route 53
Lisle, IL 60532-3198
(800) 832-2463, FAX: (630) 960-6300
e-mail: teachme@telecodia.com
Internet: http://www.800teachme.com

Software Radio
September 4-5 (Oxford, England)
University of Oxford
Department of Continuing Education
Oxford OX1 1PT, England
Dee Broquard
0-1865-286959, FAX: 0-1865-286956
e-mail: dee.broquard@conted.ox.ac.uk

2000 International Symposium on Advanced Radio Technologies
September 6-8 (Boulder, CO)
National Institute of Standards and Technology
John J. Lemmon
(303) 497-3414
Electronic registration: https://sales.nist.gov/conf/secure/CONF217/conf_register.htm
Internet: http://www.its.bldrdoc.gov

Local Area Networks: Implementation and Configuration
September 12-15 (Washington, DC)
Learning Tree International
Reston, VA 20190-5630
(800) 843-8733, FAX: (800) 709-6405
Internet: http://www.learningtree.com

Embedded and Real-Time Systems
September 19-21 (Los Angeles, CA)
UCLA Extension, Department of Engineering
109995 Le Conte Ave.
Los Angeles, CA 90024-2883
(310) 825-3344, FAX: (310) 206-2815
e-mail: mhenness@unex.ucla.edu
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Fundamentals of Data Communication
October 2-4 (Madison, WI)
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Meetings

ISSSTA 2000 IEEE Sixth International Symposium on Spread Spectrum Techniques & Applications
September 6-8 (Parsippany, NJ)
IEEE Communications Society and Center for Communications
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Clare Naporano
New Jersey Institute of Technology (NJIT)
Newark, NJ 07102
(973) 596-8474, FAX: (973) 596-8473
e-mail: Clare@megahertz.njit.edu.

15th Annual Symposium and Mini-Show: Microwaves in Wireless and Satellite Communication in the New Millennium
September 7 (Hanover Marriott Hotel, Whippany, NJ)
The Institute of Electrical and Electronics Engineers, North Jersey Section, MTT/AP-S Chapter
Kirit Dixit
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(973) 492-0371
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EMC Europe 2000—4th European Symposium On Electromagnetic Compatibility
September 11-15 (Brugge, Belgium)
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Type	Case	Ratings					Characteristics, typical					
		V_{CEO} (V)	I_C (mA)	P_{TOT} (mW)	f_T (GHz)	I_T (mA)	F (dB)	G_{um} (dB)	@ (MHz)	F (dB)	G_{um} (dB)	@ (MHz)
PMBTH10	SOT23	25	40	400	0.6	1-20						
PMBTH81	SOT23	20	40	400	0.6	1-20						
BFS17W	SOT323	15	50	300	1.6	2-20	4.5		500			
BFR92AT	SC-75*	15	25	300	5	3-30	2	14	1000	3	8	2000
BFT92W	SOT323	15	35	300	4	3-30	2.5	17	500	3	11	1000
BFR93AT	SC-75*	12	35	300	5	5-40	1.5	13	1000	2.1	8	2000
BFQ67T	SC-75*	10	50	300	8	3-30	1.3	13	1000	2.2	8	2000
PBR941	SOT23	10	50	360	8	3-30	1.4	15	1000	2	9.5	2000
PRF947	SOT323	10	50	250	8	3-30	1.5	16	1000	2.1	10	2000
PRF949	SC-75*	10	50	150	8	3-30	1.5	16	1000	2.1	10	2000
PRF957	SOT323	10	100	270	8	5-50	1.3	15	1000	1.8	9.2	2000
BFR505T	SC-75*	15	18	150	9	1-10	1.2	17	900	1.9	10	2000
BFR620T	SC-75*	15	70	300	9	3-30	1.1	15	900	1.9	9	2000
BFC520	SOT353	8	70	1000	9	3-30	1.3	31	900	1.5	19	2000
BFE520	SOT353	8	70	100	9	3-30	1.2	17	900	1.9	10	2000
BFM520	SOT363	8	70	100	9	3-30	1.1	15	900	1.9	9	2000
BFG520W/X	SOT343	15	70	500	9	3-30	1.6	17	900	1.8	11	2000
BFG540W/X	SOT343	15	120	500	9	10-60	1.9	16	900	2.1	10	2000
BFG11W/X	SOT343	8	500	760	9	50-150					7	1900
BFG403W	SOT343R	4.5	3.6	16	17	5-5	1	20	900	1.6	22	2000
BFG410W	SOT343R	4.5	12	54	22	2-15	.9		900	1.2	22	2000
BFG425W	SOT343R	4.5	30	135	22	3-30	.8		900	1.2	20	2000
BFG480W	SOT343R	4.5	250	360	18	30-150	1.2		900	1.8	16	2000
BFG21W	SOT343R	4.5	200	600	18	50-250					12	1900

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CIRCLE NO. 292

Combining technologies can improve wireless performance

Wideband code-division multiple access (WCDMA) is the front-running technology for future third-generation (3G) wireless applications, but certain types of interference could hamper its performance. A possible solution offered by Kwang-Cheng Chen and Shan-Tsung Wu of the National Taiwan University (Taipei, Taiwan) involves combining CDMA with orthogonal frequency-division multiplexing (OFDM). Multicarrier transmission, known as OFDM, is often used to combat frequency-selective multipath fading. The researchers claim that an OFDM-CDMA union has a number of benefits such as permitting finer partitioning of radio resources in the time, frequency, and code domains to provide more effective radio-resource allocation. Three different methods are presented to combine the technologies with an eye toward developing a programmable architecture that could be based on a single hardware and software platform. This could serve as the foundation for the design of a software radio. See "A Programmable Architecture for OFDM-CDMA," *IEEE Communications Magazine*, November 1999, Vol. 37, No. 11, p. 76.

Micromechanical filter design is explained

High-Q bandpass filters used in RF and IF stages of heterodyne receivers generally use off-chip, mechanically resonant components such as crystal filters and surface-acoustic-wave (SAW) devices. Being off-chip, however, increases the size and reduces the performance of receivers. Frank D. Bannon III, John R. Clark, and Clark T.-C. Nguyen of the University of Michigan (Ann Arbor, MI) show how new micromachining technologies make it possible to miniaturize and integrate highly selective filters alongside transistors with the eventual goal of miniaturizing superheterodyne receivers. They begin with a qualitative description of micromechanical filter structure and operation and move on to the details of a high-frequency micromechanical resonator. A step-by-step description of a filter design is presented, and implemented by building a surface-micromachined bandpass filter in silicon (Si). The fabrication reveals the problems in making such components and the novel technique to solve it. See "High-Q HF Microelectromechanical Filters," *IEEE Journal of Solid-State Circuits*, April 2000, Vol. 35, No. 4, p. 512.

Test technique measures two parameters at once

A problem in microwave source-pull measurements is making simultaneous measurements of the source- and device-under-test (DUT) reflection coefficients. Researchers GianLuigi Madonna, Andrea Ferrero, Marco Pirola, and Umberto Pisani of the Dipartimento di Elettronica, Politecnico di Torino (Torino, Italy) developed a novel solution to the problem that permits the parameters to be measured together. Currently, the measurement is performed in two steps that involve switching the microwave signal source, whereas the new method requires no switching or disconnection of the source. The novelty of the technique is based on an input reflectometer, a linear and invariant microwave network having six ports. It can be shown mathematically that any wave incident or reflected by the reflectometer is a function of only three measured quantities. Thus, the input from a generator need not be disconnected to make the measurement. A calibration procedure is required to determine error coefficients used to calculate reflection coefficients. Experimental results proved the accuracy of the method compared to traditional techniques. It is useful for fast and accurate device characterization. See "Testing Microwave Devices Under Different Source Impedance Values—A Novel Technique for On-Line Measurement of Source and Device Reflection Coefficients," *IEEE Transactions on Instrumentation and Measurement*, April 2000, Vol. 49, No. 2, p. 285.

Mobile antenna arrays can act in strange ways

Antenna arrays in mobile communications experience effects not present in other environments due to complicated and random scattering events where received fields can come from many different directions. This means that antenna characteristics such as gain, diversity and channel capacity must be considered in new ways according to Jorgen Bach Andersen of the Center for Personkommunikation at Aalborg University (Aalborg East, Denmark). The fundamental question for Andersen is that since the information- carrying signal is scattered in many directions, how can the antenna elements be organized to maximize the power transfer. The various approaches depend on the scattering angles (which are unpredictable), but the author suggests that power transfer can be maximized by jointly adjusting the antenna weights in the arrays. See "Antenna Arrays in Mobile Communications: Gain, Diversity and Channel Capacity," *IEEE Antennas and Propagation Magazine*, April 2000, Vol. 42, No. 2, p. 12.

Analyze VCOs And Fractional-N Synthesizers

A method of analyzing phase noise, and spurious in oscillators and synthesizers takes into account loaded and unloaded quality-factor (Q) values.

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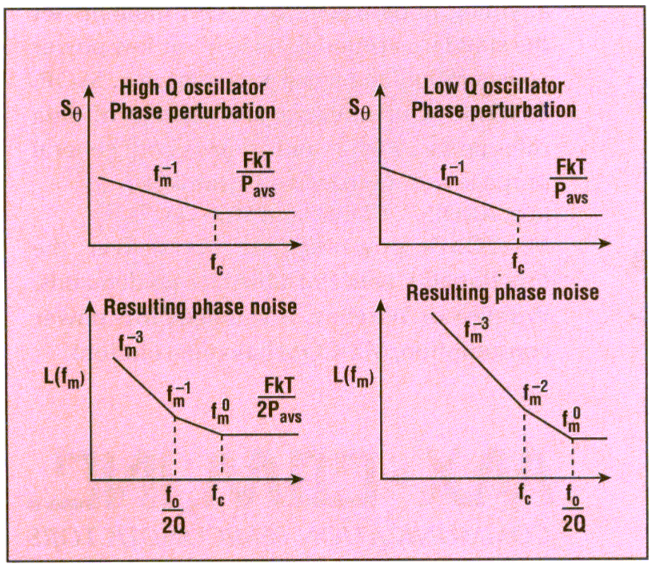
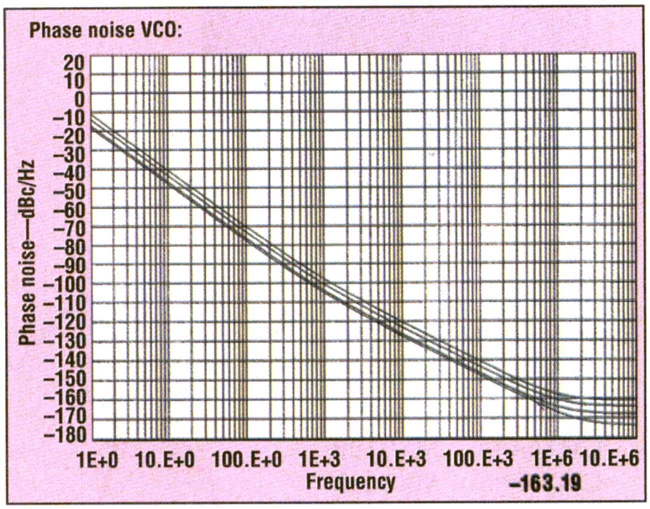
LINEAR analysis of frequency synthesizers and voltage-controlled oscillators (VCOs) can lead to improved phase-noise performance. In the analysis technique that follows, loaded and unloaded quality-factor (Q) values are being used together for the first time. This new analysis method shows that there is some optimum value for phase noise between the values of loaded and unloaded Q, and deviation for either side of this optimum value will result in deterioriation of the phase-noise performance. In addition to exploring this new analysis technique, this article will explore the performance limitations of sigma-delta-converter-based fractional-N frequency synthesizers. These limitations are set mainly by the performance of the dividers, phase detectors, and the available minimum-division ratios of the fractional divider.

Fractional-N frequency synthesizers are capable of low phase-noise performance in modern communications systems. Noise inside the loop bandwidth is controlled mainly by the noise of the active components and the upward multiplication of the reference source's noise. The chief contributors to noise are the phase detector, the reference standard, the reference dividers, and the operational amplifier. The loop bandwidth



1. These spectral plots show oscillator phase noise for high- and low-Q resonators.



2. Phase noise is shown as a function of loaded Q over unloaded Q (Q_{load}/Q^0). The lowest curve represents a Q of 200, the next-highest curve is a Q of 300, then 350 and finally 380.

depends on the reference frequency. A good combination is a reference frequency of more than 10 MHz and loop-filter bandwidth of 15 kHz. This bandwidth cleans up the noise of the VCO, specifically microphonic events. In fractional-N synthesizers, it is not uncommon to use a dual-time-constant filter where the wider bandwidth helps with frequency acquisition while the lower bandwidth is effective for phase acquisition. The loop bandwidth of the filter cannot be made infinitely wide and must be set at the crossover point, where the phase noise of the active components, as well as the reference, is equal to the VCO noise by itself. If a sigma-delta converter is used its noise-corner frequency also becomes relevant.

In designing a low-noise oscillator or synthesizer from 400 to 4000 MHz, the main resonators of choice are ceramic resonators and high-Q printed microstrip/stripline resonators. For conventionally sized oscillators requiring the best phase noise, the ceramic-resonator-oscillator (CRO) approach is preferred. In most cases, a Colpitts oscillator configuration provides good results. Grounded-base circuits exhibit severe stability problems and even the Colpitts oscillator, based on its emitter circuitry, occasionally shows unwanted resonances to 6 GHz with high-gain transistors. Once oscillation has been established at the proper frequency, the immediate question is how much phase noise does the transistor possess. The phase noise is determined by the large-signal flicker corner frequency of the oscillator device, with typical values for transistors being AF = 2, KF = 1×10^{-5} to 1×10^{-9}. (These are the equivalent SPICE parameter values which determine the corner frequency.)

The flicker-corner frequency is defined as the frequency where the first break point on the phase-noise plot occurs if a low-Q resonator is

Table 1: Gauging flicker corner frequency versus collector current

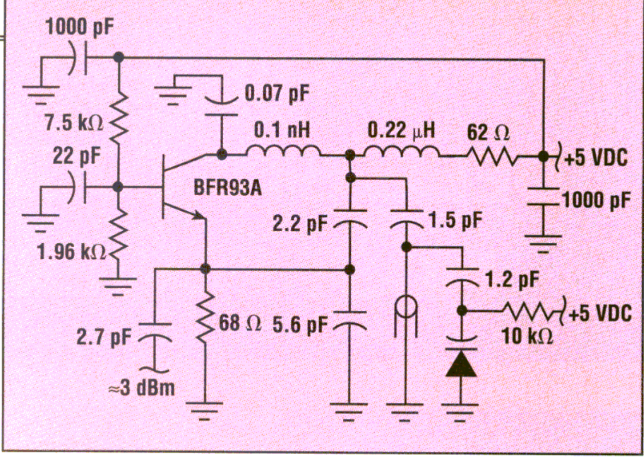
I_C (mA)	f_C (kHz)
0.25	1.0
0.5	2.74
1.0	4.3
2.0	6.27
5.0	9.3

Source: Motorola

Table 2: Scanning phase noise versus loaded Q

Q_{load}	50	100	200	300	400	500	550	580
L(dBc/Hz)	−127	−133	−139	−140	−141	−142	−141	−134

used. Figure 1 shows the high- and low-Q cases with phase noise plotted in Fig. 2. While the authors still believe that the accurate way of determining the actual phase noise is the use of a nonlinear simulator with good active-device modeling, first-order linear analysis is also legitimate approach. There is no linear equation for the bias-dependent flicker-corner frequency unless parameters AF and KF are used in a nonlinear simulator, but the linear expression allows the use of the "expected" flicker-corner frequency in the linear equation legitimately. The third component is the tuning diode, operated in a reverse-bias condition. The tuning diode also has AF and KF values, with AF being approximately 2 and KF being about 1×10^{-15}. In linear terms, it is possible to assign the diode an equivalent noise resistance which, dependent on the technology, varies between 200 Ω and 20 kΩ.



3. This schematic diagram represents a high-performance CRO that can be used from 500 to 2000 MHz (or 804.6 MHz as in the example).

In 1966, Leeson introduced a linear approach for the calculation of oscillator phase noise. His formula[1] was extended by Scherer of Hewlett-Packard Co. (Palo Alto, CA), who added the flicker-corner frequency calculation. Rohde subsequently added the VCO term.[2] As a result, the phase noise of a VCO is:

$$L(f_m) = 10 \log \Big(\{ 1 + [f_0^2 / (2 f_m Q_{load})^2] \} \times [1 + (f_c / f_m)] \times \{ FkT / 2P_{sav} \times [1 - (Q_{load} / Q_o)] \} + (2kTRK_0^2 / f_m^2) \Big) \quad (1)$$

where:

$L(f_m)$ = the ratio of sideband power in a 1-Hz bandwidth at f_m to the total power in decibels (spectral density),

f_m = the frequency offset (in Hz),

f_0 = the center frequency,

f_c = the flicker frequency,

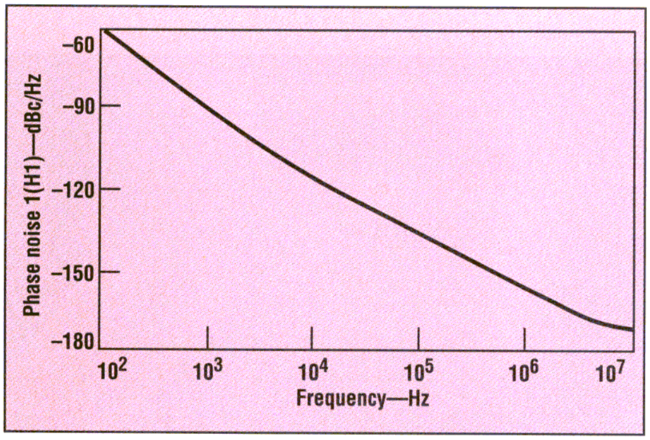
Q_{load} = the loaded Q of the tuned circuit,

Q_0 = the unloaded Q of the tuned circuit; $Q_0 > Q_{load}$,

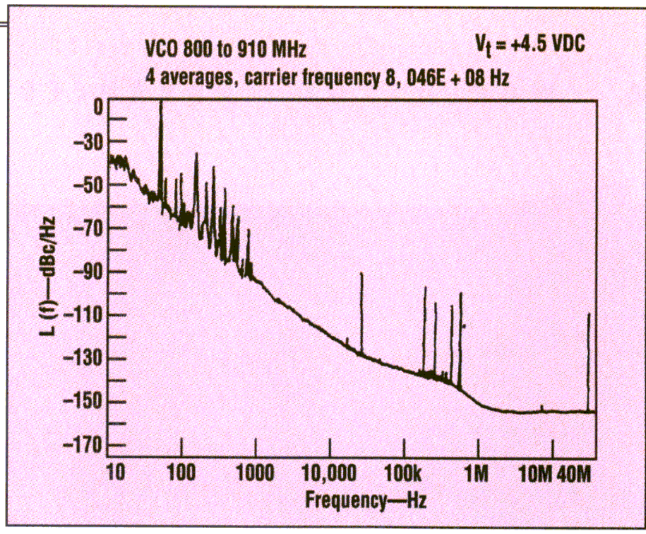
F = the noise factor,

$kT = 4.1 \times 10^{-21}$ at 300 K (room temperature),

P_{sav} = the average power at the



4. This is the simulated phase noise for the ceramic-resonator oscillator of Fig. 3.



5. This is the measured phase noise for the oscillator of Fig. 3. The pedestal above 100 kHz results from the reference oscillator, a model HP 8662 signal generator from Hewlett-Packard Co. (Palo Alto, CA).

oscillator's output,

R = the equivalent noise resistance of the tuning diode (typically 200 Ω to 10 kΩ), and

K_0 = the oscillator voltage gain.

When adding an isolating amplifier, the noise of an inductive-capacitive (LC) oscillator is determined by:

$$S_\phi(f_m) = \left[a_R F_0^4 + a_E(F_0/2Q_L)^2\right]/f_m^3 + \left[(2GFkT/P_0)(F_0/2Q_L)^2\right]/f_m^2 + (2a_R Q_L F_0^3)/f_m^2 + a_E/f_m + 2GFkT/P_0 \quad (2)$$

where:

G = the compressed power gain of the loop amplifier,

F = the noise factor of the loop amplifier,

k = Boltzmann's constant,

T = the temperature (in K),

P_0 = the carrier-power level (in W) at the output of the loop amplifier,

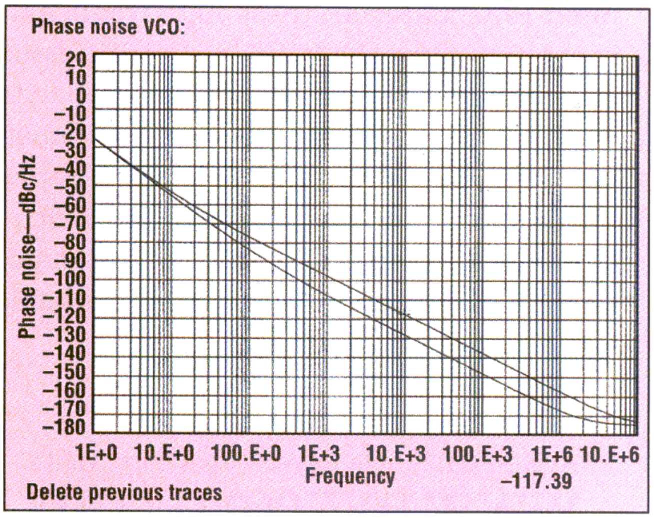
F_0 = the carrier frequency (in Hz),

$Q_L(= \pi F_0 \tau_g)$ = the loaded Q of the resonator in the feedback loop, and

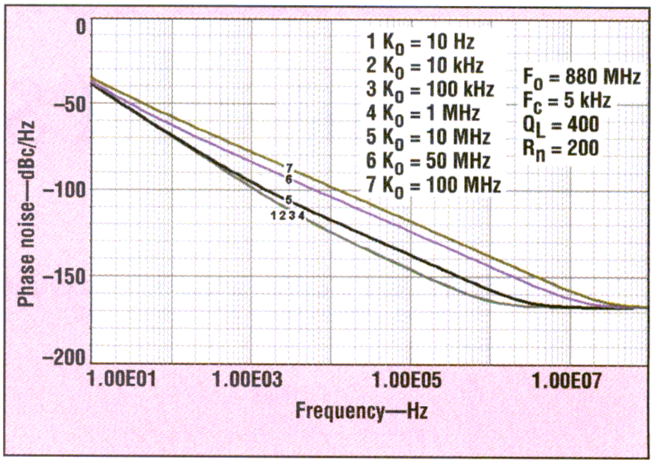
α_R and α_E = the flicker-noise constants for the resonator and loop amplifier, respectively.[1,3]

Table 1 shows the flicker-corner frequency, f_C, as a function of I_C for a typical small-signal microwave bipolar-junction transistor (BJT).

Figure 3 shows a typical CRO. Using the result of a nonlinear simulator, it is possible to compare measured and predicted phase noise (Figs. 4 and 5). From the phase noise, it is possible to determine the loaded Q and, within limits, the corner-flicker frequency. The flicker-corner frequency will shift for a high-Q resonator. For a loaded-Q (Q_{load}) value of 200 at a center frequency of 800 MHz, and a flicker frequency (f_C) of 1000 Hz, similar results are measured as obtained with the linear equation (Fig. 6). If the diode is disconnected, the phase noise improves, which means that all of the noise comes from the tuning diode. In this case, the unloaded Q (Q_0) was set to 400. As Q_{load} approaches Q_0, there is a degradation of the phase noise starting at Q_{load} = 300, and degrading further as it approaches 400, where it will be infinite (Fig. 6). This is due to the new term $[1 - (Q_{load}/QR_0)]$. Given the conditions that the output power (P_{output}) is 0 dBm, F_0 = 880 MHz, R_n



6. The linear equation for phase noise can be used to predict the performance of the oscillator with (top curve) and without the tuning diode (bottom curve).

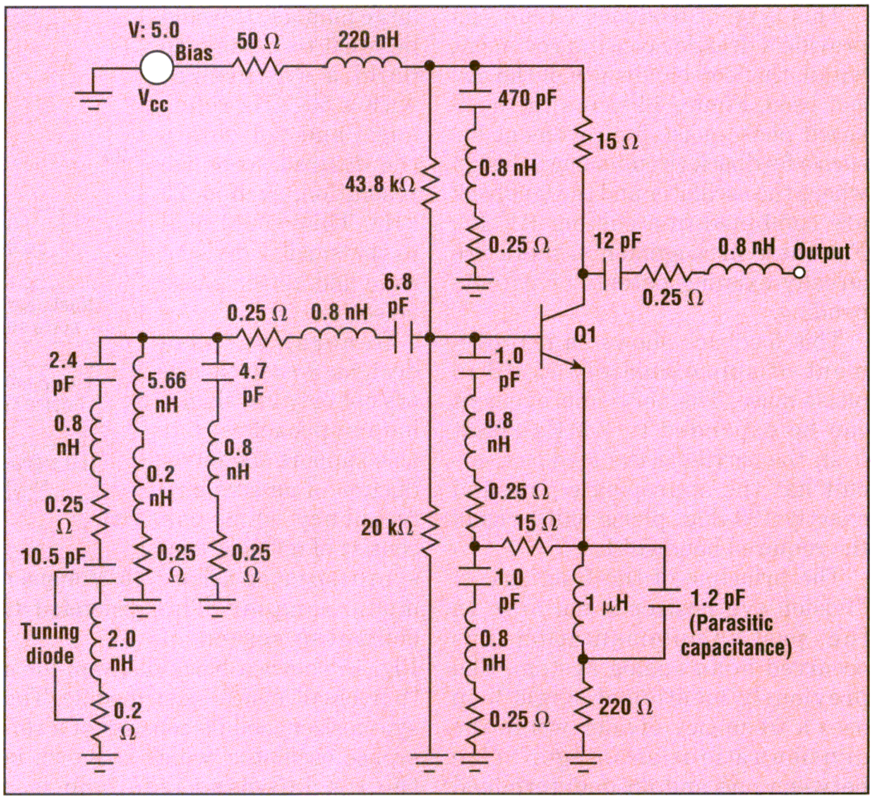


7. According to these predictions of phase noise for the 880-MHz VCO at sensitivities from 10 Hz/V to 100 MHz/V, above a certain sensitivity, the phase noise is determined only by the oscillator circuit's tuning diode(s).

$= 200\ \Omega$, $Q_0 = 600$ at 100-kHz offset from the carrier, the values shown in Table 2 are obtained. The noise remains fairly flat for Q values between 200 and 550. This is due to the losses of the resonator and the variation of the output level, which is driven by these losses.

The influence of the tuning diode, which is shown in the last term of the phase-noise equation, dominates the phase noise if the tuning sensitivity gets too high (Fig. 7).

Many authors also assume that all the components are ideal. In reality, passive devices such as capacitors and inductors have "lead inductors" and the connecting points or solder joints also have resistive elements. Therefore, a capacitor is correctly modeled as a series connection of the capacitor itself in series with a small inductor in the vicinity of 0.2 to 1.0 nH and a contact resistor of approximately 0.2 to 0.3 Ω. Figure 8 illustrates to what level of detail the modeling must be performed so that measured and predicted phase noise agree. The particular oscillator is not an award-winning design, but measured and simulated results for this oscillator (Fig. 9) agree closely.[4]



8. This Colpitts oscillator uses RF negative feedback between the emitter and capacitive voltage divider. Real, rather than ideal, components have been used, including some typical parasitic values for the capacitors and inductors.

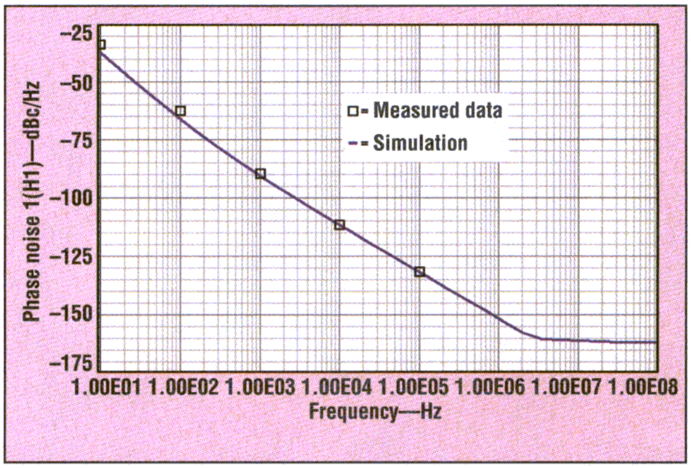
The linear model is somewhat optimistic since it "assumes" many events to be linear and easy to model. In reality, the noise, as superimposed on the ideal noiseless carrier, comes from many contributors (Fig. 10).

The noise performance can be improved by having a resonator circuit with the highest possible Q, provided the tuning diode is not the main villain. When using a transmission line-based resonator instead of a coil inductor, the Thompson formula $F = 1/(2\pi LC)^{0.5}$ is no longer valid and the inductance L must be replaced with a hyperbolic tangent (tanh) function, $X_L = Z_0 \tanh(2\pi l/\lambda)$, with $l < \lambda/4$. As a result, the circuit has a sharper resonance or appears to have a higher Q. Figure 11 shows the impact on noise by varying the DC source.

The tuning diode, which due to phase-modulation (PM) conversion adds strongly to the noise, also influences the Q. The standard circuit uses two diodes instead of one in an anti-series connection which is chosen to avoid DC rectification in self-biasing at RF voltages higher than DC voltages.

Many circuits with medium tuning range rely on a single diode for tuning as seen in Fig. 3. If the tuning range increases, the RF-voltage swing across the tuning diode increases. A remedy for this is the use of two diodes connected in a series where the opposite polarity of the diode cancels the DC rectification. The drawback of this circuit is that there are loss resistors in series. The circuit of choice uses at least two diodes in each series arm. In the example of Fig. 12, the inductor of the left circuit is replaced with a shorter stub element, which can be used to model a ceramic resonator.

Further improvement in the noise can be achieved by DC/RF feedback. The two circuits of Fig. 13 sample the DC current as well as the low-frequency components in either the emitter or the collector current. By using feedback circuits, one can combine a DC stabilization as well as a noise compensation/cancellation.



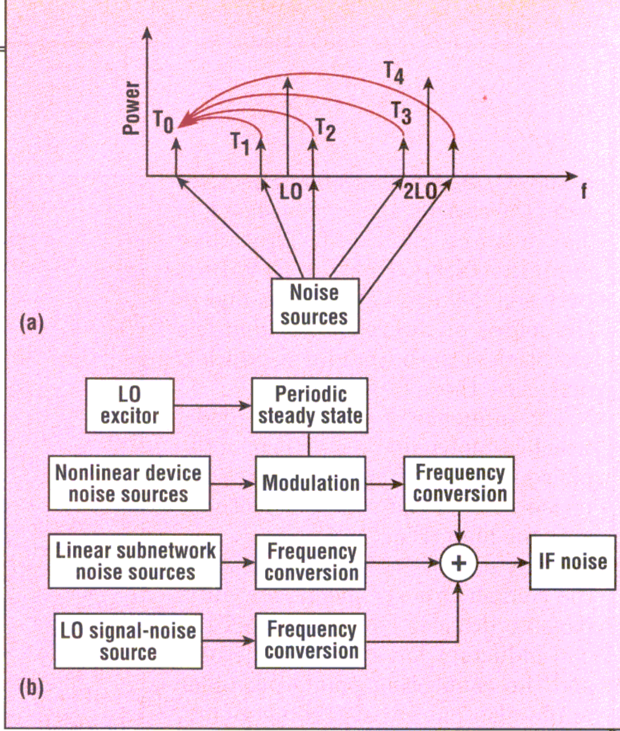
9. These plots compare the predicted and measured performance of the oscillator of Fig. 8.

This type of feedback circuit can provide a drastic noise improvement within the loop bandwidth of the circuit used. Figure 14 shows the measured phase-noise improvement for such a feedback circuit using a design where the oscillator and the feedback are combined in a custom RF IC. For these arrangements, there are already existing patents or patents pending.

This feedback shows an improvement of approximately 15 dB in phase noise. The noise improvement can be expanded to 1-MHz offset from the carrier if the feedback circuit has the appropriate gain and exactly 180-deg. phase shift within the required bandwidth.

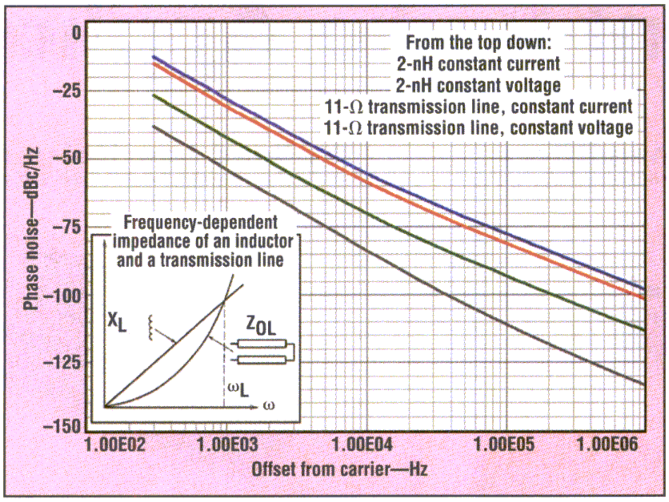
The principle of the fractional-N division synthesizer is not new. In the past, implementation has occurred in an analog system. The previously mentioned single loop uses a frequency divider where the division ratio is an integer value between one and an undetermined large number. It would be ideal to be able to build a synthesizer with a 50-MHz reference, and yet obtain the desired step-size resolution, such as 25 kHz. This would lead to the smaller division ratio and better phase noise.

General-purpose designs for synthesizers not only use single-loop approaches, but also support the introduction of an auxiliary frequency which can consist of a single-loop synthesizer or a harmonic sampler. The computer screen of Fig. 15 shows a block diagram for a "universal" synthesizer module. The synthesizer module consists of a reference oscillator, which is shown in the top left-hand side, and is a marked reference source. Typically, a high-performance crystal oscillator is used here. Reference frequencies vary from low values such as 10 kHz to high frequencies like 100 MHz. In

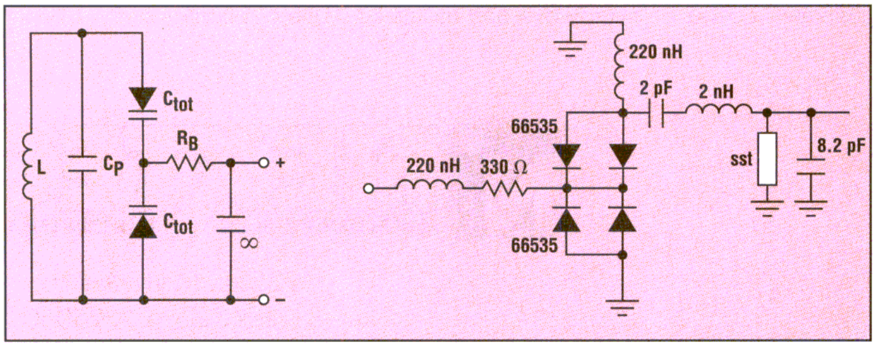


10. Multiple noise sources account for noise generated during mixing to IFs.

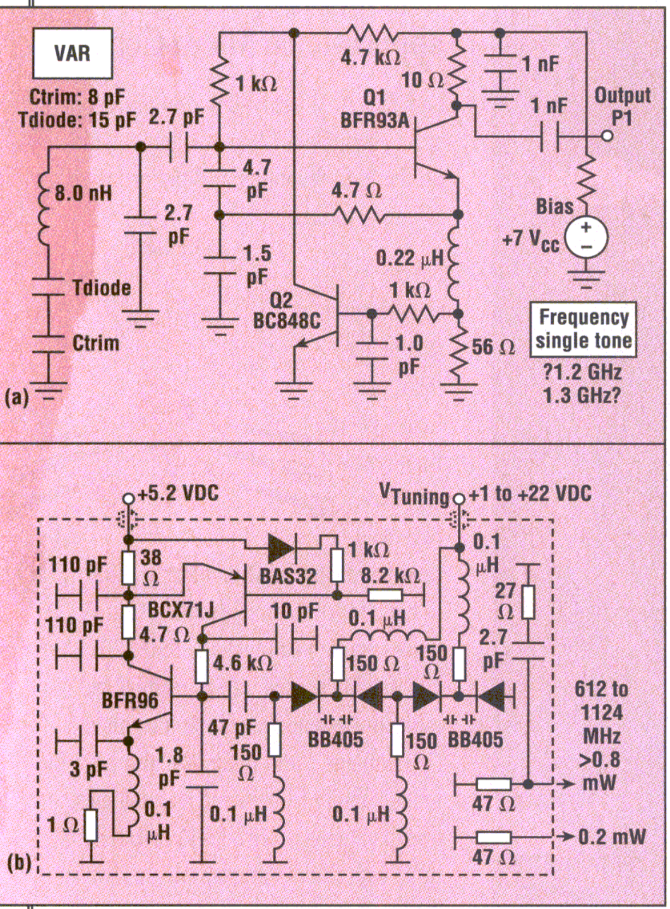
the analysis for this case, a data file was chosen which represents a 10-MHz frequency standard. This 10-MHz frequency is then divided by 1 and connected to the input of the phase detector. For conventional synthesizers, the division ratio would be chosen so the input to the phase detector is equal to the phase size. On the top right-hand side of Fig. 15, a VCO is set at 200 MHz—after being divided by 20, this provides a 10-MHz input to the phase detector. Since the auxiliary loop is in the "off" position, this portion no longer contributes to the system. The modeling software will accept inputs such as the noise floor of the dividers, the phase-detector sensitivity, the noise voltage of the active filter, and the loop bandwidth. For this single-loop synthesizer, the linear-analysis software calculates the total noise and displays the various noise contributions. For a 50-kHz loop bandwidth, the solid line illustrates the resulting total phase noise. At approximately 200 Hz, it exhibits the improvement due to the loop, which has a slight peak of around 50 kHz, with a phase noise of about –117 dBc/Hz. The free-running oscillator examined previously has a predicted phase noise of approximately –130 dBc/Hz. This means that the loop bandwidth is too wide, since it now reveals the upward multiplication of the noise sources which results in unfavorable phase noise. A more appropriate loop bandwidth would have been approximately 5 kHz where the free-running oscillator has a phase noise of about –110 dBc/Hz. The dashed line in Fig. 15, which peaks at approximately 50 kHz, illustrates the noise improvement of the VCO due to the loop. Outside of the loop bandwidth, the Q of the oscillator dominates the noise performance. The horizontal dashed line at approximately –130 dBc/Hz represents the phase noise of the mixer (not used) and the main divider contribution. The line underneath is the noise level of the operational amplifier (if used) and the phase detector at the VCO



11. Constant-current and constant-voltage biasing have been differentiated here to illustrate the sensitivity of transistor oscillators to bias networks and resonator circuits.



12. Two tuning diodes are used in the parallel-resonant circuit (left). To reduce losses, an improved circuit uses two diodes in parallel on each leg (right). The shorted stub (SST) is a model element for a ceramic resonator.



13. For these bipolar-transistor-based oscillators, noise sampling was performed at (a) the transistor emitter and (b) at the transistor collector.

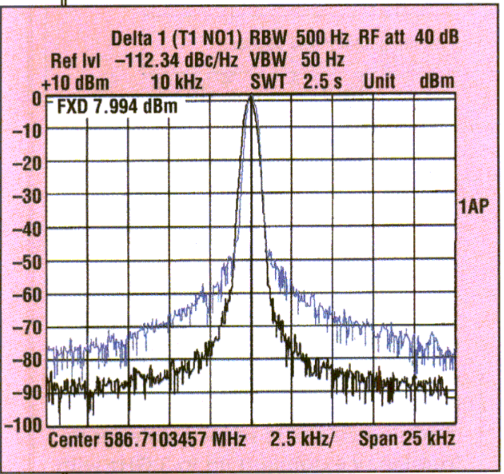
frequency. The current configuration also only supports a step size of 10 MHz.

An alternative to this design approach would be for N to take on fractional values. The output frequency could then be changed in fractional increments of the reference frequency. Although a digital divider cannot provide a fractional-division ratio, ways can be found to accomplish the same task effectively. A frequently used method is to divide the output frequency by N + 1 every M cycles and to divide by N the rest of the time. The effective division ratio is then N + 1/M, and the average output frequency is given by:

$$f_o = \left(N + \frac{1}{M}\right) f_r \qquad (3)$$

This expression shows that f_o can be varied in fractional increments of the reference frequency by varying M. The technique is equivalent to constructing a fractional divider, but the fractional part of the division is actually implemented using a phase accumulator.

Consider the problem of generating 899.8 MHz using a fractional-N loop with a 50-MHz reference frequency, 899.8 MHz = 50 MHz [N + (K/F)]. The integral part of the division N must be set to 17 and the fractional part, K/F, must be 996/1000 [the fractional part (K/F) is not a integer], and the VCO output must be divided by 996 × every 1000 cycles. This can be easily implemented by adding the number 0.996 to the contents of an accumulator every cycle. Every time the accumulator overflows, the divider divides by 18 rather than by 17. Only the fractional value of the addition is retained in the phase accumulator. When moved to the lower band or an attempt is made to generate 850.2 MHz, N remains 17 and K/F becomes 4/1000. This method of using fractional division was first introduced by using analog implementation and noise cancellation, but today it is implemented totally as a digital approach. The necessary resolution is obtained from the dual-modulus prescaling, which supports a well-established method for achieving a high-performance frequency synthesizer operating at ultra-high frequency (UHF) and higher frequencies. Dual-modulus prescaling avoids the loss of resolution in a system compared to a simple prescaler—it supports a VCO step equal to the value of the reference frequency to be obtained. The dual-modulus prescaler then divides by N or N+1



14. A novel RF IC oscillator circuit, which includes a tuning diode, provides improved phase-noise performance. The tuning-diode coupling is only approximately 10 MHz/V and does not add to the modulation noise. For test purposes, the circuit is considered as an LC circuit with a loaded Q of 50 measured at approximately 500 MHz.

depending upon the state of its control. The only drawback of prescalers is the minimum-division ratio of the prescaler for approximately N^2. The dual-modulus divider is the key to implementing the fractional-N synthesizer principle. Although the fractional-N technique appears to have a favorable potential for solving the resolution limitation, it is not free of having its own complications. Typically, an overflow from the phase accumulator (which is added along with the output feedback to the input after being latched) is used to change the instantaneous-division ratio. Each overflow produces a jitter at the output frequency, caused by the fractional division, and is limited to the fractional portion of the desired-division ratio.

In the case of the example, a step size of 200 kHz had been chosen, and yet the discrete sidebands vary from 200 kHz for K/F = 4/1000 to 49.8 MHz for K/F = 996/1000. It will become the task of the loop filter to remove those discrete spurious. While in the past the removal of the discrete spurs has been accomplished by using analog techniques, various digital methods are now available. The microprocessor must solve the following equation:

$$N^* = \left(N + \frac{K}{F}\right) = [N(F - K) + (N + 1)K] \qquad (4)$$

As an example, consider the case where F_0 equals 850.2 MHz:

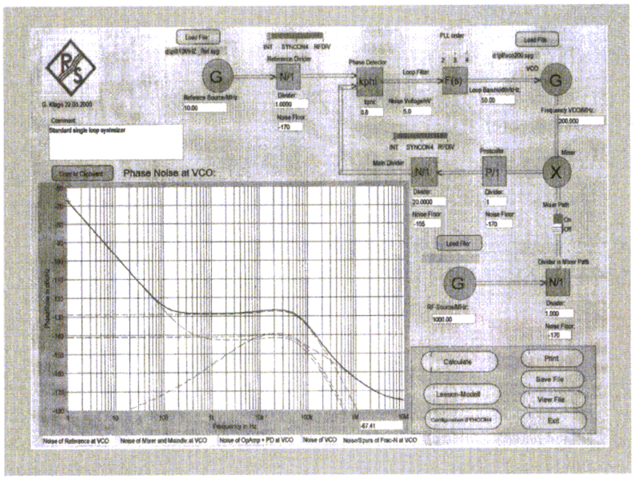
N^* = 850.2 MHz/50 MHz = 17.004

Following the formula:

$$N^* = \left(N + \frac{K}{F}\right) = \frac{[17(1000 - 4) + (17 + 1) \times 4]}{1000} = \frac{[16932 + 72]}{1000} = 17.004$$

$$F_{out} = 50MHz \times \frac{[16932 + 732]}{1000} = 846.6\ MHz + 3.6\ MHz = 850.2\ MHz \qquad (5)$$

The mechanism that generates the digital compensation is illustrated in Fig. 16. The approach was patented by Marconi (Stevenage, England) in European Patent No. 0125790B2 in 1995. It consists of three sigma-delta converters which are connected in

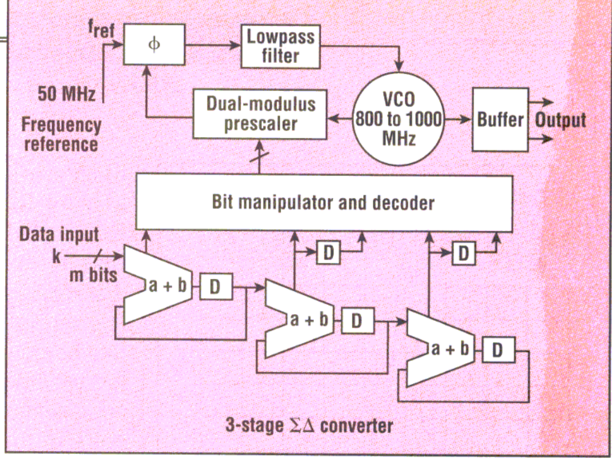


15. This computer screen displays an image of the synthesizer-simulation software which can handle single- and multiple-loop designs.

series and are responsible for bit manipulation. The approach requires a programmable divider in the loop; a dual-modulus prescaler is not sufficient.

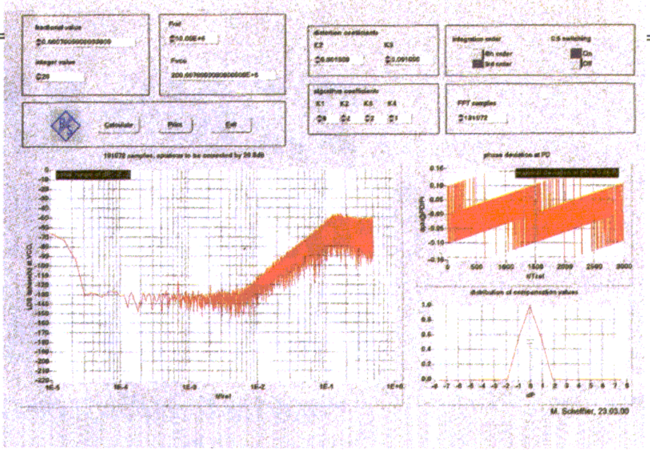
Advanced fractional-N synthesizers not only have a long accumulator, but also an efficient spurious-cancellation mechanism based on proprietary mathematical algorithms. This is the area of greatest research and patent application. Software, which has been developed for the internal fractional-division chip, supports programming "software simulation" which will then be activated in the actual hardware. The screen image of the control software is shown in Fig. 17. The software accepts all the necessary values as inputs. On the top left, it is fed the fractional and integer values that result in an output frequency of 200.007 MHz using the 10-MHz reference. The correcting mechanism requires the entry of distortion coefficients and algorithm coefficients. The values shown correspond to a particular compensation scheme. The display in the left-hand corner illustrates the equivalent phase noise generated by this internal circuitry. The phase noise, which is relative to the 20-MHz reference, is shown to be approximately –130 dBc/Hz which then increases at 5×10^{-2} relative to the reference frequency. This means that the loop bandwidth should be at least of order 3 in order to suppress the quantization noise caused by digital phase jitter. The loop filter itself will attenuate the spurious on the right side of the picture. A Type-2, fourth-order filter is used. Since the linearity of the phase detector is crucial, the software checks if the chosen coefficients do not overdeviate the phase detector. This is shown in the software's right-hand window. Finally, the distribution of the compensation values also needs to be monitored as it determines overall performance.

The system-level simulation software can now be applied to examine the overall results (Fig. 18). The solid



16. This block diagram represents the fractional-N synthesizer using the custom IC for reference frequencies to 100 MHz. The use of smaller modulus values is responsible for its frequency extension up to 3 GHz with ripple or asynchronous counters, and supports the implementation of dual-modulus counts to M/N+1.

curve in Fig. 18 shows the overall phase noise including unwanted spurious results. Using the coefficients as previously chosen, including the loop filter, the ultimate noise floor up to 1.05 MHz is better than –140 dBc/Hz, reaching –180-dBc/Hz offset 4.5 MHz from the carrier. By "playing" with these coefficients, it is possible to achieve a trade-off between the amplitude of the spurious signals and the cutoff of the loop. It is also possible to add additional filtering to increase the spurious suppression above 100 kHz. In the



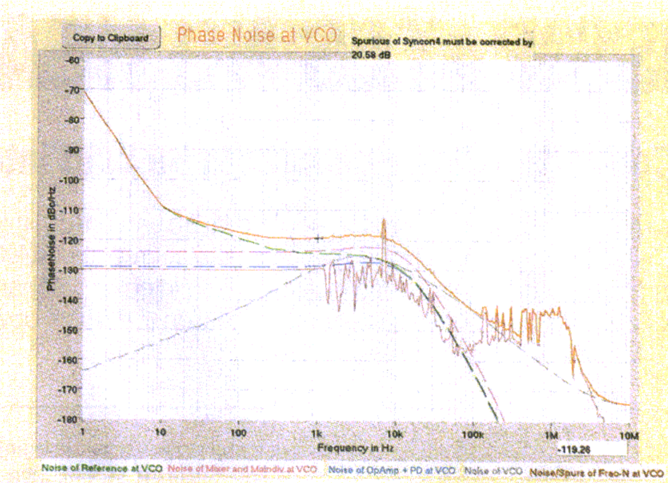
17. This plot shows the noise output of the three-stage sigma-delta converter and compensation circuitry for the fractional-N synthesizer.

case shown, the loop bandwidth was set at 10 kHz. The noise floor of the buffer amplifier following the reference was –165 dBc/Hz and, finally, the fractional divider had a noise floor of – 150 dBc/Hz. The output frequency generated was 200.007 MHz. This means that the fractional portion had three digits of resolution.

As has been seen, it is necessary to incorporate loaded Q, as well as unloaded Q, in all oscillator and synthesizer phase-noise design considerations. Due to this, it was necessary to expand the Leeson equation. If not dominated by other factors, tuning diodes (with tuning sensitivity of MHz/V) can add significantly to the phase noise. Since various approaches exist for the design of a fractional-N synthesizer, this study has focused on what influence the digital portion of the synthesizer chip has on the resulting phase noise of the system.

Given the existence of a compensation circuit with arbitrary digital compensation coefficients, it was possible to select a practical case to demonstrate the impact of the various choices. In most cases, the designer is "stuck" with a particular hardware design and can rarely achieve fractionality beyond 1/32. ICs which provide higher resolution [such as the fractional-N IC offered by Philsar (Nepean, Ontario, Canada)] achieve resolution down to 100 Hz with a 200-kHz reference at the expense of a minimum-division ratio, which is larger than 20. In the previous example, the fractional offset made it possible to achieve frequency resolution of 7 kHz. The result is that the fractional offset is inside the loop bandwidth. Standard fractional syn-

thesizers do not face the problem of "channel spacing" which is smaller than the loop bandwidth. In these cases, a complicated compensation scheme is necessary, as well as a complete spurious analysis. This is due to the fact that the dual-modulus prescaler forces a minimum-division ratio to be approximately N^2. In the case of a prescaler set at 10, the minimum-division ratio would be 100. ●●



18. This is the composite phase noise of the fractional-N synthesizer system, including all noise and spurious signals generated within the system.

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Modulation Schemes Affect The Linearity Of An HBT Amp

This article investigates the effect of several digital modulation schemes on the linearity of an HBT amplifier.

Shuo-Yuan Hsiao and Nan-Lei Wang

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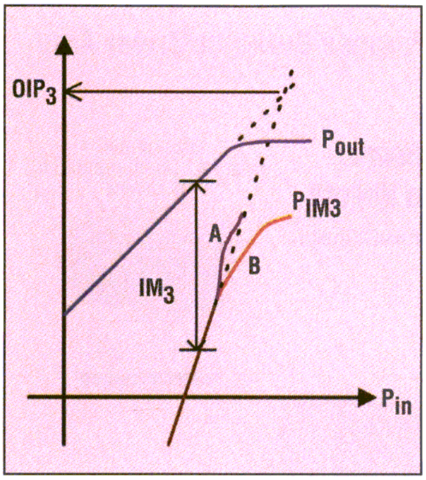
WIRELESS communication equipment growth has led to an increasing demand for high-performance RF components, including high-linearity amplifiers. In this article, a high-linearity, Class A heterojunction-bipolar-transistor (HBT) amplifier, model EC1089, is tested to investigate the behavior of its linearity characteristics in response to four modulation schemes: two-tone intermodulation, forward code-division-multiple-access (CDMA), reverse-CDMA, and North American Digital Cellular (NADC) signals. Also, to investigate the sensitivity of the linearity measurements, the authors tested six samples of the EC1089 from wafers with different epitaxial structures.

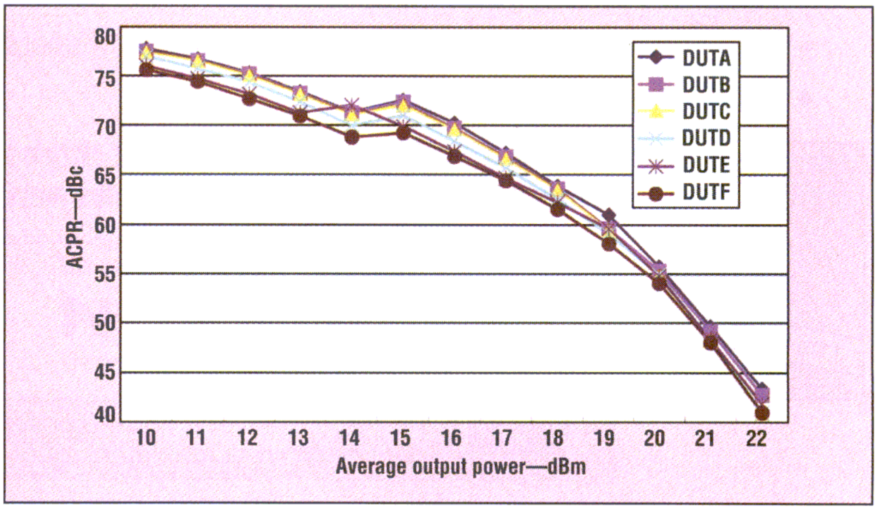
To address the linearity of an amplifier, its manufacturer usually tests and specifies its output third-order intercept point (OIP_3). By definition, OIP_3 is obtained by calculating the intercept point of extrapolated output-power (P_{out}) versus the input-power (P_{in}) curve and extrapolated third-order intermodulation (IM) power (P_{IM3}) versus P_{in} curve. Figure 1 illustrates these characteristics. The relationship between P_{out}, OIP_3, and third-order IM ratio (IM_3) is:

$$OIP_3 = P_{out} + \frac{IM_3}{2} \qquad (1)$$

Since the OIP_3 is an extrapolated value, Eq. 1 is valid only in the region of power level where the slope of



1. This graph shows the typical P_{out} versus P_{in} and P_{IM3} versus P_{in} curves of amplifiers.



2. This graph shows the measured OIP_3 versus output power for the six samples.

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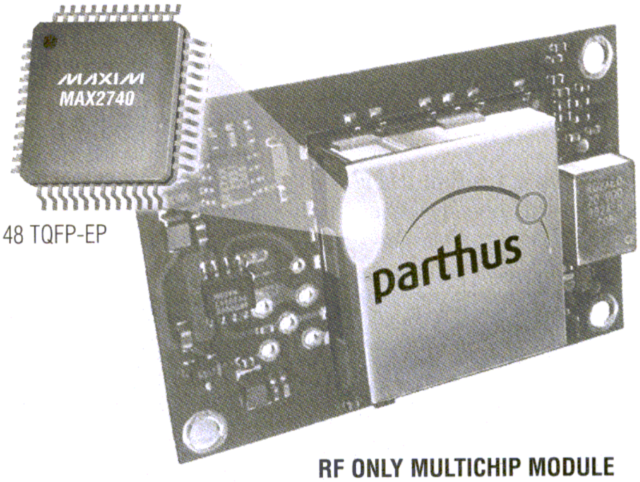

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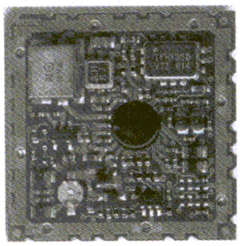

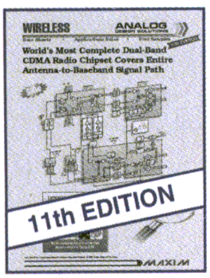

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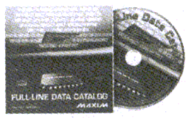

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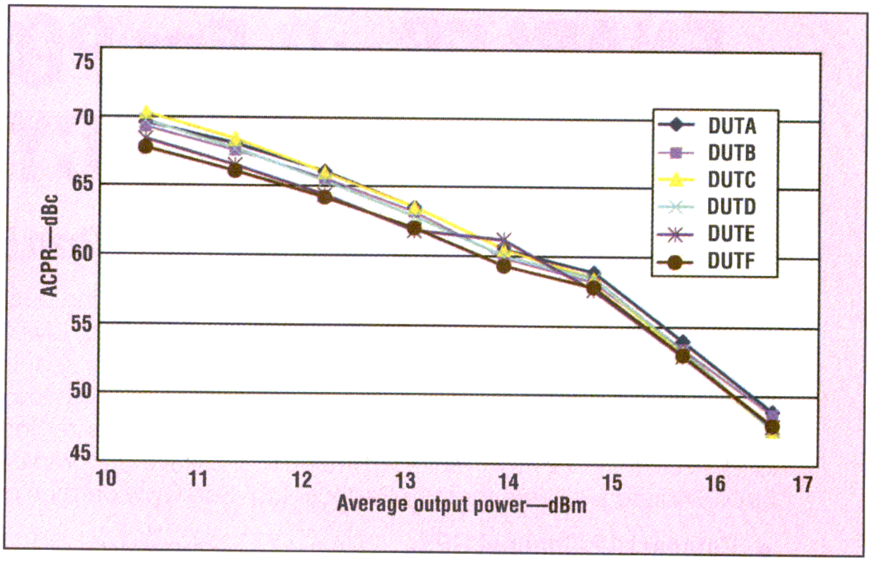
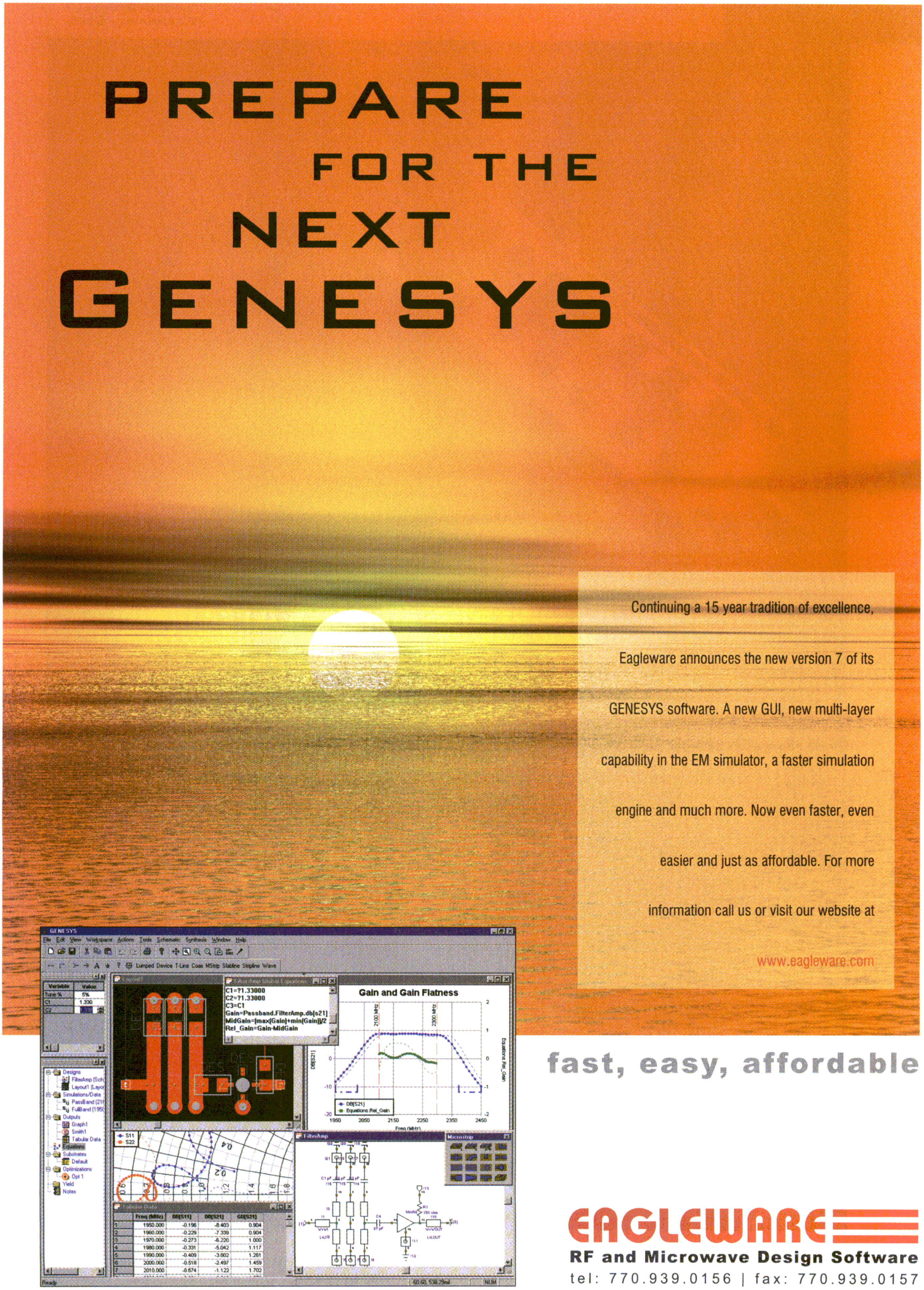
P_{IM3} versus P_{in} is 3. Outside this region, the calculated IM_3 will be incorrect. This situation is illustrated in Fig. 1, where the slopes of two curves—A and B—are no longer 3 above a certain power level. Since a single value of OIP_3 cannot predict the IM_3 over the entire power range, another way is to plot OIP_3 versus P_{out} based on Eq. 1 (extending the OIP_3 definition). If the extended OIP_3 versus P_{out} of curves A and B are plotted, one can see that the OIP_3 of curve A degrades with P_{out}, and that of curve B goes in the opposite direction.

In receiver (Rx) applications, several signals at different frequencies are applied to the amplifier. Each signal is usually at a quite low power level (the 3:1 slope range) and OIP_3 can be used directly in the calculation of the intermodulated distortion (IMD).

However, in a transmitter (Tx), there is often only one signal at a high power level. The signal can be a modulated one—having a time-domain waveform with noise-like amplitude variation, such as CDMA and NADC. Since OIP_3 over the modulated signal-amplitude range may not be constant, OIP_3 is not the proper parameter to characterize a linear amplifier. The linearity of the amplifiers in these applications is characterized by the adjacent-channel power ratio



3. This graph shows the measured 1.9-GHz CDMA 9-code forward link ACPR at 750-kHz offset for the six samples.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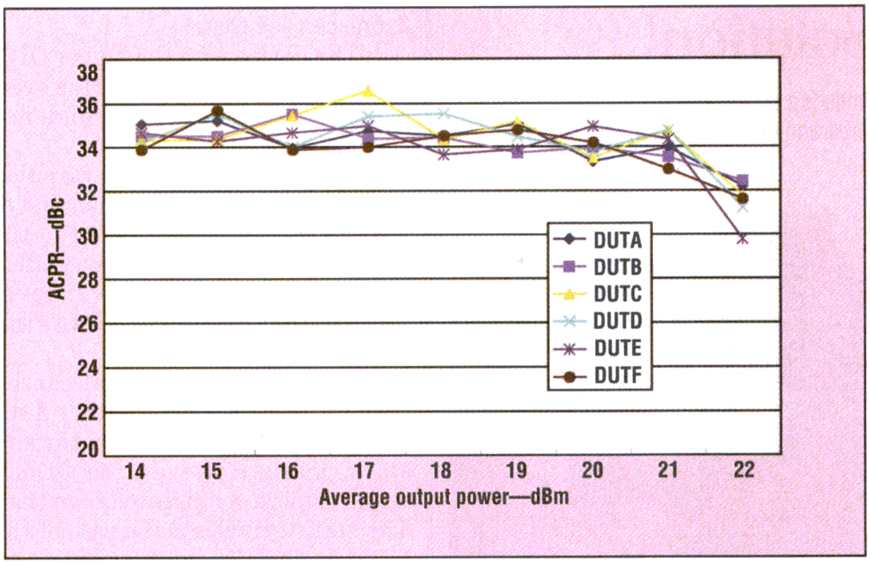


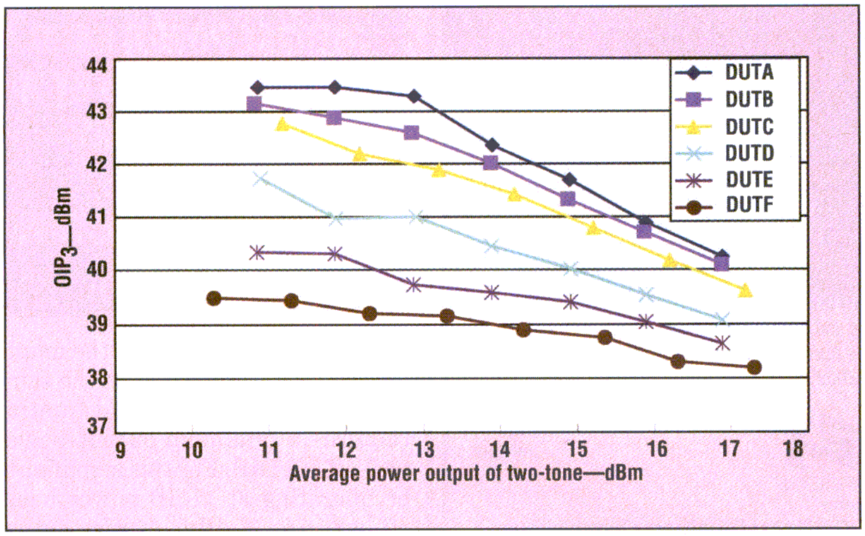
4. This graph shows the measured 1.9-GHz CDMA reverse link ACPR at 885-kHz offset for the six samples.

(ACPR), which depends not only on the IMD of the amplifier, but also on the amplitude-to-amplitude (AM-AM) and amplitude-to-phase modulation (AM-PM) characteristics of the amplifiers.[1-4]

The typical OIP_3 of the model EC1089 amplifier is +42 dBm using a +5-VDC supply voltage. The measurements are performed with an EV1089 evaluation board at the center frequency of 1.9 GHz. Figure 2 shows the measured OIP_3 versus P_{out} of these samples. Table 1 summarizes the measured two-tone characteristics. Also shown in Table 1 is the linearity figure of merit (LFOM), which is defined by the ratio of OIP_3 to DC power and is a common index of the two-tone linearity of an amplifier.

As seen from Table 1, the maximum-to-minimum OIP_3 variation of these samples is +4 dBm, corresponding to 8-dBc IM_3 deviation at an output power of +11 dBm. The maximum LFOM difference is 17.5. On the other hand, the $P_{1\,dB}$ is almost constant.

Figure 3 shows the ACPR test results using an IS-95, CDMA-forward, nine-code channel signal centered at 1.9 GHz. After investigating the differences between Figs. 2 and



5. This graph shows the measured 1.9-GHz NADC ACPR at 30-kHz offset for the six samples.

3, one can see that:

1. The device with the higher OIP_3 also shows higher ACPR. However, a 1-dBc-to-1-dBc correspondence cannot be found.

2. The sample-to-sample variation of ACPR is much smaller than that of OIP_3. This can be attributed to the fact that the modulations have different peak-to-average power level and different phase-changing characteristics.

The two-tone IM signal can be thought of as a signal with a 3-dB peak-to-average power ratio and static phase. However, the CDMA signal is more complicated than that. The statistical characteristics of the CDMA signal's peak-to-average power ratio can be described by a complementary-cumulative-distribution-function (CCDF) graph. In the nine-code-channel CDMA application, the probability of the signal's peak power reaching 9.7 dB above the average level is 0.01 percent.

THE TWO-TONE IM SIGNAL CAN BE THOUGHT OF AS A SIGNAL WITH A 3-dB PEAK-TO-AVERAGE POWER RATIO AND STATIC PHASE, WHILE THE STATISTICAL CHARACTERISTICS OF THE CDMA SIGNAL'S PEAK-TO-AVERAGE POWER RATIO CAN BE DESCRIBED BY A COMPLEMENTARY-CUMULATIVE-DISTRIBUTION-FUNCTION (CCDF) GRAPH.

Making the most use of the amplifier in this system while, at the same time, maintaining signal quality requires a minimum output-power backoff of 11 dB. For this amplifier to be operating at 11-dB power back-off, which is 13-dBm average output power, the measured maximum and minimum ACPR of the sample group

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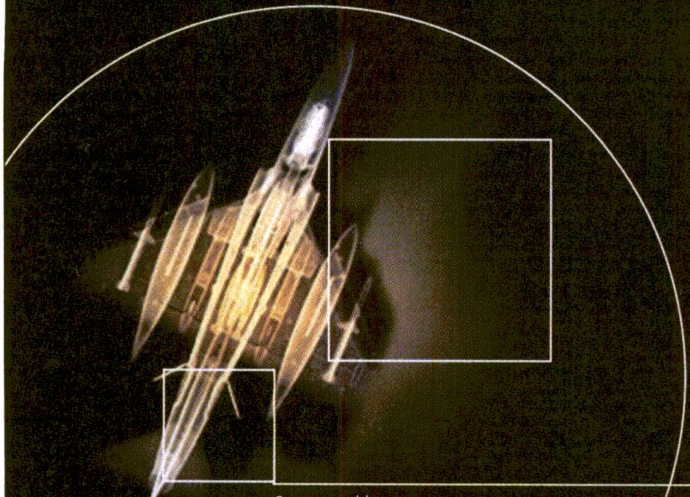

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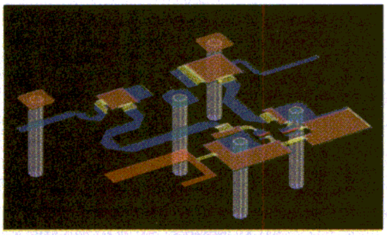
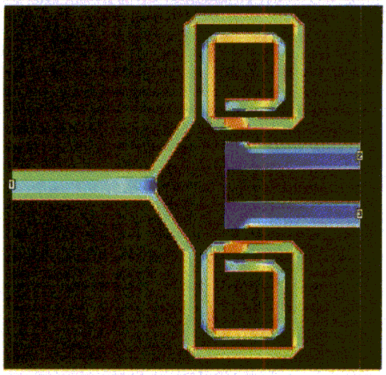
Sponsored by: Microwaves & RF ELECTRONIC DESIGN are 63.5 dBc and 61.8 dBc, respectively. This 1.7-dBc deviation is smaller than the 8-dBc IM_3 deviation in the two-tone test at the same average output power.

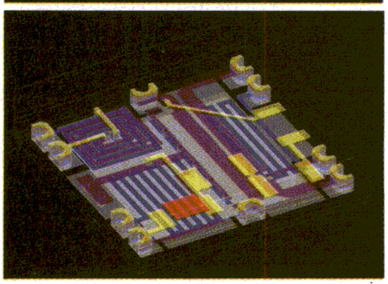
Figure 4 shows the ACPR testing results with a 1.9-GHz reverse-CDMA signal. The statistical characteristics of the reverse-CDMA signal show that the probability of the signal reaching 5-dB peak-to-average power ratio is 0.01 percent. So in the reverse mode, a similar ACPR can be obtained at an average power level of +18 dBm. The maximum and minimum ACPR are 63.9 dBc and 61.5 dBc, respectively. The deviation is 2.4 dB.

The NADC ACPR measurements are performed by setting the center frequency to 1.9 GHz and measuring the 30-kHz offset ACPR. The ACPR of the signal source is 35 dBc. The results are provided in Fig. 5. The statistical characteristics of the NADC signal show that the probability of the signal reaching 3.1-dB peak-to-average power ratio is 0.1 percent. Since the peak-to-average power is close to the two-tone IM, it is interesting to compare the test results in the two modulations. As can be seen from Fig. 5, there is no obvious ACPR degradation in low power levels below +20 dBm, and the device shows that higher OIP_3 does not necessarily mean higher ACPR in this modulation. This suggests that the distortion introduced by the EC1089 sample is much smaller than 35 dBc. The sample-to-sample variation in NADC ACPR is very low compared to the ACPR of two-tone

THE SAMPLE-TO-SAMPLE VARIATION IN NADC ACPR IS VERY LOW COMPARED TO THE ACPR OF TWO-TONE AND CDMA MODULATIONS. THIS REVEALS THE WEAKNESS OF TWO-TONE TESTING IN CDMA AND NADC APPLICATIONS.







and CDMA modulations. This reveals the weakness of two-tone testing in CDMA and NADC applications.

One way to increase the ACPR is to increase $P_{1\,dB}$. This can be accomplished by increasing the bias voltage and current of the device. At a V_{CC} of +6 VDC and an I_{CC} of 170 mA, the DUT-C has a measured $P_{1\,dB}$ of +25.1 dBm. Table 2 summarizes the measurement results before and after increasing the DC power and provides the ACPR that is measured at different power levels. There is almost no improvement in the OIP_3 performance of the high-power version. Actually, the LFOM is degraded due to higher DC power. But from the ACPR measurements, three interesting things can be found:

1. The ACPR is very similar at low output power for both DC-bias conditions.
2. At high output power, the high $P_{1\,dB}$ version has much better ACPR.
3. At the same backoff level (defined by $P_{1\,dB} - P_{out}$), the ACPR is similar.

The ACPR improvement due to the increased OIP_3 and $P_{1\,dB}$ can be compared quantitatively through the comparison of their improvement ratios, $IR(OIP_3)$. The improvement ratio, as a function of OIP_3, is defined as:

$$IR(OIP_3) = \frac{\Delta ACPR}{\Delta OIP_3} = [ACPR(DUTx) - ACPR(DUTy)] / [OIP_3(DUTx) - OIP_3(DUTy)] \; dBc/dBm \tag{2}$$

and $IR(P_{1\,dB})$, the improvement ratio as a function of $P_{1\,dB}$, is defined as:

$$IR(P_{1dB}) = \frac{\Delta ACPR}{\Delta P_{1dB}} = [ACPR(DUTx) - ACPR(DUTy)] / [P_{1dB}(DUTx) - P_{1dB}(DUTy)] \; dBc/dBm \tag{3}$$

The measurement results of DUT-A and DUT-F are used to calculate the $IR(OIP_3)$, and that of DUT-C at +5 VDC and +6 VDC are used to calculate $IR(P_{1\,dB})$. For example, the OIP_3 difference between DUT-A and DUT-F at +13-dBm P_{out} can be found from Fig. 2 to be +4 dBm, and their forward ACPR difference at the same power level is 1.5 dBc, so the

THE ACPR IMPROVEMENT DUE TO THE INCREASED OIP_3 AND P_{1dB} CAN BE COMPARED QUANTITATIVELY THROUGH THE COMPARISON OF THEIR IMPROVEMENT RATIOS, $IR(OIP_3)$.

Table 1: Summary of the testing results of six EC1089 samples from wafers with different expitaxial structures

DUT	V_{CC} (V)	I_{CC} (mA)	OIP_3 (dBm)	P_{1dB} (dBm)	LFOM
A	5	155	43.5	24.1	28.9
B	5	154	43.2	24.2	27.1
C	5	146	42.8	23.8	26.1
D	5	154	41.7	23.9	19.2
E	5	148	40.3	23.9	14.5
F	5	157	39.5	24.0	11.4

$IR(OIP_3)$ is 1.5/4 = 0.38 (dBc/dBm). For the $IR(P_{1\ dB})$, the forward ACPR difference of DUT-C at +6 VDC and DUT-C at +5 VDC and +13-dBm P_{out} is 1.3 dBc. This value divided by their $P_{1\ dB}$ difference +1.3 dBm, is 1 (dBc/dBm). Through similar calculations at the other power levels, a plot of IR versus P_{out} is obtained and shown in Fig. 6. From the figure, it is obvious that increasing the $P_{1\ dB}$, or by examining the situation in an equivalent way, increasing the backoff power level, is a more-efficient way to improve the ACPR than increasing the OIP_3.

TEST RESULTS

RF tests of the EC1089 reveal that the amplifier has distinctive linearity characteristics for each modulation. Two-tone IM testing shows the highest sensitivity to sample differences. The maximum sample-to-sample variation of OIP_3 of +4 dBm corresponds to a LFOM deviation of 60 percent. The other modulation schemes show less-sensitive results. The maximum deviation of ACPR measured at minimum backoff power level (P_{out} at +13 and +18 dBm) in forward CDMA and reverse CDMA are 1.7 and 2.4 dBc, respectively. In the NADC ACPR testing, no noticeable difference was found among the samples. Moreover, by comparing the CDMA ACPR measurement results of the devices with a different OIP_3 and output 1-dB compression point ($P_{1\ dB}$), it is shown that increasing the $P_{1\ dB}$ is a more-efficient way to improve the ACPR than increasing the OIP_3.●●

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(continued on p. 178)

Digital Predistortion Linearizes RF PAs

Predistortion techniques allow designers to improve power, linearity, and efficiency of RF power amplifiers.

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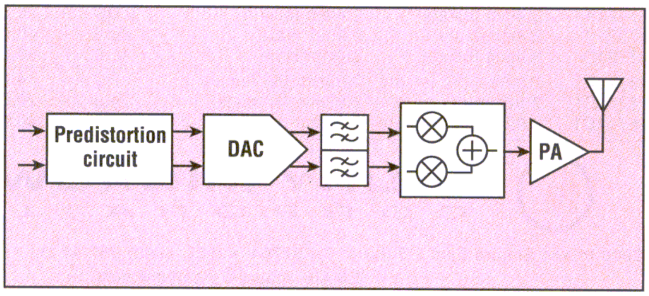
RAPID enhancements in digital signal processors (DSPs) have made it possible for designers to use digital predistortion to perform real-time correction of distortion in RF power amplifiers (PAs). Unfortunately, unlike purely analog-based solutions, this technique puts stringent requirements on several components in the transmission chain, including predistortion circuits, data converters, reconstruction filters, and the RF transmitter (Tx) components (Fig. 1). These constraints arise because there must be a perfect transmission path between the predistorter and the nonlinear PA, since any errors in this path will result in increased adjacent-channel power (ACP). Although an adaptive digital predistortion system can compensate for some of the nonlinearities in the forward path not attributable to the PAs, many other imperfections cannot be overcome. Therefore, a successful implementation of digital predistortion requires an understanding of potential system-level limitations and the means of overcoming them. This article presents an empirical overview of circuit errors in non-adaptive systems. Since many of the components in the correction loop exhibit characteristics similar to those in the forward path, the information presented is also applicable to a fully adaptive, digital predistortion system.

The predistortion algorithms developed to date are based on either a mapping-type predistorter or a gain-based predistorter.[1-2] Mapping predistorters map an input-in-phase/quadrature (I/Q) signal vector into an output-signal vector, essentially mapping the complex plain into itself.[3] The advantage of this technique is its ability to compensate for analog-modulator errors and PA nonlinearities. The two-dimensional (2D) look-up table used in this method, however, can increase memory requirements and slow convergence.[1-2] The complex gain predistorter uses a one-dimensional, complex-word, look-up table to store the level-dependent complex gain of the predistorter, which is computed from the complex gain characteristics of the amplifier.[2,4] This method requires smaller memory and exhibits faster convergence time when compared



1. This is a block diagram of a digital predistortion system.

to the mapping technique. However, these improvements come at a cost of increased signal processing.

This article describes an algorithm based on the complex gain predistorter. The complex predistortion coefficients are computed from interpolated, empirical, amplifier-gain characteristics by using piecewise cubic splines.[5] This technique has yielded the best results measured to date for laterally diffused metal-oxide semiconductor (LDMOS) and gallium-arsenide (GaAs) RF PAs. The test signal selected for the analysis of the digital predistortion system is based on the IS-95 code-division-multiple-access (CDMA) specification. This signal is the nine-channel forward model composed of pilot, paging, sync, and traffic codes 8 through 13. The peak-to-average ratio of the signal selected is 10.1 dB.

All implementations of digital predistortion use a look-up table that stores the predistortion coefficients. This may take the form of a 2D table of I and Q predistortion coefficients, or a one-dimensional complex-gain table in polar or Cartesian format. Since in all cases, the look-up table is stored in read-only memory (ROM), the size of the table is a constraint that can limit system performance. Large table size is costly and can result in slow convergence in adaptive systems. If the table size is too small, however, quantization noise will be added to the predistortion coefficients and will degrade the performance of the predistortion algorithm.[6] The level of quantization noise that can be tolerated depends on dynamic-range requirements of the system. The described system uses a one-dimensional, complex-gain table. Figure 2 illustrates the output power-spectral density (PSD) of the amplifier as a function of look-up table size. The table size is shown as 2^N for varying N. Note that for table sizes of 256 or larger, the ACP response converges to its optimum value. This indicates that for the algo-

ALL IMPLEMENTATIONS OF DIGITAL PREDISTORTION USE A LOOK-UP TABLE THAT STORES THE PREDISTORTION COEFFICIENTS. SINCE IN ALL CASES, THE LOOK-UP TABLE IS STORED IN READ-ONLY MEMORY (ROM), THE SIZE OF THE TABLE IS A CONSTRAINT THAT CAN LIMIT SYSTEM PERFORMANCE.

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rithm used here, the noise floor of the system will not be constrained by the look-up table size if more than 256 entries are used.

In most cases, the predistortion coefficients are stored in ROM as finite-length digital words. The finite precision of the coefficient values adds an additional source of quantization noise to the system. As in the case of the look-up table size, this noise degrades system performance. Figure 3 illustrates the output PSD of the amplifier as a function of predistortion-coefficient output wordlength. Note that the system is less

THE NEXT COMPONENT IN THE SYSTEM THAT CAN DEGRADE PERFORMANCE IS THE DIGITAL-TO-ANALOG CONVERTER (DAC). THE DAC's CONTRIBUTION TO SIGNAL DISTORTION IS IN THE FORM OF QUANTIZATION ERROR.

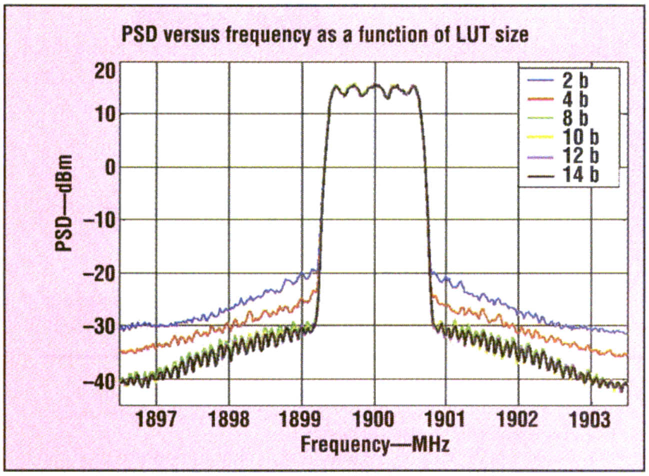
sensitive to this parameter than it is to the table size. A wordlength of 8 b or greater results in negligible performance degradation.

The next component in the system shown in Fig. 1 that can degrade performance is the digital-to-analog converter (DAC). The DAC's contribution to signal distortion is in the form of quantization error introduced due to its finite resolution. This finite resolution results in sharp edges in the DAC output signal, which imply the presence of undesired high-frequency components. These frequency components are aliased within the Nyquist band and appear as discrete spurs in the DAC output spectrum. The signal-to-noise ratio (SNR) of the DAC is given as:

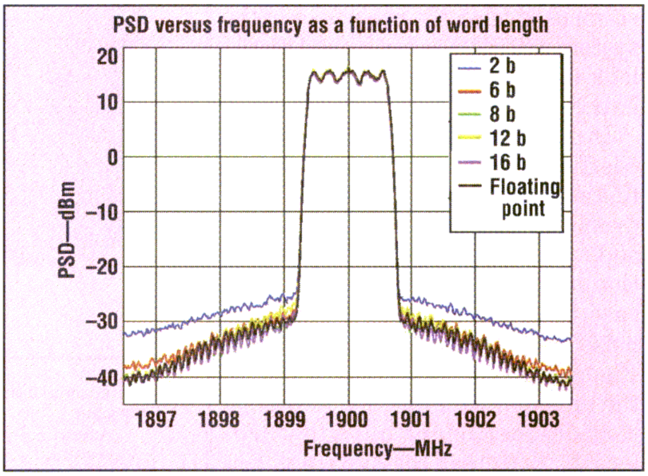
$$SNR = 6.02N + 1.73 + 20\log(FFS) + 10\log(F_{SOS}/F_S) \quad \text{(dB)}$$

where:

N = the number of bits in the DAC



2. This graph shows the PA-output PSD versus frequency as a function of look-up-table size (IS-95 signal, 10.1 dB peak-to-average ratio).



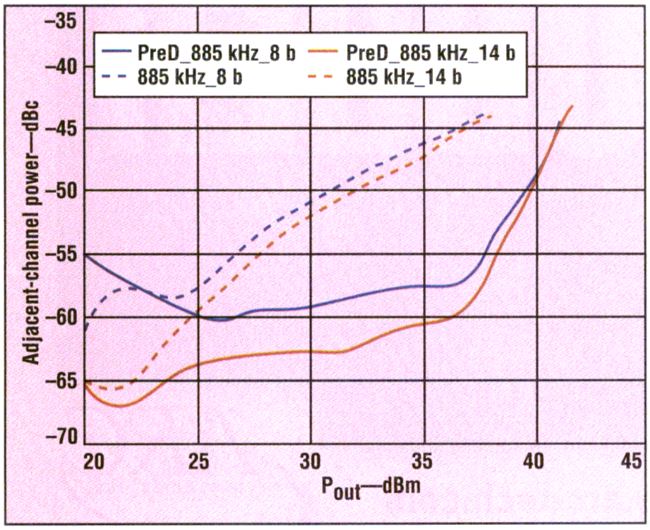
3. This graph shows PA-output PSD versus frequency as a function of word length (IS-95 signal, 10.1-dB peak-to-average ratio).

resolution,

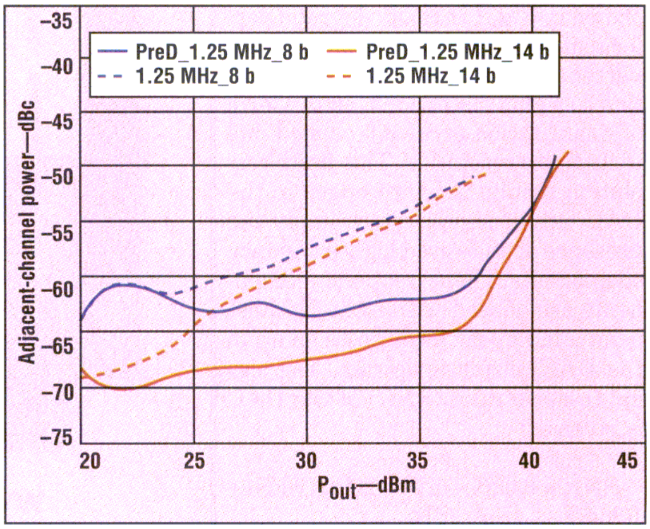
FFS = the fraction of fullscale at which the DAC operates,

F_S = the Nyquist sampling rate, and F_{SOS} = the oversampling rate.

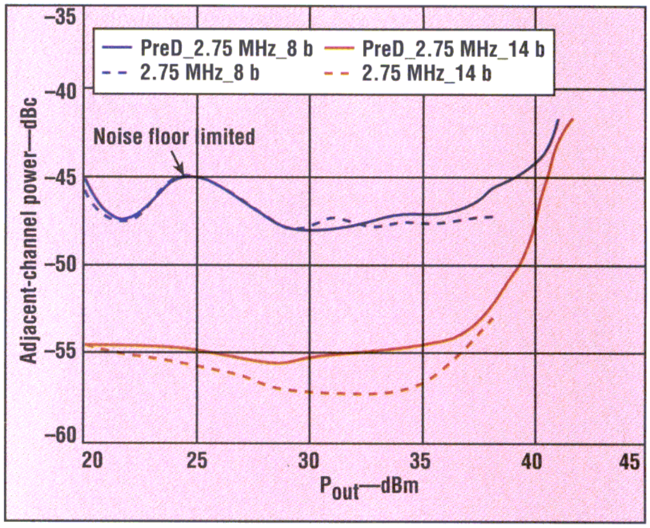
If this SNR is less than the level required by the system dynamic range, the improvements in ACP due to predistortion will be degraded by the DAC. To illustrate this effect, two DACs are considered in this system. The first DAC has a resolution of 8 b (N = 8), and the second DAC has a 14-b (N = 14) resolution. In both cases, the DAC full-scale range is used and 4-times oversampling is applied. Figures 4 to 6 show the adjacent-, first-alternate-, and second-alternate-channel powers for the system with the 8- and 14-b DACs, with and without the predistortion applied. For the adjacent- and first-alternate channels, the results without predistortion clearly show the reduction in noise-floor level when the DAC with higher resolution is used. As the input power is increased, the distortion due to amplifier nonlinearity begins to raise the noise floor and the DAC is no longer the limiting factor. In the case where predistortion is applied, a larger improvement in the noise floor is achieved for the 14-b DAC relative to the 8-b DAC. This occurs because some of the improvements attained through predistortion are canceled out in the 8-b DAC due to the large quantization noise. In both cases, the curve has a knee at approximately P_{out} = +37 dBm. This is the point at which the predistortion can no longer compensate for the amplitude modulation/amplitude modulation (AM/AM) of the PA, and the system noise floor degrades rapidly due to PA nonlinearity. In the case of second alternate channel, DAC resolution again plays a substantial role in reducing the noise-floor level, but clearly does not play a major role in limiting the performance of the pre-



4. This graph shows adjacent-channel power versus output power as a function of DAC resolution (IS-95 signal, 10.1-dB peak-to-average ratio).



5. This graph shows the first alternate-channel power versus output power as a function of DAC resolution (IS-95 signal, 10.1-dB peak-to-average ratio).

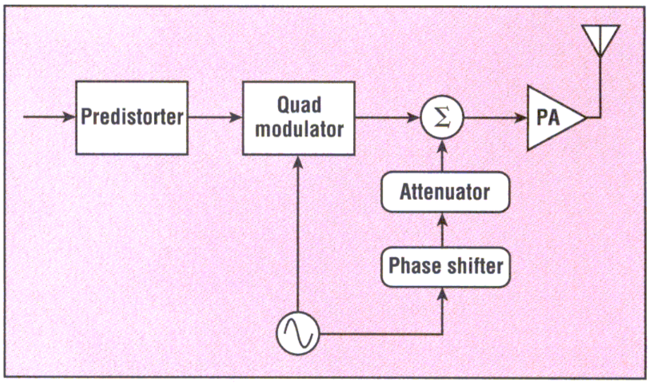


6. This graph shows the second alternate-channel power versus output power as a function of DAC resolution (IS-95 signal, 10.1-dB peak-to-average ratio).

distortion algorithm. This can be observed from the fact that the power levels for a particular DAC resolution are the same with and without predistortion.

The next element in the system shown in Fig. 1 contributing to distortion is the reconstruction filter. A reconstruction filter can be a major source of memory effects. Since most predistortion algorithms do not account for memory effects in the system, this can reduce system performance, particularly in wide-bandwidth applications. One approach to reducing the impact of the filter is to increase the sampling rate.[7] This, however, will increase cost and power consumption on the DSP side. Therefore, a combination of appropriate oversampling in conjunction with the proper reconstruction-filter type and filter-order selection will substantially reduce distortion at minimal cost. An analysis of reconstruction-filter effects on predistortion and guidelines in selecting the appropriate filter can be found in ref. 7.

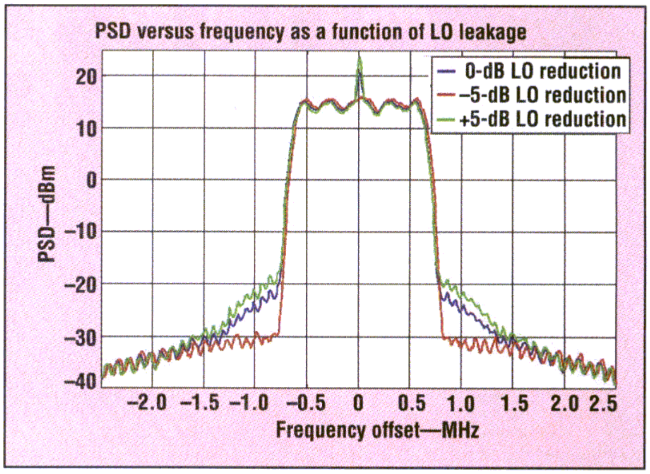
Quadrature modulators (quad mod) and demodulators (quad demod) are critical components of modern digital-communication transceivers. Although this func-



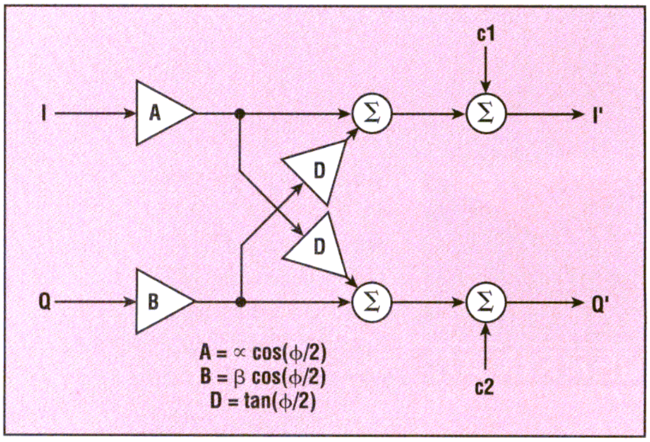
7. This is a block diagram of a digital predistortion system with an LO leakage-compensation circuit.

tionality can be implemented digitally, all-DSP quad mod and demod designs suffer from increased power consumption and low intermediate-frequency (IF) interface compared to analog implementations.[8] Therefore, analog modulators and demodulators are commonly used in most systems. The analog implementations suffer from several deficiencies, however, most notably amplitude and phase imbalance and local-oscillator (LO) leakage. These deficiencies can substantially degrade the performance of digital predistortion systems by eliminating most of the out-of-band emission reduction achieved through the predistortion algorithm.[8-14] LO leakage has a particularly negative impact on the performance of predistortion and must be suppressed by careful design of the modulator or by adding a compensation circuit to the system. A simple and effective method to overcome LO carrier feedthrough is to employ an attenuator and phase shifter in parallel with the quadrature modulator, as shown in Fig. 7. The phase shifter and attenuator are adjusted to cancel out the carrier-leakage component from the modulator output. A feedback mechanism using an envelope detector at the output of the summing junction can be added for adaptive compensation. To illustrate the efficacy of this technique, Fig. 8 illustrates the amplifier-output PSD for three different states of the LO leakage compensator. The 0-dB reduction refers to an amplifier without a compensation circuit. The −5-dB LO reduction occurs when the LO level relative to signal level is reduced by 5 dB (destructive interference), and +5-dB LO reduction occurs when the LO level relative to signal level is increased by 5 dB (constructive interference) from the no-compensation case. As illustrated, the ACP level is greatly impacted by the level of LO leakage present.

Even though analog compensation circuits can be used for LO suppression, they are not easily adaptable to changing system conditions and tend not to be very precise. Furthermore, gain and phase imbalances are very difficult to overcome with any analog solution. Digital compensation cir-



8. This graph shows the PA-output PSD versus frequency as a function of LO leakage (IS-95 signal, 10.1-dB peak-to-average ratio).



9. This block diagram shows a symmetric quadrature-modulator compensation circuit.

cuits can be used to overcome modulator imperfections with greater precision and adaptability.[10,12] If a mapping predistortion algorithm is implemented, the modulator compensation can be simply added to the predistortion coefficients. In the case of complex gain predistortion, a separate circuit is needed. This compensation circuit can take a number of symmetric or asymmetric forms. A common symmetric form of modulator compensator is shown in Fig. 9.[12] The digital modulator-compensation circuit can nearly eliminate any modulator defects and, if applied to an adaptive predistortion system, can remove nearly all constraints due to the modulator imbalance.

Although digital predistortion can be an effective method for linearizing RF PAs, imperfections in the transmission path between predistortion and amplifier can nearly eliminate all ACP reductions achieved by the system. Therefore, system components, including the look-up table, DAC, reconstruction filters, and quadrature modulators must be carefully chosen to achieve the desired effect. Some limitations can be overcome by redefining component specification. However, as in the case of the modulator, elaborate compensation circuits may be needed to overcome component limitations. Since compensation solutions are best implemented using digital circuits, this can easily be added to the overall digital predistortion system. ••

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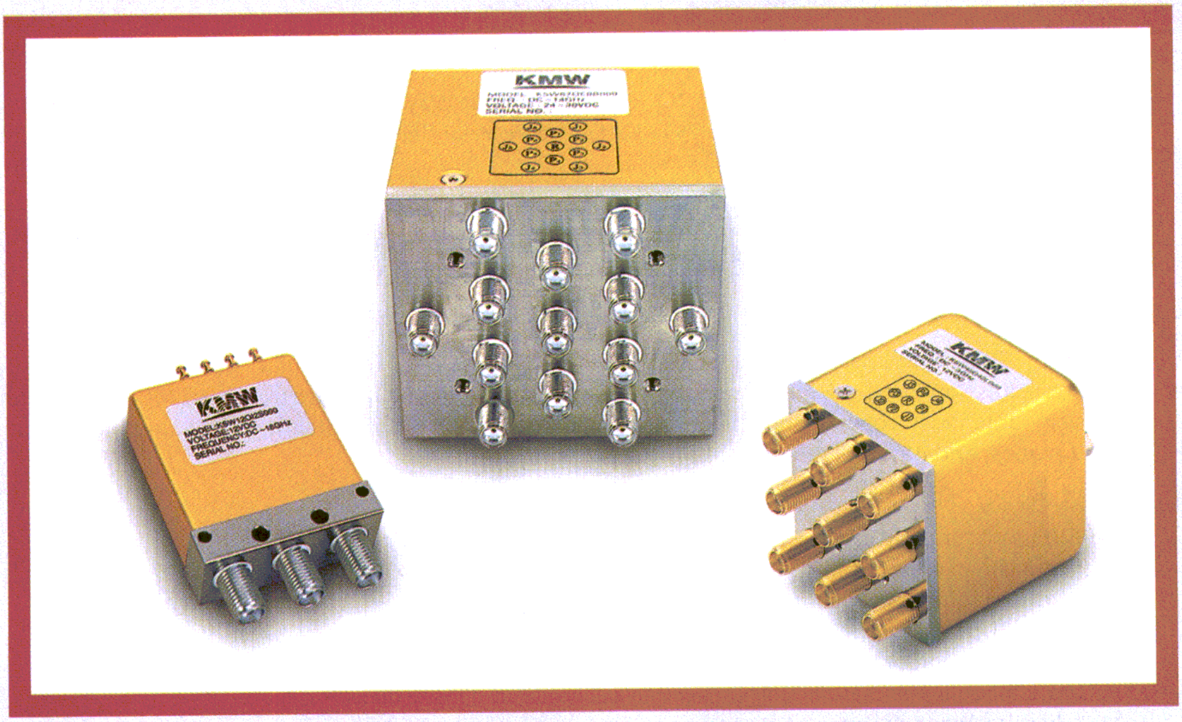

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Actuating Voltage /Current (Max.)	12Vdc ± 10% /240mA (@12Vdc, 25°C)	20 ~ 30Vdc /95mA (@24Vdc, 25°C)	24 ~ 30Vdc /85mA (@26Vdc, 25°C)
I/O Port Connector	SMA(F) / SMA(F)	SMA(F) / SMA(F)	SMA(F) / SMA(F)
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CIRCLE NO. 276

Measure INL And DNL For High-Speed ADCs

These two key parameters can be measured to provide an evaluation of the linearity and accuracy of high-speed analog-to-digital converters (ADCs) for test and communications applications.

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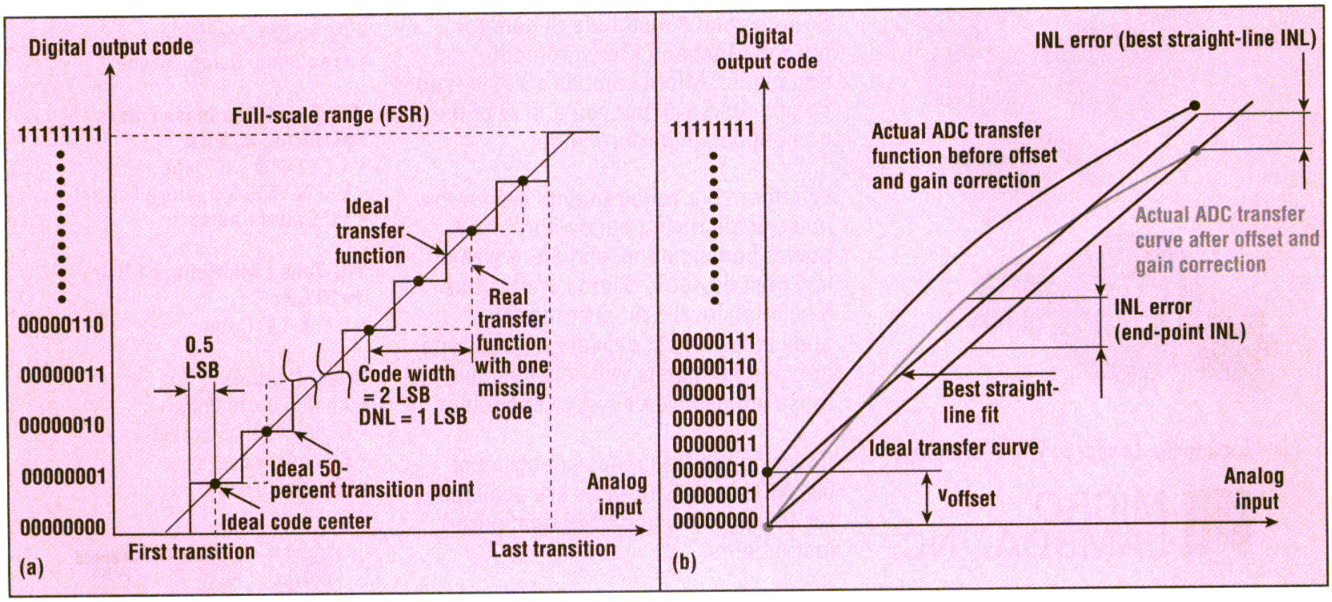
ANALOG-TO-DIGITAL converters (ADCs) are essential components in communications and test equipment. Their speeds have evolved to the realm of RF and microwave rates, capable of capturing more than one billion samples per second (1 GSamples/s). To evaluate the accuracy and linearity of these ADCs, two important parameters—integral nonlinearity (INL) and differential nonlinearity (DNL)—must first be measured.

Although these are not the most important electrical characteristics for highperformance data converters used in communications or fast data-acquisition (DAQ) applications, INL and DNL gain significance in imaging applications. The intent of this paper is to provide insight on the generation of powerful typical operating characteristics (TOCs), using tools and techniques that are simple but also smart and precise.

DNL error is defined as the difference between an actual step width and the ideal value of 1 least-significant bit (LSB) [Fig. 1a]. For an ideal ADC, where the DNL coincides with a DNL of 0 LSB, each analog step equals 1 LSB and the transition values are spaced exactly 1 LSB apart. A DNL error specification of less than 1 LSB guarantees a monotonic-transfer function—a monotonic ADC produces a non-decreasing/increasing digital out-



1. To guarantee that an ADC has no missing codes and a monotonic transfer function, its DNL must be less than 1 LSB (a). Best straight line and endpoint INL are two possible ways to define the linearity characteristic of an ADC (b).

UNDERSTANDING TRANSFER FUNCTION

The transfer function for an ideal analog-to-digital converter (ADC) is similar to a staircase where each tread represents a particular digital output code, and each riser represents a transition between two adjacent codes. The input voltages corresponding to these transitions must be located to specify many of the ADC's performance parameters. This chore can be complicated—especially for the noisy transitions found in high-speed converters, and particularly for digital codes that are near the final result and changing slowly.

Transitions are not as sharply defined as shown in Fig. 1b, but are more realistically presented as a probability function. As the slowly increasing input voltage passes through a transition, the ADC converts more often into the next adjacent code. By definition, the transition corresponds to that input voltage for which the ADC converts with equal probability to each of the flanking codes.

put code, when stimulated with a monotonically increasing/decreasing analog input, neglecting the effects of random phenomena—with no missing codes. DNL is specified after the static-gain error has been removed. Its definition is:

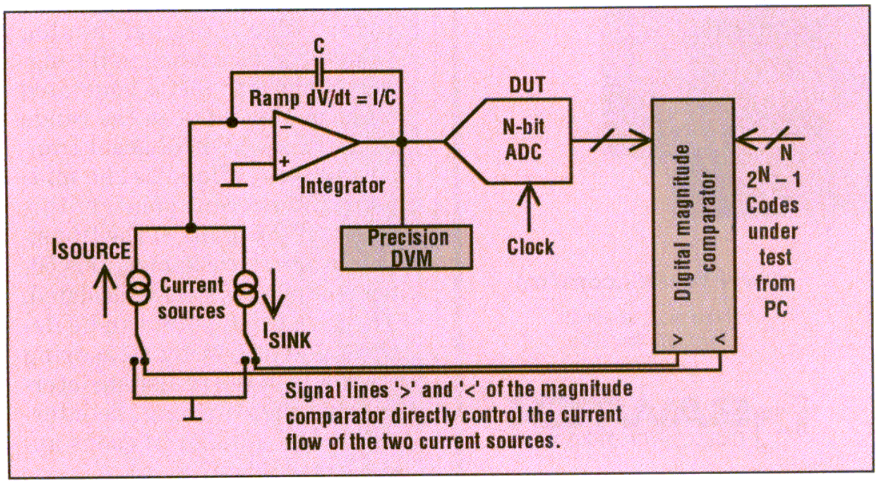
$$DNL = |[(V_{D+1} - V_D)/V_{LSB\text{-}IDEAL} - 1]|,$$
where $0<D<2^N-2$.

where:
V_D = the physical value corresponding to the digital output code D,
N = the ADC resolution, and
$V_{LSB\text{-}IDEAL}$ = the ideal spacing for two adjacent digital codes.

By adding noise and spurious components beyond the effects of quantization, higher values of DNL usually limit the ADC's performance in terms of signal-to-noise ratio (SNR) and spurious-free dynamic range (SFDR).

INL error is described as the deviation [in LSB or percent of full-scale range (FSR)] of an actual transfer function from a straight line (see sidebar, "Understanding Transfer Function"). The INL-error magnitude then depends directly on the position chosen for this straight line. At least two definitions are common: "best-straight-line INL" and "endpoint INL" (Fig. 1b):

The best-straight-line approach provides information about offset (intercept) and gain (slope) error, plus the position of the transfer function. It determines, in the form of a straight line, the closest approximation to the ADC's actual transfer function. The exact position of the line may not be clearly defined, but this approach yields the best repeatability, and it

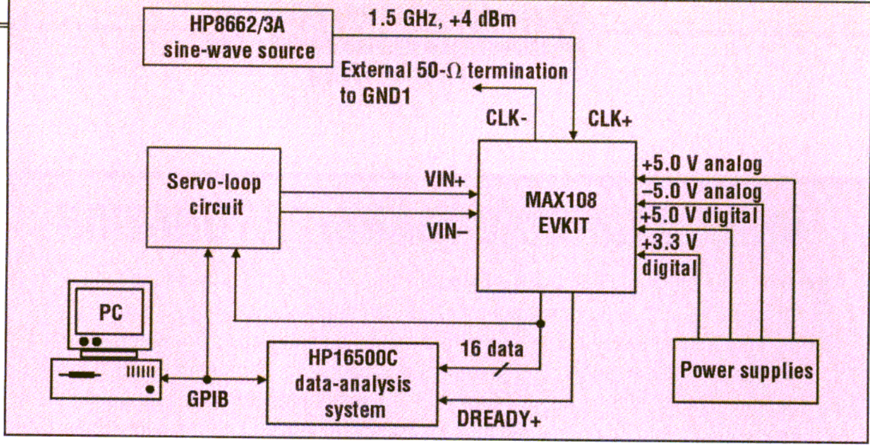


2. This circuit configuration is an analog integrating servo loop.

serves as a true representation of linearity.

Endpoint INL passes the straight line through the endpoints of the converter's transfer function, thereby defining the precise position for the line. Thus, the straight line for an N-bit ADC is defined by its zero- (all zeros) and full-scale (all ones) outputs.

The best-straight-line approach is preferred because it produces better results. The INL specification is measured after static offset and gain errors have been nullified and can be described as follows:



3. With the aid of the MAX108EVKIT and an analog integrating servo loop, this test setup determines the MAX108's INL and DNL.

$$INL = |[(V_D - V_{OFFSET})/V_{LSB\text{-}IDEAL}] - D|,$$
where $0<D<2^N-1$.

where:

V_{OFFSET} = the minimum analog input corresponding to an all-zero output code.

INL and DNL can be measured with either a quasi-DC voltage ramp or a low-frequency sine wave as the input.

TRACKING TRANSITIONS

A transition voltage is defined as the input voltage that has equal probabilities of generating either of two adjacent codes. The nominal analog value—corresponding to the digital output code that is generated by an analog input in the range between a pair of adjacent transitions—is defined as the midpoint (50-percent point) of this range. If the limits of the transition interval are known, this 50-percent point is easily calculated. The transition point can be determined at test by measuring the limits of the transition interval, and then dividing the interval by the number of codes appearing within it.

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CIRCLE NO. 236

A simple DC (ramp) test may incorporate a logic analyzer, a high-accuracy digital-to-analog converter (DAC) [optional], a high-precision DC source for sweeping the input range of the device under test (DUT), and a control interface to a nearby personal computer (PC) or X-Y plotter.

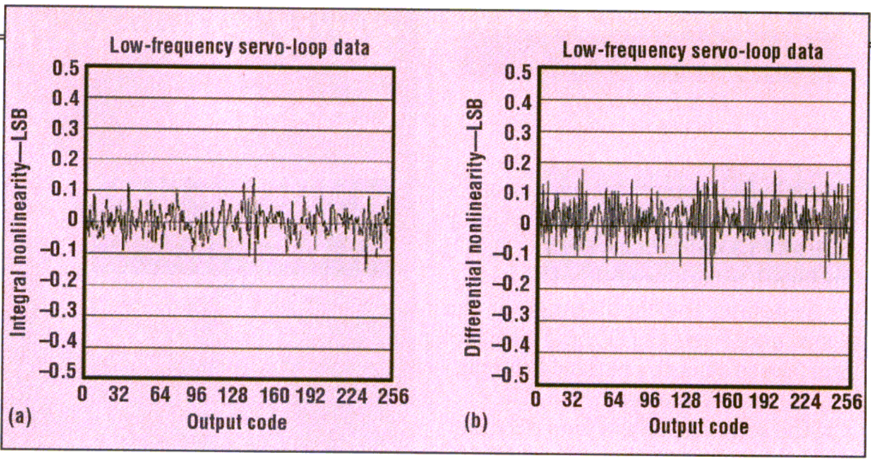
If the setup includes a high-accuracy DAC (higher than that of the DUT), the logic analyzer can monitor offset and gain errors by processing the ADC's output data directly. The precision signal source creates test voltages for the DUT by slowly sweeping through the input range of the ADC from zero-scale to full-scale. Once it is reconstructed by the DAC, each test voltage at the ADC input is subtracted from its corresponding DC level at the DAC output, producing a small voltage difference (V_{DIFF}) that can be displayed with an X-Y plotter and linked to the INL and DNL errors. A change in quantization level indicates DNL, and a deviation of V_{DIFF} from zero indicates the presence of INL.

Another way to determine static linearity parameters for an ADC, which is similar to the preceding example but more sophisticated, is an integrating analog servo loop. This method is usually reserved for test setups that focus on high-precision measurements rather than speed.

A typical analog servo loop (Fig. 2) consists of an integrator and two current sources connected to the ADC input. One source forces a current into the integrator, and the other serves as a current sink. A digital magnitude comparator connected to the ADC output controls both current sources. The other input of the magnitude comparator is controlled by a PC, which sweeps it through the $2^N - 1$ test codes for an N-bit converter.

If the polarity of feedback around the loop is correct, the magnitude comparator causes the current sources to servo the analog input around a particular code transition (see sidebar "Tracking Transitions"). Ideally, this action produces a small triangle wave at the analog inputs. The magnitude comparator controls the rate and direction of these ramps. The integrator's ramp rate must be fast when approaching a transition, yet be sufficiently slow to minimize peak excursions of the superimposed triangular wave when measuring with a precision digital voltmeter (DVM).

For INL/DNL tests on the MAX108, which is a high-speed ADC capable of sampling at 1.5 GSamples/s with 8-b resolution (see *Microwaves & RF*, March 1999, p. 128), the servo-loop board connects to the evaluation board through two headers (Fig. 3). One header sets a connection between the MAX108's primary (or auxiliary) output port and the magnitude comparator's latchable input port (P). The second header ensures a connection between the servo loop (the magnitude comparator's Q port) and a computer-generated digital reference code.



4. The left-hand plot shows typical INL for the MAX108 ADC, captured with the analog integrating servo loop (a), while the right-hand plot shows DNL for the MAX108, captured with the analog integrating servo loop (b).

The fully decoded decision resulting from this comparison is available at the comparator output P>QOUT, and is then passed on to the integrator configurations. Each comparator result controls the logic input of the switch independently, and generates voltage ramps, as required, to drive succeeding integrator circuits for both inputs of the DUT. This approach has its advantages, but it also has drawbacks:

- The triangular ramp should have low dV/dt to minimize noise. This condition generates repeatable numbers, but results in long integration times for the precision meter.
- Positive- and negative-ramp rates must be matched to arrive at the 50-percent point, and the low-level triangular waves must be averaged to achieve the desired DC level.
- Integrator designs usually require careful selection of the charge capacitors. To minimize potential errors due to the capacitors' "memory effect," for instance, select integrator capacitors with low dielectric absorption.
- Accuracy is proportional to the integration period and inversely proportional to the settling time.

A DVM connected to the analog integrated servo loop measures the INL/DNL error versus output code (Figs. 4a and b). Note that a parabolic

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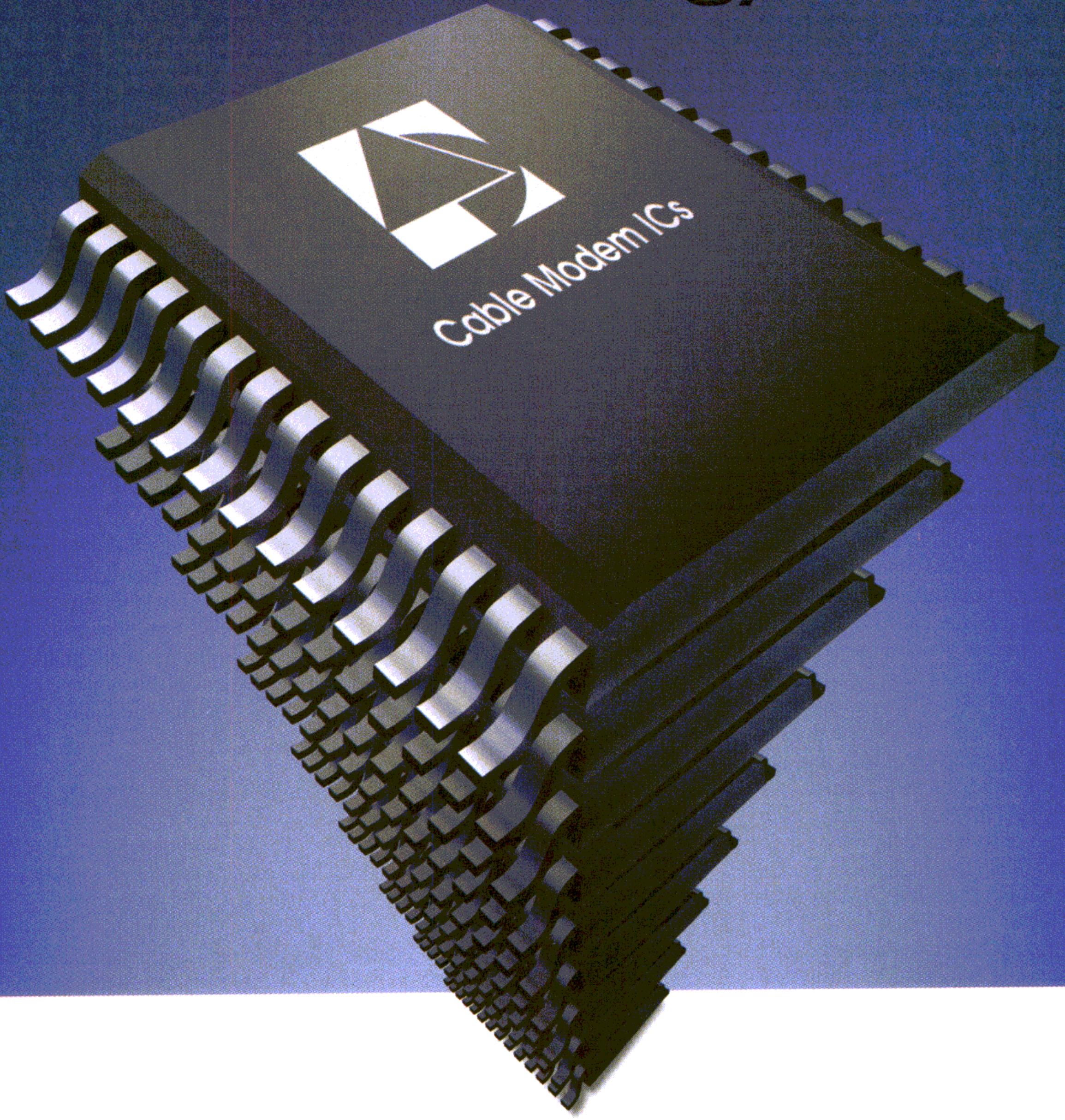

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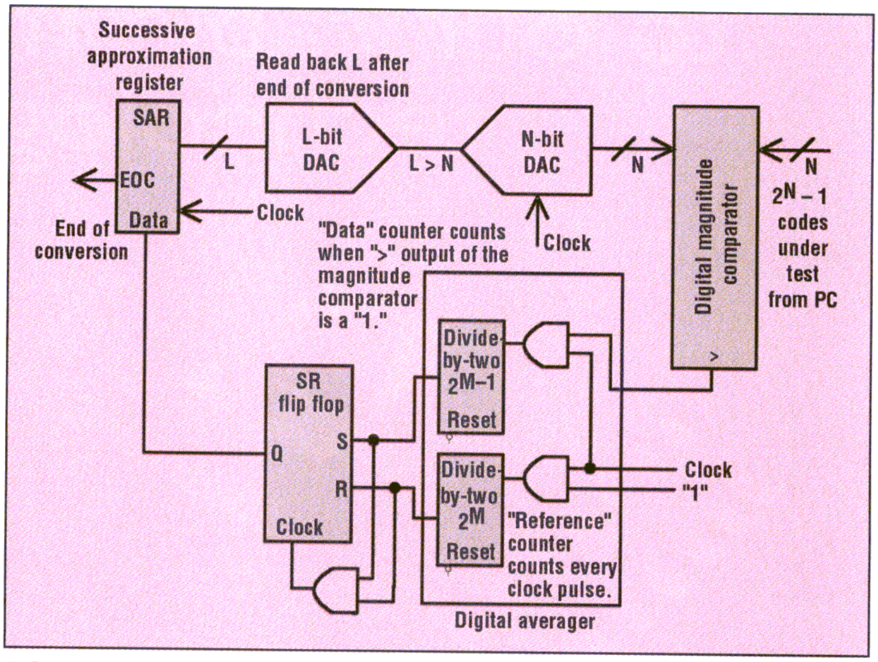
or bow shape in the plot of "INL versus output code" indicates the predominance of even-order harmonics, and an "S shape" indicates the predominance of odd-order harmonics.

To eliminate negative effects in the previous approach, one can replace the servo loop's integrator section with an L-bit successive-approximation register (SAR) that captures the DUT's output codes, an L-bit DAC, and a simple averaging circuit. Together with the magnitude comparator, this circuit forms a SAR-type converter configuration (Fig. 5) [see sidebar, "Using An SAR Converter"], where the magnitude comparator programs the DAC, reads its outputs, and performs a successive approximation. Meanwhile, the DAC presents a high-resolution DC level to the input of the N-bit ADC under test. In this case, a 16-b DAC was chosen to trim the ADC to 0.13-LSB accuracy and obtain the best possible transfer curve.



5. Successive approximation and a DAC configuration replace the integrator section of the analog servo loop.

The advantage of an averaging circuit is apparent when noise causes the magnitude comparator to toggle and become unstable, as it does on approaching its final result. Two divide-by counters are included in the averaging circuit. The "reference" counter has a period of 2^M clock cycles, where M is a programmable integer governing the period (and, hence, the test time). A "data" counter, which increments only when the magnitude comparator output is high, has a period that is equal to one-half of the first 2^{M-1} cycles.

The data counter executes a count only when the magnitude counter's output is high. Together, the reference and data counters average the number of highs and lows, store the result in a flip flop, and pass it on to the SAR register. This procedure is repeated 16 times (in this case) to generate the complete output-code word. Similar to the previous method, this one has advantages and disadvantages:

- The test setup's input voltage is defined digitally, supporting easy modification of the number of samples over which the result is to be averaged.
- The SAR approach provides a DC level rather than a ramp at the DUT's analog input.
- As a disadvantage, the DAC in the feedback loop sets a finite limit on resolution of the input voltage.

USING AN SAR CONVERTER

A successive-approximation-register (SAR) converter works similar to the old-fashioned chemist's balance. On one side is the unknown input sample, and on the other side is the first weight generated by the SAR/digital-to-analog-converter (DAC) configuration (the most significant bit, which equals half of the full-scale output). If the unknown weight is larger than one-half of the full-scale reading (FSR), this first weight remains on the balance. If the unknown is smaller, the weight is removed.

The SAR converter then determines the desired output code by repeating this procedure N times, progressing from the most-significant bit (MSB) to the least-significant bit (LSB). N is the resolution of the DAC in the SAR configuration, and each weight represents one binary bit.

To assess an ADC's dynamic nonlinearity, it is possible to apply a full-scale sinusoidal input and measure the converter's SNR over its entire full-power input bandwidth. The theoretical SNR for an ideal N-bit converter (subject only to quantization noise, with no distortion) is:

$$\text{SNR (in dB)} = N * 6.02 + 1.76.$$

Embedded in this figure of merit are the effects of glitches, INL, and sampling-time uncertainty. It is possible to obtain additional linearity information by performing the SNR measurement at a constant frequency and as a function of the signal amplitude. Sweeping the entire amplitude range, for example, from zero-scale to full-scale and vice-versa, produces large deviations from the source signal as the source amplitude approaches the converter's full-scale limit. To determine the cause of these deviations—while ruling out the effects of distortion and clock instability—a spectrum analyzer should be used to analyze the quantization-error signal as a function of frequency. ●●

For Further Reading

MAX108 data sheet, Rev. 1, 5/99, Maxim Integrated Products, Sunnyvale, CA.

MAX108EVKIT datasheet, Rev. 0, 6/99, Maxim Integrated Products, Sunnyvale, CA.

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

E. Sanchez-Sinencio and A.G. Andreou, *Low-Voltage/Low-Power Integrated Circuits and Systems(Low-Voltage Mixed-Signal Circuits*, IEEE, Piscataway, NJ, 1999.

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

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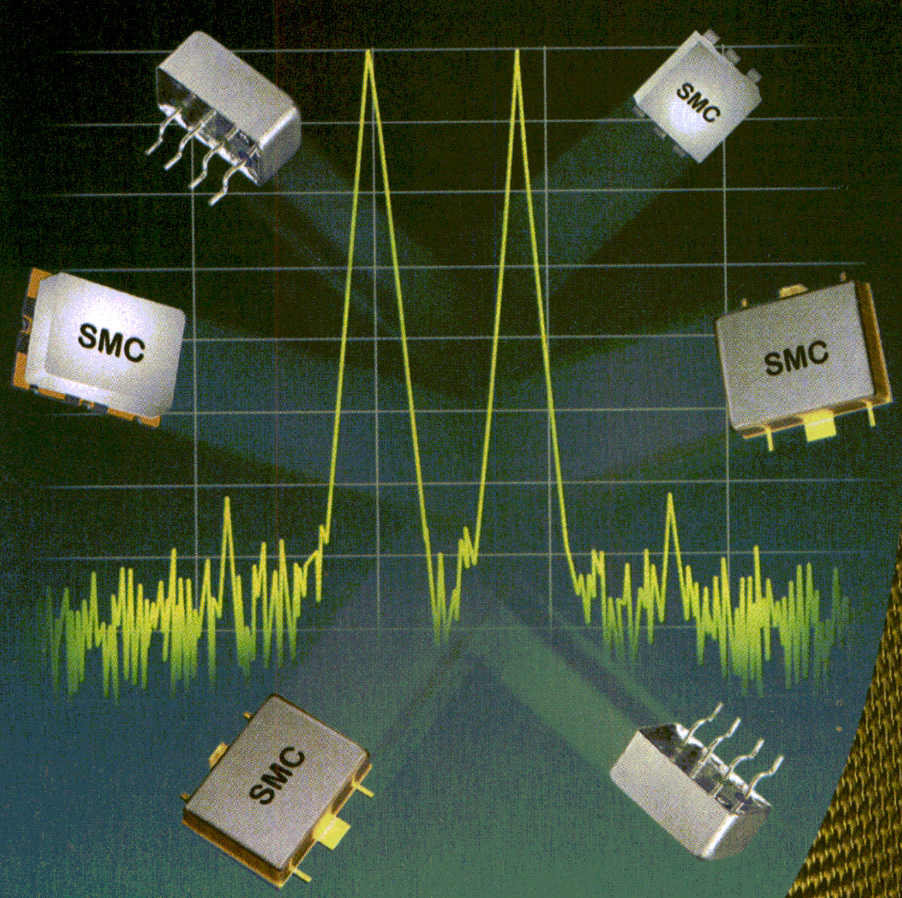

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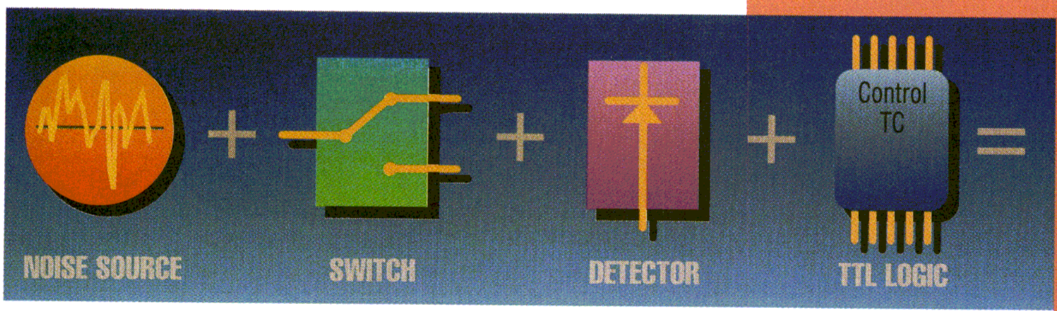

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DESIGN FEATURE

Wireless Connectivity

Modeling Systems Based On Bluetooth Wireless Connectivity

Accurate channel models and precise propagation simulations are needed to reasonably predict the performance of Bluetooth systems.

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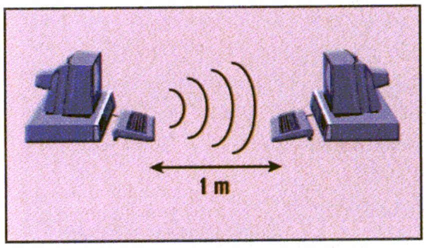
SHORT-RANGE wireless connectivity solutions for the home and office, such as Bluetooth, will simplify the transfer of control and data signals between electronic devices. As the number of devices equipped with wireless connectivity capability increases, it makes sense to evaluate the effects of multipath and Doppler shifts, due to reflectivity and mobility, on voice- and data-transmission quality. Identifying the sources of channel impairments on the signals transmitted and received with this enabling technology is important for proper use of any new devices which are designed for the underlying applications. In this article, residential homes and their short-range communication channel characteristics will be discussed. With small Doppler shifts and the absence of large delay spreads, testing and performance evaluations of the short-range wireless connectivity inside our homes may prove to be plain and simple when performed using designated RF faders and channel simulators.

Driven by declining prices for personal computers (PCs), free wireless phones equipped with the third-generation (3G) wideband-cellular-telephony technology, the boom in Internet usage, on-line trading, banking and shopping, recent trends in higher education that require college students to purchase and use their own computers, rapid increases in the number of home offices, and a drop in the percentage of corporate employees commuting to work, a typical home in the new millennium will be poised to house laptop and desktop computers, printers, scanners, pagers, and cordless wired and wireless telephones. Connectivity among these devises will become a necessity to share resources, such as scanners and printers, transfer e-mail, Internet, and video data among wired and wireless units. Connectivity must incorporate the mobility of pagers, wireless and cordless phones, as well as the immobility of PCs and the portability of laptop computers.



1. A typical application for IrDA devices is the exchange of data between two closely spaced laptop computers.

Home local-area networks (LANs) using telephone lines as a tool for connectivity are faced with difficult challenges due to variations in the lines' voltage levels and impedance, poor shielding, and improper line termination.[1] Furthermore, the use of phone lines, in general, limits mobility and does not lend itself to future integration of all electrical and electronic devices at home. It is foreseen that home appliances will soon be equipped with wireless transmission to communicate household information to PCs or wireless devices. The alternative to wired connectivity is the wireless short-range connectivity offered by Bluetooth technology at 2.4 GHz.

Short-range wireless connectivity between computer and telecommunication devices at home is within reach.[2] As more devices become available, this technology must be tested and evaluated using RF-channel faders and simulators which generate proper signal impairments for this application. As discussed in the following sections, due to the nature of human occupancy in residential homes and home sizes, Doppler and delay spread are minimal, causing the wireless indoor-communication channel to be either stationary or a very slowly varying Rician flat-fading channel.

Bluetooth[2] is a standard for short-range voice and data transfer through low-power RF transmission and reception. Bluetooth can transmit through solid, non-metal objects. Its nominal link range is from 10 cm to 10 m, but it can be extended to 100 m by increasing the transmitted power. It is based on a low-cost, short-range radio link, and facilitates ad-hoc connections for stationary and mobile-communication environments. This technology will enable users to connect to a wide range of computing and telecommunications devices without the need to buy, carry, or connect cables. It can be used in phones, pagers, modems, LAN-access devices, headsets, as well as notebook, desktop, and hand-held computers.

Bluetooth generally operates in the license-free 2.4-GHz industrial-scientific-medical (ISM) band. It offers the capability of non-line-of-sight transmission and reception through walls and other solid object with only moderate attenuation. This is in contrast to infrared (Ir) devices built according to the Infrared Data Association (IrDA) standards, which require a short direct nonobstructed path from the transmitter (Tx) to the receiver (Rx). Figure 1 shows a typical application of IrDA in exchanging business cards between two closely spaced laptop computers. The line-of-sight requirement limits connectivity, as it lacks the flexibility to connect a wireless device to a wired network, and thus reduces the options in placing a LAN-access point within the premises.

In contrast to IrDA, which requires that the Rx be within 30 deg. of a direct line of sight with the Tx, Bluetooth is omnidirectional. That is, the Tx signal is radiated with equal power in all directions, implementing point-to-multipoint propagation. In addition, Bluetooth offers 10 times the range of IrDA devices. Bluetooth uses frequency-hopping, spread-spectrum (FHSS) techniques to combat potential interlopers from baby monitors, garage-door openers, Specs are what you are given...

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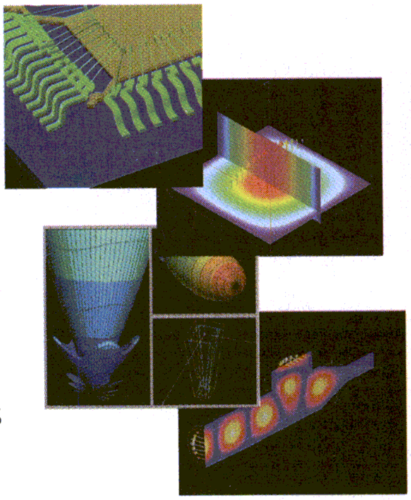

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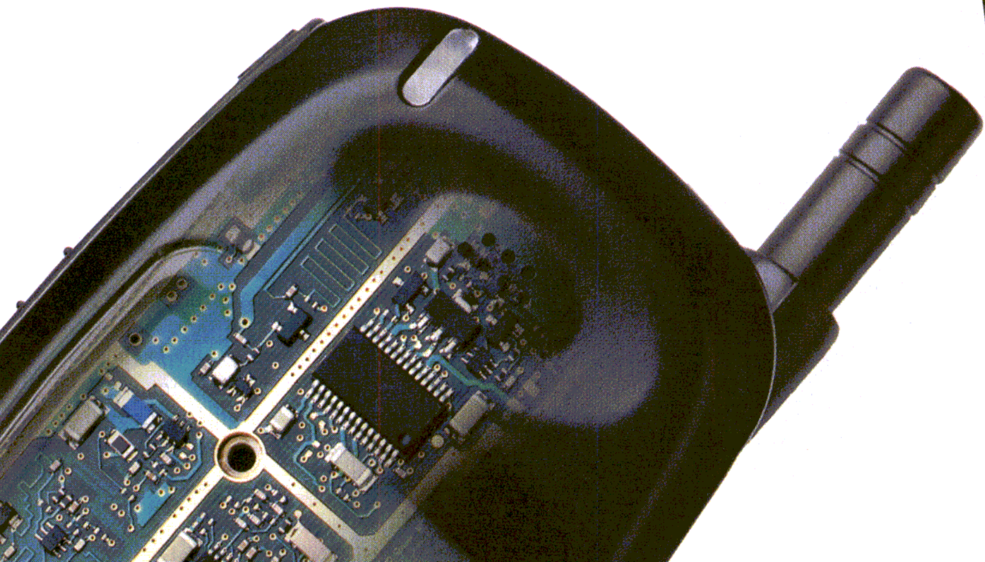

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cordless phones, and microwave ovens, all of which operate in the same frequency band. Bluetooth can support up to eight devices in a piconet, and more devices can be added by linking piconets. It has built-in security, and supports isochronous and asynchronous services and easy integration of transmission-control protocol/Internet protocol (TCP/IP) for networking.[2]

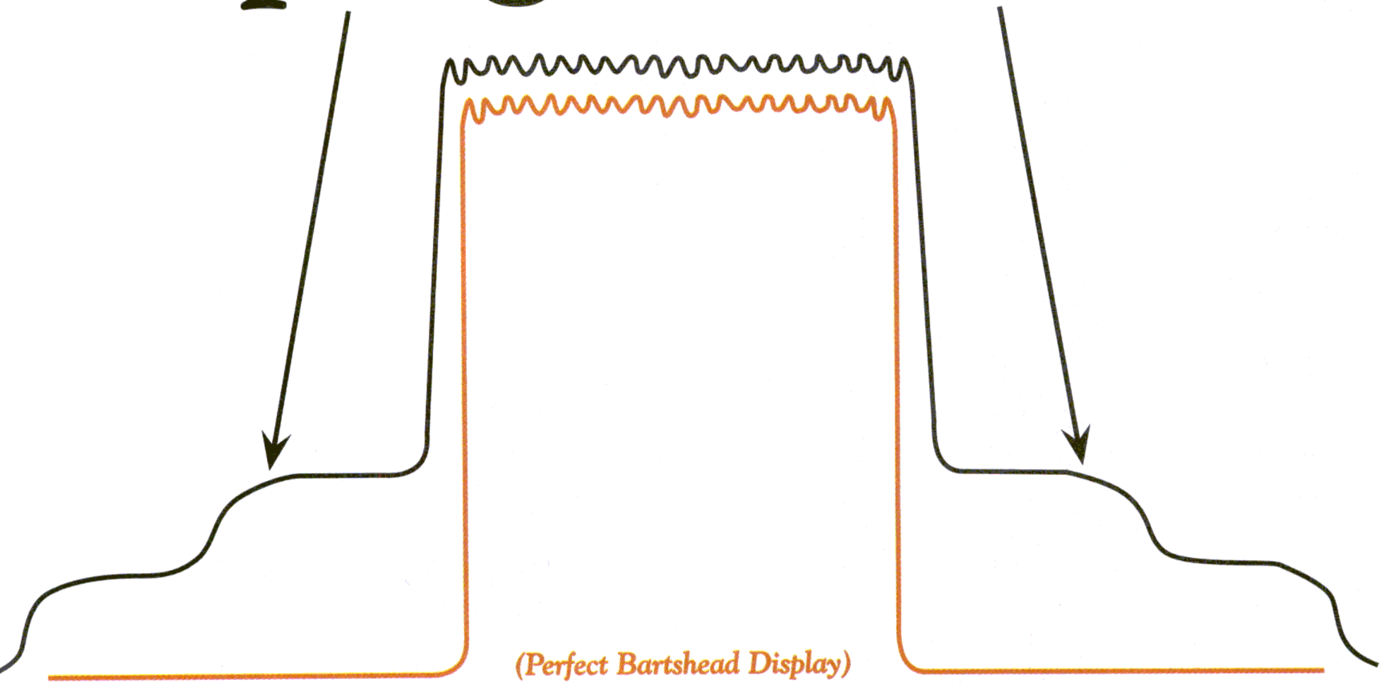
Wireless communications channel-modeling and signal-fading characteristics in residential homes are guided by three fundamental factors—the nature of human occupancy, home design, and structural materials. While the first factor mainly determines ambient motions and subsequently sets the Doppler shifts and defines the time-varying characteristics of the communication channel, the other two factors are mostly responsible for the multipath-profile signature, delay spread, and path loss between the Tx and the Rx. Spatial information regarding the correlation of signals among multiple antennas, direction of arrivals, and multipath angle spread are key parameters describing the spatial channel for smart-antenna technology. At present, these parameters do not play an important role in single, omnidirectional Rx/Tx communications associated with short-range connectivity.

There are fewer occupants in any residential home compared to the two cases of an office building or a commercial area. The latter includes manufacturing floors, shopping malls, storage places, and transportation stations. Often, people inside their homes are either sitting down or standing up. Furthermore, since it is not a standard practice to buy new furniture or move furniture around the house, home furniture and appliances should be considered in constant stationary modes and viewed as fixed reflectors of electromagnetic (EM) waves.

In indoor settings, PCs, laptop computers, printers, and other computer devices and equipment are not typically used while in motion. Subsequently, the only potential mobile Rxs or Txs in residential areas are typically cordless and wireless telephones and personal digital assistants (PDAs). Within the home, it is unlikely that a person would continuously walk during an entire call. Their movements would most likely be characterized by repeated patterns of moving and then being stationary. Therefore, Doppler spreads associated with the use of wireless units, caused by holding these mobile units or reflecting their signals, are smaller in residential homes when compared to other indoor-propagation environments.

In fact, Doppler spreads in homes will hardly reach the maximum frequency recommended by the Personal Communications Services (PCS) Joint Technical Committee for indoor pedestrian-communication environ-

Danger. Sloping Shoulders.



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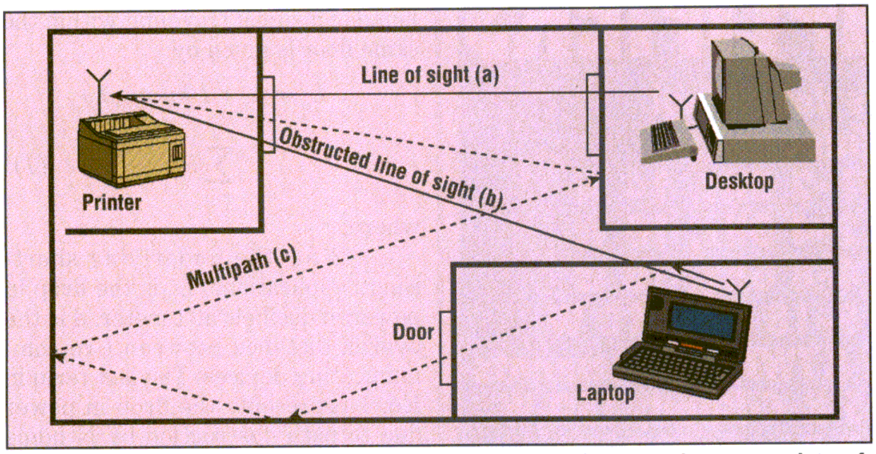

2. In a typical home environment, RF transmissions can experience a variety of multipath conditions.

ments. This frequency is 9.6 Hz for services operating at 2 GHz. By examining the nature of human occupancy, it is safe to assume negligible Doppler spreads in a typical residential home over long periods of daytime, and zero Doppler spread during the night.

Residential homes represent the smallest sites for indoor communications. For short-range connectivity and by the virtue that the Tx and Rx are contained in the same home, there is modest path loss due to the distance traveled in free-space at home compared to outdoor propagation. If the loss in the first meter is 30 dB,[3] and a free-space loss of 6 dB per octave is factored in, then a house that is 24 m long will attenuate the signal by 54 dB, as it propagates from one end of the home to the other. The signal-power attenuation is also caused by losses associated with signals traveling through exterior and interior building structures. Residential homes are typically wooden-framed, low-ceilinged, single-family units with one or two stories. The interior walls are usually covered with a thin layer of plaster inside cardboard. The exterior frame is often filled with insulation and covered by layers of plywood and wooden or aluminum (Al) siding or brick. In contrast to indoor-office areas, houses are not usually built using metallic studs and concrete frames and their floors often do not contain large amount of metal and concrete. There are almost no, or very few, large metal objects or cubicles, for which there is no application or use at home. The table shows typical partition losses that were measured in office buildings.[4] Comprehensive tables for average signal-loss measurements reported by various researchers for radio paths obstructed by common building material are given in Ref. 5. The first four measurements in the table were gained at 815 MHz, whereas the last two measurements in the table were performed at 9.6 MHz.

The last three values in the table apply to residential-home indoor communications and show losses possibly up to a maximum of 7 dB per interior partition. Signals penetrating the exterior walls are sufficiently attenuated to be considered a distressing source of interference to nearby homes. Conversely, it is maintained that, unlike commercial areas, residential homes have several glass windows that may remain open most of the summer and spring seasons. Signals escaping one home to a neighboring home in very close proximity through open windows can be considered a worrisome source of interference that can be eliminated only if FHSS-communication protocols are different in the two homes. More typical and fundamental interference sources in short-range communications in residential areas are caused by the microwave oven and garage-door openers.

Detailed statistical modeling and computer simulation of indoor radio channels can be found in refs. 6-8. The formula for path loss in a residential home of a signal traveling within the same floor is given by:

$$L_p = L_o + 10n\log_{10}d + \sum_{m=1}^{N} P_m \qquad (1)$$

where:

n = 2 for the underlying short-range applications, L_o is the drop in power in decibels at 1 meter, d is the overall distance between two communicating devices. The last term in Eq. 1 represents the drop in power due to path obstruction by N home interior walls. The terms in the summation in Eq. 1 may be considered equal, as the interior walls do not vary much in texture and thickness.

Figure 2 shows one possible short-range connectivity setting that includes a laptop, PC, and a printer, each in a separate room. With the room doors open, the PC has a line of sight to the printer (a). In this case, only the first term of the path loss in Eq. 1 applies. Conversely, the laptop and the printer can only communicate either through an obstructed line of sight or multipaths. With the transmitted signal power dropping 6 dB per octave, and considering typical room and hallway sizes in residential homes, it is easily realized that the transmitted signal along path (b) obstructed by a plywood wall is received at the printer with higher power than the signal along multipath (c). The latter looses power due to the long distance traveled and also due to the attenuation incurred from the reflections against the interior walls.

It should be mentioned that there are several other possible multipaths in the specific home structure depicted in Fig. 2 that could connect the printer to the two computers through reflections from walls, floors, and ceilings. Many of those multipaths, after a certain number of reflections and transmissions through walls are sufficiently attenuated and their effects on signal fading are therefore negligible. It is also important to point out that Fig. 2 only shows one example of specular multipath. This is because at approximately 2-GHz center frequency, most surfaces of reflection in homes

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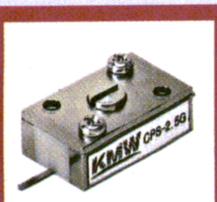
VSWR (Max.)	1.25:1	1.25:1	1.25:1
Incremental Phase Shift	90 degree min. @ 2GHz		
Electrical Delay	125 psec min.		
Nominal Impedance	50 ohm		
I/O Port Connector	SMA(F) / SMA(F)		
Average Power Handling	20W @ 2GHz		
Temperature Range	-30°C ~ +60°C		
Dimension (inch)	A type : 1.496*1.102*0.457 B type : 1.225*1.102*0.457		



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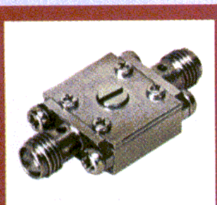

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Insertion Loss (Max.)	0.15dB	0.25dB	0.35dB	0.15dB	0.25dB	0.35dB
VSWR (Max.)	1.3:1	1.3:1	1.3:1	1.25:1	1.25:1	1.25:1
Incremental Phase Shift	30 degree min. @ 2GHz			35 degree min. @ 2GHz		
Electrical Delay	41.7 psec min.			48.6 psec min.		
Nominal Impedance	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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are relatively smooth and the scattered media that produces diffuse reflections are minimal. Therefore, diffuse multipaths for short-range connectivity applications in homes can be ignored.

As mentioned previously, the bit rate used in Bluetooth is 10 Mb/s (i.e., the bit duration is 100 μs). EM propagation travels at the speed of light, or 3×10^8 m/s, so one bit is approximately 30 m long. Frequency-selective fading responsible for intersymbol interference would occur if the path differences inside the residential home were significant portions of 30 m. Propagation experiments conducted for small buildings show that the maximum delays are in the order of 100 ns (this value will become smaller when specifically considering residential homes). The distance traveled over the aforementioned maximum delay spread is 30 m. Not counting the loss due to reflections, this 30-m distance causes an approximate 30-dB drop in power relative to a signal that is received one meter away from the Tx. This drop is significant and causes any resolvable multipath to be weaker than the first signal arrival. As a result, Bluetooth-signal propagation inside residential homes is primarily guided by a flat fading channel, where all effective multipaths arrive within the information symbol.

An indoor wireless channel is often described by Rician fading, where the probability-density function of received signal envelope x is given by:

$$f(x) = \frac{x}{\sigma^2} e^{-(x^2+A^2)/2\sigma^2} I_0\left(\frac{Ax}{\sigma^2}\right),$$

$$\textit{with } K = \frac{A^2}{2\sigma^2} \qquad (2)$$

where:

$I_0(.)$= the modified Bessel function of the first kind and zero order. Parameter A is greater or equal to zero and denotes the peak-to-zero value of the specular radio signals comprised of the superposition of the dominant line-of-sight signal and the time-invariant scattered signals reflected from walls, ceilings, and stationary objects. Parameter σ^2 represents the average power of the signal received over paths that vary with time due to people moving within the house. Parameter K is the Rician distribution. When K equals infinity, the channel is Gaussian, whereas K = 0 defines the Rayleigh channel. A value of 3 dB for K is typical for modeling indoor radio-channel amplitude fluctuations.

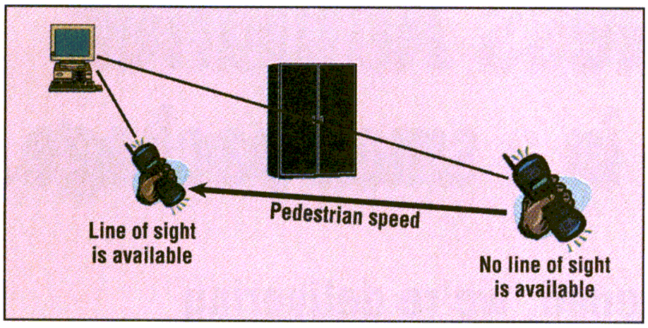
If the Tx and Rx are stationary with no pedestrian movements, as in the case depicted in Fig. 2, the communication channel is constant with time-invariant impulse response and zero Doppler. In this case, the channel is deterministic and its multipath characteristics remain constant over a long period of time. Depending on the locations of the Tx and the Rx antennas inside the home, the multipath components may add constructively or destructively. As such, the received signals may be amplified by a constant factor or remain in a deep fade. It is therefore important in Bluetooth applications to find "blind spots" as well as optimum locations. An example of this is a laptop computer in each room of the house which communicates with a fixed printer that is stationary in a specific room. These locations should be determined in most likely propagation scenarios, that is, they should not be influenced by Doppler shifts but rather should be identified in stationary settings when most of the house occupants are away, or in "stand-still" positions.

With no randomness induced in the communication channel, the ratio of the square of the mean to the variance in Eq. 2 becomes extremely high and K can be simply approximated by a positive infinity.

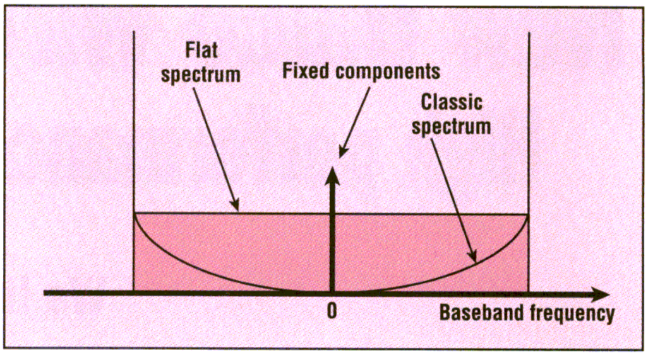
In the second type of propagation environment, the Tx and Rx are stationary, but there are some ambient motions of people within the house. This causes the communication channel to be slowly time-varying. In this case, the multipath components can be divided into two different categories. The first category is comprised of the dominant line of sight signal, if it exists, in addition to all time-invariant scattered signals reflected from the stationary objects such as walls, ceilings, furniture, and appliances. The other category consists of the multipaths whose first- or higher-order reflection patterns include at least one path that bounces from a nonstationary source in the house. While the lifetime of the multipath signals in the first category is very long and may last until the signal transmission is terminated, multipath signals in the second category may cease to exist shortly after they became established. The contribution of the first category of multipaths to the signal-fading environment is deterministic with zero Doppler, whereas the effect of the second multipath category is stochastic and changes the signal correlation and frequency characteristics. The combined categories yield Rician fading described by Eq. 2 with σ assuming nonzero values.

BECAUSE OF POTENTIAL MUTIPATH CONDITIONS WITHIN INDOOR HOME AND OFFICE ENVIRONMENTS, IT IS IMPORTANT IN BLUETOOTH APPLICATIONS TO FIND "BLIND SPOTS" AS WELL AS OPTIMUM LOCATIONS.

In the third type of propagation environments, either the Rx and/or the Tx are in motion. This, indeed, introduces Doppler effects on the transmitted signal with frequency changing up to approximately 10 Hz. The movement, as well as displacement of the antennas inside the house, may create a new line of sight or obstruct an already existing one (Fig. 3). It may also change the signal, scattering propagation profiles near the Txs and the Rxs, and may consequently give rise to random scatter-



3. When a Tx and/or Rx moves within the home environment, new conditions of multipath can be created and line-of-sight signal paths can be obstructed.



4. The classic of U-shaped Doppler spectrum is shown in contrast to a flat-shaped Doppler spectrum.

ing that defines the probability-density function of Eq. 2.

The correlation of the fading envelope is determined by the relative power carried by the different scatterers and their corresponding angles of arrival. In isotropic scattering, the power is uniformly distributed over 360 deg. This leads to a U-shaped spectrum (Fig. 4), which is known as the classic Doppler spectrum. While the classic spectrum is often assumed in outdoor propagation, a flat Doppler spectrum is shown to be characteristic of indoor communications.[3] In the flat Doppler spectrum, the uniformity is encountered across frequencies rather than over the angles of arrivals. This type of spectrum is caused by random movements of scattering elements in the area of the communication path, from random movement of the Tx or Rx causes. A flat Doppler spectrum is also shown in Fig. 4, where the impulse represents the presence of the line-of-sight signal path that can be associated with the classic and flat Doppler spectra. The impulse location at zero frequency indicates, among other possibilities, the orthogonalilty between the direction of motion of the Rx and the line connecting the Rx and transmitting antennas.

FADING SIMULATORS

New RF techniques in channel simulation can work well for testing and characterizing short-range indoor-radio modules. These techniques employ combinations of programmable phase shifters, attenuators, and vector modulators to achieve the desired fading profiles. The RF-channel simulators are simpler in design complexity than their digital-signal-processing (DSP)-based counterparts and are typically less costly and easier to program and operate. In contrast to DSP solutions, an RF-channel simulator does not require signal downconversion and upconversion and, as such, does not introduce distortion products and frequency-conversion errors into a channel which otherwise would not exist. There are several factors that favor a channel simulator where the signal processing is performed at RF.

A flat-fading channel with one or two resolvable multipaths can be achieved easily and cost effectively. An RF-channel simulator requires that the incident signal be split into a number of resolvable multipath components appropriate for simulating a particular operating environment. Each signal component is faded independently and undergoes its own signal processing. The multipath signals are then recombined. While this can become cumbersome with twelve multipaths, for the underlying indoor applications, it can be performed using a simple power splitter, combiner, and two signal-processing components.

The short propagation delays associated with indoor communication can be easily handled by a channel simulator using RF signal processing. Delays are typically short when they can be achieved using simple, commercially available delay lines.

The demand for bandwidth in wireless LAN (WLAN) systems is increasing. In addition, techniques that employ direct-sequence spread spectrum (DSSS) with frequency hopping are being developed to combat potential interference generated from home appliances and shared spectrum users. Effective interference suppression requires large spreading of the transmitted signal bandwidth. A channel simulator which performs signal processing at RF is only band-limited by the RF control components processing the signal. The extent of signal bandwidth at the 2.4- and 5.8-GHz ISM bands may become unsuitable for even the most-powerful DSP-based commercially available channel simulators.

In conclusion, new short-range wireless connectivity technologies that are cable-free and allow electronic devices to talk to each other with little user intervention are within reach. The home environment is seen as a major growth area and the principal host of these technologies. This is mainly due to the increase in information appliances and Internet-borne services. Testing and evaluating the performance of present and future short-range wireless connectivity can be achieved by a new generation of simple RF faders and channel simulators designed for this purpose.●●

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Determine IMD Products For Digital Communication Standards

A physical-statistical investigation into the frequency offsets of IMD products correlates IMD and ACPR for several digitally modulated communication standards.

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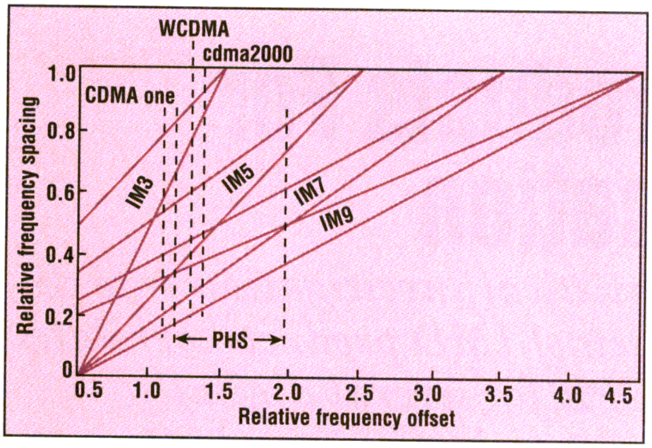
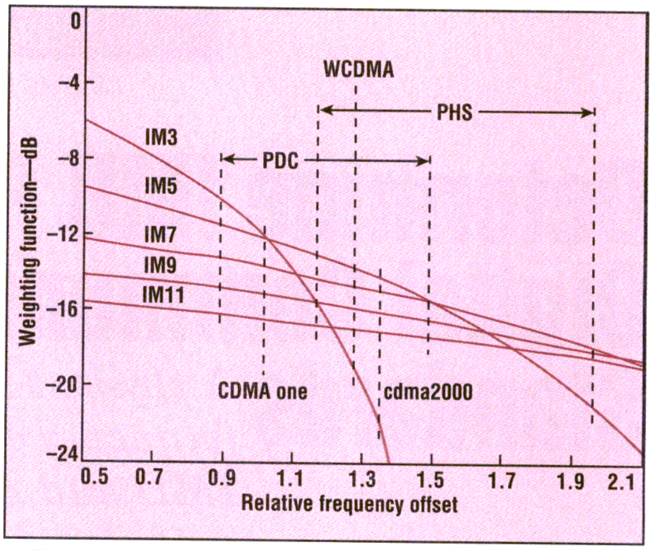
NONLINEAR effects of RF power amplifiers (PAs) in digital wireless communications systems often result in spectral regrowth—the generation of noise outside the system's allocated RF spectrum. This noise can create interference in a carrier's own receiver or in the receivers of nearby or co-located competing carriers. To analyze the results of these effects, many engineers use discrete-Fourier-transform analysis while driving a PA with pseudorandom data streams.[1] Another approach is to drive a PA with a statistical power distribution and evaluate distortions using techniques such as crest-factor, cumulative-distribution function, and clipping effects.[2,3] All of these approaches require single-tone, amplitude-modulation (AM)/AM, and AM/phase-modulation (PM) waveforms and sophisticated software. This article investigates the frequency offsets of intermodulation-distortion (IMD) products to find a correlation between IMD and adjacent-channel-power ratio (ACPR) for several digitally modulated communication standards having flat-power spectra. It proposes a physical-statistical approach based on the two-tone analysis proposed in Ref. 4, which supports easy analysis of ACPR on the basis of IMD. Representing a signal in a statistical manner in the frequency domain helps to introduce weighting functions for each IMD product, as well as demonstrating the contribution of all IMD products to system performance as a whole. This approach clarifies the tuning procedures for a system and helps it meet a desired specification.

Consider the case where a PA is driven by a statistical signal with an idealized, rectangular, flat-shaped average-power spectrum over a frequency band (Δf) that is a close approximation of quadrature-phase-shift-keying (QPSK), quadrature-amplitude-modulation (QAM), and orthogonal-frequency-division-multiplexing (OFDM) signals. One can assume that all of the useful information of a signal is restricted by margins $\pm\Delta f/2$ from the carrier frequency, and that the entire input power is placed inside these margins in the frequency domain. The power spectrum is solid due to the non-periodic driving of the PA. Taking into account the physical entity of the statistical signal's average-power spectrum, one can represent this signal in the frequency domain statistically with a flat probability-distribution function within the margins of Δf. This implies that each average-power spectral component inside the flat shape is composed of the instantaneous input signal on the same frequency, with an additional probability of the appearance of a signal on that frequency. Thus, the probability of the appearance of the input-signal



1. This plot shows relative frequency spacing versus relative frequency offset.



2. This plot illustrates weighting function versus frequency offset.

spectral component inside the limits df_1 is:

$$g(P_{in})\frac{1}{\Delta f}df_1 dP_{in} = g(P_{out})\frac{1}{\Delta f}df_1 dP_{out} \qquad (1)$$

where:

$g(P_{in})$ and $g(P_{out})$ are densities of the probability-distribution functions for the PA's input and output power, respectively. At the same time, the spectral component inside the limits df_2 exists with a similar probability. For the flat spectrum shape, one can suppose that the spectral components at df_1 and df_2 are not correlated. These simultaneous spectral components cause instantaneous IMDs of the PA-output signal at frequency offsets determined by a difference between df_1 and df_2. In this case, the probability of the appearance of IMD products at different frequency offsets can be represented as:

$$g(IMD)\frac{1}{\Delta f} * \frac{1}{\Delta f}df_1 df_2 dIMD = g(IMD)\left(\frac{1}{\Delta f}\right)^2 df_1\ df_2 dIMD \qquad (2)$$

where:

$g(IMD)$ is the density of a probability-distribution function for IMD imposed by an instantaneous input signal. Assuming that the main contribution to IMD is produced by frequency components with equal power levels, the average value of IMD at different offsets can be expressed as:

$$IMD = \int_{-\infty}^{\infty}\int_{\Delta f_1}^{\Delta f_2}\int_{\Delta f_1}^{\Delta f_2} IMDg(IMD)\left(\frac{1}{\Delta f}\right)^2 df_1 df_2 dIMD \qquad (3)$$

The frequency spacing where IMD is defined is given by the following equations:

$$\frac{N-1}{2}\Delta f_1 = f_0 - \frac{\Delta f}{2}$$

$$and\ \frac{N+1}{2}\Delta f_2 = \Delta f_1 + \Delta f \qquad (4)$$

where:

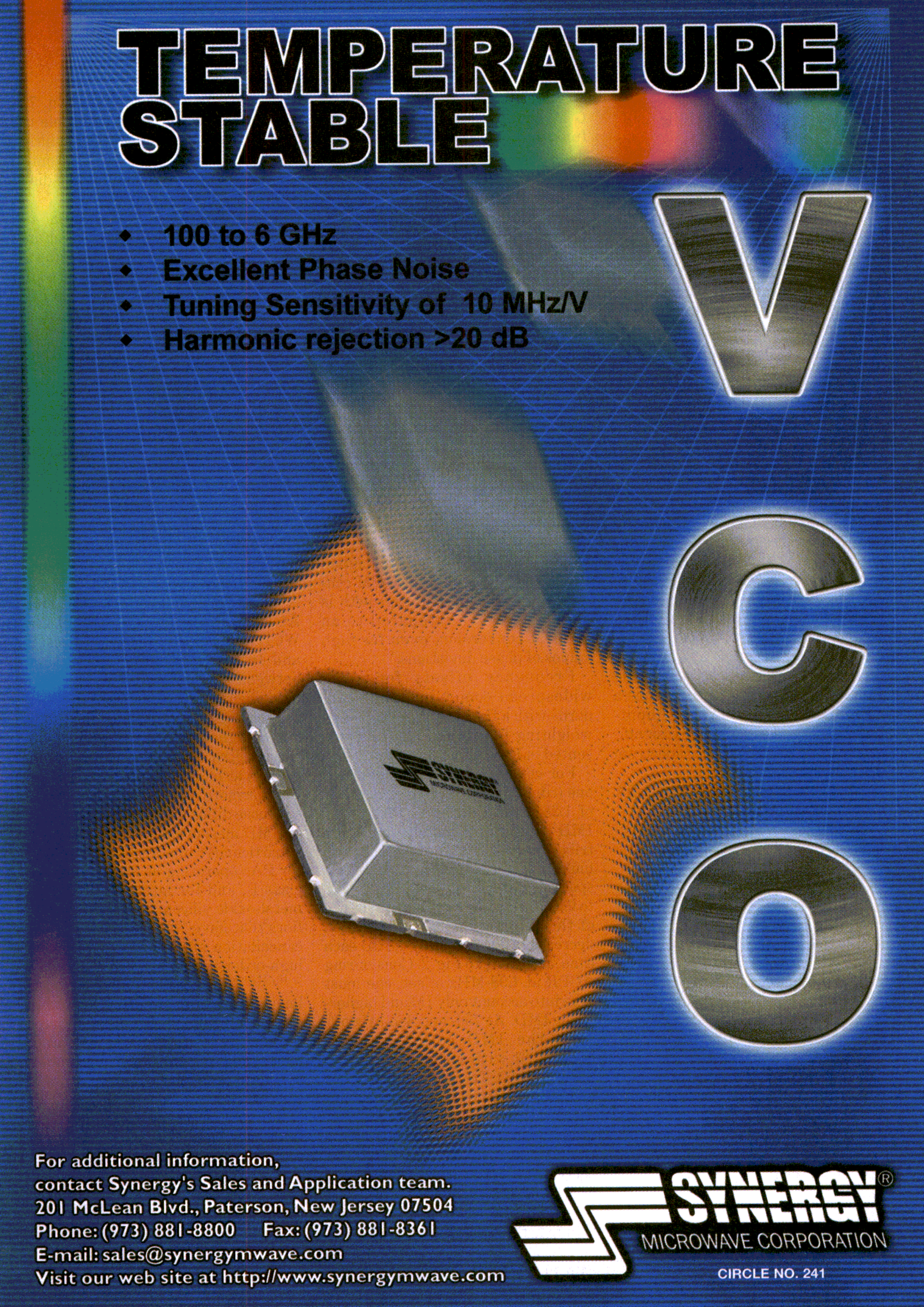
f_0 is the offset from the center frequency of the spectrum where one calculates the spectral regrowth, and N is the order of the IMD considered. Figure 1 shows these offsets and spacing margins at relative values. It can be seen that IMD must be measured at different spacing for diverse frequency offsets. Furthermore, this spacing depends on the IMD order. Moreover, even for a particular offset at a fixed IMD order, one must consider IMD through a wide range of frequency spacing depending on the offset value. This range increases with decreasing f_0. At offsets $f_0/\Delta f > 1.5, 2.5, 3.5$, etc., one should exclude IM_3, IM_5, IM_7, etc., from consideration. All communication standards with flat-power spectra are narrowband, and the difference in IMD at a fixed offset is quite small through the range of frequency spacing. In this case, one can consider the mean values and rewrite Eq. 3 as:

$$IMD = \int_{-\infty}^{\infty}\int_{\Delta f_1}^{\Delta f_2}\int_{\Delta f_1}^{\Delta f_2} IMDg(IMD)\left(\frac{1}{\Delta f}\right)^2 df_1 df_2 dIMD = \left(\frac{\Delta f_2 - \Delta f_1}{\Delta f}\right)^2 \int_{-\infty}^{\infty} IMDg(IMD)dIMD \qquad (5)$$

The function $F(\Delta f) = [(\Delta f_2 - \Delta f_1)/\Delta f]^2$ plays the role of a weighting coefficient for the IMD average value at different offsets. Figure 2 shows its shape in decibel units for several orders of IMD. By such consideration, the spectral components at different offset f_0 can be represented in the simple additive form, including the noise N_0 at the output of a PA:

$$IMD_\Sigma = IM_3 + IM_5 + ... + N_0 \qquad (6)$$

Considering Fig. 2 and taking into account Eq. 6, one can see that the contribution of higher-order IMD to the total distortion value increases with increasing offset f_0. Usually, most digitally modulated, wireless-communications systems have an approximately logarithmic, normal power distribution.[5] Assuming this distribution for a PA-driving signal, one may write: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$$IMD = F(\Delta f)\int_0^\infty IMD[mW]g(IMD[mW])dIMD[mW]$$
$$= 4.342945\ F(\Delta f)\int_0^\infty \frac{IMD[mW]}{P_{in}[mW]}g(P_{in}[dBm])dP_{in}[mW]$$
$$= 4.342945F(\Delta f)\int_0^\infty T_a dP_{in}[mW] \qquad (7)$$

The function $T_a = g(P_{in})IMD/P_{in}$ is the auxiliary transfer function for IMD.[4] The physical meaning of T_a is that the integral of this function is proportional to the IMD at a particular frequency offset f_0. The function T_a is convenient for the graphical consideration of Eq. 7. It clearly shows the contribution of the input-signal statistics and the IMD characteristics of a PA at each instantaneous drive level to the spectral regrowth at f_0. The square restricted by the curve T_a and the P_{in}-axis is equal to the IMD value at f_0, with an accuracy of a coefficient. Considering curves T_a versus P_{in}, one can tune a circuit to decrease IMD at input-power levels that contribute significantly to the output-power spectrum distortions.

The average output power of a PA is defined as Ref. 4:

$$P_{out} = \int_0^\infty P_{out}[mW]g(P_{out}[mW])dP_{out}[mW]$$
$$= 4.342945\int_0^\infty \frac{P_{out}[mW]}{P_{in}[mW]}g(P_{in}[dBm])dP_{in}[mW]$$
$$= 4.342945\int_0^\infty G(P_{in}[mW])g(P_{in}[dBm])dP_{in}[mW]$$
$$= 4.342945\int_0^\infty G_a(P_{in})dP_{in}[mW] \qquad (8)$$

G is the two-tone gain and $G_a = G(P_{in})g(P_{in})$ is the auxiliary gain function.[4] Its physical meaning is defined as the integral of this function is proportional to the average output power of a PA.

Finally, the power-spectral components at different offset f_0 are calculated as:

$$P(f_0) = 10\log(IMD_\Sigma / P_{out})\ [dB] \qquad (9)$$

IMD_Σ is defined by Eq. 6, and P_{out} is defined by Eq. 8. By the accepted assumptions, the phase-characteristic considerations have been eliminated because they are implied by the IMD measurements. No filtering characteristics were considered at the PA output, and Eq. 9 represents the spectral regrowth inherent only for a PA. The approach is considered valid for the AM/PM conversion inherent only for the nonlinear part of a PA. With a properly designed PA, the contribution of AM/PM conversion to the spectral regrowth is negligible when compared to an AM/AM conversion, even for high values of phase deviations.[3]

Frequency offsets for some widespread communication standards with a flat spectrum shape and weighting functions $F(\Delta f)$ are presented in Figs. 1 and 2.

For code-division multiple-access one (cdma One) at a frequency offset of 1.25 MHz, third- and fifth-order IMD have approximately equal contribution to ACPR. But for IM_3, one must consider 0.6356- to 0.9322-MHz spacing. For IM_5, it should be 0.3178 to 0.6215 MHz. For a further increase of the IMD order, its contribution decreases approximately 1.5 dB each time, according to Fig. 2. Considering ACPR at 1.98-MHz offset, one must take into account only IM and higher-order IMD. At 750- and 885-kHz offsets, the IMD contribution is decreased with an increase of an order of IMD. To consider ACPR for an entire adjacent or alternate channel, one must integrate Eq. 7 at frequency limits determined by this channel. That is why one can see a correlation between IM and ACPR at an adjacent channel, and between IM_5 and ACPR at first alternate channel, for a single-user signal.[8, 9] But, according to Fig. 2 and Eq. 7, this correlation is less perceptible if one drives an amplifier with a multicarrier signal at fairly high output-power levels due to the higher dispersion of signal statistics.

For the single-carrier, 3.6864-Mchip/s, CDMA-2000 standard at an offset of 5 MHz, the situation differs from cdma One. In this case, the contribution of IM to ACPR is 8.5 dB less than for IM. Using that as a model, IM would be measured from 3.1568 to 3.4216 MHz and IM from 1.5784 to 2.2811 MHz.

For the 4.096 Mchip/s, wideband CDMA (WCDMA) system at 5.23-MHz offset, the situation is similar to CDMA-2000. IM_3's contribution to ACPR is 5 dB less than that for IM_5. IM_3 is measured from 3.182 to 3.639 MHz and IM_5 at 3.452 to 3.774 MHz. For this standard, the approach proposed has been verified experimentally in Ref. 4 with good agreement, even for hard-driven, nonlinear PAs. A good correlation between IM_3 and ACPR at an adjacent channel for WCDMA PAs is observed in Refs. 9 and 10. However, the correlation between IM_5 and ACPR at an alternate channel is not as favorable. At the alternate channel, the contribution of higher-order IMD products to the spectral regrowth can become evident at high power levels (Fig. 2). Furthermore, one should not forget to choose different frequency spacing by IMD measurements (Fig. 1). This is particularly important for transistor PAs operating at modes close to class B, and for laterally-diffused-metal-oxide-silicon (LDMOS) amplifiers that have "sweet spots" due to cancellation effects imposed by the high-order nonlinearity in a circuit.[7, 8, 10, 11, and 12] Circuit tuning for high linearity requires different load conditions for in-band, out-of-band (harmonics and sub-harmonics), and baseband.[12, 13] Moreover, one must take into account the noise parameters of a circuit, considering ACPR (see Eq. 6). Usually, the wider the frequency bandwidth of a signal, the higher the relative noise floor of a system. This can significantly alter the ACPR's value and shape through the power sweep.[4]

(concluded on p. 155)

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Portable by Design

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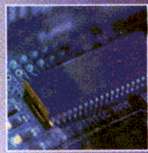

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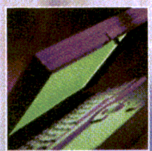

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Conference Director

Gauging The Effects Of Nuclear Radiation On Microstrip Antennas In Space

Nuclear radiation can degrade the performance of the dielectric materials in space-based antennas.

A. Kumar
President
AK Electromagnetique, Inc., P.O. Box 240, 30 Rue Lippee, Les Coteaux, Quebec J7X 1H5, Canada; (514) 620-3717, FAX: (450) 267-1144, e-mail: kumar@colba.net.

RESISTANCE to radiation is critical for the long-term use of dielectric materials in space-based applications. In these applications, devices and antennas can be exposed to substantial doses of nuclear radiation. Not only must dielectric materials used in antennas for space-based satellite-communications systems endure the physical stress of the launch, they must also withstand prolonged exposure to intense radiation without drastic changes in mechanical and electrical properties.

In space applications, microstrip antennas are subject to radiation in the Van Allen Belt,[1] which consists of electrons with substantial densities at energies up to 100 meV. Bombardment on microstrip antennas etched on dielectric materials, such as polytetrafluoroethylene (PTFE), with particles of this high energy results in a continual degradation of mechanical and electrical properties.

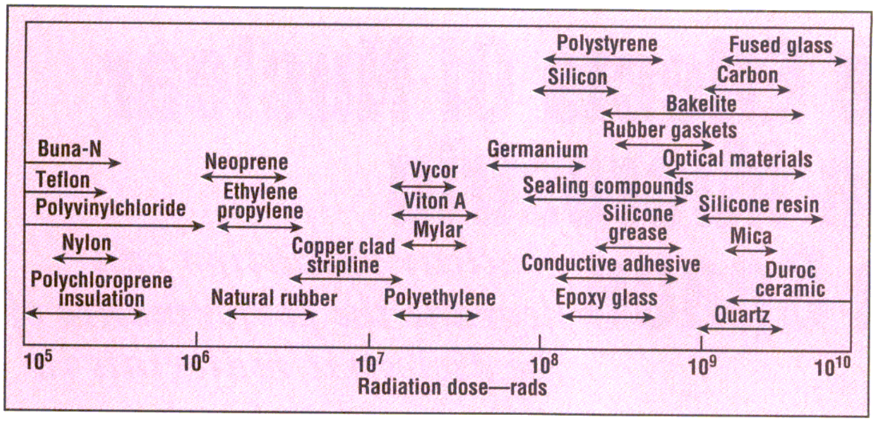
The effects of nuclear radiation on the electrical and mechanical properties of many dielectric materials were investigated by Pascale, Herrmann, and Miner.[2] These researchers used a turntable device as the mounting rack for the test specimens.

In a dielectric material, most electrons are restricted to the valance band, with only a negligibly small number being free to move. The electronic excitiation processes are responsible for increasing the number of conducting electrons in the dielectric substrate. Two main phenomena occur in these materials when irradiated—transient effects, which disappear upon removal from the radiation field, and permanent effects, which remain after the radiation field is removed.

In most cases, failure due to structural changes occur before any significant permanent change in dielectric properties has taken place. These structural changes generally cause the fracture and cracking of material, a particularly serious problem if the irradiated component is under stress, and continued irradiation always leads to progressive deterioration. Such properties as tensile strength, elasticity, viscosity, transparency, chemical reactivity, electric-field strength, and resistivity become permanently changed. The primary effect of nuclear radiation on materials based on PTFE is the reduction of molecular weight by breaking the large polymer molecule into smaller parts. There is an increase in brittleness and this reaction is maximized in the presence of air. It has been reported[2,3] that changes in the mechanical properties of PTFE depend on the total radiation dose and to be independent of dose rate. The dielectric properties are affected by an electrical-charge distribution in the resin which decays with time and, therefore, the dose rate is an impor-

Threshold radiation doses of PTFE in air and vacuum

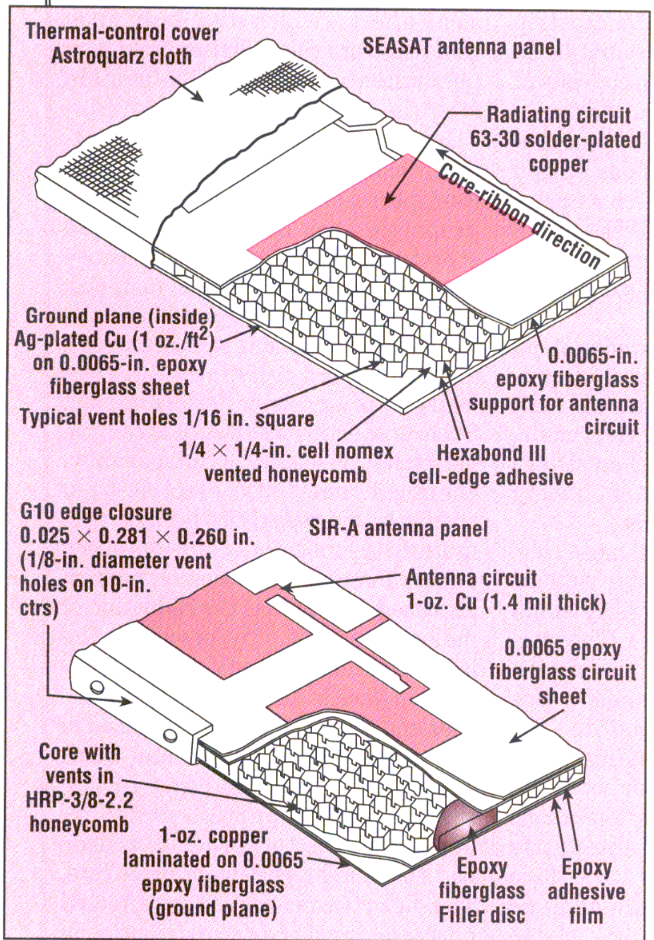
Parameter	In air	In vacuum
Threshold	$2\text{-}7 \times 10^4$	$2\text{-}7 \times 10^4$
50-percent tensile strength	10^6	10^7 or more
40-percent tensile strength	10^7 or more	8×10^8 or more
Retain 100-percent elongation	$2\text{-}5 \times 10^5$	$2\text{-}5 \times 10^6$



1. Dielectric susceptibility varies with radiation dose.

tant parameter.

During irradiation, the real and imaginary parts of the permitivity of the material are temporarily increased. The effect of space radiation on these parameters is decreased at elevated frequencies which would be encountered in microwave applications, such as satellite-communications systems. The degree to which PTFE is affected is essentially a function of the amount of energy absorbed, regardless of the identity of the radiation (i.e., beta, gamma, and X-ray energies all have the equivalent effect). It has been reported[4-6] that during solar flares (solar storms), the radiation intensity in space increases many thousands of times when compared to periods of non-solar-flare activity. The radiation dose during solar storms can damage materials which are susceptible to ordinary levels of space radiation. The table offers a summary of the radiation doses (in rads) related to the damage levels of PTFE materials. Figure 1 shows the susceptibility levels of dielectric materials as a function of total nuclear-radiation dose in space. Usually, a radiation-dose rate of 10 rads/h is quoted for the Van Allen Belt.[3] At this rate, PTFE could operate for approximately five years before a threshold level of damage would be electronically and mechanically detectable. However, during solar storms, the operational age of PTFE could depend upon the intensity of nuclear radiation in space. This was discussed during the design phase of the ERS1-IDHT antenna.[7] The European Space Agency did not use microstrip feeder and dielectric support for the ERS1-IDHT antenna and a new IDHT antenna was designed without dielectric material.[8,9]



2. These sketches show the SEASAT antenna panel (a) and the SIR-A antenna panel (b).

The SEASAT and SIR-A microstrip antennas are among the largest arrays that have been flown in space. The deployed radiating surface of the SEASAT antenna was 10.74 × 2.16 m and the measured antenna gain was 34.9 dB at 1.275 GHz. The SIR-A microstrip-array radiating antenna panel's surface was 9.4 × 1.25 × 0.006 m, with 128 microstrip elements per panel. The SEASAT antenna consisted of eight panels supported on an extendable graphite epoxy structure. The SIR-A antenna consisted of seven panels supported by an Al-tube-truss structure.[10] The substrate materials for these microstrip arrays were chosen on the basis of the lightest weight, the lowest-loss microstrip dielectric, the highest efficiency, and good repeatability in electrical and mechanical performance. The honeycomb structure shown in Fig. 2 was selected because it provided the required structural strength and, while the loss tangent of the epoxy fiber-glass sheets and honeycomb dielectric material is relatively high, the substrate loss is small due to the large percentage of voids in the honeycomb. The relative permittivity of the SIR-A microstrip-antenna substrate was measured as 1.18. The nuclear-radiation dose of the material in space was more than 10^8 rad, which is 1000 times more than PTFE-type substrate. ••

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Wireless Event Readies For Chicago

The autumn version of the Wireless Symposium/Portable By Design Conference & Exhibition promises a full lineup of educational workshops and technical sessions.

JACK BROWNE
Publisher/Editor

WIRELESS-EQUIPMENT designers may want to make plans to be in Chicago this fall. The Windy City is this year's site for the autumn Wireless Symposium/Portable By Design Conference & Exhibition, scheduled for September 26-29 at McCormick Place. With a variety of workshops and a full slate of technical sessions, Wireless/Portable forms a key educational event for engineers and engineering managers seeking to know more about key wireless technologies, from Bluetooth to wireless local-area networks (WLANs). Co-located with Wireless/Portable is the huge PCIA GlobalXChange event (see sidebar), which draws more than 20,000 attendees to its more than 600 exhibitor booths and market-oriented education sessions.

Workshops at a glance

Tuesday September 26	Wednesday September 27	Thursday September 28	Friday September 29
9:00 AM — 4:30 PM	9:00 AM — 4:30 PM	9:00 AM — 4:30 PM	9:00 AM — 4:30 PM
DSP Made Simple (I) Rick Lyons Besser Associates	**DSP Made Simple (II)** Rick Lyons Besser Associates	**Wireless Made Simple (I)** Al Scott Besser Associates	**Wireless Made Simple (II)** Al Scott Besser Associates
RF Fundamentals (I) Les Besser Besser Associates	**RF Fundamentals (II)** Les Besser Besser Associates	**Oscillator Design** Randy Rhea Eagleware Corp.	**Antennas and Propagation for Wireless Communications** Dr. Steven Best Cushcraft Corp.
CDMA Technology for 2G/3G Wireless Systems Darryl Schick Linear Lightwave	**3G WCDMA and cdma2000 Technology** Darryl Schick Linear Lightwave	**RF Wireless System Fundamentals** Rick Fornes Besser Associates	
Short Range/LAN & Bluetooth Rob Morrow Besser Associates			
3G Made Simple Al Scott Besser Associates			

The Wireless/Portable technical program was assembled under the guidance of the Conference Advisory Committee, a learned group that includes Don Brown, director of the IWPC (Warrington, PA); Tim Carey, director of marketing for Anritsu Co. (Morgan Hill, CA); Dr. Samuel Horowitz, marketing manager for DuPont Microcircuit Materials (Research Triangle Park, NC); Robert Morrow, an instructor for Besser Associates (Mountain View, CA); Dave Osika, chief scientist for Strategic Technologies with ANADIGICS (Warren, NJ); and Darryl Schick, president of Linear Lightwave, Inc. (Lafayette Hill, PA).

The technical program organized under the conference direction of the well-known educational organization Besser Associates. The expertise of the Besser instructors is evident from the breadth of workshops offered during the Wireless/Portable event (see table), including several of the "Made Simple" tutorial courses that help bring engineers and their managers up to speed on different technology areas, such as third-generation (3G) wireless systems and digital-signal processing

(DSP). For example, the DSP Made Simple workshop taught by Rick Lyons of Besser Associates provides an accessible introduction to DSPs. Striving to minimize the amount of mathematics contained in the course, Lyons nonetheless unveils the fundamental equations of DSP and how they are applied in modern dedicated microprocessors. The workshop highlights some of the key algorithms of DSP technology, including discrete Fourier transforms, finite-impulse-response (FIR) digital filters, and infinite-impulse-response (IIR) digital filters.

Additional workshops include several strong sessions from Darryl Schick, former chief engineer of RDL (Conshohocken, PA) and now president of Linear Lightwave. One of these workshops, a full-day seminar on "CDMA Technology for 2G/3G Wireless Systems," provides a thorough introduction to code-division-multiple-access (CDMA) technology, and how CDMA is implemented in the TIA/EIA-95 standards. The course covers Walsh codes, Gold codes, error protection and correction, synchronization, the CDMA RAKE receiver, and how to predict CDMA traffic capacity. Schick's other workshop, "3G WCDMA and cdma2000 Technology," is a one-day examination of wideband CDMA (WCDMA) and discusses how 3G-CDMA is implemented in the TIA cdma2000 and ETSI/ARIB WCDMA standards. The second workshop is an excellent followup course for those who are taking the introduc-

A BRIEF LOOK AT PCIA GLOBALXCHANGE

CHICAGO is this year's site for one of the wireless industry's largest events—the PCIA GlobalXChange. Formerly the PCS Show (such as PCS '99), the event is co-located in the McCormick Place with the Wireless Symposium/Portable By Design Conference and Exhibition and is scheduled for September 26-29, 2000.

The exhibition portion of the PCIA GlobalXChange is vast, and includes an impressive list of equipment and service providers. Key Internet companies, such as Amazon.com and America Online, Inc. will be in attendance on the exhibit floor. Additional exhibitors include Agilent Technologies; Analog Devices, Inc.; Andrew Corp.; Anritsu Co.; Ansoft Corp.; Chomerics; Ericsson Mobile Communications AB; Hewlett-Packard Co.; RangeStar Wireless, Inc.; Richardson Electronics; and Sun Microsystems.

The conference portion of the PCIA GlobalXChange features eight different tracks of market-oriented panel sessions and presentations, a special symposium, and a one-day workshop. The workshop, "Winning in Wireless; Chasm Strategies for Market Success," is scheduled for Tuesday, September 26th. The workshop will explore ways to ensure revenue growth and profits in wireless markets through the year 2001 and beyond.

The special symposium, "WAP Technology and Development Symposium," is scheduled for September 25-26, 2000 at the Hyatt Regency Chicago. An official event of the WAP Forum, the two-day course provides a detailed look at this important technology. Targeted at engineers, developers, and wireless professionals, the symposium explains how to use the wireless-application-protocol (WAP) development environment and the client/server programming model to write network- and device-independent applications and services using WML and WML Script languages. The symposium also covers WAP programming, protocol layers, and interoperability issues.

The main conference's eight tracks include coverage of the wireless web (Internet), wireless voice, interactive messaging and wireless data, 3G, emerging technologies, fixed broadband services, and international-market perspectives. For example, the track on emerging technologies includes a Tuesday session on in-vehicle wireless applications, including access to the Internet from the automobile.

The track on wireless Internet should draw large crowds. The wireless Internet offers a new frontier for the industry, with new revenue streams, new competition, and new customers. What are the technological content and service developments that will drive the growth of this portion of the industry? This track will explore the alliances, strategies, and business models that carriers should be examining today to ensure their future on the new wireless Internet frontier.

The track on content for the wireless Web covers what is currently available in terms of content for WAP devices, and what will be done to improve the types and amount of content that will be available to wireless Internet subscribers. A Tuesday presentation from this track, for example, highlights the need for partnerships, no matter how small, in order to build the specialized Internet content.

The track on wireless voice notes that, in spite of the number of cellular subscribers passing 100 million in the United States and more than 500 million globally, many opportunities still exist in the wireless voice arena. These opportunities rely on targeting new markets and developing new business models. One session within the track looks at the current state of the wireless voice industry, and provides key trends and statistics and identifies the drivers of the future.

For those who still have time, there are the many workshops and technical sessions on wireless technology to be found in the Wireless Symposium/Portable By Design Conference and Exhibition co-located with the PCIA GlobalXChange. For more information on the PCIA GlobalXChange, contact its sponsor at: **PCIA, 500 Montgomery St. Suite 700, Alexandria, VA 22314; (800) 759-0300, (703) 739-0300, FAX: (703) 836-1608, Internet: http://www.pciaglobalxchange.com.**

tory workshop on CDMA.

Other workshops include the popular Oscillator Design workshop by Randy Rhea, founder of software innovator Eagleware Corp. (Stone Mountain, GA), the popular two-day course, "Wireless Made Simple," taught by Al Scott of Besser Associates, a one-day course on short-range wireless technologies, such as Bluetooth and WLANs, taught by Rob Morrow of Besser Associates, and the popular workshop "Antennas and Propagation for Wireless Communications," taught by Dr. Steven Best, president of Cushcraft Corp. (Manchester, NH).

TECHNICAL SESSIONS

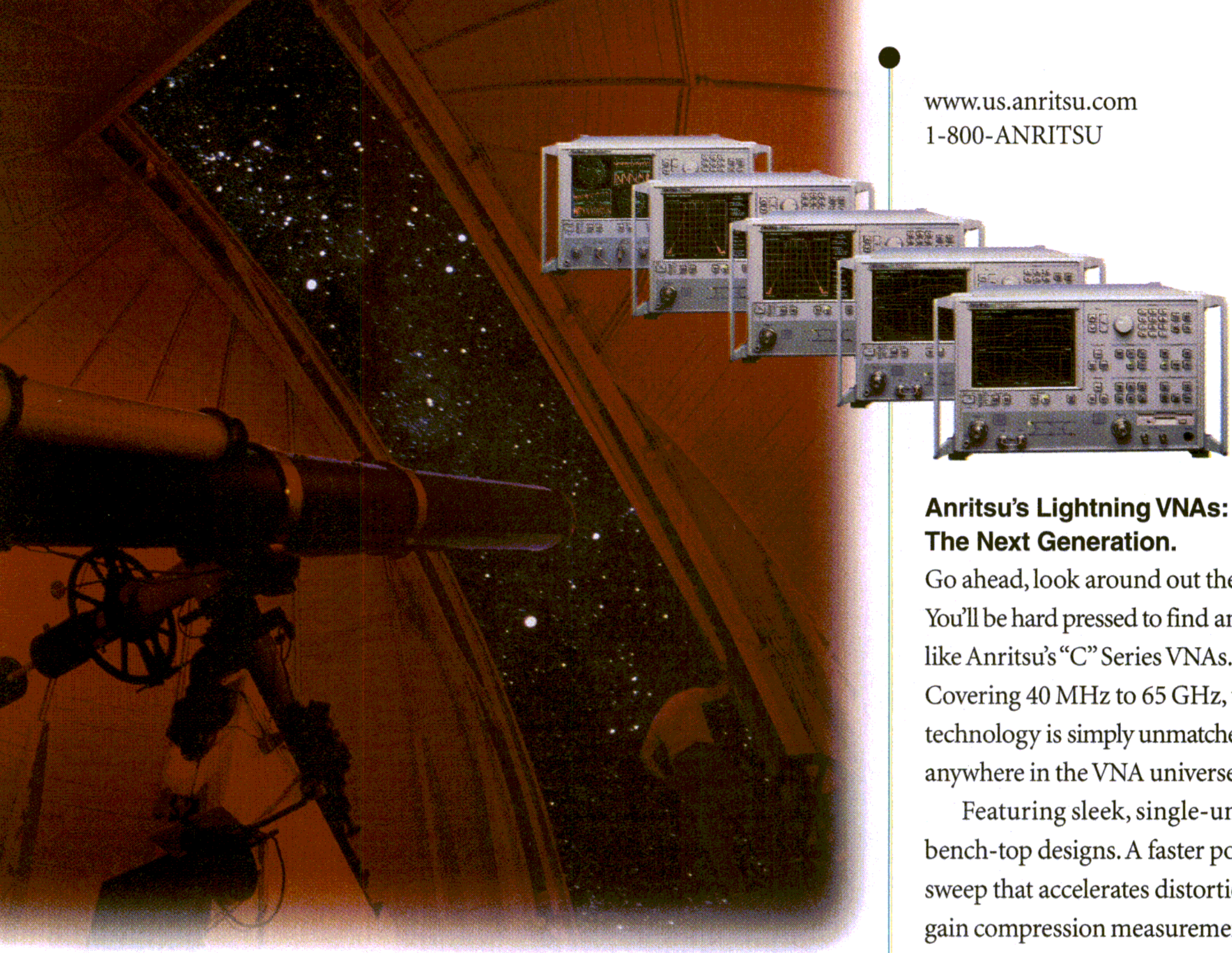
A wide range of technical presentations and mini-tutorial sessions are also available at Chicago's Wireless/Portable event. As with the PCIA's companion event, presentations are grouped by technology tracks, such as wireless packaging, short-range (Bluetooth) technologies, 3G technologies, embedded devices for wireless Internet applications, and high-power design for base stations. The Bluetooth track, chaired by Tim Carey, director of marketing for Anritsu Co. compares the technologies of Bluetooth and WLAN systems and where each can be effectively used, and how integrated-circuit (IC) designers are responding to the need for lower-cost Bluetooth devices operating at 2.4 GHz.

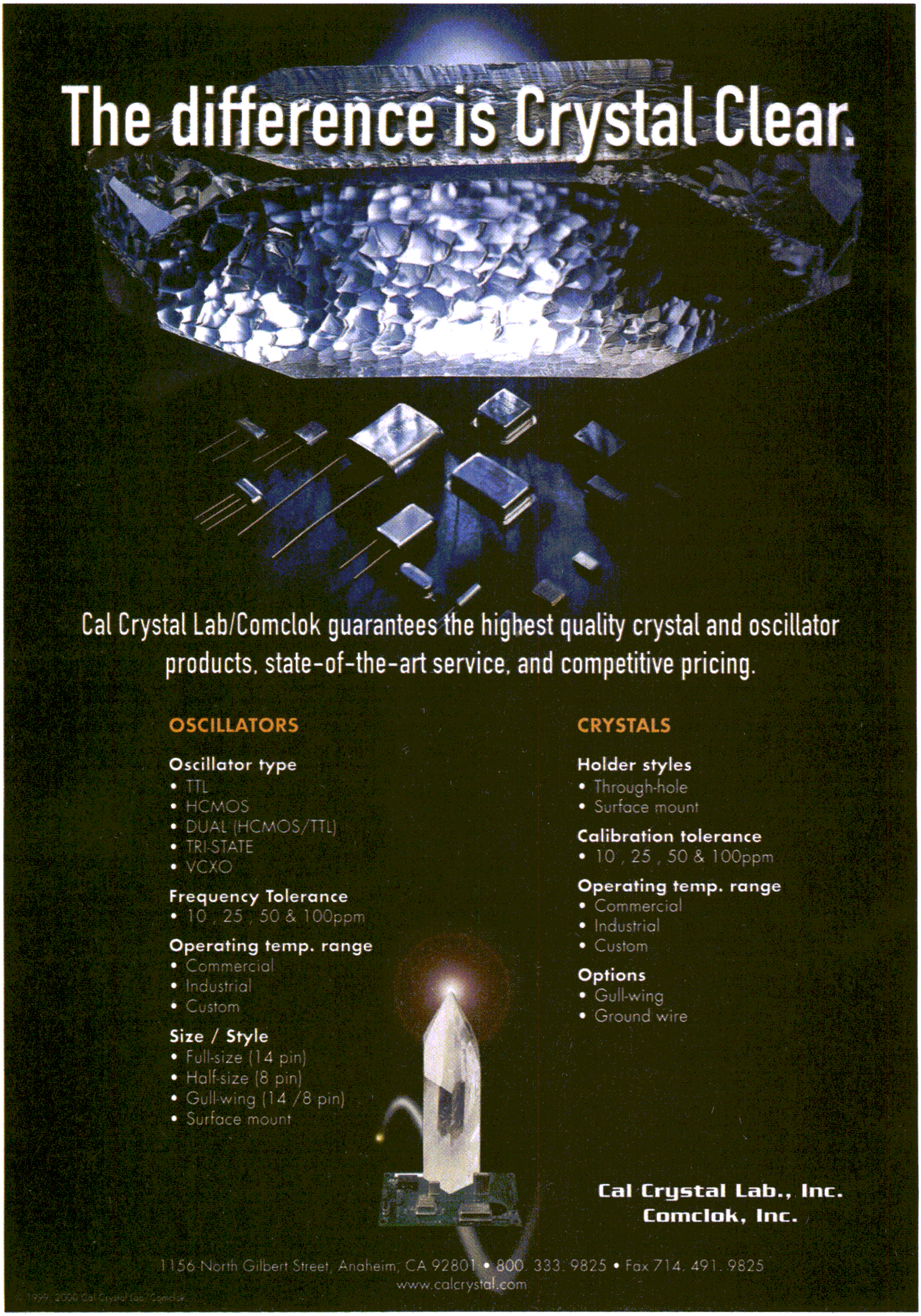
HIGH-POWER AMPS

Several sessions on high-power amplifier designs, chaired by Dave Osika of ANADIGICS, explore the latest technologies and techniques to help amplifier designers reach their target specifications for output power and intermodulation distortion (IMD) for base-station amplifiers. In contrast, a series of sessions organized by Rob Morrow of Besser Associates focuses on new technologies for low-bit-rate, low-power applications, and how to design devices for low-power use, in low-data-rate RF identification (RFID) as well as remote-keyless-entry (RKE) applications.

For those who remain throughout the length of the conference, their reward is a powerful mini-tutorial session on Friday by Barry Herbert of Nortel Networks (Dallas, TX) on "Enabling the Wireless Internet." Herbert notes that this may prove to be one of the largest of all wireless applications, and it one that will certainly change the way that people live and revolutionize the way that they do business. The session will focus on the challenges facing wireless Internet providers in the years to come.

The Wireless Symposium/Portable By Design Conference & Exhibition is managed by Penton Media. For more information on the upcoming Wireless/Portable event, visit the website at http://www.WirelessPortable.com. ••





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(continued from p. 140)

For the personal-digital-cellular (PDC) standard, the specification requires ACPR measurements at the first alternate channel. One can see in Fig. 2 that the IM_5 contribution to ACPR exceeds the IM_3 contribution for nearly the entire frequency bandwidth considered. This note is especially important at high output levels. That is why one can observe the correlation between IM_5 and ACPR for PDC systems.[6]

The situation for the personal-handyphone-system (PHS) standard is similar to that for PDC, but with a higher correlation between IM_5 and ACPR. Moreover, one should observe some noticeable correlation between IM_7 and ACPR as well.

For cellular standards where ACPR is defined as a spurious emission and channels are spaced very tightly, the low-frequency spacing may play a considerable role in ACPR at low-frequency offsets (Fig. 1). Due to this, IMD products can rise significantly.[14]

The proposed approach allows one to consider IMD's contribution to ACPR for communication standards with a flat spectrum. If there is any unpredictable spectral regrowth at some frequency offset, one must check IMD at the frequency spacing determined in Fig. 1 for the IMD order specified in Fig. 2. After that, one can find out what part of a circuit creates this distortion and eliminate it by adding a tuning circuit. In addition, it is senseless to look for a strong correlation between fixed values of IMD and ACPR, even at the correctly chosen output-power level. One must consider the IMD behavior through the entire power range where the signal exists due to the appropriate statistics. This is especially important at high power levels, where the cancellation effect may be perceptible[7] and different-order IMD products have a variable shape. The signal statistics play a decisive role for ACPR. For example, failing to take this into account for the hybrid phase-shift-keying (PSK) signal in the WCDMA system used in Ref. 15, it is easy to receive false correlation data, especially in a "sweet spots" environment.

The two-tone, IMD-measurement approach discussed here predicts and analyzes ACPR in RF/ microwave PAs used in modern digitally modulated communication systems with flat-power spectra. This method works best in multicarrier environments and for higher quantities of signal-constellation states. But even for a single-carrier signal with some correlation between the spectral components, the method can help engineers analyze and tune circuits for the best linearity. ••

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Pondering the effects of 1/f noise

Transistor 1/f noise has a major influence on the ultimate phase noise of a voltage-controlled oscillator (VCO). To better understand how 1/f noise changes with bias conditions and from device to device, a study was performed by California Eastern Laboratories (Santa Clara, CA) on the 1/f noise of two different bipolar transistors, and nonlinear models were developed and compared. The study is contained in a six-page application note, "1/f Noise Characteristics Influencing Phase Noise."

In a communications system, the phase noise of an oscillator determines the system's ability to separate adjacent channels. Given the increase in wireless signal traffic, system designers are seeking oscillators with improved phase noise. And selecting the proper active device for a low-noise oscillator is very critical. Phase noise, which is generally measured in a single 1-Hz sideband at some specified offset frequency from the carrier, is essentially random frequency variations in an oscillator. These random variations are caused by thermal noise, shot noise, and flicker noise. Thermal noise is a function of the temperature, operating bandwidth, and resistance. Shot noise is a function of the DC bias current to the active device. And flicker noise is a function of the active device's characteristics. Oscillator phase noise has two components: phase noise resulting from direct upconversion of white noise and 1/f noise, and phase noise resulting from the changing phase of the noise sources (the white noise and 1/f noise) modulating the oscillation frequency.

The application note describes a test setup to measure device 1/f noise. In order to eliminate noise, the supply voltages are provided by batteries, and the entire test setup is contained in an electromagnetic-interference (EMI) shielded enclosure. Measurements using this setup were performed on the company's NE68819 and NE68519 bipolar transistors. From these measurements, it was possible to calculate the intrinsic base-flicker noise-corner frequency.

The note goes on to examine the bias dependency of the flicker-corner frequency for the two bipolar devices, and develop a 1/f noise simulation for each transistor. Application note AN1026, "1/f Noise Characteristics Influencing Phase Noise," is available upon request from the company, or can be downloaded free of charge from their website. Contact: **California Eastern Laboratories, Inc., 4590 Patrick Henry Dr., Santa Clara, CA 95054-1817; (408) 988-3500, FAX: (408) 988-0279, Internet: http://www.cel.com.**

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Evaluating a PCS phase shifter

Phase control is critical to achieving low distortion in modern cellular and personal-communications-services (PCS) base stations. Compensation techniques generally rely on voltage-controlled or digitally controlled phase shifters to make phase adjustments across the operating frequency range. Application note APN1009 from Alpha Industries, Inc. (Woburn, MA), "A Varactor Controlled Phase Shifter for PCS Base Station Applications," describes the design of a high-performance phase shifter that is well suited for PCS band base-station applications. The phase shifter is based on one of the company's varactor diodes and a model HY19-12 90-deg. hybrid.

The application note provides a schematic diagram for a typical phase shifter using a quadrature coupler or 90-deg. hybrid circuit. Varactor diodes are incorporated in the circuit as reactive loads which can shift between open-circuit conditions (a reflection coefficient of 0 deg. for zero varactor capacitance) and short-circuit conditions (a reflection coefficient of 180 deg. for infinite varactor capacitance).

For users of computer-aided-engineering (CAE) tools, the application note provides a phase-shifter circuit model of the 90-deg. hybrid for use with the Libra circuit simulator, as well as a SPICE model of the company's model SMV1245-011 varactor diode. In addition, a full circuit diagram is presented for the PCS-band phase-shifter design, along with a detailed listing of required circuit elements and circuit-board materials.

Copies of the four-page application note, "A Varactor Controlled Phase Shifter for PCS Base Station Applications," can be downloaded free of charge from the company's website. Contact: **Alpha Industries, Inc., 20 Sylvan Rd., Woburn, MA 01801; (781) 935-5150, FAX: (617) 824-4579, e-mail: sales@alphaind.com, Internet: http://www.alphaind.com.**

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Thermal Conductivity (% of Cu)	.05	.01	20	0.1	50
Large Format Processing	Yes (Shrinkage)	Yes	No	Yes (Shrinkage)	Yes
TCE Matched To GaAs/Si	Poor/Fair	Poor	Good	Fair	Good
Applicability To High Frequency Applications	Good	Good To 2 GHz	Yes (Materials Dependent)	Good/ Excellent	Excellent

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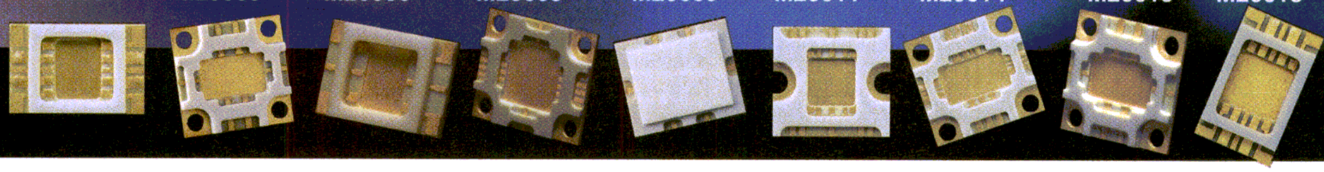

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Digital Upconverter IC Tames Complex Modulation

An improved 14-b architecture, simplified synchronization, and enhanced power-saving circuitry are a few of the features of this quadrature digital upconverter.

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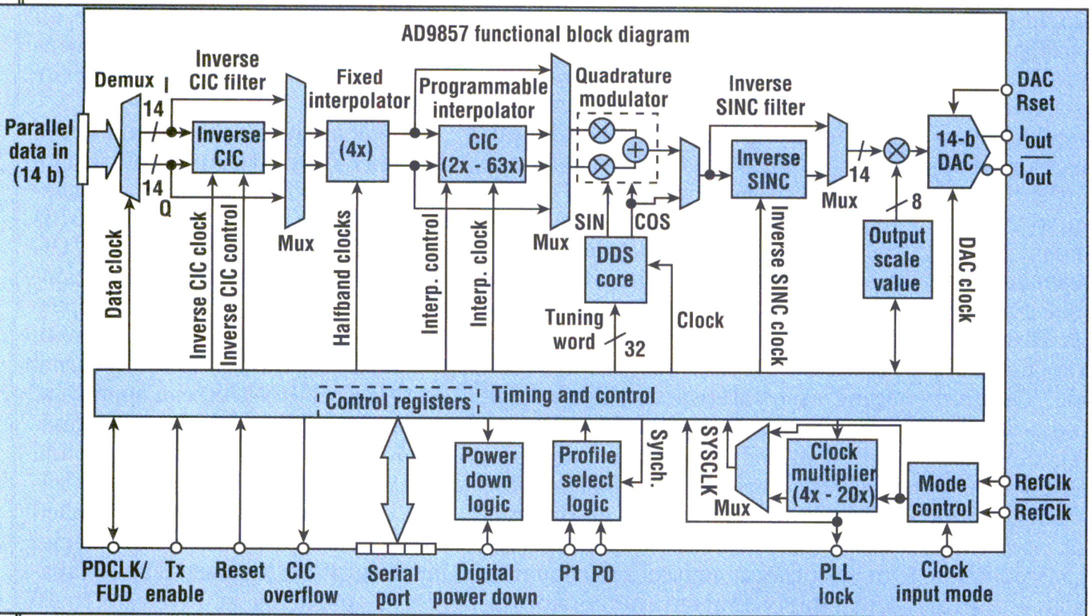
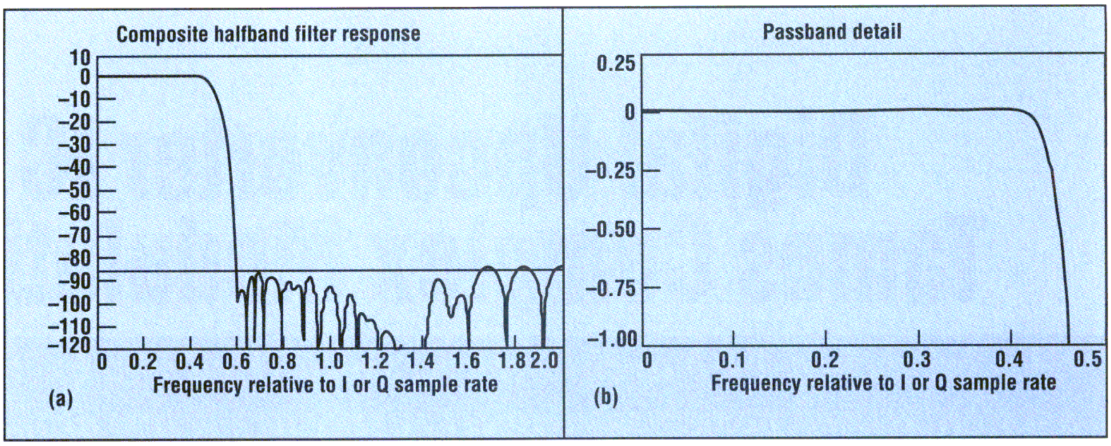
DIGITAL technology continues to replace many analog functions in modern receiver architectures. The latest contributor to this trend is the AD9857 quadrature digital upconverter from Analog Devices, Inc. (Norwood, MA), which can replace several front-end components, including a mixer and local oscillator (LO). The 14-b integrated circuit (IC) provides improved dynamic range and sensitivity when compared to its 12-b predecessor, with enhanced power-conservation circuitry.

As the demand for higher data rates continues to grow, system designers are finding it more attractive to design carrier transmission systems instead of baseband transmission systems. There are several reasons for this. First, many data-transmission systems are broadcast over the "air," such as mobile phones, satellite communications, and digital radio and television. The transmission of signals through free space is not practical at low frequencies. The main reason is due to the physical size of the required antenna, which is inversely related to the transmitted frequency. Thus, placing baseband information on a high-frequency carrier makes the use of a



1. The AD9857 quadrature digital upconverter employs a 14-b architecture for improved sensitivity and dynamic range compared to its 12-b predecessor, the AD9856.



2. Note that the stopband attenuation is 85 dB (a). In this plot, passband ripple is virtually nonexistent (b).

smaller antenna possible. Secondly, a modulated carrier is capable of carrying a complex spectrum, which is not possible with a baseband signal. A complex spectrum results when data are encoded into real and imaginary parts and mixed onto the carrier orthogonally. This offers an automatic factor-of-two reduction in the bandwidth required to carry the data (relative to a signal confined to baseband). Thirdly, a modulated carrier can be easily shifted in frequency to offer multiple channels of operation.

At the heart of any data-transmission system resides a bit stream, which is a sequence of ones and zeros that represents the information being transmitted. The bit stream can be converted to a sequence of pulses through a variety of methods, but the end result in any case is the generation of a baseband signal. A baseband signal has a spectrum that starts at 0 Hz (DC) and extends to some positive frequency. The frequency range thus spanned defines the bandwidth of the baseband signal. As mentioned earlier, it is often desirable to shift the baseband signal to a higher frequency to take advantage of the benefits of carrier transmission. The process of shifting a baseband signal to a higher frequency is known as upconversion.

In the past, upconversion has been accomplished through analog mixers, oscillators, and filters. However, upconversion is now possible through digital methods. The first digital upconverter from Analog Devices, the model AD9856 (see *Microwaves & RF*, February 1999, p. 125), provided 12-b resolution and the capability to work at clock rates up to 200 MHz. With the introduction of the AD9857 quadrature digital upconverter, enhanced performance is possible by virtue of the AD9857's 14-b architecture. The additional bits of resolution yield increases in dynamic range and signal-to-noise ratio (SNR). Improvements have also been made to simplify the synchronization process between the device's baseband data port and the user's controlling hardware. Additionally, improvements to the signal-processing chain make the AD9857 a more robust and flexible device than its predecessor. The AD9857 also incorporates a sophisticated power-conservation architecture that automatically shuts down functional blocks that are not configured for use.

Similar to the AD9856, the AD9857 incorporates baseband signal-processing, quadrature mixing, and oscillator functions—all implemented in digital technology. There are a number of significant advantages to implementing these functions digitally. 1. A digital quadrature modulator supports precise matching of the in-phase (I) and quadrature (Q) channels, a challenging task when implemented with analog circuits. 2. The digital circuits used to implement the signal-processing functions do not suffer the effects of thermal drift and aging associated with their analog counterparts. 3. Digital circuitry provides the ability to control and program the device, which means more flexibility for the system designer. 4. The digital nature of the device yields precise control of frequency and nearly instantaneous tuning speed, greatly expanding the range of system capabilities available to the designer. 5. The implementation of digital functional blocks makes it possible to achieve a high degree of system integration. This implies a major reduction in the printed-circuit-board (PCB) area required, which is relative to an equivalent analog design.

Even though the AD9857 incorporates a largely digital architecture, it nonetheless delivers analog output signals by virtue of its integrated 14-b digital-to-analog converter (DAC). This saves the user the burden of having to provide an external DAC to make the conversion to analog signals. Despite the presence of high-speed digital circuitry, the AD9857 IC achieves analog output signals with excellent noise performance, all on a single silicon (Si) complementary-metal-oxide semiconductor (CMOS) chip.

A look at the block diagram of the AD9857 helps to appreciate the design's blending of digital and analog circuitry (Fig. 1). The AD9857 is designed to operate with a +3.3-VDC supply (with ± 5-percent regulation) over the extended industrial temperature range of –40 to +85°C. As with the AD9856, the AD9857's internal system clock (SysClk) can operate at rates to 200 MHz, which is the maximum sample rate of the onboard DAC (200 MSamples/s). The quadrature digital upconverter is controlled through an SPI-compatible serial input/output (I/O) port that is capable of operating at rates up to 10 MHz. The architecture is designed so that the device may be used in any

one of three operating modes: single-tone generator mode, interpolating DAC mode, and quadrature upconverter mode.

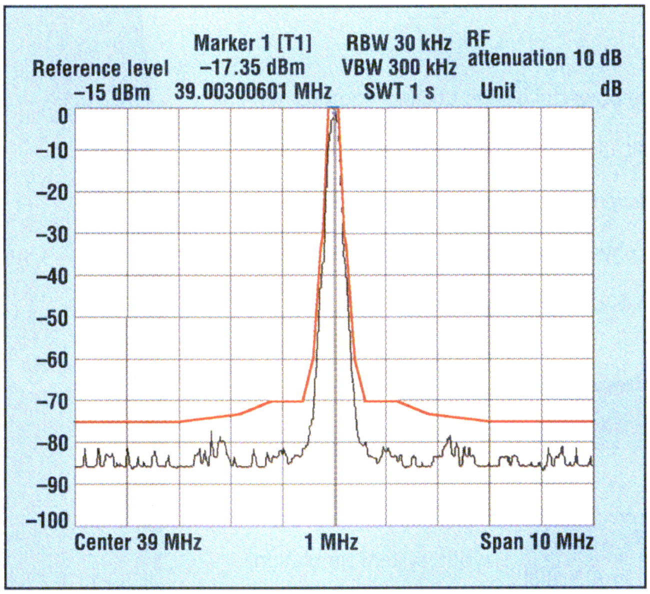
The focus here will be on the last mode of operation, which makes use of all of the AD9857's functional blocks. The other two operating modes are simply subsets of the quadrature upconverter mode. Briefly, in the interpolating DAC mode, the quadrature modulator is bypassed. In the single-tone mode, the quadrature modulator and the entire baseband signal-processing section are bypassed.

Many of the features of the AD9857 are directly controlled or programmed by a set of internal control registers, which are accessed by the user through the serial I/O port. The serial I/O port is limited to a maximum clock rate of 10 MHz, which sets an upper limit on the rate at which registers may be updated. The control registers have write and read access, which can be beneficial during the prototype phase of a design because it provides the designer with the ability to verify the contents of the control registers after they have been programmed.

SYNCHRONIZATION

The AD9857 derives its timing through a user-supplied external reference clock (RefClk) from which the internal SysClk signal is derived. The RefClk input will accept either a differential or single-ended clock source at frequencies up to 200 MHz. The SysClk signal is either a directly buffered version of RefClk or it is derived from RefClk through an internal phase-locked-loop (PLL)-based frequency multiplier (programmable in integer values from 4 to 20 through the control registers). The PLL is bypassed or enabled with a control bit in the control registers. This allows the user to employ a low-frequency crystal oscillator for the RefClk signal (e.g., 10 MHz), yet still obtain 200-MSamples/s performance by enabling the PLL at a multiplication factor of 20. Of course, any phase noise inherent in the RefClk signal will be amplified by the PLL multiplication factor. For this reason, very-low-noise clock sources are preferred when the PLL is employed.

The PLL also provides an external-lock indication signal (the PLL lock pin). When the PLL lock signal registers high, it indicates to the user that the PLL has acquired a lock condition. In addition, this signal is routed internally to provide certain internal housekeeping functions within the baseband signal-processing sections of the device. A control bit in one of the control registers will determine whether the PLL lock indicator triggers the internal housekeeping functions.



3. This plot shows GMSK modulation for a 39-MHz carrier with a sampling rate of 156 MSamples/s.

The heart of the AD9857 is its direct-digital-synthesizer (DDS) core. The DDS consists of a 32-b accumulator that operates at the SysClk rate of f_S. The frequency of the DDS is tunable from DC to $f_S/2$ by means of programming registers, which are accessed through the I/O port. With its 32-b tuning word, the DDS offers a tuning resolution of $f_S/2^{32}$, which is equivalent to 0.047 Hz when operating at the maximum clock rate of 200 MHz.

In the AD9857, the primary role of the DDS is the simultaneous generation of a digital sine and cosine waveform. Each waveform is represented as a series of 14-b numbers that are samples of the tuned frequency taken at the f_S rate. In the quadrature upconverter mode, the digital sine and cosine samples serve as a digital quadrature local oscillator (LO) operating at the carrier frequency, f_C. The carrier is simply the user-specified frequency to which the DDS is tuned.

The quadrature modulator (or digital mixer) multiplies the quadrature carrier samples from the DDS core by the baseband I and Q samples. A fundamental requirement must be established at this point, which is that the sample rate at the input to the digital mixer must be the same as the sample rate of the quadrature carrier signal. For the AD9857, the quadrature carrier signal is sampled at f_S, the frequency of SysClk. This means that the I and Q samples must also arrive at the digital mixer at the f_S sample rate. The quadrature modulator also handles the merging of the multiplication results by either adding or subtracting the two products. By default, the Q results are subtracted from the I results. However, the user can specify a spectral inversion, in which case the products are summed. This feature makes it possible to select either the upper or lower sideband when the device is used as a single-sideband (SSB) transmitter. The presence or absence of the spectral inversion function is handled through the control registers.

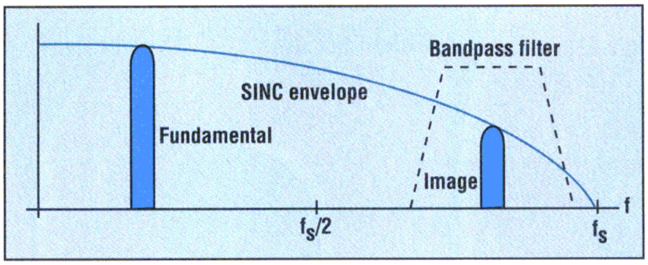
The output of the quadrature modulator may be optionally routed to an inverse SINC filter. When selected, this digital FIR filter compensates for the inherent sin(x)/x (SINC) rolloff characteristic introduced by the DAC (an unavoidable artifact of the sampling process) by imposing an x/sin(x) response on the samples before they are routed to the DAC. This compensation is effective over the frequency range of DC to $0.45f_S$.

The inverse SINC filter is enabled or disabled through the control registers.

Additionally, the digital samples may be scaled just prior to the DAC input. This is accomplished through a programmable scaler. The scaler multiplies the samples by an 8-b positive number that is programmed by the user through the control registers. The scaler is designed so that the maximum multiplier value (all digital ones) corresponds to a scale factor of 510/256 (1.9921875). The digital scaler makes it possible to generate amplitude-modulated (AM) signals or to ramp the analog output signal level up or down at will.

The AD9857's most enabling attribute is its integrated 14-b DAC. With 14-b resolution, there is sufficient dynamic range to satisfy multicarrier and spread-spectrum applications. Additionally, there is the benefit of a reduced noise floor over devices with fewer bits of resolution. The DAC can handle sample rates up to 200 MSamples/s, while still offering a spurious-free dynamic range (SFDR) of better than –50 dBc with an 80-MHz output signal. Though this specification does not seem to be overly impressive, it should be noted this represents a worst-case scenario. Since the largest spurious components are harmonics of the fundamental output signal (a DAC-induced artifact), it is often possible to select a clock and signal plan that minimizes harmonic artifacts, thereby making it possible to achieve better SFDR performance numbers than those specified.

The DAC is implemented as a current-output DAC with complementary output pins. An externally connected resistor (Rset) establishes the full-scale current generated by the DAC. Typically, Rset is chosen for 10-mA full-scale output current. When the input data to the DAC are at positive full scale, the normal DAC output pin sources 10 mA while the complementary DAC output pin sources 0 mA. The opposite is true when the input data to the DAC are at negative full scale. This complementary output architecture makes it possible to easily interface to a differentially coupled pulse transformer in order to reduce common-mode spurious signals.



4. This diagram shows the technique of using a spectral image of the output signal to eliminate an upconversion stage in a Tx.

BASEBAND PROCESSING

The preceding sections have described the functional blocks of the AD9857 related to generating and conditioning a carrier signal for the purpose of modulating a baseband signal. The following sections describe the baseband signal-processing chain. Baseband information is delivered to the AD9857 over a 14-b parallel data bus. The data presented to the device are assumed to consist of sampled I and Q data pairs that arrive consecutively. That is, a 14-b "I word" which is followed by a 14-b "Q word." The device provides an output clock (PDCLK) to assist the user in synchronizing the transfer of data. The PDCLK signal is a square wave at a frequency determined by the programmed interpolation rate of the digital cascaded-integrator-comb (CIC) filters. Specifically, the output data-clock frequency, f_{DClk}, is equal to $f_S/2M$, where M is the programmed interpolation rate of the CIC filters. Each rising edge of the PDCLK signal causes the AD9857 to read in an I or Q word. The user provides a synchronization signal to the AD9857 (T_x Enable), the rising edge of which signifies that the next rising edge of PDCLK represents an I word. The AD9857 continues to accept alternating I and Q words with each rising edge of PDCLK as long as the user holds T_x Enable high. This input timing architecture makes it very easy to implement data transmission in either burst mode or continuous mode.

The I and Q words are demultiplexed at the front end of the device and fed down parallel paths, which constitutes the baseband signal-processing pathway. The I and Q data are first routed to an optional inverse CIC filter. Next, it is passed through two consecutive halfband finite-impulse-response (FIR) filters that result in the I and Q data being upsampled by a factor of 4. The images normally produced by the upsampling process are effectively eliminated by the frequency-response characteristic of the halfband filters (Fig. 2). The frequency scale is relative to the sample rate of either the I or Q samples (which equates to half the frequency of the PDCLK signal). Note in Fig. 2a that the stopband attenuation is 85 dB. In Fig. 2b, it can be seen that the passband ripple is virtually nonexistent. Close inspection of Fig. 2b indicates that some of the input bandwidth at the high end must be sacrificed as part of the transition region of the filter. Approximately 10 percent of the bandwidth needs to be sacrificed to retain flatness within 1 dB. This implies that the user must upsample the raw baseband data by at least a factor of 2 prior to delivering the data to the AD9857 to avoid the bandwidth limitation imposed by the halfband filters.

Following the halfband filters is an optional CIC filter, which may be bypassed through the control registers. There are two advantages to using a CIC filter. First, the architecture is easy to implement in hardware since there are no multiplication operations required—only additions, negations, and unit delays. Second, a single CIC filter can be made to interpolate data at different rates. This is a unique characteristic of CIC filters and makes it possible to have a programmable interpolation rate. In the AD9857, the CIC filter can be programmed (through the control registers) to interpolate at any integer rate from 2 to 63. Programming the CIC filter with a value of 1 causes it to be completely bypassed.

A CIC filter is not without its shortcomings, however. There are three distinct disadvantages associated with the use of a CIC filter, but each of these shortcomings has been overcome in the AD9857 through careful system design. The first disadvantage imposed by a CIC filter is that the loss through the filter is a function of the programmed interpolation rate (an unavoidable consequence of the architecture). The magnitude of the loss ranges from no loss up to 6-dB loss. To combat this problem, the output stage of the CIC filter incorporates a digital scaler that automatically corrects the rate-related loss. A second disadvantage of a CIC filter is that the passband response is not flat, but follows a form very similar to a SINC response. This problem can be overcome by enabling the optional Inverse CIC filter mentioned earlier. This FIR filter modifies the data before they arrive at the halfband filters to compensate for the frequency response of the CIC filter. This effectively flattens the CIC response over the bandwidth of the passband of the halfband filters. In most applications, the inverse CIC filter is not required because the halfband filters limit the spectrum of the incoming data to a nearly flat region of the CIC response curve. However, some applications cannot tolerate the 0.5 to 0.7 dB of droop imposed by the CIC filters, in which case, the user has the option of enabling the inverse CIC filter to flatten out the offending droop. A third disadvantage of a CIC filter is that the integrator section incorporates feedback. This is not a problem in and of itself, but if the device is not used properly, the CIC filters can be forced into an overflow condition (from which a soft recovery is not possible). As a fail-safe against this potential problem, a control bit is provided in the control registers that can force the CIC filters to be cleared. In addition, the user is provided with a CIC overflow indicator pin, which serves as an alarm, indicating that the CIC filters have entered an overflow condition.

The AD9857 has been designed with both power conservation and noise reduction in mind. A separate clock domain controls each major functional block. When a functional block (e.g., the CIC filters) is disabled or bypassed, the clock to that section is stopped, thereby reducing power consumption and any associated digital switching noise.

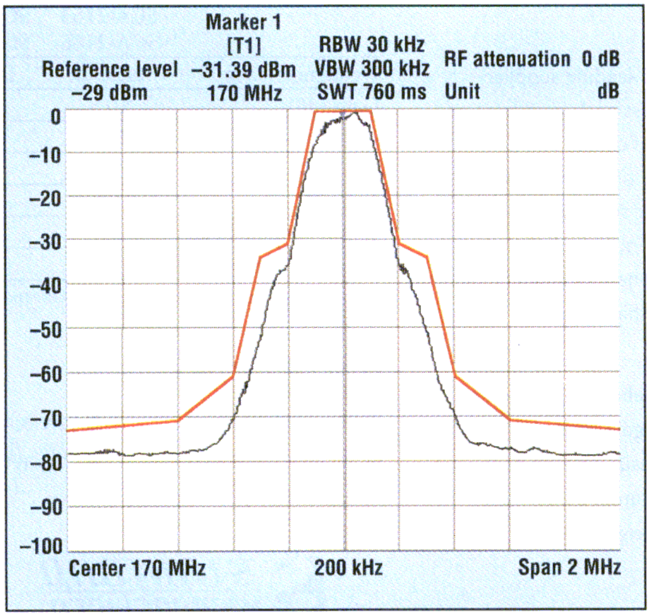
CONFIGURATION PROFILES

The AD9857 has another feature that adds to its versatility—its ability to switch between four configuration profiles. A profile is a specific grouping of control registers. Each profile consists of an identical grouping of registers. However, the individual registers within each profile are independently programmable, allowing four different device configurations to be programmed. The registers associated with each profile support programming of a tuning-word value (frequency), a CIC interpolation rate, spectral inversion (or not), an output scale value, and bypassing of the inverse CIC filter (or not). To select a specific profile, the user simply applies the appropriate logic levels to the two profile-select pins (P0 and P1). This provides the user with the ability to rapidly switch between configurations through external hardware control. Furthermore, the profile-select pins are synchronized internally with the rising edge of the PDCLK signal to ensure predictable latency through the device.

The advantages of using profiles become apparent when, for example, a form of modulation known as frequency-shift keying (FSK) is employed. FSK modulation is accomplished by switching between two predefined frequencies in sympathy with the 0's and 1's of an input data stream. FSK can be easily implemented by loading frequency number one in to Profile 0 and frequency number two into Profile 1. Then, with the P1 pin held low, the user simply drives the P0 pin with the FSK data stream. Thus, whenever the FSK data stream presents a logic zero to the P0 pin, then frequency number one is transmitted. Likewise, when a logic one is presented to the P0 pin, then frequency number two is transmitted. Many other scenarios are feasible using this flexible architecture.

The AD9857 is well-suited for wireless base-station applications. Base stations in Global System for Mobile Communications (GSM), for example, offer a significant design challenge in terms of noise floor and spurious requirements. Especially those systems that employ Gaussian minimum-shift-keying (GMSK) modulation. The stringent noise and spurious requirements can make it difficult for the system designer to find an adequate digital solution. However, with its 14-b integrated DAC and 14-b data pathway, the AD9857 yields sufficient resolution and dynamic range to provide the designer with a digital alternative to base-station transmitter (Tx) design.

Given that the basic GSM data rate is 270.833 kb/s, it is a simple matter to convert the basic GSM data rate to a higher sample rate with the AD9857's baseband signal-processing chain. For exam-



5. This plot shows GMSK modulation for a 170-MHz carrier at a sampling rate of 195 MSamples/s.

ple, assume that the user elects to oversample the basic GSM data by a factor of 2 prior to presenting it to the AD9857. This alleviates the bandwidth-sacrificing problem pointed out in the halfband-filter section. The sample rate is now 541.667 kSamples/s. With a CIC interpolation rate of 63 (the maximum value) and the 4 × upsampling of the two halfband filters, the sample rate becomes 136.5 MSamples/s, well within the 200-MSamples/s limit of the AD9857. Since it is now established that the sample rate is 136.5 MSamples/s, in this case, the GMSK signal can be broadcast on an intermediate-frequency (IF) carrier at any frequency up to 40 percent of this rate. For this example, any carrier frequency up to 54.6 MHz may be used.

To demonstrate the AD9587's capability in generating a GMSK-modulated carrier, consider Fig. 3. It shows a spectral plot of the AD9857 transmitting a GMSK-modulated signal on a 39-MHz carrier (156 MSamples/s). The basic GSM data have been oversampled by a factor of 4 and the CIC interpolation rate set to 36. Note that the AD9857 meets the GSM spectral mask with margin to spare.

In some systems, multiple upconversions are required to shift the IF up to the final broadcast frequency, which can be in the 900- or 1800-MHz range depending on the GSM system employed. Thus, a 39-MHz IF signal may not be sufficiently high enough to eliminate an upconversion stage. However, the exceptional performance of the AD9857 makes the use of an image of the output signal possible instead of the fundamental. Figure 4 demonstrates the technique of using a spectral image.

Figure 5 shows a spectral plot of the AD9857 transmitting on a 170-MHz carrier (195 MSamples/s), which is the first image of the 25-MHz fundamental carrier. Similar to the setup of Fig. 3, the basic GSM data rate has been oversampled by a factor of 4, but the CIC interpolation rate is set to 45 instead of 36. Note that the AD9857 is still able to meet the GSM spectral mask despite the reduced SNR imposed by the SINC rolloff characteristic. In Figs. 3 and 5, the spectral mask specified in the GSM 05.05 document for GSM-900 transmitters is superimposed on the AD9857 output spectrum.[1] This demonstrates the AD9857's performance relative to the GSM specification.

In short, the AD9857 is an extremely flexible, low-cost DDS-based solution for a wide range of frequency-synthesis applications. It can be used as a single-tone clock source or LO, as a rate-programmable interpolating DAC, or as a quadrature digital upconverter in a wide variety of applications. Its 14-b DAC and 14-b baseband data path yields sufficient resolution and dynamic range to make it suitable for wireless (and wired) base-station and spread-spectrum applications. A low-cost evaluation board is available for checking the performance of the AD9857, complete with a Windows™-based software package with intuitive graphical user interface (GUI). The evaluation board connects to the parallel port of a personal computer (PC), which serves as the link between the user and the AD9857's serial I/O port. The register programming protocol is handled by the software package, thereby sparing the user the task of learning the details of communicating with the AD9857. P&A: $15.40 (1000 qty.) and $250 (evaluation boards); stock. **Analog Devices, Inc., One Technology Way, P.O. Box 9106, Norwood, MA 02062-9106; (800) 262-5643, (781) 329-4700, FAX: (781) 326-8703, Internet: http:// www.analog.com.**

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Reference

1. "Digital Cellular Telecommunications System (Phase 2+); Radio Transmission and Reception", GSM Technical Specification, GSM 05.05, Version 5.2.0, July 1996, European Telecommunications Standards Institute (ETSI).

Acknowledgments

The author gratefully acknowledges the technical and editorial suggestions of Jim Surber, Nelson Pascal, and John Kornblum of Analog Devices, Inc. (Greensboro, NC).

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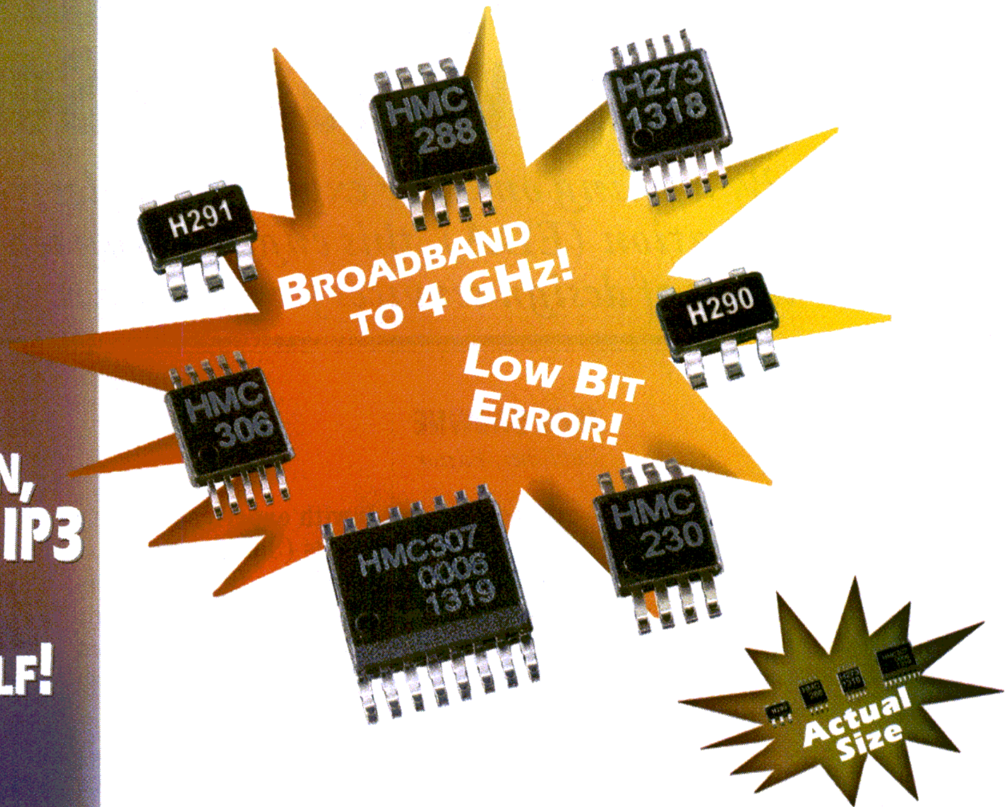

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ICs Help Implement A Trim-Free VCO

A new family of integrated circuits can ease the task of developing compact, fixed-frequency voltage-controlled oscillators (VCOs) for IF applications.

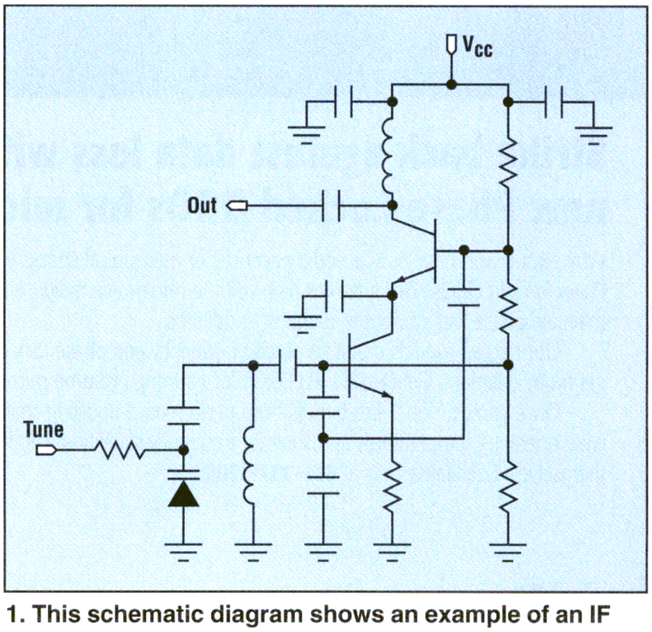
Chris O'Connor

Member of Technical Staff

Maxim Integrated Products, 120 San Gabriel Dr., Sunnyvale, CA 94086; (408) 737-7600, FAX: (408) 737-7194, Internet: http://www.maxim-ic.com.

DESIGNING a voltage-controlled oscillator (VCO) for fixed intermediate-frequency (IF) use can be daunting. Fortunately, the MAX2605-2609 VCO integrated circuit (IC) from Maxim Integrated Products (Sunnyvale, CA) can simplify the task while saving money and printed-circuit-board (PCB) space when compared to conventional discrete-device VCO solutions.

In a traditional IF VCO design, discrete transistors, resistors, capacitors and inductors form the oscillator core and output-buffer stage (Fig. 1). The tank is built from a network of the frequency-setting inductor, varactors, coupling capacitors, and feedback capacitors. The output stage uses reactive elements to impedance match the output to the particular load impedance. The proper component values must be found to ensure a successful design. These include considering the needs to establish the desired nominal oscillation frequency, guaranteed oscillator startup under all conditions, adequate tuning range, proper biasing, and proper output-stage performance. Even with a good first-order design, problems can occur—trade-offs exist between current consumption, start-up margin, frequency-tuning range, and phase noise.



1. This schematic diagram shows an example of an IF VCO implemented with discrete-circuit elements.

A major disadvantage with discrete IF VCO designs is the amount of PCB area needed. Much effort must be expended in optimizing the layout to below 6 × 10 mm. Furthermore, the PCB layout is critical in affecting the performance and design accuracy of the VCO. The layout contains parasitic capacitances and inductances that affect the frequency of oscillation. These must be taken into account in order to properly implement the oscillator. Parasitic elements often cause an undesired shift in the nominal oscillation frequency, which negatively contributes to design-centering errors and ultimately forces the need for greater tuning range to account for the errors.

In order to simplify the design of IF VCOs, as well as reduce cost and size, Maxim Integrated Products has developed the MAX2605-2609 family of IF VCO ICs. The five ICs in the family are designed specifically for low-power, fixed/single-frequency portable wireless applications from 45 to 650 MHz. The ICs include much of the circuitry needed on chip—only the tank inductor (which establishes the oscillation frequency) is left off-chip. Once the correct value of external inductance is chosen, the IC is guaranteed to tune to the corresponding frequency over the tuning-voltage range of +0.4 to +2.4 VDC. The IC's tuning-voltage input can be driven directly from the output of a phase-locked-loop (PLL) loop filter. The MAX2605-2609 ICs are designed for supply voltages of +2.7 to +5.5 VDC. The supply-voltage connection does not require special regulation for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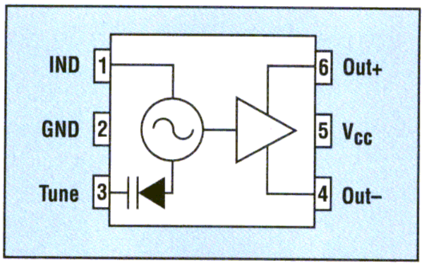
proper operation. Each IC is supplied in a tiny six-pin plastic SOT-23 package (Fig. 2).

The product family includes the MAX2605, MAX2606, MAX2607, MAX2608, and MAX2609. The MAX2605 tunes from 45 to 70 MHz with phase noise of –117-dBc/Hz offset 100 kHz from the carrier. The MAX2606 tunes from 70 to 150 MHz with phase noise of –112-dBc/Hz offset 100 kHz from the carrier. The MAX2607 tunes from 150 to 300 MHz with phase noise of –107-dBc/Hz offset 100 kHz from the carrier. The MAX2608 tunes from 300 to 500 MHz with phase noise of –100-dBc/Hz offset 100 kHz from the carrier. The MAX2609 tunes from 500 to 650 MHz with phase noise of –93-dBc/Hz offset 100 kHz from the carrier.

The frequency-tuning range, biasing, startup, and other oscillator characteristics are managed within the IC, eliminating the design headaches typically associated with a VCO design. On-chip varactors and capacitors eliminate the need for external tuning elements, and thus simplify the design of an IF VCO.

The MAX2605-2609 was conceived to provide several important new value elements to RF designers. The IC was designed to create VCOs that would be trimless, without the need for external adjustments. It was also designed to cover a wide range of application frequencies to accommodate the anticipated range of system IFs found in dual-conversion systems. Additionally, it was designed to have a flexible output interface, help reduce the cost of building an IF VCO, and help shrink the size of the final design.

Since the MAX2605-2609 represents a new VCO concept, a fundamentally new circuit approach was needed to achieve the product objectives. Maxim devised an oscillator scheme that could use the reliable and flexible Colpitts oscillator structure. The topology was adapted so that all oscillator-circuit elements except the inductor would be integrated into the IC. Integrating nearly the entire oscillator on-chip



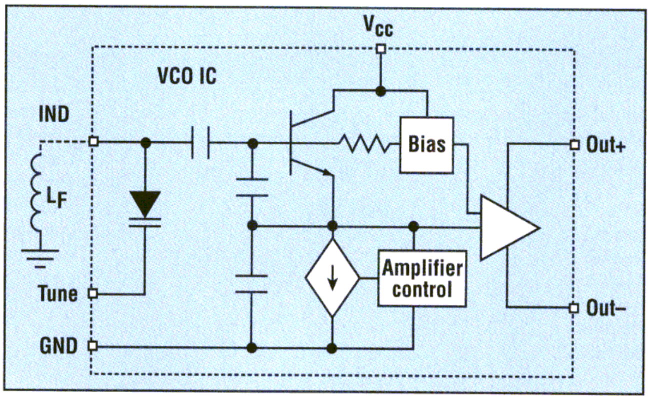
2. The MAX2605-2609 IF VCO IC is supplied in a six-pin surface-mount SOT package designed to occupy minimum PCB space.

created the opportunity to design the IC to simultaneously provide all the desired operating objectives of a good VCO design. The circuit could ensure proper oscillator startup, wide frequency range, required tuning characteristics for trimless operation, controlled current consumption, as well as power-supply and temperature-independent biasing.

Using an off-chip inductor allows the VCO to be applied over a wide range of operating frequencies. The on-chip capacitance remains the same, but changing external inductance values modifies the resonant frequency of the oscillator-tank circuit. If the inductor has a minimum quality factor (Q), then the phase-noise and startup behavior can be guaranteed (Fig. 3).

In order to implement this new circuit approach, the IC technology used needed a full complement of active and passive elements to support construction of the oscillator circuit shown. Specifically, the IC-process technology had to provide high-frequency transistors, high Q, high-capacitance-ratio varactor diodes, high-Q capacitors, and positive-negative-positive (PNP) or p-channel-metal-oxide-semiconductor (PMOS) devices. The MAX2605-2609 IC is fabricated with a silicon (Si) bipolar-complementary-metal-oxide-semiconductor (BiCMOS) process developed specifically for RF ICs, including monolithic oscillator structures. This process features negative-positive-negative (NPN) transistors with a transition frequency (f_T) of 25 GHz; PNP, N-type-metal-oxide-semiconductor (NMOS), and PMOS devices; low-series-resistance varactor diodes with better than 2:1 capacitance ratio for tuning voltages from +0.4 to +2.4 VDC; very high-Q metal-insulator-metal (MIM) RF capacitors; precision thin-film resistors; and three layers of metal.

Finally, in order to absolutely guarantee that the oscillator possessed sufficient frequency-tuning range to account for the shift in operating frequency caused by component tolerances, Maxim elected to perform production testing on the devices and guarantee a set of frequency limits. The limits provide the user of the MAX2605-2609 with an absolute guaranteed set of high and low frequency-tuning limits (f_{MAX} and f_{MIN}, respectively), where a passing IC has a frequency of oscillation (f_{OSC}) that is less than or equal to the minimum tuning frequency (f_{MIN}) at a tuning voltage (V_{TUNE}) of +0.4 VDC and $f_{OSC} \leq f_{MAX}$ at V_{TUNE} = +2.4 VDC. This testing guarantees that the VCO will always tune to the inductor-selected operating frequency with no adjustments to the external inductance value—assuming a ±2-percent tolerance-external inductor, temperature drift, and a small design centering error (less than 0.5 percent)—thereby achieving a trimless VCO design.



3. This circuit diagram of the MAX2605-2609 VCO IC shows how changing the value of the external inductor changes the resonance of the on-chip tank circuit.

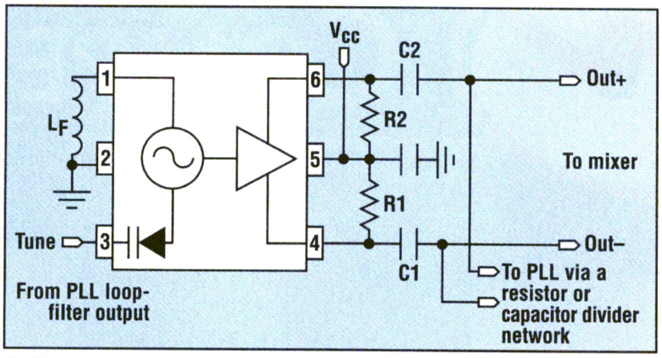
The application of the MAX2605-2609 was designed to be highly simplified and conceptually easy to understand. The application of the MAX2605-2609 involves two simple steps:

1. Selecting and imple-

menting an external inductance to set the desired oscillation frequency.

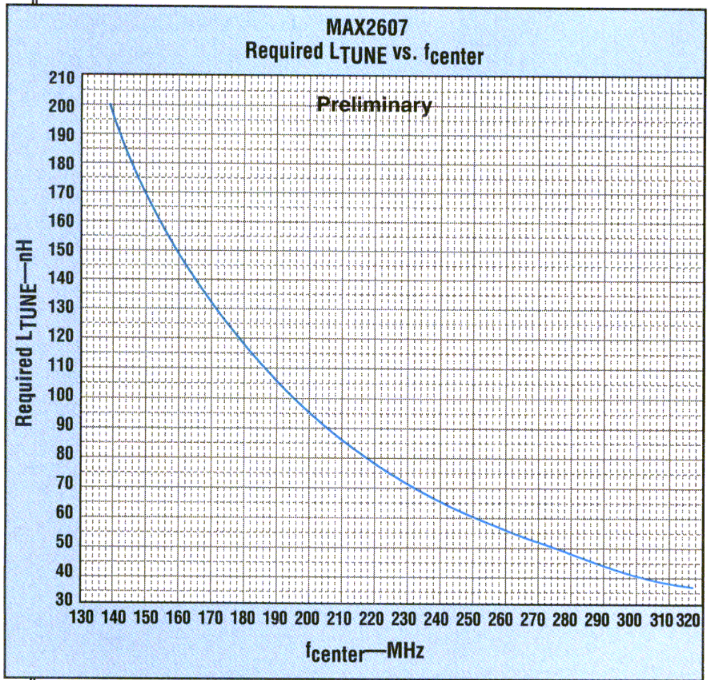
2. Resistively or reactively matching the output stage to the load (Fig. 4).

The nominal desired operating frequency of the VCO (f_{NOM}) is selected solely by the value of external inductance. The required external inductance versus the desired operating frequency is described through a detailed curve (Fig. 5). The required inductance value is taken directly from this curve. The inductance is the effective inductance at pin 1 (IND).



4. This simple schematic diagram represents a typical application for the MAX2605-2609 VCO IC.

The inductance value (L_F) that is required for the desired operating frequency may not necessarily coincide with a standard value surface-mount-technology (SMT) inductor, which typically increases in approximately 1.2× steps. In these cases, the inductance must be constructed from two inductors—L_{F1} and L_{F2}—in order to achieve the desired value. Inductor L_{F1} should be chosen as the nearest standard-value inductor just less than the required total inductance. Inductor L_{F2} should be chosen as the nearest-standard value just less than $L_F - L_{F1}$. Inductor L_{F1} should adhere to the minimum-Q requirements, but inductor L_{F2} may be implemented as a lower-cost thin-film SMT inductor. Its lower Q has only a small impact on the overall Q of the inductance because it is less than 20 percent of the total inductance. It is also permissible to use PCB traces in order to provide a small amount of inductance, thereby adjusting the total inductance value. On the MAX2608 and MAX2609, the inductance values for inductor L_{F2} are sometimes precisely implemented as a PCB trace (shorted to ground), rather than as an SMT inductor. Once the value of inductance is established at pin IND, the VCO is guaranteed to tune to this oscillation frequency over all component variations, operating temperatures, and supply voltages.



5. This plot contains values of required total-tuning inductance (L_F as a function of desired-oscillation frequency for the 150-to-300-MHz MAX2607 VCO IC.

The MAX2605-2609 VCO includes a differential-output amplifier after the oscillator core. The amplifier stage provides valuable isolation and offers a flexible interface to the IF functions, such as a mixer and/or PLL prescalar. The output can be taken single-ended or differentially. However, the maximum output power and lowest harmonic output are achieved in the differential output mode. Outputs OUT– and OUT+ are open-collector style and require a pullup element to the collector voltage V_{CC}. The output stage may be applied with a pullup resistor or inductor. A pullup resistor is the most straightforward method of forming an interface to the output and works well in applications that operate at lower frequencies or only require a somewhat modest voltage swing.

For frequencies of operation above the 3-dB bandwidth and/or when a greater voltage swing or output power is desired, a reactive-power match is required. The matching network is a simple circuit with shunt inductor and series capacitor. The inductors are connected from pins OUT– and OUT+ to pin V_{CC} to provide DC bias for the output stage. The series capacitors are connected from pins OUT– and OUT+ to the load. The values for LM and CM are chosen according to the operating frequency and load impedance. In general, the differential output may be applied in any manner like conventional differential outputs. The only constraints are the need for a pullup to VCC and a limit to the voltage swing at the output pins OUT– and OUT+.

A comparison of the design time needed to apply each approach reveals a dramatic difference in the development time. The classical/discrete approach shown is very design-intensive to implement. Conversely, with the MAX2605-2609, it is possible to design the VCO solution in minutes and verify and test it in an afternoon. The MAX2605-2609 dramatically reduces the required development time. **Maxim Integrated Products, 120 San Gabriel Dr., Sunnyvale, CA 94086; (408) 737-7600, FAX: (408) 737-7194, Internet: http://www.maxim-ic.com.**

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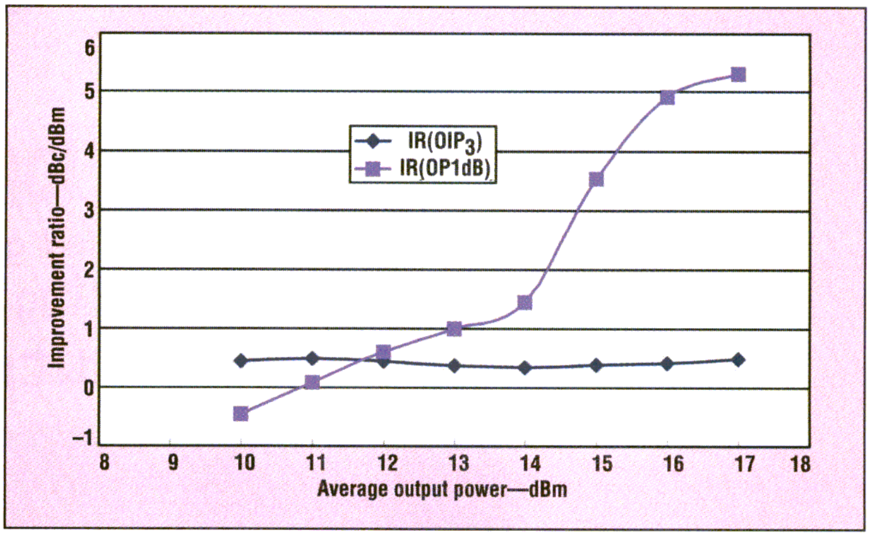

6. This graph shows the ACPR improvement ratios IR(OIP_3) and IR($P_{1\,dB}$) in the forward CDMA modulation. Increasing the backoff power level is a more-efficient way for improving ACPR than increasing the OIP_3.

(continued from p. 92)

RF Power Amplifier Design For CDMA Signals," *1996 IEEE International Microwave Theory And Techniques Symposium Digest*, June 1996, pp. 851-854.

3. Hsin-Chin Chang, Seng-Woon Chen, Doug Moe, and Scott Hackman, "CDMA ACPR Prediction Using Two-Tone Intermodulation Spectrum By Modeling Device Non-Linearity With Tchebyshev Polynominals," *1997 Wireless Communications Conference*, pp. 117-120.

4. Nuno Borges de Carvalho and Jose Carlos Pedro, "Compact Formulas To Relate ACPR And NPR To Two-Tone IMR And IP3," *Microwave Journal*, Vol. 42, No. 12, December 1999, pp. 70-75.

Table 2: Summary of the testing results of the DUT-C biased at +5 VDC and +6 VDC

DUT		C		C		Delta
V_{CC} (V)		5		6		1
I_{CC} (mA)		146		170		24
OIP_3 (dBm)		42.8		43.2		0.5
P_{1B} (dBM)		23.8		25.1		1.3
LFOM		28.9		20.4		–8.5
	P_{out} (dBm)	ACPR (dBc)	Backoff (dBm)	ACPR (dBc)	Backoff (dBm)	
Forward CDMA	12	66.0	12	66.8	13.3	0.8
	13	63.5	11	64.8	12.3	1.3
	14	60.5	10	62.4	11.3	1.9
	15	58.4	9	63.0	10.3	4.6
	16	53.1	8	59.5	9.3	6.4
Reverse CDMA	17	66.6	7	69.0	8.3	2.4
	18	63.4	6	66.1	7.3	2.7
	19	59.4	5	63.5	6.3	4.1
	20	54.8	4	61.3	5.3	6.5
	21	48.2	3	57.0	4.3	8.8
NADC	22	31.8	2	33.2	3.3	1.4

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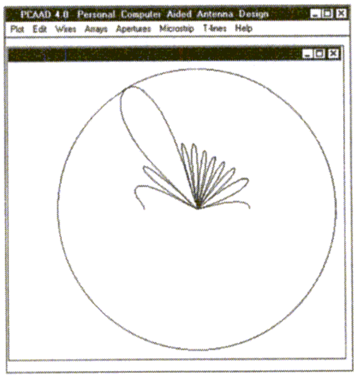

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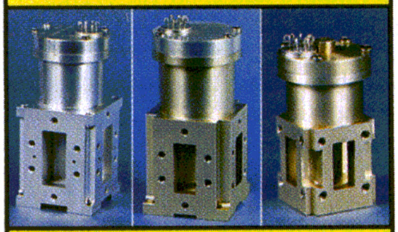

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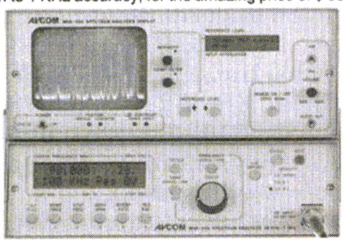

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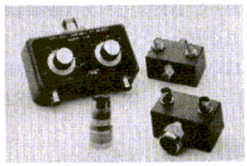

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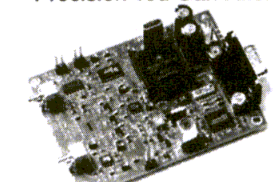

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The model JCA23-W03 amplifier for wireless-local-loop (WLL) applications covers 2.3 to 2.5 GHz with a minimum gain of 24 dB and a gain flatness of ±0.5 dB. The amplifier boasts a power output of +30 dBm minimum at the 1-dB compression point. It offers a typical noise figure of 1.3 dB and typical output intercept compression point of +40 dBm. Customized versions are available, with options for alternative gain specification, temperature compensation, and drop-in packages. **JCA Technology, Inc., 4000 Via Pescador, Camarillo, CA 93012; (805) 445-9888, FAX: (805) 987-6990, e-mail: jca@jcatech.com, Internet: http://www.jcatech.com.**

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Attenuator suits cellular environments

Model 3-A-MFB-XX is a convection-cooled attenuator with a 3-W continuous power rating and ambient temperature range of operation of −40 to +40°C. With a maximum VSWR of 1.1:1 from DC to 1 GHz and 1.25:1 VSWR from 1 to 4 GHz, this unit is ideal for cellular, personal-communications-services (PCS), laboratory, and field environments. The product's standard attenuation values are 1, 2, 3, 6, 10, 20, and 30 dB. **Bird Component Products, 10950 72nd St. N., Suite 107, Largo, FL 33777-1527; (727) 547-8826, FAX: (727) 547-0806, e-mail: sales@birdfla.com, Internet: http://www.birdfla.com.**

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VCO reaches 2300 MHz

Model VCO-2200AT is a voltage-controlled oscillator (VCO) that generates frequencies from 2100 to 2300 MHz with control voltages from 0.5 to 4.5. The unit typically requires 12-mA current from a +5-VDC power supply. Typical phase noise at 100-kHz offset is −116 dBc/Hz, and typical output power is 0 dBm. Second harmonic suppression is typically −14 dBc and third harmonic suppression is typically −20 dBc. The unit is housed in a 0.50 × 0.50 × 0.18-in. (1.27 × 1.27 × 0.46-cm) surface-mount package. **Vari-L, 4895 Peoria St., Denver, CO 80239; (303) 371-1560, FAX: (303) 371-0845, Internet: http://www.vari-l.com.**

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Sync module filters jitter

The model SY-0001 Stratum-3 clock module is a semicustom subsystem that offloads costly design engineering tasks and boasts a turnkey solution to many communication-systems design challenges. The unit features its own oven-controlled crystal oscillator (OCXO), two independent inputs of any reference frequency from 8 kHz to 77.76 MHz, and a digital-signal-processor (DSP) phase-locked loop (PLL) with a bandwidth of 0.1 Hz that eliminates jitter in any incoming signal. DSP coefficients are held onboard in Flash memory and provide a high degree of user flexibility. Flexibility features include two external reference frequencies, control inputs through header pins, J-TAG or SCI connectors, and three packaging options. The packaging options include a hermetically sealed metal box for severe conditions, an electromagnetic-interference (EMI)-protected metal enclosure, or an open circuit board for low-profile installations. **Raltron Electronics Corp., 10651 N.W. 19th St., Miami, FL 33172; (305) 693-6033, FAX: (305) 594-3973, Internet: http://www.raltron.com.**

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Analyzer's reach is extended

The models 6846 and 6847 microwave system analyzers enhance a system's flexibility and ability to measure microwave signals to 26.5 GHz. Model 6846 measures transmission characteristics of 7-GHz radio feeds and spectral purity to the third harmonic. The instrument includes an 8.4-GHz swept source and scalar analyzer with a 24-GHz spectrum analyzer. Model 6847 has a source and scalar analyzer that extends to 20 GHz and a 26.5-GHz spectrum analyzer. This supports feeder measurements on 7-GHz, 13-GHz, and 15-GHz radios with spectrum- and modulation-envelope measurements on 23-GHz radios. **IFR Systems, Inc., 10200 W. York St., Wichita, KS 67215-8999; (800) 835-2352, (316) 522-4981, e-mail: info@ifrsys.com, Internet: http://www.ifrsys.com.**

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Filter isolates critical frequencies

The model 12086-2 waveguide filter boasts very low loss, 0.04 dB maximum, across the 3.625-to-4.2-GHz receive band. It maintains a return loss of greater than 25 dB (VSWR 1.12:1) in the passband, and achieves a rejection of greater than 70 dB in the transmit band across the 5.85-to-6.425-GHz frequency range. WR-229G flanges are standard, with other flange and mechanical configurations available upon request. **Microwave Filter Co., 6743 Kinne St., East Syracuse, NY 13057; (800) 448-1666, (315) 438-4700, FAX: (888) 411-8860, (315) 463-1467, e-mail: mfcsales@microwavefilter.com, Internet: http://www.microwavefilter.com.**

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CIRCLE NO. 387

Bypass switch serves microstrip applications

The 409B series of miniature bypass switches operates from DC to 3 GHz and is designed to be mounted on a printed-circuit board (PCB). This 0.75 × 1.00 × 0.45-in. (1.91 × 2.54 × 1.14-cm), 50-Ω impedance series of switches with +24-VDC actuating voltage is designed for microstrip applications. The unit can be hot switched at 10 W CW and can handle up to 30 W CW. The insertion-phase repeatability from port to port is within 0.01 dB for a 100-MHz bandwidth. Maximum VSWR is 1.20:1, maximum insertion loss is 0.20 dB, and minimum isolation is 70 dB. The switches have positive-intrinsic-negative (PIN) connectors. **Dow-Key Microwave Corp., 4822 McGrath St., Ventura, CA 93003-7718; (805) 650-0260, FAX: (805) 650-1734, Internet: http://www.dowkey. com.**

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Antennas suited for RFID applications

The 0015 series of antennas is suited for RF-identification (RFID) applications in license-free, ultra-high-frequency (UHF) bands. The 50-Ω impedance antennas are available in linear polarization as well as lefthand circular polarization (LHCP). They offer 7-dBi nominal gain, a beamwidth range of 65 to 70 deg., and a 1.25:1 maximum VSWR. The antennas operate at temperatures from −20 to +80°C. They include a wall-bracket mounting and SMA female RF connectors. The LHCP units feature a maximum axial ratio of 2.5 dB. The antennas weigh approximately 1 lb. and are manufactured of aluminum (Al) 6061-T6 with an ABS cover. **Seavey Engineering Associates, Inc., 28 Riverside Dr., Pembroke, MA 02359; (781) 829-4740, FAX: (781) 829-4590, e-mail: info@seaveyantenna.com, Internet: http://www.seaveyantenna.com.**

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Combline filter targets WLAN applications

Model QCB-5575M2P5SCS is a low-profile, combline filter designed specifically for the wireless local-area-network (WLAN) market. The filter operates at 5.8 GHz and boasts rejection performance to 26 GHz. The filter measures 2.50 × 0.66 × 0.40 in. (6.35× 1.55 × 1.02 cm) and weighs less than 1 oz. Its operating temperature ranges from −55 to +95°C. SMA-type male connectors or pins are included. **Q Microwave, Inc., 1331 N. Cuyamaca St., Suite G, El Cajon, CA 92020; (619) 258-7516, FAX: (619) 258-7516, Internet: http://www.qmicrowave. com.**

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Buffer amps boast high isolation

Models SGA-1163 and SGA-1263 silicon-germanium, heterojunction-bipolar-transistor (SiGe HBT) amplifiers claim to offer as much as 40-dB isolation at 2 GHz. The model SGA-1153 operates across the DC-to-4-GHz frequency range, and the model SGA-1263 operates from DC to 6 GHz. The two-stage buffer amplifiers are fabricated using the latest 50-GHz Ft process and are designed for applications at frequencies to 6 GHz. They are housed in surface-mount, SOT-363 plastic packages and are ideal for use in oscillator applications covering cellular, industrial-scientific-medical (ISM), and narrowband, personal-communications-services (PCS) bands. **Stanford Microdevices, Inc., 522 Alanor Ave., Sunnyvale, CA 94086; (800) 746-6642, FAX: (408) 739-0970, Internet: http://www. stanfordmicro.com.**

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Terminations cover military equipment

Models CT-53-BNC/M and CT-53-SMA/M represent a series of miniature 50-Ω coaxial terminations suitable for military and commercial equipment. The VSWR averages 1.1:1 over the frequency range of DC to 2 GHz (BNC) and 4.2 GHz (SMA), with a maximum VSWR of 1.3:1. Power dissipation is 0.5 W CW, 1 kW peak over the temperature range of −25 to 85°C. The design uses male gold (Au)-, silver (Ag)-, or nickel (Ni)-plated connectors and resistor elements mounted in Ag-plated housings. **Elcom Systems, Inc., PMB 255, 20423 State Rd. 7 No. F6, Boca Raton, FL 33498; (561) 883-1945, FAX: (561) 883-1945, e-mail: sales@elcomsystems.com, Internet: http:// www.elcomsystems.com.**

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Capacitors focus on high-Q applications

The model 1500 Eco-Trim® variable air capacitor is designed for applications where high-quality factor (Q) and cost are design criteria. The capacitor is ideal for impedance matching, filter tuning, interstage coupling, and antenna tuning. It has a capacitance range of 1 to 10 pF, a rated voltage of +250 VDC, and an operating temperature range of −65 to +125°C. It is available in five mounting styles for personal-computer (PC) and surface-mount-technology (SMT) applications. **Johanson Manufacturing Corp., 301 Rockaway Valley Rd., Boonton, NJ 07005; (973) 334-2676, FAX: (973) 334-2954, Internet: http://www. johansonmfg.com.**

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Generator is ideal for compact analysis

The model MG3681A signal generator boasts the performance capabilities to measure the upcoming Third-Generation-Partnership-Project (3GPP) wideband-code-division-multiple-access (WCDMA) signals. Covering 1.6-Hz-to-16.5-MHz and 30-MHz in-phase/quadrature (I/Q) modulation bandwidth, the generator is suitable for component analysis and base stations that are designed for 3GPP signals. Measuring 250 kHz to 3 GHz, the unit boasts four downlink 3GPP channels—dedicated physical channel (DPCH), primary common-control physical channel (P-CCPCH), primary synchronization channel (P-SCH), and secondary synchronization channel (S-SCH). **Anritsu Co., 1155 E. Collins Blvd., Richardson, TX 75081; (800) ANRITSU, (972) 644-1777, FAX: (972) 644-3416, Internet: http://anritsu. com.**

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Coaxial probes

A catalog describes a line of coaxial probe designs. Features that are unique to these probe designs are included, along with various specialized designs. **Interconnect Devices, Inc.;** *(913) 342-5544, FAX: (913) 342-7043, e-mail: info@idinet.com, Internet: http://www.idinet.com.*

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Adapters and terminations

Waveguide products that are designed to address military and commercial microwave applications up to 40 GHz are described in a catalog. Adapters, transitions, terminations, attenuators, couplers, harmonic filters, horn antennas, mismatches, transverse-electromagnetic (TEM) cells, and water-cooled loads are offered. Information on the company's capabilities is also provided. **Advanced Microtek Ltd.;** *01256-355771, FAX: 01256-355118, e-mail: sales@linkmicrotek.co.uk, Internet: http://www.linkmicrotek.co.uk.*

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Wireless communications

Wireless systems and components are available from a 1000-page catalog. Site infrastructure products, radio-communications equipment, power systems products, and wireless equipment are offered. Specifications are provided. **Hutton Communications, Inc.;** *(972) 417-0100, FAX: (972) 417-0180.*

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Wireless links

Laser technology and wireless links are covered in a 24-page quarterly magazine. Interviews with design leaders and conference reporting are included. **Philips Research;** *+31-40-27 42204, FAX: +31-40-27 44947, e-mail: prpass@natlab.research.philips.com, Internet: http://www.research.philips.com/password.*

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Mobile antennas

Dual-band Advanced Mobile Phone Service personal-communications-services (AMPS PCS) antenna solutions are introduced in a four-page brochure. Application information is provided. Antenna features are also included. **Richard Hirschmann of America, Inc.;** *(800) 255-0524, (973) 830-2000, FAX: (973) 830-1470, e-mail: ant-sales@hirschmann-usa.com, Internet: http://www.hirschmann-usa.com.*

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3G communications

An eight-page newsletter deals with solutions for the global mobile communications market. Bluetooth™ projects, third-generation (3G) mobile communications, and the teaming of CETECOM with T-Mobil are featured articles. **CETECOM, Inc.;** *(510) 252- 0559, FAX: (510) 252-0559, Internet: http://www.cetecom.de.*

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Radio ICs

A four-page brochure details a line of devices for personal wireless connectivity. Multipurpose RF transceivers, fractional-N frequency synthesizers, Bluetooth radio solutions, and integrated Global Positioning System (GPS) receiver RF front-end applications are discussed. Information on the company's capabilities is provided. **Philsar Semiconductor;** *(800) 551-2319, (613) 274-0922, FAX: (613) 274-0915. e-mail: info@philsar.com, Internet: http://www.philsar.com.*

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Instrumentation amplifiers

Instrumentation amplifiers (in-amps) are examined in a 66-page application book. Applying in-amps effectively, input-protection basics, and issues affecting DC accuracy are covered. Real-world in-amp application examples are included. **Analog Devices, Inc.;** *(800) 262-5643, FAX: (781) 937-1021, Internet: http://www.analog.com/in-amp-book.*

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Varactor diodes

Varactor diodes are examined in an 80-page catalog. Specifications include reverse voltage, reverse current, diode capacitance, series resistance, and capacitance ratio. Application information is provided, along with outline drawings. **Toko America, Inc.;** *(800) PIN TOKO, (847) 297-0070, FAX: (847) 699-7864, e-mail: info@tokoam.com, Internet: http://www.tokoam.com.*

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Leaded inductors

Leaded inductors are the subject of an eight-page engineering bulletin. RF-interference (RFI) suppression, filters, decoupling, as well as energy-storage chokes in switching-mode power-supply applications are offered. Features, specifications, outline drawings, along with a section of application notes is provided. **Sprague-Goodman Electronics, Inc.;** *(516) 334-8700, FAX: (516) 334-8771, e-mail: info@spraguegoodman.com.*

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Wireless products

High-performance components, networks, and instruments are listed in a 648-page catalog. Wireless products, electromechanical RF switches, passive components, active components, RF radiation safety products, as well as power meters and monitors are featured. Specifications, outline drawings, performance charts, and a secton of application notes are provided. Articles that help engnieers select the most suitable product for every requirement are presented. **Narda Microwave-East;** *(631) 231-1700, (631) 231-1390, FAX: (631) 231-1711, e-mail: nardaeast@ L-3 COM.com, Internet: http:// www.nardamicrowave.com.*

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PCB packaging

Printed-circuit board (PCB) hardware assembly and packaging products are highlighted in a catalog. Card guides, card ejectors, screw pacers and insulators, universal mounts and spacers, light-emitting-diode (LED) mounting products, as well as permanent mounts are offered. A cross-reference guide and specifications are provided. **Bivar, Inc.;** *(949) 951-8808 ext. 312, FAX: (949) 951-3974, e-mail: Bivar@Bivar.com, Internet: http://www.bivar.com.*

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RF/MICROWAVE CIRCUIT DESIGN FOR WIRELESS APPLICATIONS

Ulrich L. Rohde and David P. Newkirk

Ulrich Rohde has made a career of making an impact on the RF/microwave industry. Under his guidance during the late 1980s and 1990s, the computer-aided-engineering (CAE) firm Compact Software developed a suite of innovative design tools that included nonlinear simulation and system simulation, before the company's acquisition by Ansoft (Pittsburgh, PA). Rohde's family business, Rohde & Schwarz (Munich, Germany) is a global leader in test-and-measurement solutions. His efforts for Synergy Microwave (Paterson, NJ), where he is Chairman, have helped to raise the engineering excellence of that organization, resulting in the development of a line of fractional-N synthesizers with unparalleled performance (see *Microwaves & RF*, April 1998, p. 151).

Rohde's efforts in publishing have included an extensive bibliography of scholarly papers on low-phase oscillator and synthesizer design, as well as a series of authoritative books on frequency synthesizers and communications receivers. His latest effort in publishing, *RF/Microwave Circuit Design For Wireless Applications*, may be his finest moment. No single text could possibly address all of the complexities of RF design for wireless systems, but *RF/Microwave Circuit Design For Wireless Applications* comes close. As co-author, David Newkirk, a professional editor, did a marvelous job in transforming highly technical and sophisticated text and graphics into a quite readable book.

The text covers the design and specification of all major RF portions in a wireless system. The book only contains six chapters and two appendices, but these are information-packed chapters that could serve as the basis for separate multi-day technical workshops.

For example, Chapter 1 includes a block diagram of a handheld cellular-telephone receiver as a starting point. This chapter, which opens with discussions on radio-channel characteristics, such as fading and multipath, and the development of a radio-channel transfer function, focuses largely on modulation and access schemes, with a brief comparison of frequency-division-multiple-access (FDMA), time-division-multiple-access (TDMA), and code-division-multiple-access (CDMA) techniques. Details are presented concerning the generation of digitally modulated carriers for transmitters, how data are mapped onto baseband waveforms, and how information is demodulated from a carrier at the receiver. The "Building Blocks" section provides a miniature selection guide of RF front-end components. This chapter provides brief overviews of key system specifications, including amplitude linearity, gain compression, linearity, dynamic range, and error vector magnitude (EVM). In addition, specifications are provided for proven test instruments, such as the CMD 65 radiocommunications tester from Rohde & Schwarz.

Chapter 2, entitled "Models For Active Devices," details the usefulness of nonlinear circuit and SPICE CAE simulation tools to predict bias points for active devices in small-signal and power amplifiers (PAs) for wireless systems. Since diodes are also key semiconductor devices (for switching and varying capacitances) in wireless designs, the chapter begins by presenting a basic diode model. The model is extended to include mixer and detector diodes as well as PIN diodes and varactor-tuning diodes. The physics of varactor diodes are detailed, along with discussions on diode quality factor (Q), loss, and tuning drift. Several diode-tuned resonant circuits are presented in Chapter 2, along with supporting design equations. About one-third of the way through the chapter, the discussion turns to bipolar transistors and their large-signal behavior. Operating characteristics are reviewed, and several data sheets for the BPF420 NPN silicon (Si) bipolar from Infineon, formerly Siemens, (Munich, Germany) are included to illustrate typical performance characteristics. Small-signal models of bipolar transistors are also covered, along with a useful list of key bipolar-junction-transistor (BJT) equivalent model parameters, such as the base-to-emitter capacitance (CBE) and the collector-to-emitter capacitance (CCE). An overview of current BJT technologies is shown in the form of tabular data, complete with information on bipolar-complementary-metal-oxide-semiconductor (BiCMOS) and Si-germanium (SiGe) devices.

The second half of Chapter 2 reviews field-effect transistors (FETs) and their key parameters, again using manufacturers' data sheets to show parameter values for commercially available devices. The large- and small-signal behaviors of junction FETs (JFETs) are discussed, as well as the operating characteristics and small- and large-signal behaviors of metal-oxide-semiconductor FETs (MOSFETs). Gallium-arsenide (GaAs) metal-epitaxial-semiconductor FETs (MESFETs) are also included in the FET device coverage. The final part of Chapter 2 explores parameter extraction for active devices, including the development of noise models and scalable device models. As with the other chapters in this book, detailed information is provided on reference materials, along with a listing of publications for further reading.

Chapter 3 highlights amplifier design with FETs and BJTs, covering low-noise amplifiers (LNAs), high-gain amplifiers, and medium-to-high-power amplifiers. This is by far the largest chapter in *RF/Microwave Circuit Design For Wireless Applications*, essentially equaling the information found in some texts dedicated to amplifier design. For LNAs, extensive information is provided on noise-figure characterization, analysis, and prediction, the use of noise circles, noise correlation in two-port networks, and measurement equipment. For PAs, analysis techniques are shown for distortion with digitally modulated waveforms, and for amplifiers with automatic-gain-control (AGC) functions. Chapter 3 features design guidelines for LNAs, narrowband amplifiers, high-gain amplifiers, as well as single-stage and multistage amplifiers with feedback. This chapter also covers associated circuitry, such as filters and frequency doublers, biasing considerations, and development of matching networks. Several examples are presented for each type of amplifier, with advice on various analysis techniques, such as small-signal AC analysis and stability-analysis approaches.

Chapter 4 provides 80 pages of insights into one of the more difficult portions of an RF design, the mixer. It reviews the properties of mixers, such as conversion loss or gain, noise figure, isolation, linearity, and local-oscillator (LO) drive levels. It also points out the differences among mixer types, such as single-balanced mixers and double-balanced mixers, while comparing and contrasting diode-based mixers with transistor mixers. This chapter is generously supported with manufacturers' data sheets and schematic diagrams to help novice as well as veteran engineers understand the complex behavior of high-frequency mixers.

Chapter 5 also addresses a complex component, the high-frequency oscillator. Although this is not a simple subject, it is one in which Rohde is well-versed and his command of the subject material is apparent in the clarity of his presentation. The chapter covers basic oscillator circuits, such as the Hartley, Colpitts, and Clapp-Gouriet configurations, along with an equation for predicting a quick estimate of an oscillator's noise performance. The chapter details the use of different resonator types, such as ceramic resonators, microstrip inductors, and crystal resonators, and how to implement tuning circuits in practical oscillator designs. A detailed discussion on noise in oscillators includes noise-generation mechanisms and methods for optimizing phase-noise performance.

The final chapter in *RF/Microwave Circuit Design For Wireless Applications* explores frequency synthesizers for wireless applications. The chapter reviews the basics of phase-locked loops (PLLs) and the use of phase/frequency comparators, flip-flops, filters, and charge pumps. It includes several design examples and practical design techniques when using CAE tools. This chapter also offers some of Dr. Rohde's insights into fractional-N frequency synthesizers, as well as some thoughts on direct-digital-synthesizer (DDS) sources.

The two appendices provide information on modeling heterojunction bipolar transistors (HBTs) and on nonlinear microwave circuit design using multiharmonic load-pull simulation techniques. The latter is applied to the design of a narrowband amplifier and a frequency multiplier (doubler).

No doubt, the sheer size of *RF/Microwave Circuit Design For Wireless Applications*, at approximately 1000 pages, will intimidate some readers. But in terms of information per dollar, the book represents a real value. Many texts are available on various aspects of wireless RF engineering, but usually on narrowly defined topics. *RF/Microwave Circuit Design For Wireless Applications* is the only text that provides a complete education for wireless circuit designers, by covering wireless RF engineering from semiconductor through system levels. It can be used on many different levels, from basic tutorial to advanced design concepts, depending upon the background of the reader. The book should be particularly useful for those designers tasked with learning more about an unfamiliar portion of the RF signal chain, such as antenna designers who need to know more about LNAs. (2000, 954 pp., hardcover, ISBN: 0-471-29818-2, $125.00.) **John Wiley & Sons, Inc., 605 Third Ave., New York, NY 10158-0012; (800) CALL-WILEY, (212) 850-6011, FAX: (212) 850-6008, e-mail: PERMREQ@WILEY.COM, Internet: http://www.wiley.com.**

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Seventeen years ago, a special report by Associate Editors Mike Kachmar and Barry Manz featured an interview with Stephen Levy, senior vice-president of Motorola (Phoenix, AZ) and head of operations in Japan. Motorola became one of the first US companies to successfully compete in Japan, winning a $9 million contract from Nippon Telegraph and Telephone (NTT) for pagers.

Microwaves & RF September Editorial Preview

Issue Theme: Dual-Use Technology

News

With approximately 3000 researchers and engineers, the US Naval Research Laboratory (Washington, DC) serves as the US Navy's corporate laboratory. In many ways, NRL initiates the engineering research and development (R & D) that is eventually transformed by the electronics industry into military and commercial products. The September issue will feature an inside look at one small part of the work being conducted at NRL, where the vacuum-electronics group pursues ever-higher output-power levels with improved efficiency.

Design Features

A key high-frequency design area that has benefitted from the dual-use transition of technology from military to commercial applications is in millimeter-wave communications. September will offer several articles highlighting this technology. In addition, a method to calculate the pulse response of a dipole antenna will be discussed.

Product Technology

September's Product Technology section will introduce a radical departure for a series of microwave vector network analyzers (VNAs), a product concept where a high-powered personal computer (PC) has been seamlessly integrated with the measurement hardware to yield a new approach to vector network analysis.