A PENTON PUBLICATION MARCH 2000

FOR DESIGNERS AT HIGHER FREQUENCIES

Microwaves & RF

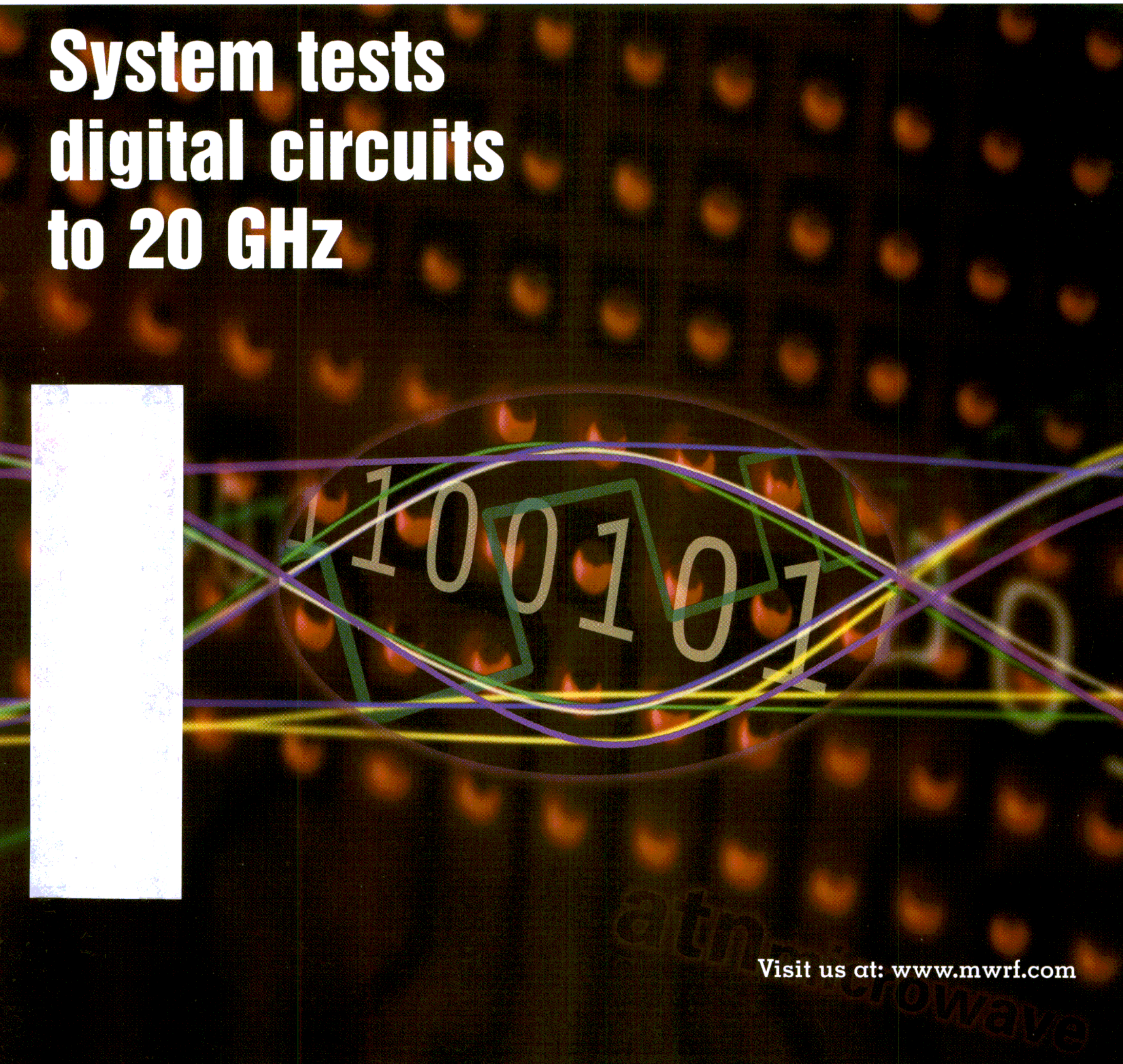
Wireless Technology/ Show Wrapup Issue

NEWS
Surveying products from the Wireless Show

DESIGN FEATURE
Review the basics of microstrip lines

PRODUCT TECHNOLOGY
Prematching tuners ease high-power device testing

System tests digital circuits to 20 GHz



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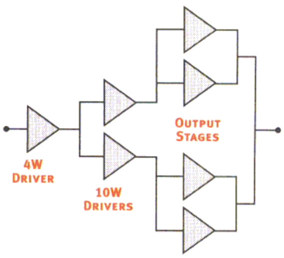

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40% Efficiency
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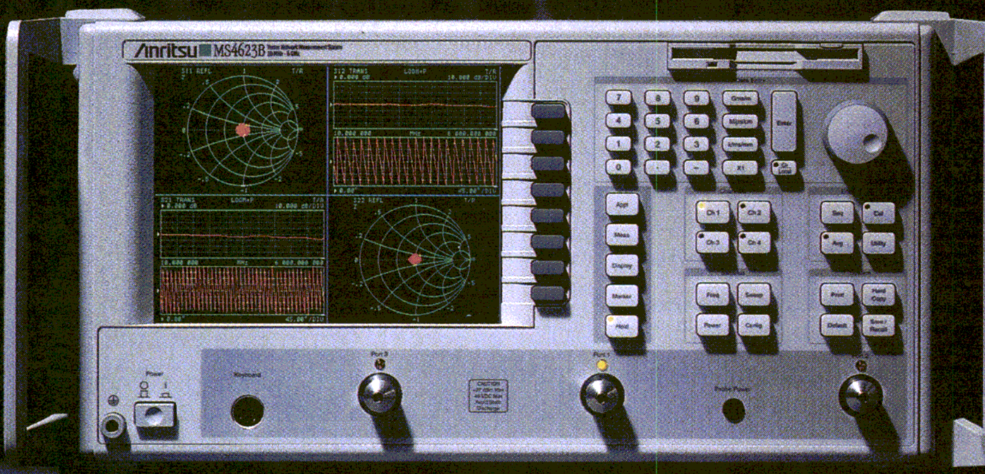
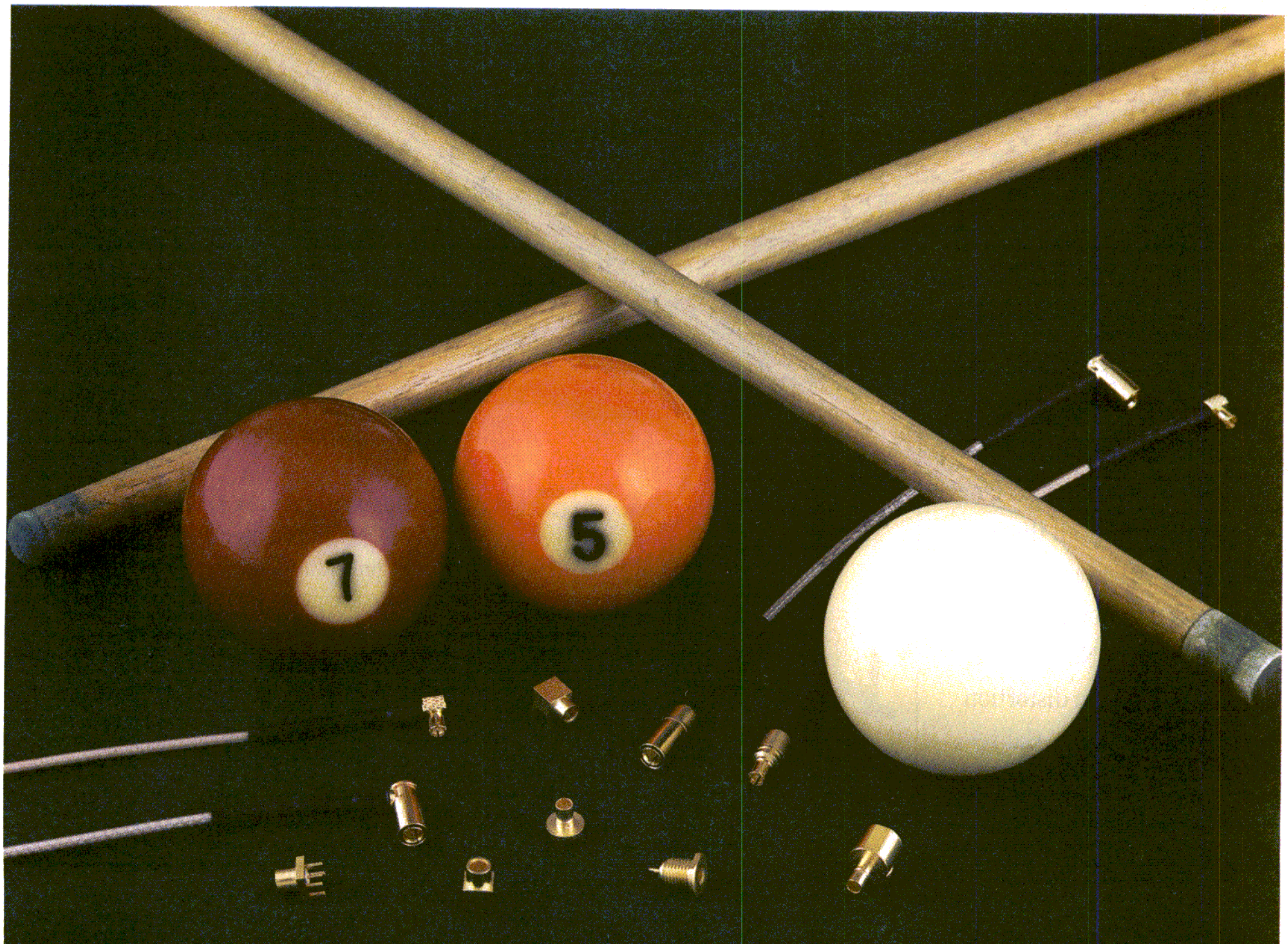
At Least, Until They Build Theirs.





INDEX LISTINGS

To the editor:

I have at least 40 years of experience (as of now) generating phrase indexes. One form of a phrase index—a phrase-index dictionary—could be published periodically as a follow-on to the listing of phrases in a paper or article. This could occur once a year depending on the amount of material that is indexed.

One section could serve as a dictionary with each word or phrase being used as a lead-in to a reference on a second page. The second page would provide a complete set of phrases for the article in question. The important thing about the phrases that are used would be that each idea expressed, whether in line with the main theme of the article or not, would be indexed if there is any possibility that a reader might have an interest in its use.

The suggested listings with their corresponding articles could be used as the basics for the more extensive listings. The listings would increase the usefulness of the magazine at the time it is received because they would help readers find unexpected items in papers or articles that could prove useful. The result would be a more useful publication for the reader.

Why do this? As previously mentioned, one looks up a subject word that can lead to a reasonably well-defined idea. From that, one goes to the referenced documents—the "article cards" on the second page. In doing this, one can determine which idea fits the desired search point most closely. Ideally, the editor keeping up the data base of articles, their key phrase words, and listings could track all significant ideas in the articles—directly important and peripheral. Catching the peripheral ones is critical.

One can find a single word, then a phrase, note the various references to these phrases, and look up the sets where one can find the article or paper that he or she is looking for. The listing could be easily prepared by the author of the article, an editor, or anyone who is knowledgeable in the field.

As mentioned before, these listings would simplify and speed the search for subjects or ideas that are of particular interest. Readers and editors would be more interested in the magazine and advertisers may also benefit.

It is important to remember that one needs to find a listing from any word in the phrase to go to the specific reference. At that location, the phrase can be Boolean added with the other phrases to evaluate the document prior to the reading decision. Instructions on using this system of indexing should be available on request and published occasionally. The obvious objective is to make the publication more useful to the reader.

Keats Pullen Jr.
Kingsville, MD

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We have the capacity to handle product volume & design complexity requirements of any dimension...at astonishing speed.

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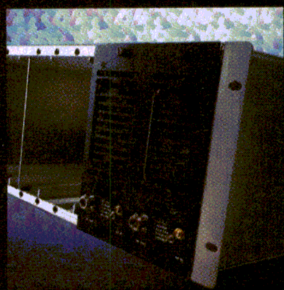
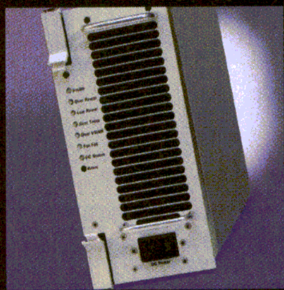
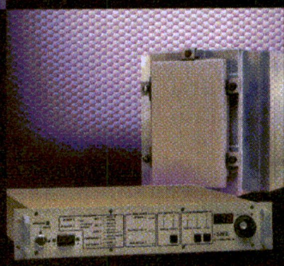






ADDING FIBER TO A MICROWAVE DIET

Strange as it may seem, fiber-optic technology has become an increasingly important part of the Wireless Symposium/Portable By Design Conference & Exhibition. Numerous fiber-optic manufacturers were on hand at the eighth annual conference and exhibition to report on practical solutions to meet the needs of increasing analog bandwidths and digital data rates. While kilometers of glass fibers may seem out of place at a meeting devoted to wireless technology, it is now not uncommon for wireless system designers to be working with photodetectors and laser diodes.

Fiber-optic technology, of course, is not new to the readers of this magazine. In October 1983, a story on a small company in Alhambra, CA named Ortel Corp. reported on laser diodes with microwave-like amplitude-modulated (AM) bandwidths approaching 7 GHz.

Now, of course, commercially available laser diodes and photodetectors operating at rates above 10 Gb/s (the data equivalent of an analog 10 GHz) are available from a number of suppliers, including Fujitsu Microelectronics (Santa Clara, CA) and Oki Semiconductor (Sunnyvale, CA). What is their connection to wireless communications systems?

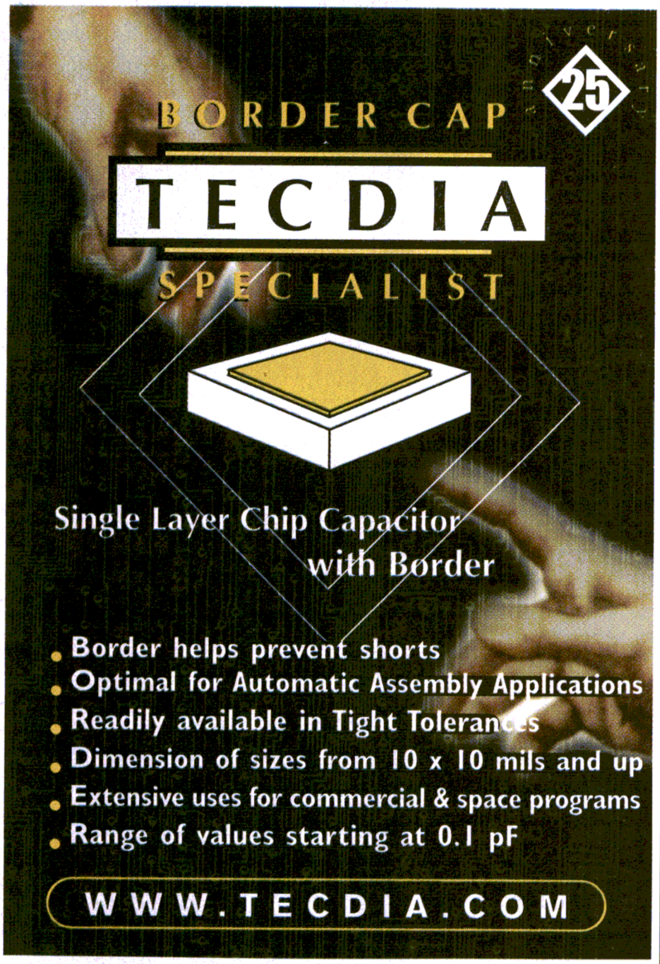
Well, one of the most oft-heard topics during the Wireless Symposium (San Jose Convention Center, San Jose, CA, February 22-25, 2000) was wireless data. In the early days of this conference (1993), the majority of wireless activity had to do with voice communications. But the emergence of the Internet, e-mail, e-commerce, and other forms of data communications in recent years have forced wireless service providers to take a hard second look at their infrastructures. The wireless community has responded in kind with a score of air-interface standards for the second- and third-generation (2G and 3G) wireless communications systems. But, as Motorola learned with its IRIDIUM satellite communications system, an entirely wireless network may not always present the most cost-effective solution. In many cases, the best solution is a blend of technologies that includes traditional copper (Cu) cables, wireless links, and, yes, fiber-optic links.

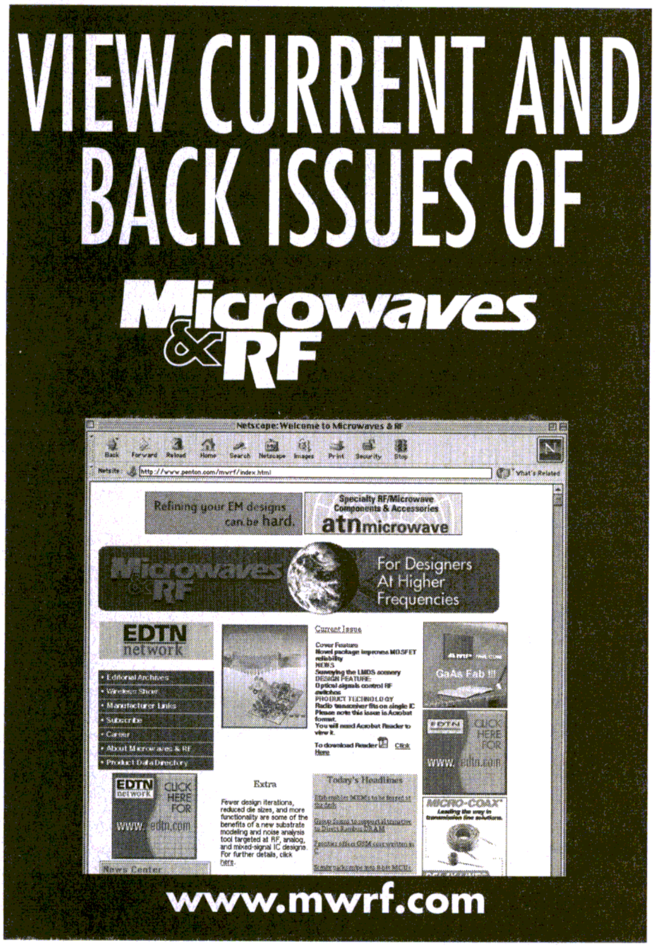
Fiber-optic technology can be considerably more expensive to install than building a few wireless antenna towers and base stations, but in many ways it is superior to wireless technologies. Fiber offers tremendous bandwidth, without the government-regulated transmission channels of wireless systems. And fiber-optic links, because they use modulated light, are not subject to RF interference (RFI) and electromagnetic interference (EMI) in the manner of a wireless system.

Fiber-optic hardware suppliers are already talking about operating at rates of 40 Gb/s, which may be a few years away in terms of commercial products. But the potential for bandwidth is enormous, and this is a technology that works quite well alongside existing wireless technologies. It is no wonder that so much of the discussions at the Wireless Symposium on wireless data and Internet access also included mentions of fiber optics.

Jack Browne

Publisher/Editor





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DIVERSITY

COMPONENTS

RF Semiconductor Market In Cell Phones To Hit $7.7 Billion In 2004

MOUNTAIN VIEW, CA—The worldwide market for RF semiconductor devices in cellular telephones will reach $7.7 billion in 2004, up from $3.9 billion in 1999, according to a recently released report from Strategies Unlimited, a Silicon Valley market-research firm. The report, *RF Semiconductors for Cellular/PCS Handsets*, Market Review and Forecast 2000, examines the semiconductor technology and trends behind the evolution to third-generation (3G) cellular chip sets. It projects worldwide demand for handsets through the year 2004 and the accompanying RF semiconductor market.

Led by strong global demand for mobile communications, the number of cellular and personal-communications-services (PCS) subscribers is forecast to reach $1.3 billion in 2004. Annual handset demand is projected to grow from 240 million units in 1999 to 600 million in 2004. New services such as instant messaging, wireless data, and Internet access will provide continual momentum for strong growth over the next five years.

According to the report, Global System for Mobile Communications (GSM) handsets are the largest market for RF semiconductors with device shipments of $2 billion in 1999, followed by code-division multiple access (CDMA), IS-136 time-division multiple access (TDMA), Personal Digital Cellular (PDC), and analog chip sets with $1.9 billion. Small-signal amplifiers, frequency-conversion devices, and intermediate-frequency (IF) chip sets represented the largest part of the total market by chip type. Power-amplifier (PA) and RF control chips also showed strong growth over 1998 levels.

MEMS-Based Filter Achieves High Efficiency And Low Insertion Loss

RESEARCH TRIANGLE PARK, NC—A newly developed RF frequency-agile filter using micro-electromechanical-systems (MEMS) technology matches or exceeds several performance criteria of traditional RF filters, but it is in a much smaller assembly using fewer parts. Using MEMS relays supplied by Cronos Integrated Microsystems, Inc., Raytheon Co. (Fort Wayne, IN) developed an experimental 25-W tunable power RF bandpass filter for potential use in a variety of military and commercial telecommunications applications where compact size and high efficiency are needed. The Raytheon filter is the first demonstration of a high-watt MEMS-based tunable RF device. Previous RF filter developments using MEMS technology centered on low-power circuits suitable for receiving, and not for high-power transmitting applications.

By employing MEMS relays in the tuning process, the board area and volume of the RF filter-tuning components can be reduced by greater than 50 percent compared with RF filters using positive-intrinsic-negative (PIN)-diode switches (and their associated drive electronics) for tuning. The small size comes with no sacrifice in performance. Characteristics of the MEMS-based filter include low insertion loss and performance comparable to or better than standard RF filters tuned with PIN-diode switches. MEMS relays have the significant advantage of not exhibiting the nonlinearity associated with PIN diodes and should offer much improved intermodulation (IM) performance under high-power conditions.

"These characteristics demonstrate that the MEMS-based RF filter can result in improved design and performance for telecommunications applications," says Ramaswamy Mahadevan, a Cronos researcher involved in the development of the MEMS relay used in the RF filter. "An RF filter using MEMS switches instead of PIN-diode switches not only meets the necessary performance specifications with a smaller footprint, it actually offers improvement in some areas," he says. "Total insertion loss and power loss are lower than some of the best RF filtering devices available today."

Kudos

Sonoma Scientific, Inc. has been awarded a Certificate of Compliance for ISO-9000 certification as an ISO-9001 company...Motorola has officially received QS-9000 certification of its Compound Semiconductor-1 wafer fabrication facility (CS-1). All of Motorola's gallium-arsenide (GaAs) semiconductor devices are fabricated in the CS-1 facility...RF Monolithics, Inc. has accepted the InfoVision99 Award for the TR1000 916.50-MHz hybrid transceiver. This honor, awarded by the International Engineering Consortium (IEC), recognizes technologies, applications, products, and services judged to be the most unique and beneficial to the industry.

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MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMPR. (dBm, Min.)	Price	DC POWER @ +15 V (mA, Nom.)
OCTAVE BAND AMPLIFIERS									
JS2-00500100-035-5A	0.5 – 1	35	1	0.45*	2:1	2:1	5		250
JS2-00500100-10-5A	0.5 – 1	35	1.2	1	2:1	2:1	5		250
JS2-01000200-035-5A	1 – 2	33	1	0.45*	2:1	2:1	5		250
JS2-01000200-10-5A	**1 – 2**	**33**	**1.2**	**1**	**2:1**	**2:1**	**5**	**$395**	**250**
JS2-02000400-035-5A	2 – 4	28	1	0.45*	2:1	2:1	5		175
JS2-02000400-10-5A	**2 – 4**	**28**	**1.2**	**1**	**2:1**	**2:1**	**5**	**$450**	**175**
JS2-04000800-070-0A	4 – 8	22	1	0.7	2:1	2:1	0		150
JS2-04000800-15-0A	**4 – 8**	**22**	**1.2**	**1.5**	**2:1**	**2:1**	**0**	**$495**	**150**
JS3-04000800-060-5A	4 – 8	30	1	0.6	2:1	2:1	5		175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5		175
JS2-08001200-09-5A	8 – 12	15	1	0.9	2:1	2:1	5		150
JS2-08001200-15-5A	**8 – 12**	**15**	**1.2**	**1.5**	**2:1**	**2:1**	**5**	**$495**	**150**
JS3-08001200-080-5A	8 – 12	25	1	0.8	2:1	2:1	5		175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5		175
JS2-12001800-16-5A	12 – 18	15	1	1.6	2:1	2:1	5		100
JS2-12001800-30-5A	**12 – 18**	**15**	**1.5**	**3**	**2:1**	**2:1**	**5**	**$495**	**100**
JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5		175
JS3-12001800-30-5A	**12 – 18**	**23**	**1**	**3**	**2:1**	**2:1**	**5**	**$495**	**175**
JS4-12001800-12-5A	12 – 18	30	1	1.2	2:1	2:1	5		200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5		200
JS2-18002600-20-5A	18 – 26	14	1	2	2:1	2:1	5		100
JS2-18002600-30-5A	18 – 26	14	1	3	2:1	2:1	5		100
JS3-18002600-20-5A	18 – 26	22	1	2	2:1	2:1	5		175
JS3-18002600-30-5A	18 – 26	22	1	3	2:1	2:1	5		175
JS4-18002600-16-5A	18 – 26	27	1	1.6	2:1	2:1	5		200
JS4-18002600-26-5A	18 – 26	27	1	2.6	2:1	2:1	5		200
JS2-26004000-35-5A	26 – 40	12	2	3.5	2:1	2:1	5		100
JS2-26004000-45-5A	26 – 40	12	2	4.5	2:1	2:1	5		100
JS3-26004000-35-5A	26 – 40	18	2	3.5	2.5:1	2.5:1	8		175
JS3-26004000-45-5A	26 – 40	18	2	4.5	2.5:1	2.5:1	8		175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	8		200
JS2-26004000-100-20A	26 – 40	17	1.25	10	2.3:1	2.3:1	20		**
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0		175

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMPR. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS								
JS2-00500200-05-5A	0.5 – 2	32	1	0.5	2:1	2:1	5	250
JS2-00500200-20-5A	0.5 – 2	32	1	2	2:1	2:1	5	250
JS2-01000400-07-5A	1 – 4	27	1	0.7	2:1	2:1	5	200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5	200
JS2-02000600-07-5A	2 – 6	24	1	0.7	2:1	2:1	5	125
JS2-02000600-20-5A	2 – 6	20	1	2	2:1	2:1	5	125
JS2-02000800-08-0A	2 – 8	22	1	0.8	2:1	2:1	0	125
JS2-02000800-20-0A	2 – 8	18	1	2	2:1	2:1	0	125
JS3-02001800-25-5A	2 – 18	21	2	2.5	2.5:1	2.5:1	5	150
JS3-02001800-50-5A	2 – 18	21	2	5	2.5:1	2.5:1	5	150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5	200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5	200
JS3-02002600-30-5A	2 – 26	21	2	3	2:1	2:1	5	150
JS3-02002600-40-5A	2 – 26	21	2	4	2:1	2:1	5	150
JS3-06001800-18-5A	6 – 18	23	1.3	1.8	2:1	2:1	5	125
JS3-06001800-30-5A	6 – 18	23	1.3	3	2:1	2:1	5	125
JS4-06001800-135-5A	6 – 18	31	1	1.35	2:1	2:1	5	200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5	200

* Noise figures to 0.35 dB available on a limited basis.
** This unit requires +8V @ 500 mA and -8V @ 90 mA.

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MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMPR. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS (Continued)								
JS3-08001800-17-5A	8 – 18	24	1.2	1.7	2:1	2:1	5	125
JS3-08001800-30-5A	8 – 18	24	1.2	3	2:1	2:1	5	125
JS4-08001800-13-5A	8 – 18	32	1.5	1.3	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	1.5	3	2:1	2:1	5	200
JS3-08002600-30-5A	8 – 26	21	2	3	2:1	2:1	5	150
JS3-08002600-40-5A	8 – 26	21	2	4	2:1	2:1	5	150
JS3-12002600-25-5A	12 – 26	22	2	2.5	2:1	2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2	3.5	2:1	2:1	5	150
JS4-12002600-22-5A	12 – 26	30	1.7	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	30	1.7	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-06-5A	0.1 – 2	32	1	0.6	2:1	2:1	5	250
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	250
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-18-5A	0.1 – 10	26	1.5	1.8	2:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	26	1.5	3.5	2:1	2:1	5	150
JS3-00101200-19-5A	0.1 – 12	25	1.5	1.9	2:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	25	1.5	3.5	2:1	2:1	5	150
JS3-00101800-26-5A	0.1 – 18	23	1.5	2.6	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.5	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-32-5A	0.1 – 26	20	1.8	3.2	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	1.8	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2	5	2.5:1	2.5:1	5	200
JS4-00103000-35-5A	0.1 – 30	20	2.5	3.5	2.5:1	2.5:1	5	200
JS4-00103000-45-5A	0.1 – 30	20	2.5	4.5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	3.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	3.5	8.5	2.75:1	2.75:1	5	200

NOTE: Higher 1 dB compression levels are available on many designs.

For additional information or technical support, please contact either Rosalie DeSousa at (516) 439-9458, e-mail rdesousa@miteq.com or Rizwan Syed at (516) 439-9267, e-mail rsyed@miteq.com.

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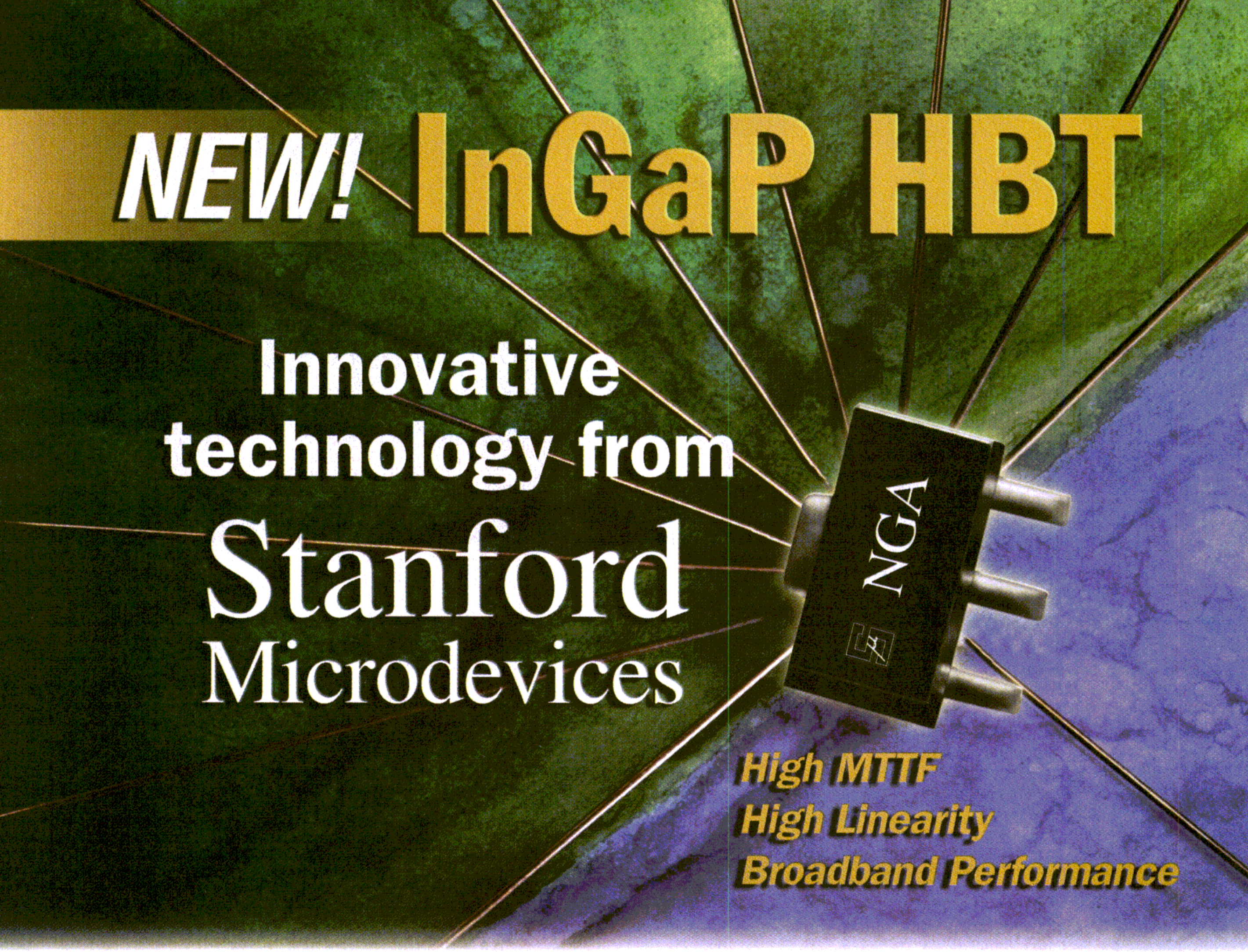

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NGA-489 DC-8 GHz

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

SPECIFICATION MATRIX

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

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

NGA-589 DC-6 GHz

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

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

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Alliance Provides Telephone-Based Emergency Warning System

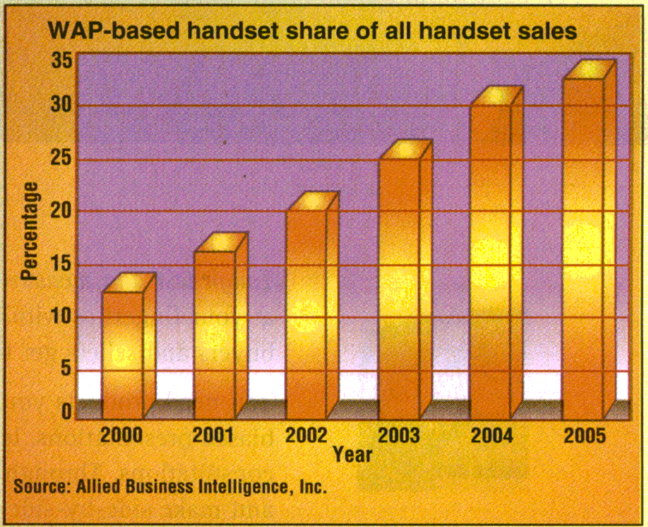
BOULDER, CO—SCC Communications, in alliance with U S WEST, has announced the nation's first advanced telephone-based emergency warning service—U S WEST Emergency Preparedness Network (EPN). Counties and cities with the service will have the ability to deliver vital information to residents through the telephone during life-threatening crises such as natural disasters, chemical spills, or hostage situations.

On October 13, 1999, SCC and U S WEST signed an agreement that incorporates SCC's Early Warning and Evacuation℠ (EWE) service with U S WEST technology to develop and market EPN. According to George Heinrichs, CEO at SCC Communications, "We are very pleased to announce our expanded relationship with U S WEST to provide EPN throughout the U S WEST region. Our technology in this area provides an important new tool for local public-safety officials to help citizens in times of crisis."

EPN is able to inform citizens of impending dangers within minutes. In the event of a natural disaster, wild fire, chemical spill, or hostage crisis, authorities can activate EPN to call residents in a designated population. The pre-recorded message may then provide details about the situation and potentially life-saving instructions. Thousands of calls can be made within minutes using the U S WEST data base of geographically coded telephone numbers—published and unpublished. The data base is strictly controlled and the telephone numbers are made available only for emergency notification. U S WEST EPN can be programmed to leave messages on answering machines, call numbers back that are not answered, and retry busy numbers.

One-Third Of All New Handsets To Be WAP-Enabled

OYSTER BAY, NY—The movement to Internet-based services will provide another lucrative market for handset manufacturers and component producers as wireless-access-protocol (WAP)-based handsets account for one-third of all user terminals by 2005, according to new findings from Allied Business Intelligence (ABI).



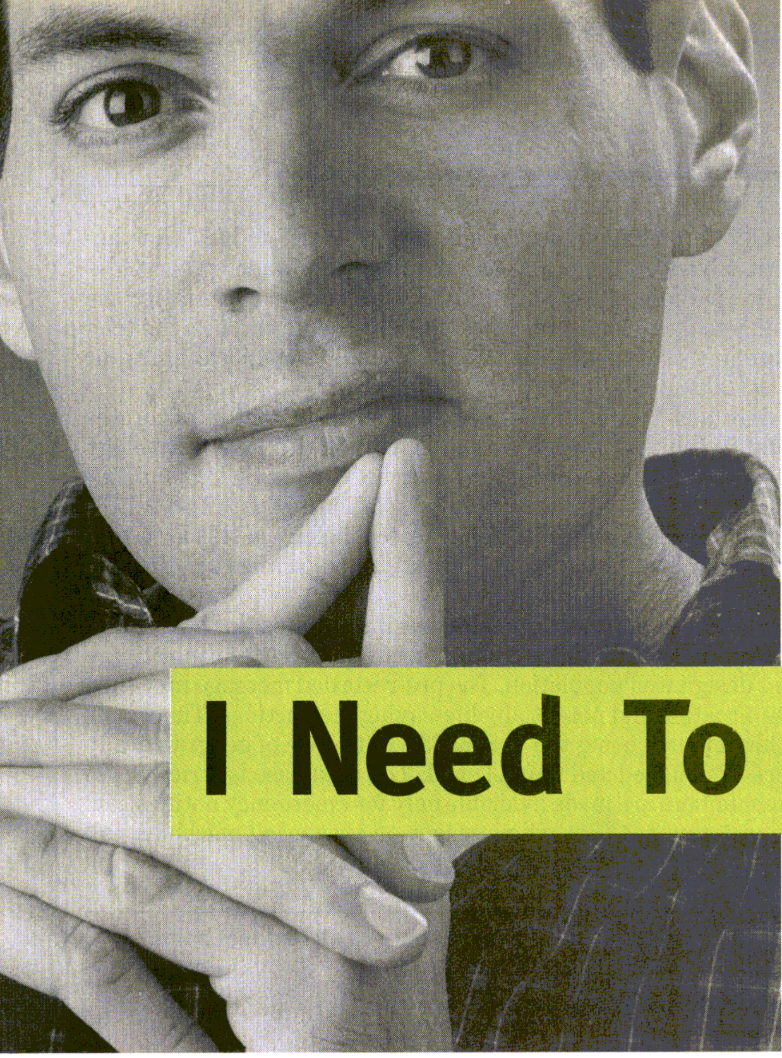
There will be more than 600 million WAP-based handsets shipped from 2000 to 2005. WAP-based handsets will grow from 12 percent of all handsets produced in 2000 to 33 percent in 2005 (see figure) as wireless operators begin to define data offerings which will suit the user. "It's just beginning to get exciting. We have already seen some early movement in Japan and Western Europe which we believe demonstrates thirst for simple data access," says report author Larry Swasey, vice president of communications research at ABI. "The only real hurdle may be the overselling of these services as something they are not," Swasey adds.

Handset and device makers will still be able to realize growth in other handset market segments as well, according to the report "Wireless Access Devices." The digital/dual-mode handset marketplace will have a compound annual average growth (CAAG) of 29 percent from 2000 to 2005.

Analog-only handset users will dwindle from 34 percent of all users in 1998 to 5 percent in 2005.

Complementing the wireless movement are the wireless personal/commercial vehicle-access markets, which comprise fleet-management systems (FMS), in-vehicle navigation systems (IVNS), and in-vehicle information systems (IVIS).

"Wireless Access Devices" covers the global market for WAP-based, dual-mode, analog, and digital handsets.



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University of Houston Unveils State-Of-The-Art Laboratory

HOUSTON, TX—The University of Houston's Department of Electrical and Computer Engineering unveiled a state-of-the-art telecommunications networks laboratory recently. The lab will support students who are working on a Master of Electrical Engineering degree with a specialization in telecommunications.

"The lab, which is still expanding, is built around the high-speed backbone that is central to modern wide-area networks. It provides the hands-on experience that our students need to be successful in the telecommunications industry," says Fritz Claydon, chair of the Department of Electrical and Computer Engineering.

Claydon stresses that the lab would not have been possible without the support of area telecommunications companies. The lab was designed in collaboration with Southwestern Bell and contains equipment donated by several companies, including ADC Communications; Fujitsu Network Communications, Inc.; and Lucent Technologies.

The telecommunications program embraces a wide variety of areas from analog and digital telephony to high-speed digital data networks. It supplements engineering instruction with courses on the regulatory, legal, and management issues in the fast-changing industry.

iForums Present Education Over The Internet

CHICAGO, IL—The International Engineering Consortium (IEC) has unveiled plans for a new method of providing continuing education to the information industry that is unlike anything currently available. Christened the "iForums," the program offers private, classroom-style instruction over the Internet through the Consortium's website (http://www.iec.org/iforums).

The iForums provide all of the elements of a traditional classroom-style setting, including lectures from a university professor, one-on-one virtual discussions with the professor, as well as question-and-answer sessions with fellow students. The iForums are available on a round-the-clock basis, making them very convenient.

"In today's always-connected environment, professionals must receive high-quality information on their own timetables and at their own convenience," says Roger Plummer, IEC managing director. "To accommodate this need, the IEC developed iForums to provide today's busy professionals a compelling alternative to courses presented in a traditional classroom setting. The iForums will offer students a unique combination of video, text, and graphics that will accelerate learning."

Registrants can log onto iForums from work, home, or on the road. Once the student is logged in, he or she will choose lectures from an overview menu. When a selection has been made, the student then "attends" the lecture whenever he or she wishes. Later, the student can choose whether to review each lecture. To complete the course and gain accreditation, the student completes a short exam. Typically, students will need between 90 minutes and two hours each week to review the lecture, illustrations, and supplementary reading materials.

New Atomic Clock Placed Into Operation

BOULDER, CO—The Commerce Department's National Institute of Standards and Technology (NIST) has placed a new atomic clock into operation that will neither gain nor lose a second in approximately 20 million years.

Termed NIST F-1, the new cesium atomic clock at NIST's Boulder, CO laboratories began its role as the nation's primary frequency standard by contributing to an international pool of the world's atomic clocks that is used to define Coordinated Universal Time (known as UTC), the official world time. Since NIST F-1 shares the distinction of being the most accurate clock in the world (with a similar device in Paris), it is making UTC more accurate than ever before.

NIST f-1 recently passed the evaluation tests that demonstrated it is approximately three times more accurate than the atomic clock that it replaces—NIST-7, which is also located at the Boulder facility. NIST-7 has been the primary atomic time standard for the US since 1993 and is among the best time standards in the world.

The NIST F-1 clock's method of resolving time differs greatly from that of the NIST-7. That device—and the versions before it—fired heated cesium atoms horizontally through a microwave cavity at high speed. NIST F-1's cooler and slower atoms allow more time for the microwaves to "interrogate" the atoms.

The show's exhibitor booths, technical papers, and workshops showcased products and technologies for present and future applications.

Wireless Symposium Heralds Innovations

DON KELLER
Senior Editor

JACK BROWNE
Publisher/Editor

WIRELESS technology has been with us since the days of Marconi, and has drastically changed our communications over the years. In the midst of the information age, those changes are still taking place, and the simple cellular telephone of the past decade is quickly being replaced by multifunction devices that promise voice communications, data communications, and remote Internet access. Certainly the industry that has grown up around wireless technology is still going strong, as evidenced by attendance at the recent Eighth Annual Wireless Symposium/Portable By Design Conference & Exhibition. More than 400 exhibitors and over 8000 attendees traveled to the San Jose McEnery Convention Center (San Jose, CA) during the week of February 22-25, 2000 to learn more about the latest wireless products, trends, and technologies.

It became apparent from visitors and exhibitors at the Wireless Symposium that cellular communications is just one part of the growing wireless industry. Increased growth is expected from areas such as wireless local-area networks (WLANs), satellite communications, and Bluetooth/HomeRF applications.

The show floor was alive with new products aimed at the design of wireless handsets and base stations. Motorola (Phoenix, AZ), for example, introduced a single-chip, dual-band frequency downconverter, the MC13740, for use in time-division-multiple-access (TDMA) Global System for Mobile Communication (GSM) cellular phones that operate to 2 GHz (Fig. 1). The chip integrates two complete downconverters in a single, 24-pin, plastic, thin-quad-flat-pack (TQFP) housing.

Another product for GSM applications, exhibited by Analog Devices (Wilmington, MA), was the AD20mps430 SoftFone chip set. The random-access-memory (RAM)-based integrated circuits (ICs) allow GSM cellular-phone manufacturers to customize features and options entirely in software, and to support an entire family of low-end and high-end handsets simply by loading different software versions. The chip set is comprised of two ICs—the AD6522 digital baseband processor and the AD6521 baseband converter.

Synergy Microwave Corp. (Paterson, NJ) introduced a new C-band, double-balanced mixer and a new line of in-phase/quadrature (I/Q) modulator/demodulators. The mixer, housed in a 0.3 × 0.2 × 0.1-in. (0.762 × 0.508 × 0.254-cm) surface-mount leadless package, operates from 3.6 to 4.8 GHz. The model VMS-935 modulator/demodulator operates from 935 to 960 MHz. The model VMS-1710 modulator/demodulator operates from 1710 to 1780 MHz. The modulator/demodulators are housed in packages measuring 0.5 × 0.5 × 0.22 in. (1.27 × 1.27 × 0.56 cm).

Hittite Microwave Corp. (Woburn, MA) introduced no less than 16 new monolithic-microwave-integrated-circuit (MMIC) products at the show. Products applicable to handsets include five new positive-bias, multithrow switches with onboard, transistor-transistor-logic (TTL) decoder/drivers covering frequencies to 6 GHz.

ANADIGICS (Warren, NJ) introduced a new line of pseudomorphic-high-electron-mobility-transistor (PHEMT) single-pole, four-throw (SP4T) antenna switches with internal decoders. The gallium-arsenide (GaAs) switches, which operate on a single +3-VDC power source, combine logic control and RF-band-and-mode selection in one device to simplify digital and analog multimode, multiband, cellular handset design.

EiC Corp. (Fremont, CA) added to its line of GaAs ICs with five components (models EC-1019, EC-1119, EC-1078, EC-1178, and EC-1089) based on indium-gallium-phosphide (InGaP)-on-GaAs heterojunction-bipolar-transistor (HBT) technology. The gain-block amplifiers offer fixed-gain levels from 13 to 20 dB at cellular and PCS frequencies with 1-dB compressed output-power levels of +18 to +21 dBm.

California Eastern Laboratories (Santa Clara, CA) announced six new products for handset applications. The model NE52118 GaAs heterojunction bipolar field-effect transistor (HBT FET) low-noise amplifier (LNA) operates from a single, positive-voltage power source. It is designed for use in TDMA handsets and other L-band and S-band receivers. Four new GaAs driver amplifiers operate at low voltage and low current and feature low adjacent-channel-power-ratio (ACPR) for digital cellular TDMA applications. The sixth product is a low-loss GaAs SPDT switch that operates from 0.3 to 2.5 GHz.

Several exhibitors unveiled new or updated oscillators, resonators, synthesizers, and phase-locked loops (PLLs) for handset applications. Vectron International (Norwalk, CT) updated its line of V-type voltage-controlled crystal oscillators (VCXOs) with models . that now reach 77.76 MHz. Piezo Technology, Inc. (Orlando, FL) also extended the frequency range of its temperature-controlled crystal oscillators (TCXOs) and VCXOs. The model XO3080 is now available at frequencies to 125 MHz. Micronetics Wireless (Hudson, NH) introduced the model MW500-1089 voltage-controlled oscillator (VCO), which generates +8-dBm signals from 4 to 6 GHz.

Philsar Semiconductor (Nepan, Ontario, Canada) introduced a trio of fractional-N frequency synthesizers for cellular and other wireless applications. Featuring fine resolution (100 Hz) and fast switching speeds, the models PS-1200, PS-2500, and PS-6500 synthesizers are usable in applications to 1.9 GHz. Fabricated with a high-speed silicon-germanium (SiGe) process, the ICs are ideal for cellular-telephone, WLAN, and wireless-local-loop (WLL) applications. In addition, during the opening sessions of the Wireless Symposium, Philsar's Mike O'Neill addressed the company's deep involvement with the development of ICs and assemblies for Bluetooth personal wireless-connectivity applications at 2.4 GHz.

Micro Lambda, Inc. (Fremont, CA) displayed its MLSO series of YIG-based tunable oscillators. With an integral transistor-transistor-logic (TTL) driver, the YIG sources provide tuning steps for 50 to 200 MHz over a total frequency range of 2 to 20 GHz. Models are available for output-power levels of +12 to +20 dBm.

MTI-Milliren Technologies, Inc. (Newburyport, MA) offered its 220 series dual-in-line-package (DIP) oven-controlled crystal oscillator (OCXO), available with an AT- or SC-cut crystal is a hermetic package. With a warmup time of less than 5 min., the miniature OCXO features an outstanding phase noise of −95 dBc/Hz offset 1 Hz from the carrier with a noise floor of −155 dBc.

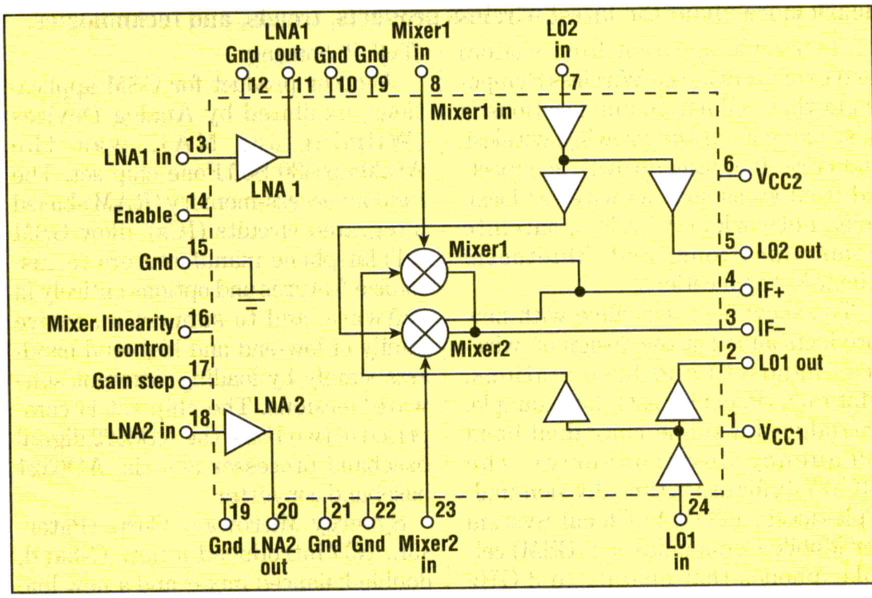
A significant number of new products for base-station applications also made their debut at the show. One product in this category that drew much attention was a new DSP introduced by Texas Instruments (Dallas, TX). The model TM320C64x has speeds that are scalable to 1.1 GHz while processing approximately 9000 million instructions per second (MIPS).

Anaren Microwave (East Syracuse, NY) announced the availability of a high-power Xinger 3-dB coupler for DCS, PCS, and Universal Mobile Telecommunications Services (UMTS) applications from 1.8 to 2.5 GHz. With better than 20-dB return loss and maximum insertion loss of 0.23 dB, the hybrid coupler can handle 100-W power when used in digital-audio-broadcast (DAB) applications at 2.4 GHz.

Stanford Microdevices (Sunnyvale, CA) introduced its model SL-1010 10-W laterally-diffused-metal-oxide-semiconductor (LDMOS) transistor for use in base stations to 1 GHz. With linearity suitable for TDMA and CDMA applications, the transistor yields a typical linear gain of 15.5 dB with 45-percent drain efficiency and −33 dBc two-tone intermodulation distortion (IMD). The firm also introduced the SL-1020 LDMOS device, with similar performance for applications to 2 GHz.

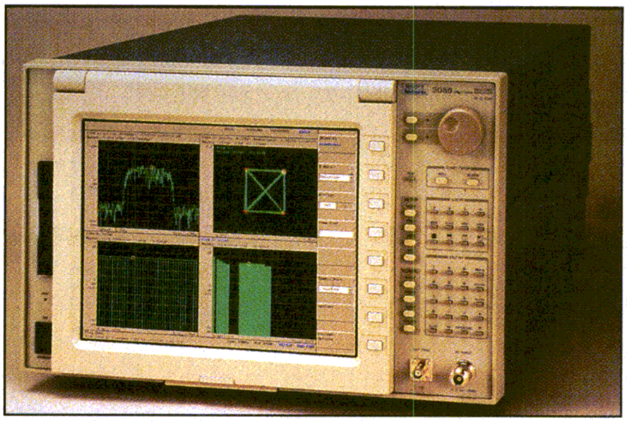
MicroWave Technology (Fremont, CA) featured a high-gain, medium-power PHEMT device for applications from 0.5 to 32 GHz. The model MwT-H7LN is designed to produce +20-dBm output power at 12 GHz with 13-dB gain. Available in chip and packaged configurations, the transistor features a low-noise figure of 1.5 dB at 12 GHz with 11-dB associated gain.

The early effects of Bluetooth were apparent throughout the show floor as several IC suppliers displayed products aimed at the new 2.4-GHz wireless-connectivity standard. One of the early adopters, National Semiconductor (Santa Clara, CA), for example, displayed their USB/UART reference



1. Model MC13740 is a dual-band downconverter for use in GSM mobile-telephone handsets operating to 2 GHz. (Photograph courtesy of Motorola, Phoenix, AZ.)

design kit, based on the firm's proven LMX3162 Bluetooth radio IC. The kit, which includes samples of the LMX3162, the LMX5001 Bluetooth link controller, baseband firmware that conforms with Version 1.0 of the Bluetooth standard, Gerber files for a 2.4-GHz radio printed-circuit board (PCB), and a control software development program, allows product developers to get to market quickly with a Bluetooth design capable of operating over distances as great as 100 m.



2. Improvements to the model 3086 real-time spectrum analyzer include code-domain power-measurement capability and built-in complementary-cumulative-distribution-function (CCDF) measurement capability. (Photograph courtesy of Tektronix, Inc., Beaverton, OR.)

TEMIC Semiconductors (St. Quentin-en-Yvelines, France) also featured a Bluetooth product, with its model T2901 single-chip transceiver. Operating with supply voltages of +2.7 to +3.3 VDC, the transceiver IC supports gross data rates of 1 Mb/s over the unlicensed 2.4-GHz band using the Bluetooth time-division-duplex (TDD) frequency-hopping scheme.

A design kit for lower-frequency applications, at very-high frequency (VHF) and ultra-high frequency (UHF), was announced by RF Micro Devices (Greensboro, NC). The firm's series of reference designs targets VHF/UHF applications in remote-keyless-entry (RKE) systems, wireless security systems, and RF remote controls. The kits include a transmitter based on the company's RF2516 IC and receiver based on the model RF2919 IC.

Renaissance Electronics Corp. (Boxborough,MA) features a wireless transceiver designed for Internet, data, and voice communications in the MMDS band. Operating at transmit frequencies of 2550 to 2610 MHz and receive frequencies of 2640 to 2686 MHz, the transceiver maintains harmonic and spurious levels of −55 dBc. It measures only 6.00 × 6.00 × 1.25 in. (15.24 × 15.24 × 3.175 cm) and draws 850-mA current from a +12-VDC supply.

Micrel Semiconductor (San Jose, CA) was one of several fiber-optic product manufacturers in attendance on the Wireless Exhibition floor (p. 17). The firm displayed their SY889XX family of fiber-optic post amplifiers and laser drivers in support of optical applications at 1.25 and 2.50 Gb/s. Operating at +3.3 VDC, the ICs are suitable for Gigabit Ethernet, Fibre Channel, OC-12, and OC-48 optical communications systems.

A wide range of materials suppliers was also in force at the Wireless Symposium & Exhibition. Rogers Corp. (Rogers, CT) spotlighted its RO3000 and RO4000 series of microwave laminates for low-loss microwave and wireless applications as well as its Ultralam 1000 series of woven-glass-reinforced laminates for matching the dielectric and loss-tangent characteristics of other woven-glass polytetrafluroethy-

SEVEN WIRELESS WONDERS

Wireless circuit and systems designers labor long hours to develop the hardware, software, and test equipment used to create wireless products, such as cellular telephones. And while the Wireless Symposium/Portable By Design Conference & Exhibition has served the wireless system designer for eight years as a key educational event, it is only now that the end products of wireless design have been acknowledged at the show. The event, held February 22nd, was an awards ceremony for the Seven Wireless Wonders of 1999. Sponsored by the new wireless e-commerce website, http://www.wirelesswonders.com (based in San Jose, CA), and The Wireless Symposium, the awards ceremony drew a large crowd eager to learn about the seven outstanding wireless consumer products of 1999.

Voting for the top products was conducted online at http://www.wirelesswonders. The website's CEO, Dr. Rahat Khanna, presided over the presentation of the awards, made to some of the leading commercial wireless equipment manufacturers in the world. For example, Ericsson won for a wireless headset based on Bluetooth technology at 2.4 GHz. while Siemens took a prize for its Gigaset 2420 cordless-telephone system, also operating at 2.4 GHz. Additional winners included 3Com for their Palm VII personal digital assistant (PDA), Motorola for their i1000plus time-division-multiple-access (TDMA) cellular-telephone/pager/two-way radio, Nokia for their 9000il cellular telephone/PDA, NeoPoint for their NP1000 PDA/Internet browser, and Samsung for their amazing Watch Phone, a fully functional code-division-multiple-access (CDMA) cellular telephone in the form of a watch.

According to Dr. Khanna, "Wireless Wonders has started this annual event to enable wireless consumers to participate in identifying the seven most exciting consumer wireless products each year by voting for their favorite product." The criteria for voting included novelty in terms of functionality, integration or miniaturization impact on productivity, and ease of use.

lene (PTFE) substrates.

Jack Gear, Marketing Manager of Emerson & Cuming Microwave Products (Randolph, MA), offered applications advice on the firm's ECCOSORB AN HR series of carbon-loaded foam-microwave absorber materials. The lightweight materials, which are ideal for reducing the back radiation of shrouded antennas, lining shielded antenna test boxes, and reducing RF interference (RFI) around antenna feeds, can be supplied in preconfigured shapes to meet specific customer requirements.

TEST DEVELOPMENTS

Several new and improved test instruments and systems were on the Wireless Show floor, including an enhanced version of the model 3086 real-time spectrum analyzer from Tektronix, Inc. (Fig. 2). The analyzer, which essentially digitizes an entire 5-MHz slice of bandwidth, has been improved with the addition of code-domain power-measurement capability and built-in complementary-cumulative-distribution-function (CCDF) measurement capability.

The dual-band VX19000 series programmable noise generator from Noise Com (Paramus, NJ) can switch between two different frequency bands in support of carrier-to-noise and bit-error-rate (BER) testing. The VXI-based generator, which is ideal for automatic-test-equipment (ATE) applications, is available from 10 Hz to 40 GHz.

Agilent Technologies (Santa Rosa, CA), dominant on the show floor with a massive booth, also loomed large in exhibition news with the introduction of the wireless industry's first wideband vector signal analyzer (VSA). The 89600 series VSA, with 36-MHz bandwidth capacity, analyzes complex (I and Q) signals to 2.7 GHz.

Anritsu Co. (Morgan Hill, CA) launched its model ML2530A precision measurement receiver at the Wireless Symposium and Exhibition. With a frequency range of 100 kHz to 3 GHz, the measurement receiver is ideal for calibrating attenuation levels and output levels in frequency synthesizers and signal generators. It features a resolution bandwidth of 1 Hz and measures power levels from −140 to +20 dBm.

One of the more intriguing instruments to debut at the Wireless Symposium & Exhibition was from Celerity Systems (Cupertino, CA). The firm's CS29010 distortion measurement test set is based on a 600-MHz Pentium III computer architecture running with customized LabVIEW measurement and control software. It can be configured with field-replacement modules for measurements from baseband to 40 GHz at instantaneous measurement bandwidths to 40 MHz.

New software products were also in abundance at the Wireless Symposium and Exhibition. In addition to the aforementioned Agilent ADS, Elanix, Inc. announced the release of its SystemView wireless design suite, a soft-

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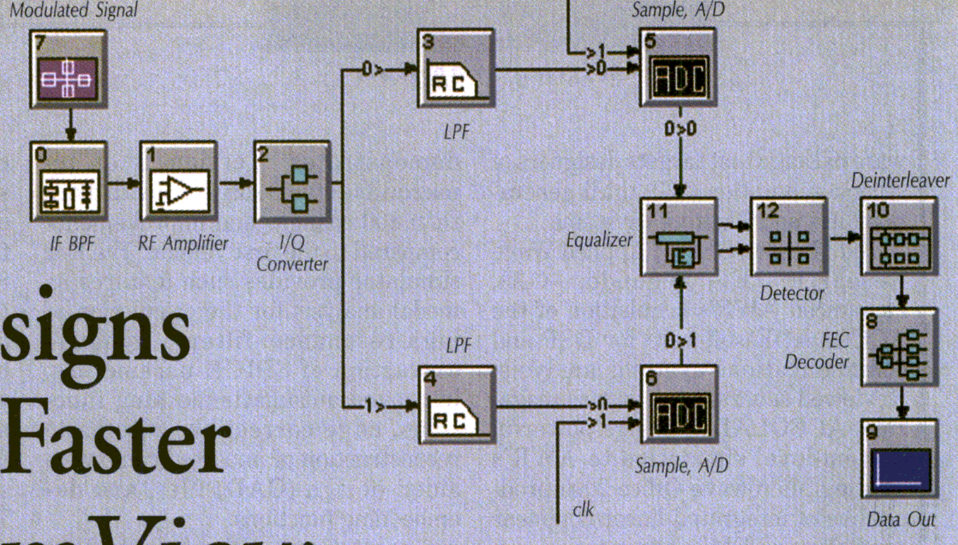

Mixed-Mode Receiver System Simulation

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Design and test digital/analog communications systems such as CDMA, GSM, DVB, QPSK, and QAM. Includes your choice of error correcting encoders/decoders and modulators/demodulators.

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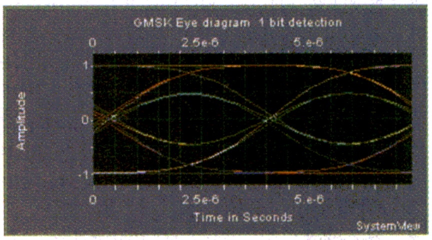
Supports all major fixed and floating-point processors. Simulates accurately to the bit level. Generates overflow, underflow and carry flags.

■ Distortion-True RF/Analog Design

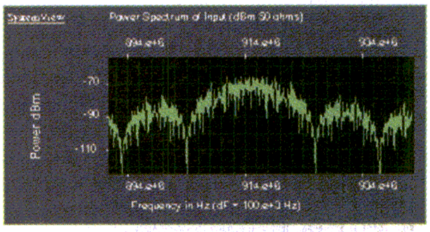
Includes circuit components such as distortion-true mixers, amplifiers, RLC circuits, opamp circuits, etc. Allows creation of complete TX/RX systems, including propagated noise figure.

■ Interface to MATLAB®

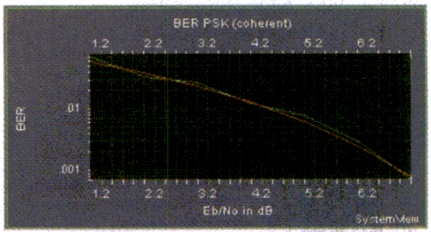
Use the new SystemView M-link™ option to incorporate MATLAB, Simulink® and third party libraries and systems into your SystemView design.



GMSK Eye Diagram (1-bit Detection)



Input Power Spectrum



Theoretical vs. Actual BER (Coherent PSK)

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What They're Saying About SystemView

"No other program with so much simulation power is as easy to use. When you finish a session with SystemView you realize that you've spent all the time applying engineering principles and none of the time struggling with the computer."
- Richard Chick, MIT Lincoln Labs -

"We use SystemView very heavily, not only in the brainstorming/prototyping stage, but also throughout the development process — right up to the point where we're actually building silicon."
- Steven Hall, CommQuest Technologies -

"For digital filter design, SystemView was indispensable due to its accurate simulation of the fixed point arithmetic mode employed on the actual DSP processor used in my design."
- Bisla Balvinder, Itron -

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CIRCLE NO. 368

ware package that targets designers of second-generation (2G), third-generation (3G), and Bluetooth systems .

Ted Miracco, CEO of Applied Wave Research (El Segundo, CA), announced AWR's acquisition of the ACCOLADE software for DSP and communications systems analysis. Reviewed several times in this magazine, ACCOLADE, brings powerful system-level simulation to AWR's existing Microwave Office 2000 product line of integrated linear/nonlinear simulator and EM-analysis tools.

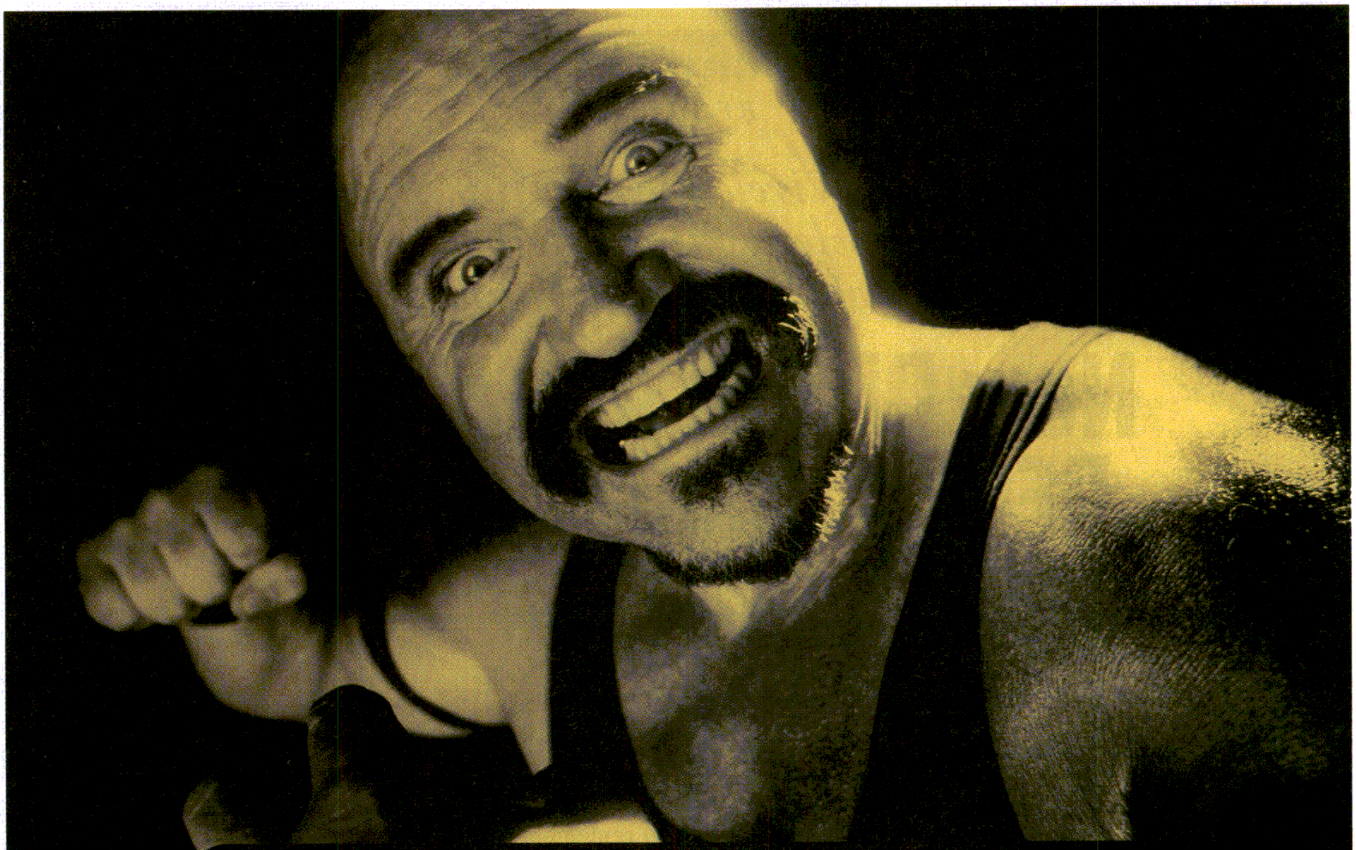
Eagleware's engineering team was in force to demonstrate Version 7 of their GENESYS computer-aided-engineering (CAE) software suite. Offering circuit simulation, synthesis, electromagnetic (EM) simulation, schematic entry, layout capabilities, and artwork generation within a single integrated framework, the GENESYS suite features an intuitive graphical user interface well matched to Windows 95/98/NT.

CST of America (Cambridge, MA) demonstrated Version 2 of its microwave studio software, with more than 800 changes and improvements compared to the first version. The EM simulator provides such features as modal analysis for the calculation of high-resonance filters, dynamic extraction of SPICE parameters, improved automatic meshing functions, edge correction, automatic reconstruction of imported computer-aided-design (CAD) files, and de-embedding functions.

Ansoft Corp. (Pittsburgh, PA) announced several new software releases, including version 7.0 of the company's well-known high-frequency structure simulator (HFSS) software, a three-dimensional (3D) EM simulation and analysis tool; Optimetrics, a parametrics, sensitivity, and optimization module for HFSS; version 6.1 of Ensemble, a planar EM simulator; and version 8.5 of Serenade, a combination circuit and system simulation tool.

In additional software news, Sonnet Software (Liverpool, NY, made free copies available of their Sonnet Lite software, a full-featured EM simulator designed for analysis of planar high-frequency structures. Xpedion Design Systems (Santa Clara, CA) discussed the use of their GoldenGate RF/microwave simulator, a tool that combines linear and nonlinear circuit-analysis capabilities. Also, DSP Development Corp.(Cambridge, MA) offered free copies of DADiSP, the firm's DSP simulation and analysis software. These fully functional copies were supplied with a 30-day time limit. Finally, COMNAV Engineering (Portland, ME) offered free copies of the "Dr. Rez" filter design and analysis software. The single-disk tool simplifies the design of dielectric and coaxial resonator filters.

The Ninth Annual Wireless Symposium/Portable By Design Conference & Exhibition is scheduled for next February at the San Jose Convention Center. For more information, visit the show website at http://www.WirelessPortable.com.●●



The start of the Wireless Symposium signaled the start of a new electronics distributor based on an old and trusted name.

New Distributor Aims At RF And Fiber Solutions

JACK BROWNE
Publisher/Editor

WYLE Electronics is a name known and respected throughout the world of semiconductors. To those old enough to remember, the company was founded in 1949 as an independent test laboratory known as Wyle Laboratories. The firm entered electronic distribution in the 1960s and took on the name Wyle Electronics in 1995. Now in 2000, the company is taking on distribution of RF/microwave and fiber-optic components through its newly formed subsidiary RF Vision.

Representatives of the new company announced their business intentions at the recent Wireless Symposium/Portable By Design Conference & Exhibition in San Jose, CA (San Jose Convention Center, February 22-25, 2000). Addressing a crowd of trade-press members and curious engineers, the new firm's new president and chief executive officer (CEO), Ken Wadors, outlined a mission that would establish RF Vision as the leading RF/microwave and fiber-optic distributor in the electronics industry. Wadors (see figure), with more than 30 years of engineering, sales, and general management experience, is no stranger to electronic distribution, coming to RF Vision from a position as vice president and director of the RF/small-signal business unit of Avnet, Inc. (San Jose, CA).

Ken Wadors brings more than 30 years of electronics engineering and management experience to his new position as president and CEO of newly formed RF/microwave and fiber-optic distributor RF Vision.

According to Wadors, "RF Vision was created to bring the Wyle model of superior demand creation and value-added fulfillment to the wireless communications segment of the electronic market. RF Vision will offer an unmatched level of engineering expertise to its customers and will work as a design team member to ensure that the best technical solution is offered for every design requirement."

The new RF/microwave and fiber-optic distributor is part of the larger Wyle Electronics which, in turn, is part of the even larger VEBA Electronics Group (Santa Clara, CA), one of the world's fastest-growing distributors of semiconductors and computer products. With more than $5 billion in global sales during 1999, employment in the VEBA Electronics Group exceeded 6000 people worldwide. So the force behind the force behind RF Vision is considerable.

RF Vision has a charter to focus on five key areas of electronic distribution, areas not typically well-served by traditional distributors: RF small-signal components, RF passive components, RF power devices, RF interconnections, and fiber-optic components. These are among the components needed by wireless and communications designers to complete current and future systems-level designs.

Within these product areas, RF Vision will bring a host of value-added services designed to ensure that customers receive products in the manner and form that they need for their specific requirements. These value-added services include parametric testing (according to specific customer requirements), solder dipping, screening, tape-and-reel packaging, product qualification, and special cabling.

For more information, contact: RF Vision, a subsidiary of Wyle Electronics, 3000 Bowers Ave., Santa Clara, CA 95051; (877) 450-4441, FAX: (408) 986-1269, Internet: http://www.rfvision.com.

"We already make ~~1,786~~ 2,107 standard connectors and cable assemblies...

SMA Connectors: cable types, microstripline, bulkhead, panel mount, field replaceables, hermetic, stripline and radius right angles.

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Adapters: in and between series; SMA, TNC, Type "N", 7mm, 3.5mm. 2.4mm and "K"™.

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...but if you need something special, let's talk."

Walt Carpenter
CDI Sales Manager

CDI

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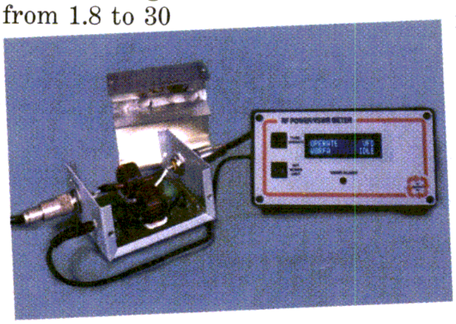
CIRCLE NO. 280

Power meters measure HF and VHF

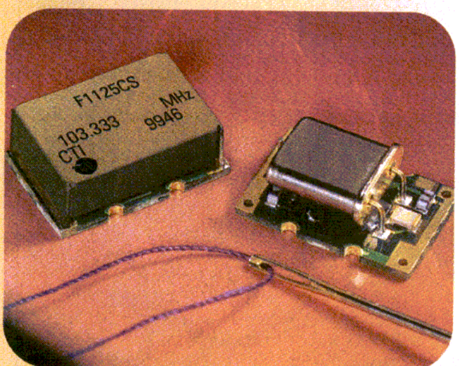
The VFD series of RF wattmeters measures power in the high-frequency (HF) and very-high-frequency (VHF) ranges. The high-frequency model covers frequencies from 1.8 to 30 MHz at power levels from 15 to 2950 W. The VHF model covers 50 to 150 MHz at 5 to 300 W. The meters connect to the RF feedline through a remote sensor and have a settable SWR alarm with an optional external relay to protect connected equipment. They feature a two-line, 16-character vacuum fluorescent display that includes a 65-segment, peak root-mean-square (RMS)-power bar graph as well as an alphanumeric power and VSWR indicator. The meters are supplied with standard SO-239 connectors, and type-N connectors are available as an option. A power-monitor option provides a relay to control external devices when the meter senses power output. The meters operate from a +12-VDC power source. **RF Applications, Inc., 7345 Production Dr., Mentor, OH 44060; (440) 974-1961, FAX: (440) 974-9506, Internet: http://www.rfapps.com.**



CIRCLE NO. 57 or visit www.mwrf.com

Clock timer boasts high stability

The F1100CS series of high-frequency clock timers boasts high stability and low jitter for applications such as phase-locked loops (PLLs), clock recovery, reference-signal tracking, and synthesizers. The timers are available at frequencies from 65 to 105 MHz. They have a typical stability of ±25 PPM and a typical jitter of 5 ps root mean square (RMS) at 100 MHz. The timers have a tri-state output, 3-ns rise and fall times, and are available with a 45/55-percent duty cycle. The devices have hermetically sealed crystals for predictably long-term aging and reliability and are housed in surface-mount packages. **Champion Technologies, Inc., 2553 N. Eddington St., Franklin Park, IL 60131; (847) 451-1000, FAX: (847) 451-7585, Internet: http://www.champtech.com.**



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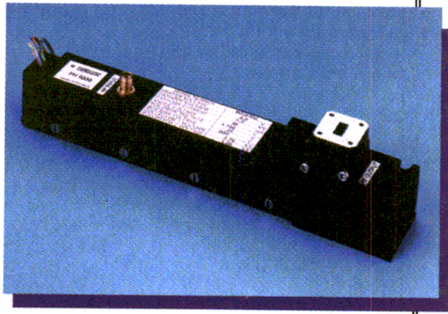
Tiny TWT serves uplinks and LMDS

Model TH 4030 conduction-cooled miniature traveling-wave tube (TWT) operates from 25.5 to 31.5 GHz and is specifically designed for applications in satellite uplinks and local multipoint distribution systems (LMDS). Its dual-stage collector is said to offer high-efficiency operation. The tube develops 40-W continuous-wave (CW) RF power while consuming a maximum of 140 W. Measuring 180 × 30 × 40 mm, it has a minimum gain of 45 dB and boasts an unrivaled power-to-size ratio. **Thomson Tubes Electroniques, 18 avenue du Maréchal Juin, F-92366 Meudon-la-foret, Cedex, France; (33-1) 30 70 35 00, FAX: (33-1) 30 70 35 35, Internet: http://www.tte.thomson-csf.com.**



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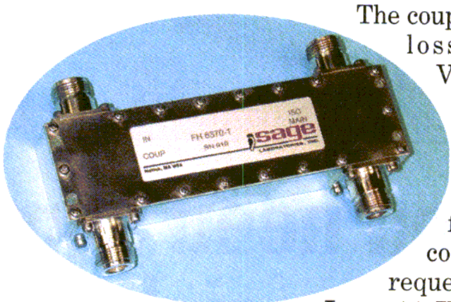
Dual-band hybrid handles high power

Model FH6370-1 quadrature hybrid coupler operates from 800 to 2400 MHz and can handle 200-W continuous-wave (CW) RF power. It is ideal for use in antenna feeds and combining amplifier power in telecommunication base stations. The coupler offers typical insertion loss of 0.2 dB, maximum VSWR of 1.22:1, and minimum isolation of 20 dB. It measures 5.6 × 1.5 × 1.0 in. (14.224 × 3.81 × 2.54 cm) and includes type-N female connectors. Other connectors are available on request. **Sage Laboratories, Inc., 11 Huron Dr., Natick, MA 01760-1338; (508) 653-0844, FAX: (508) 653-5671, Internet: http://www.sagelabs.com.**



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Communications Giants Slug It Out

Buyouts and mergers in the worldwide communications industry are turning into a real donnybrook as the heavyweight contenders continue to pound each other in an attempt to control the global business in wireless, Internet, and media. The latest blockbuster (in February) came when Britain's Vodafone AirTouch, plc. appeared ready to merge with Germany's Mannesmann AG in a $180 billion deal that ranks as the largest linking up of two companies in corporate history. Just one month earlier, however, the marriage of Internet giant America Online (AOL) to media power Time Warner for $165 billion was hailed as the biggest ever. And three months before that, MCI World Com's $115 billion buyout of Sprint Corp. topped the charts for the biggest merger.

At stake in the Vodafone-Mannesmann linkup are 42 million wireless customers around the world. Vodafone was already the largest mobile-phone business when it acquired AirTouch Communications, a major US wireless operator, last year. The Mannesmann deal would give Vodafone a market capitalization of $350 billion, lagging behind only Microsoft, General Electric, and Cisco Systems. In Europe, Vodafone would be the Continent's biggest company while controlling the first or second mobile-phone market position in 11 countries with a total of 29 million customers. European companies are waging war to determine who will win in a market expected to be worth $100 billion by 2003.

The AOL buyout of Time Warner takes the largest online service provider (AOL) with 20 million subscribers and joins it with the world's largest media company. AOL's Internet access is slow, so it needs Time Warner for the latter's vast cable networks which go out to 13 million subscribers and will be able to deliver broadband access to appliances such as desktop computers, TV sets, and handheld devices. AOL also needs the content that Time Warner provides through its publishing empire of magazines, books, and other media. Indeed, AOL seeks to control not only the way future subscribers will access the Internet, but the content that people will desire as well.

If you are getting the idea that the two key communications technologies of the 21st century—wireless and the Internet—are coming under the control of a very few hands, you are right. Not only that, but the two are merging, driven by the idea that wireless phones will soon overtake personal computers (PCs) as the main way to surf the Internet. In the view of industry and government leaders, this consolidation could lead to a dangerous situation where information sources, choices, and prices are dictated by a few giant corporations. ●●

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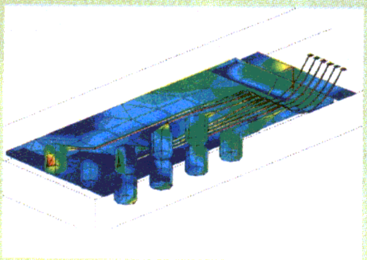
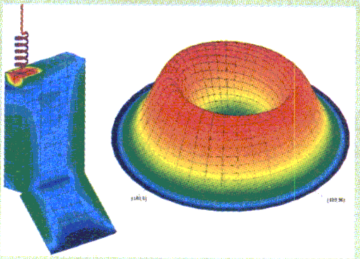
- RLC parameter extraction compatible with SPICE
- 2D and 3D display of current distribution, radiation patterns and near field
- Calculation of antenna and scattering parameters including directivity, efficiency and RCS
- Current and near field animation

IE3D Simulation Examples and Display

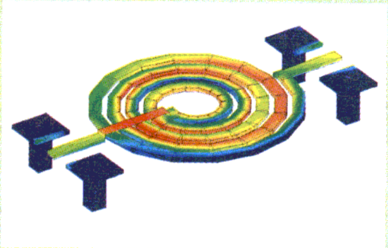
The current distribution on an AMKOR SuperBGA model at 1GHz created by the IE3D simulator



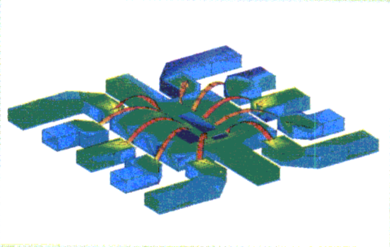
The current distribution and radiation pattern of a handset antenna modeled on IE3D



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

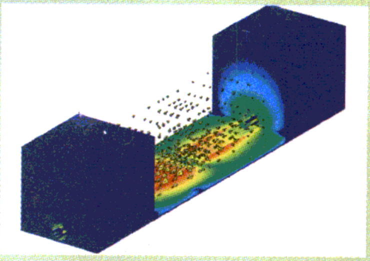


IE3D modeling of an IC Packaging with Leads and Wire Bonds

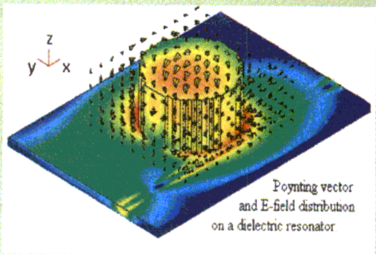


FIDELITY Examples

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



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



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CIRCLE NO. 319

Contracts

REMEC, Inc.—Has been awarded a $9.42 million contract from Raytheon Systems Co. for the initial production of a series of microwave multifunction modules used in defense electronic communication systems.

Tellabs—Has signed a multimillion dollar agreement to deploy a voice-enhancement system for Chunghwa Telecom's nationwide Global System for Mobile Communications (GSM) mobile network that will greatly improve voice quality and network efficiency.

Scientific-Atlanta, Inc.—Has reached a definitive agreement with ViaSat, Inc. where ViaSat will acquire the satellite-networking business of Scientific-Atlanta in a $75 million transaction. The transaction is subject to various regulatory and other conditions and is expected to close within 120 days.

Piezo Technology, Inc.—Has been awarded a Department of Defense (DoD) Small Business Innovation Research (SBIR) contract to develop a temperature-compensated crystal oscillator (TCXO) which will maintain accuracy after a 50,000 G launch shock. The targeted application is next-generation smart munitions.

Electro-Radiation, Inc. (ERI)—Announced the recent award of a US Army Small Business Innovative Research (SBIR) Phase I contract valued at approximately $70,000. The six-month program will explore the adaption of ERI's commercial-off-the-shelf (COTS) technology to small Precision Guided Munitions (PGMs). ERI will perform research and development (R&D) to enhance the anti-jam performance of Global Positioning System (GPS) receivers using a single antenna and associated electronics.

Harris Corp.—Announced the signing of the latest in a series of contracts with China United Telecommunications Corp. (China Unicom), the country's second-largest telecommunications provider, for a total value of $22 million (USD). China Unicom, a state-owned entity that works to promote competition within the country, will use Harris equipment to build its Global System for Mobile Communications (GSM) network across China, providing basic cellular service as a supplement to fixed lines.

ViaSat, Inc.—Has been awarded a contract, along with its teammates and subcontractors Harris Corp. and the Xetron subsidiary of Northrup Grumman, by the Department of Defense (DoD) to commence work on the Multifunctional Information Distribution System (MIDS) LVT Production Program. The initial contract commits $11.7 million to immediately begin work with an expected contract value of more than $30 million to be finalized by the end of the second calendar quarter of 2000. MIDS is a NATO-certified line-of-sight, tactical radio system that collects data from many sources and displays an electronic overview of the battlefield.

Fresh Starts

Diamond Antenna and Microwave Corp.—Announced the acquisition of the waveguide-switch product line of Micronetics Wireless of Hudson, NH.

RF Micro Devices, Inc.—Announced the opening of a new engineering design center in the Boston, MA area. The new facility is expected to eventually house a team of 35 senior-level RF engineers and technicians designing high-efficiency linear power amplifiers (PAs) and other RF integrated circuits (RF ICs) used in cellular and personal-communications-services (PCS) telephones.

Transistor Devices, Inc., Advanced Conversion Products (ACP) Division—Announced the launch of its new website, which is designed to make product, corporate news, and application information more easily accessible to current and potential customers. The website is located at http://www.tdipower.com/acp.

Wavetek Corp., Precision Measurement Division—Has launched a new website, http://www.wavetekprecision.com. The site is dedicated to precision measurement and provides up-to-date information about all of the division's products and services. Data sheets, specifications, and application notes are available to be downloaded. Press releases are available in a library of current and recent releases.

Centurion International, Inc.—Introduced a new website, which will make it easier for product designers and manufacturers to quickly and efficiently obtain information on antennas and power products. The website is located at http://www.centurion.com.

W.L. Gore & Associates, Inc.—Has developed a new online resource for designers facing electromagnetic-interference (EMI)/RF-interference (RFI) shielding problems. The website, located at http://www.gore.com/electronics/emicenter, provides a compilation of technical, product, and standards information for beginning and experienced engineers handling EMI.

Celeritek, Inc.—Announced the availability of its indium-gallium-phosphide (InGaP) heterojunction-bipolar-transistor (HBT) process, which is suitable for low-voltage (+3-VDC) linear, efficient power amplifiers (PAs). The InGaP HBT process is production qualified at Celeritek's high-volume facility.

TESSCO Technologies, Inc.—Announced an alliance with TTC of Germantown, MD to bring the T-BERD, FIREBERD, and TPI test instruments to the wireless communications market.

Motorola, Inc. and SkyTel, Inc.—Have signed a memorandum of understanding to expand two-way wireless communications solutions to Latin America.

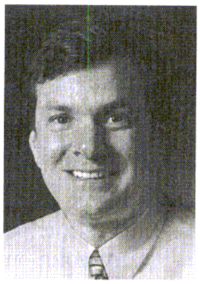
ON Semiconductor—Announced a new corporate initiative to advance its proprietary high-bandwidth integrated-circuit (IC) products. They are working on new product families that will bring a higher level of functional integration and deliver the 40 gigabit devices required to support the Super Internet.

ANADIGICS—Will expand its state-of-the-art 6-in. production facility. The company's expansion plan includes the development of additional Class 10 cleanroom space and equipment installation within its existing indium-gallium-phosphide (InGaP) heterojunction-bipolar-transistor (HBT), pseudomorphic-high-electron-mobility-transistor (PHEMT), and metal-semiconductor-field-effect-transistor (MESFET) manufacturing facility. The expanded facility is scheduled for completion by the end of the first half of 2000.

Endgate Corp.—Lionel Kirton to vice president of operations; formerly worldwide manager of communications components for Hewlett-Packard Co.

Ferronics, Inc.—Timothy Reeder to chief executive officer and president; formerly chief operating officer.

REEDER



LENIHAN

Alpha Industries—Bruce Lenihan to director of European sales; formerly regional sales director for Europe and the Middle East for RF Monolithics.

Palomar Technologies—Tom Simonson to direct sales account manager; formerly program manager at Honeywell's Commercial Flight Systems Division.

Ball Aerospace & Technologies Corp.—Art Rancis to vice president of communication and video products; formerly chief executive officer at Racom Systems, Inc.

Wyle Electronics—Jim Schaeffer to senior vice president of marketing; formerly vice president of supplier business management with Avnet.

Gabriel Electronics, Inc.—Russell Vest to vice president of sales and marketing; formerly vice president at P-Com.

CTS Corp.—John D. Watson to vice president and general manager of the hybrid microelectronics business; formerly vice president and general manager of the interconnect products business.

Techtrol-Cyclonetics, Inc.—Glenn R. Kurzenknabe to marketing manager; formerly marketing manager with Piezo Crystal Co.

Temptronic Corp.—William Stone to president; formerly chief operations officer.

TESSCO Technologies, Inc.—Kevin P. Demery to vice president and unit director of product and brand development; formerly senior account executive with Indus International.

IPC—John Riley to vice president of professional development; formerly director of education.

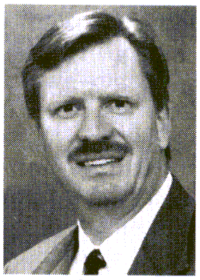
Voltronics Corp.—George J. Blanar, Ph.D. to senior vice president of sales and marketing; formerly vice president of sales and marketing for Giga-tronics, Inc.

Radiant Networks, plc.—Phil Whitehead to head of regulatory affairs; formerly marketing manager at Plextek.

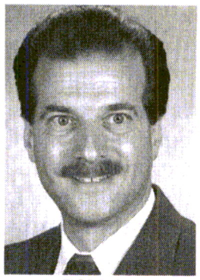
RF Monolithics, Inc.—Michael R. Bernique to chairman of the board; remains as president and chief executive officer of TelOptica, Inc. Also, David M. Kirk to president and chief executive officer; formerly vice president of marketing.

Kymata—Clive Wilson to operations director; formerly general manager of integrated circuits (ICs) and sourcing at Ericsson Components.

ShellCase, Inc.—Philip E. Rogren to North American vice president of sales and marketing; formerly employed at Nexus, Inc.



ROGREN

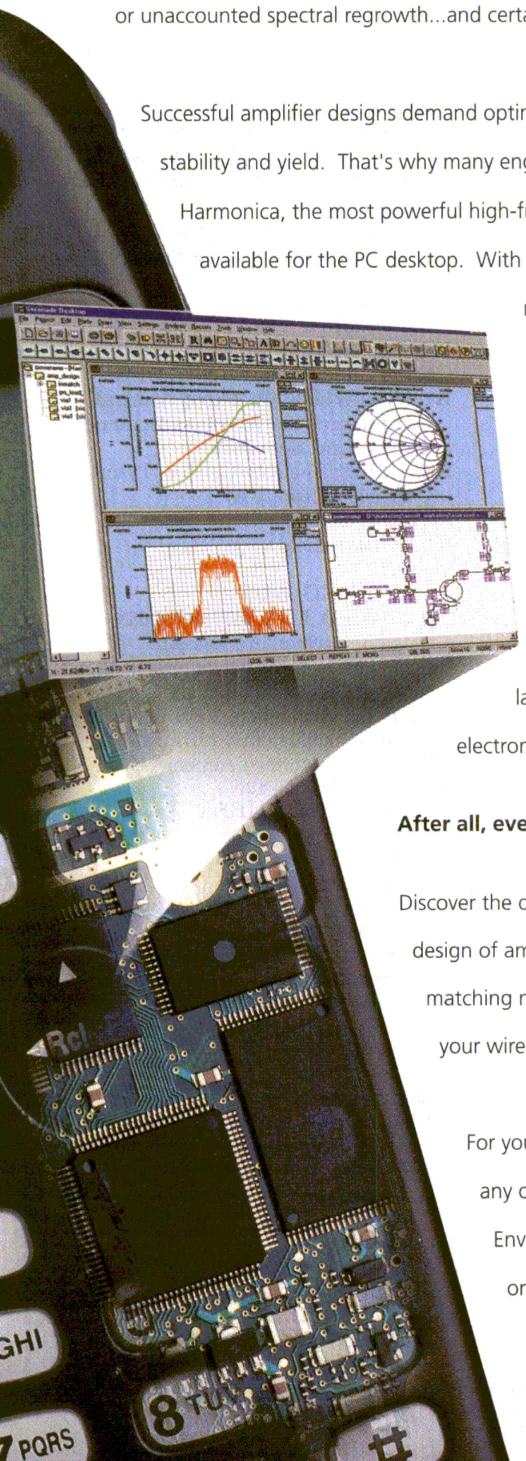


BORTON

TEGAM, Inc.—James Borton to director of global sales; formerly marketing manager for data-acquisition (DAQ) products at Keithley Instruments.

Xpedion Design Systems—Richard Curtin to senior vice president of sales and marketing; formerly employed at Simpod.

SignalSoft Corp.—Don Winters to chief operating officer; formerly vice president of operations. Also, Gerry Christensen to director of product management; formerly business unit manager at ILLUMINET.



Short Courses

RF Power Amplifiers, Classes A through S: How the Circuits Operate, How to Design Them, and When to Use Each
March 20-21 (Holiday Inn Select, Niagara Falls, NY)
R.A. Wood Associates
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Utica, NY 13501
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e-mail: rawood@rawood.com

Wireless Data Systems/Digital Cellular Systems
March 27-31 (Los Angeles, CA)
UCLA Extension, Department of Engineering, Information Systems and Technical Management, Short Courses
10995 Le Conte Ave., Suite 542
Los Angeles, CA 90024
(310) 824-3344, FAX: (310) 206-2815
e-mail: mhenness@unex.ucla.edu
Internet: http://www.unex.ucla.edu/shortcourses

Circuit Design for Wireless Communications
March 27-31 (Monterey, CA)
MEAD Microelectronics, Inc.
7100 NW Grandview Dr.
Corvallis, OR 97330
(541) 758-0828, FAX: (541) 752-1405
e-mail: valence@mead.ch or usmead@cmug.com

Fundamentals of Cellular and PCS Wireless Communications
March 29-31 (Madison, WI)
Department of Engineering Professional Development
University of Wisconsin-Madison
432 North Lake St.
Madison, WI 53706
Katie Peterson
(800) 462-0876, FAX: (608) 263-3160
e-mail: Custserv@epd.engr.wisc.edu
Internet: http://epd.engr.wisc.edu.

How To Design RF Circuits
April 5 (Savoy Place, London, UK)
The Institution of Electrical Engineers (IEE)
Savoy Pl.
London WC2R OBL, United Kingdom
+44 (0)20 7240 1871, FAX: +44 (0)20 7240 7735
Internet: http://www.iee.org.uk/Events/e05apr00.htm.

EMI Made Simple: Systems, Grounding & Shielding, and Designing for EMC
April 5-7 (Rochester, NY)
April 10-12 (Long Branch, NJ)
Kimmel Gerke Associates Ltd.
300 Christine Ln.
West St. Paul, MN 55118
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Meetings

IP 2000: The System-on-Chip Conference & Exhibition
March 20-22 (Santa Clara Convention Center, Santa Clara, CA)
Electronic Times/ EE Times
Internet: http://www.ip2000/com

Hannover Fair
March 20-25 (Hannover Fairgrounds, Hannover, Germany)
Hannover Fairs, USA, Inc.
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Princeton, NJ 08540
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The Third Annual Conference on Fiberoptics Markets In Latin America
March 21-22 (Fountainbleau Hilton, Miami Beach, FL)
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Newport, RI 02840
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Fourth Annual Army Research Laboratories (ARL) Federated Laboratory Symposium (Fed Lab 2000)
March 21-23 (The Inn and Conference Center, University of Maryland, College Park, MD)
TMC Design Corp.
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Suffolk, VA 23432
(757) 357-4011, FAX: (757) 357-5108

Wireless Web 2000-Third Annual Conference on Wireless Internet Access and Services
March 27-29 (The Ritz-Carlton, Tysons Corner, McLean, VA)
ACT Conferences
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Suite 213
Springfield, MO 65807
(888) 274-7720, FAX: (417) 882-9653
e-mail: info@actconferences.com
Internet: http://www.actconferences.com/ww2000/index.htm

2000 IEEE Emerging Technologies Symposium on Broadband, Wireless Internet Access
April 10-11 (Dallas, TX)
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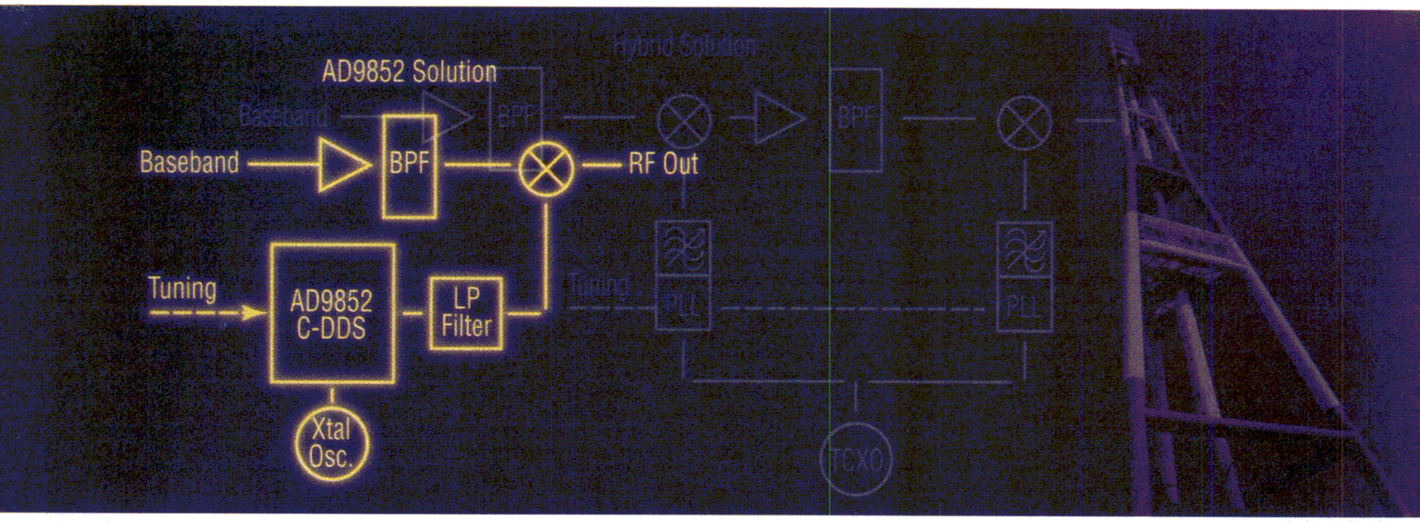

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Shielding material cuts board-level EMI

Electromagnetic interference (EMI) poses difficult problems for designers of mobile communications equipment where RF and digital components must operate without interacting with one another. Moreover, few standard technologies are available to allow designers to meet today's emission and susceptibility challenges. A new material for EMI shielding at the personal-computer (PC)-board level has been developed by Donald M. Yenni Jr., Mark G. Baker, and Curt Maynes of 3M Co.'s Electronic Handling & Protection Div. (Austin, TX) that the authors claim meets the standard tests required by the mobile-phone industry. The material is a heat-stakable plastic shielding cap composed of two layers—a 0.13-mm thick film of polycarbonate and a thermally extensible mat of tin/bismuth (Sn/Bi) alloy fibers enclosed in an ethylene-vinyl-acetate (EVA) hot-melt adhesive resin. The metal-fiber mat provides the composite material with its EMI shielding characteristics. Testing indicates that the material (in the far field) is equivalent or superior to many of the perforated and stamped metal cans in use today. The costs associated with all of the shielding techniques in use are presented by the authors. See "A New Alternative for Board-Level EMI Shielding," *ITEM Update 1999*, p. 34.

Optical antennas bend to fit curved surfaces

Fitting an antenna to a curved support such as a vehicle window, a building, or computer monitor is becoming simpler with the availability of optically transparent patch antennas. According to Richard Q. Lee of the NASA Glenn Research Center (Cleveland, OH) and Rainee M. Simmons of NYMA, Inc., these antennas are useful where the need for radio communications is dictated by a lack of room for adding separate antenna-supporting structures. An optically transparent patch antenna can be manufactured from an optically transparent, electrically conductive film deposited on one face of a polyester film or glass substrate. The deposited layer has a surface resistance of 6 to 10 Ω per square, and the patch and its feed strip can be formed simply by cutting the coated film in the required pattern. Tests of prototype antennas reveal that they exhibit radiation patterns, return losses, and input impedances similar to those of conventional copper (Cu)-conductor patch antennas. See "Optically Transparent Patch Antennas," *NASA Tech Briefs*, December 1999, Vol. 23, No. 12, p. 40.

CMOS receiver designed for WLANs

Wireless local-area networks (WLANs) meeting the IEEE 802.11 standard in the 2.4-GHz band are making strong inroads as high-speed data links in office buildings, college campuses, hospitals, and factories. The technique gaining favor for 802.11 applications is direct sequence (DS) with quadrature phase-shift keying (QPSK). It is for this technique that Behzad Razavi of the Electrical Engineering Department of the University of California, Los Angeles (Los Angeles, CA) has designed a complementary-metal-oxide-semiconductor (CMOS) receiver incorporating a low-noise amplifier (LNA), quadrature downconversion mixers, partial channel-selection filtering, AC coupling, and baseband amplification. The receiver is fabricated in 0.6-μm CMOS and produces a noise figure of 8.3 dB, third-order intercept (TOI) of –9 dBm, and a voltage gain of 34 dB. See "A 2.4-GHz CMOS Receiver for IEEE 802.11 Wireless LANs," *IEEE Journal of Solid-State Circuits*, October 1999, Vol. 34, No. 10, p. 1382.

Smart-antenna systems get a comprehensive look

Due to the rapid expansion of demand for mobile communications services, adaptive antenna-array systems (smart antennas) are critical to expanding communications capacity and quality. Two types of smart-antenna systems are in use—the tracking-beam array (TBA) and the switching-beam array (SBA). Both types have been examined in great detail by Seungwon Choi of the School of Electrical and Computer Engineering at Hanyang University (Seoul, Korea) *et al.* with the goal of determining which is the better solution. The two systems are compared in a code-division-multiple-access (CDMA) mobile communications environment. As might be expected, there is no clear-cut winner. TBA is superior when the desired signal is sufficiently larger than each interfering signal. Since the ratio of the power of the desired signal to that of the interfering signals is lower, SBA becomes the better choice. See "A Comparison of Tracking-Beam Arrays and Switching Beam Arrays Operating in a CDMA Mobile Communication Channel," *IEEE Antennas & Propagation Magazine*, December 1999, Vol. 41, No. 6, p. 10.

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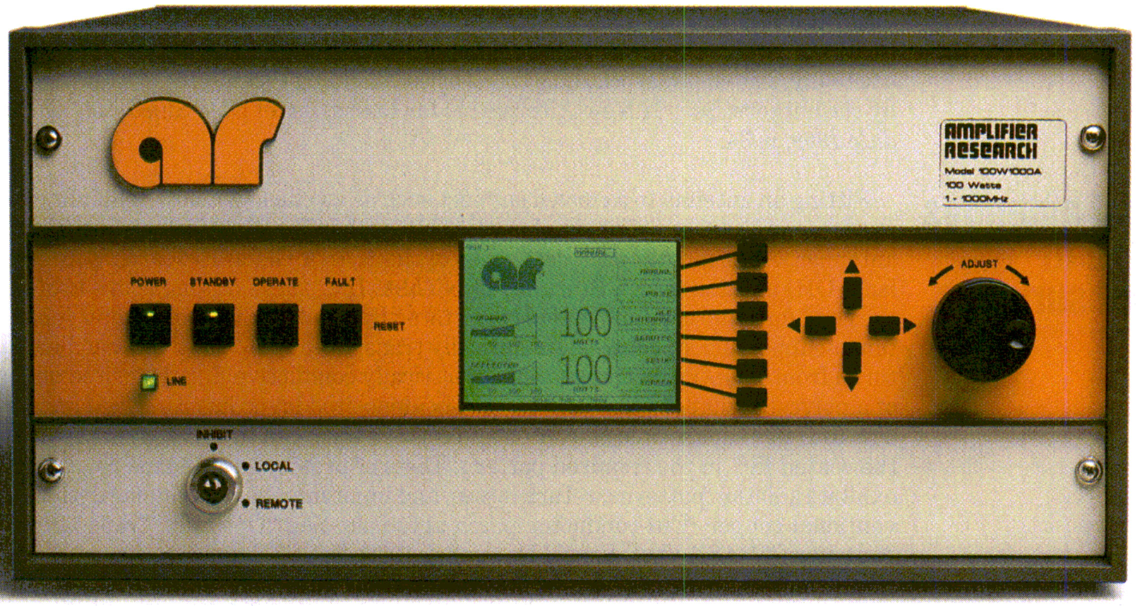

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CIRCLE NO. 235

Digital Predistortion Linearizes CDMA LDMOS Amps

The results of applying digital predistortion show that the technique has a positive impact on amplifier output power, linearity, and efficiency.

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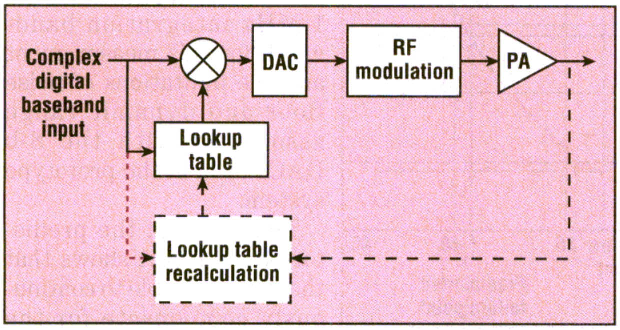
As digital signal processors (DSPs) become faster, the real-time correction of distortion effects of RF power amplifiers (PAs) by applying digital predistortion is becoming an increasingly viable solution. The concept of digital predistortion as a linearization technique for RF PAs has existed for more than a decade. Only recently, however, has the processing power become available to move the idea beyond the theoretical world and into the laboratory.

Previous reports of digital predistortion focused primarily on the mathematical derivation and simulated results with little attention to actual measurements.[1-5] In this article, measured results of linearity improvements due to predistortion will be presented. The focus is on the applications of predistortion in the IS-95 code-division-multiple-access (CDMA) format. In particular, a family of Motorola laterally-diffused-metal-oxide-semiconductor (LDMOS) PAs designed for CDMA applications was used in the measurements. For the IS-95 test signals, the signal configuration is varied as a way to determine the performance capabilities of the predistortion system. Specifically, issues related to peak-to-average power ratio, PA output power, and power-added efficiency (PAE) will be analyzed and results presented.

A simple implementation of adaptive (closed-loop) digital predistortion is shown in Fig. 1. Several variations of this system have been proposed over the last decade.[6-9] In general, the main differences are in the implementation of the predistortion algorithm or the configuration of the adaptive loop.

The predistortion algorithm used in this analysis is based on an initial measured PA amplitude modulation (AM)-to-AM and AM-to-phase-modulation (PM) response extracted from vector-network-analyzer (VNA) S-parameter measurements. The power and phase characteristics are then interpolated using splines, which are continuous piecewise cubic functions with continuous first and second derivatives. The spline technique has shown to provide the most numerically accurate results for LDMOS PAs. The interpolated power and phase characteristics are then used to compute the appropriate predistortion coefficients for the desired PA input-power level. These coefficients are then multiplied with the original IS-95 signals to generate the desired predistorted baseband signals. The predistorted signals are passed through the digital-to-analog converter (DAC), upconverted, and passed to the PA. For the measure-

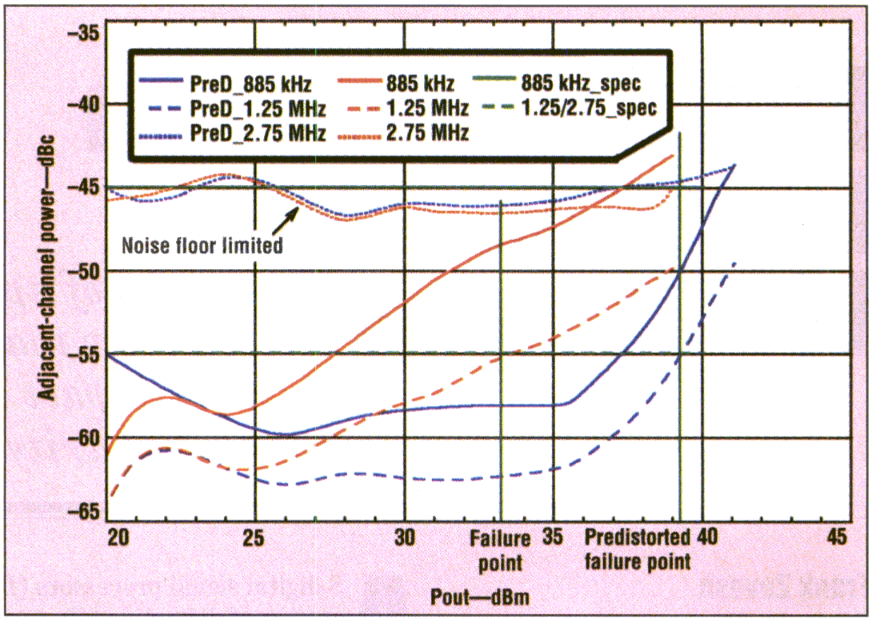


1. Adaptive digital predistortion is carried out with this type of circuit that takes predistorted signals and passes them through a digital-to-analog converter (DAC), an RF modulator, and a PA.

ment setup, the baseband-signal generation is carried out on a personal computer (PC) and uploaded to an arbitrary-waveform generator (ARB), which operates as a DAC. For the analog/RF components of the systems, off-the-shelf components are used.

When analyzing the performance of any system, it becomes essential to focus on issues related to the system standard. Previous discussions of digital predistortion have reported improvements in adjacent-channel power (ACP) but have not discussed the overall requirements of the RF PA as applied to the system.

The IS-95 specification clearly defines three bands of interest when analyzing the spectral regrowth performance of the amplifier. In the results presented in this article, the Rhode & Schwarz spectrum analyzer is used to perform the appropriate bandwidth integration for obtaining the desired out-of-band interference power levels.



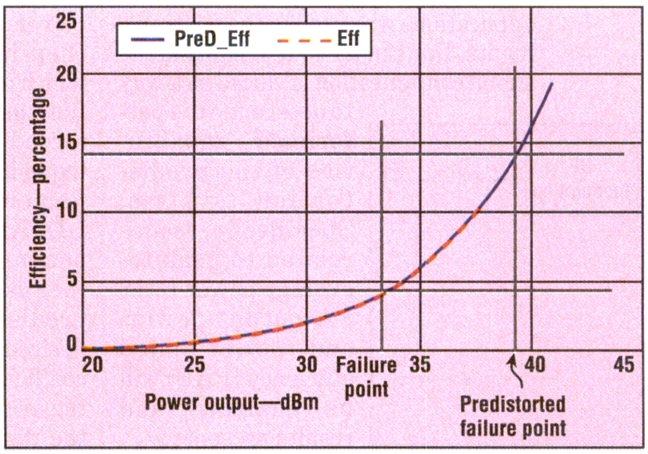
2. The three bands of interest in the IS-95 CDMA specification along with their integration bandwidths are shown in this plot. These are called the adjacent channel, first alternate channel and second alternate channel.

STIMULUS AND TEST

Test signals for the analysis of the digital predistortion system are based on the IS-95 CDMA specification. These signals are generated using an in-house Matlab-developed routine. The program provides the versatility of switching on or off any channel or series of channels to generate a signal with the appropriate peak-to-average ratio. It is important that designers understand how a device will perform in a particular application prior to any actual design activity. Thus, Motorola has begun to characterize device performance under CDMA formats. Using measurement standards, plots of power drive-up versus ACP are presented on typical RF device datasheets. It is possible to determine the improvements in linearity that can be obtained by using the baseband digital predistortion within the test methodology outlined.

Figure 2 shows output power versus ACP using the integration bandwidths definitions for IS-95 CDMA. The plot shows the performance of the MHL19936, a Class A, 14-W LDMOS module with 30-dB gain, driving an MRF19030, Class AB, 30-W LDMOS discrete device with greater than 12 dB of gain. The performance curves are for +26-VDC operation at 1.9 GHz.

There are three adjacent-channel measurements that are important when designing a CDMA base-station amplifier. The adjacent channel is indicated by the solid line, the first alternate is described by the dashed line, and the second alternate is shown by the dotted line (Fig. 2). The red plot lines show the performance of the device lineup non-corrected, while the blue lines show the predistortion-corrected performance. With the given specifications, the mask-failure point, due to spectral regrowth, occurs first at the first alternate in both cases. The lineup would fail at +33.25-dBm output power without correction and +39.25 dBm with the baseband predistortion applied. For the given peak-to-average ratio of 11.2 dB, this corresponds to an improvement of 6 dB in output power.

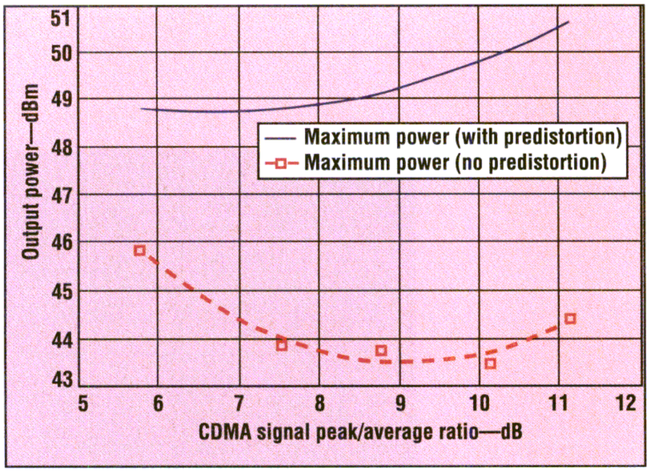


3. A curve of efficiency versus power output of the LDMOS devices with and without predistortion correction indicates that correction permits a sharp improvement in amplifier efficiency (14.2 percent versus 4.25 percent).

In all cases, the second alternate fails the specification. This is attributed to the 1-MHz integration bandwidth for this measurement and the limitations of noise floor and dynamic range associated with the 8-b DACs used in the prototype system.

The trend of the predistorted ACP plot shows that the system is able to continuously compensate for the AM-to-AM and AM-to-PM distortions of the lineup up to +36-dBm output power. This is evident by the shape

of the ACP curves, which are essentially flat versus output power, compared to the constantly increasing ACP for the non-compensated curves. Beyond +36 dBm, the system continues to compensate for AM-to-PM characteristics, but is starting to hit AM-to-AM limitations as the amplifier is driven into compression. This causes spectral regrowth to overtake the amplifier at a much-greater rate than is seen in the non-corrected plots, and at 39.25 dBm, the system is no longer able to compensate and the amplifier fails the CDMA mask specifications. With the given improvement in output power, it is expected that an improvement in efficiency would also occur. In general, driving an LDMOS device harder results in increased efficiency. Plotting the output power versus the overall efficiency of the lineup with and without the correction applied, as shown in Fig. 3, indicates that there is no degradation in efficiency as a result of the correction.

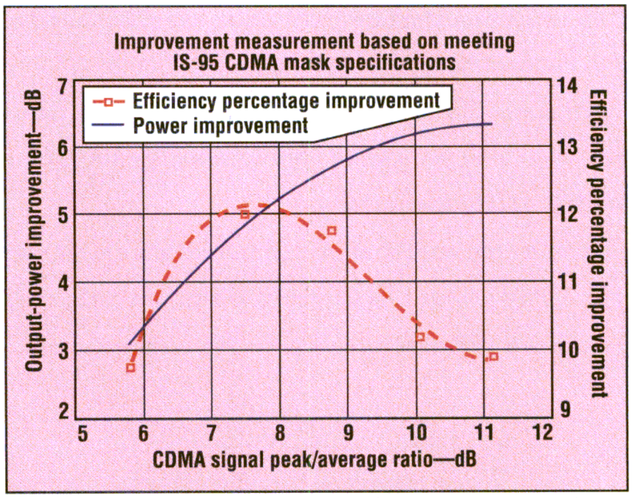


4. This plot of peak output power versus peak-to-average ratio shows that a high peak-to-average ratio need not compromise performance. In fact, since the device spends less time at the peak, it runs cooler and can operate at a higher saturated power.

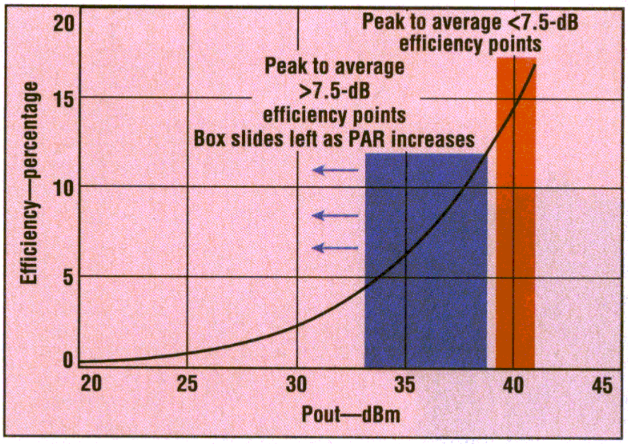
In summary, the increased output power of the lineup as a result of the predistortion has resulted in a more-efficient amplifier at the CDMA mask-failure point. For the mask-failure power levels indicated in Fig. 2, the efficiency can be compared at 4.25 percent without correction and 14.2 percent with predistortion. This is an improvement of approximately 10 percent (in efficiency percentage points), or a 237-percent efficiency improvement.

CDMA systems, due to the dynamics of channel utilization, must be able to handle varying peak-to-average ratio signals. Consequently, it is expected that the system be able to meet the mask requirements as the system changes. Performance of a predistortion system at a single peak-to-average ratio will give an idea of how the system will perform, but investigation of that performance as the signal dynamics varies is very useful.

The first task in determining the performance at various peak-to-average ratios requires



5. Predistortion techniques on LDMOS devices result in improvements in both output power and efficiency as shown by these two curves. Notice that power improvement reaches a saturation point at a peak-to-average ratio of 11 dB.



6. Efficiency improvements are divided into two groups depending on the value of the peak-to-average ratio. When peak-to-average is greater than 7.5 dB, efficiency tends to decline, while for a ratio of less than 7.5 dB, it tends to increase.

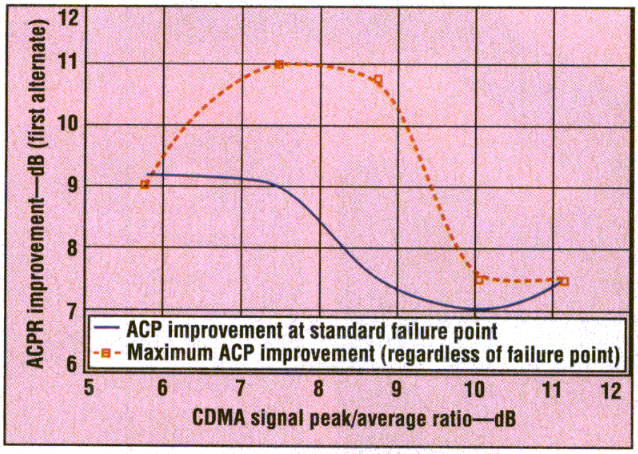
being able to generate multiple peak-to-average appropriately scaled levels of the pilot, paging, and sync channels. A prior look would indicate that this should be a fairly simple task. Matlab was used to generate an IS-95 signal. The traffic channels were scaled and summed to appropriately scaled levels of the pilot, paging, and sync channels. The resulting data were then filtered using the standard IS-95 filter.

After several valid peak-to-average-ratio signals were created, repeated generation of output power versus adjacent-channel power and the efficiency of the lineup, resulted in several useful plots, one of which is illustrated in Fig. 4.

This graph presents the peak output power versus the peak-to-average ratio. In this case, the peak output power is obtained by measuring the average output power where the device fails the IS-95 spectral mask and adding the peak-to-average ratio (in decibels). The blue line shows the maximum power with the predistortion applied, while the red line indicates the maximum power without predistortion.

The expected trend of the predistorted maximum power curve would be a straight line, or a constant independent of the peak-to-average ratio. This would indicate that a maximum amount of correction has been achieved and the theoretical limits based on amplifier saturation have been reached. In reality, it is possible to see an upward trend in this curve. This can be explained by realizing that even though the peak-to-average ratio has increased, the amount of time spent at that peak (or the probability of reaching that peak) has dropped. As such, the device tends to operate cooler (on average) which, in turn, results in higher saturated power capability.

Additionally, the same trend (a constant independent of the peak-to-average ratio) would be expected for the uncorrected maximum power curve. The inflection at the high peak-to-average ratios can be explained based on the same thermal interactions described, but this does not explain the inflection at low peak-to-average conditions. This is revealed



7. Improvements in ACPR plotted against peak-to-average ratio are indicated by these two curves. The blue curves shows where the system would fail mask specifications without the introduction of predistortion.

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by looking at the nature of the distortion in a low peak-to-average environment. In this case, PM and AM are starting to occur at approximately the same time. As a result, the amplifier is able to achieve a higher output power before the AM-to-AM and AM-to-PM distortions result in sufficient spectral regrowth to cause mask failure. If the difference of the indicated maximum powers is plotted, it is possible to see the output-power improvement afforded by implementing the predistortion technique. Additionally, plotting the efficiency improvement (in delta-efficiency percentage points) results in the plot of Fig. 5.

The blue line shows the output-power improvement using the predistortion technique. This line has a trend that would be expected. It is believed that at some point, the power improvement would reach a maximum. At this point, no matter how much the peak-to-average ratio increases, the improvement as a result of the technique has diminished returns. It should be noted however, that for an IS-95 CDMA signal, the output-power improvement for the device lineup at this point still supports a four-fold improvement (6 dB).

The efficiency-improvement characteristic (the red line) is not as obvious. The shape of the output power versus efficiency plot is shown in Fig. 6. The curve can be represented by a power function ($y = M_0 \times x^{M1}$). This shape leads to the explanation of the efficiency improvement plot.

To analyze the improvement trends, the efficiency-improvement plot is broken up into two sections. The first section includes peak-to-average ratios below 7.5 dB where the curve shows an increase in efficiency improvement as peak-to-average ratio increases. The second region, which includes peak-to-average ratios greater than 7.5 dB, shows a decrease in efficiency improvement as the peak-to-average ratio increases.

The first region (peak-to-average ratios less than 7.5 dB) is explained by noticing that the output-power improvement is approximately 3 dB. At these points, the efficiency delta is 10 percent and occurs after the inflection point in the efficiency versus output-power curve has occurred. Since the power improvement at this particular point is so small, a small increase in efficiency would be expected. As the output-power improvement increases (when the peak-to-average ratio increases), the trend is an increase in the efficiency improvement.

The second region, defined as having peak-to-average ratios greater than 7.5 dB, shows a downward trend in efficiency improvement. This is explained again by the shape of the efficiency curve. Essentially, the maximum (average) output power is decreasing as the peak-to-average ratio increases, and the delta in output powers (corrected versus uncorrected) appears to be a sliding window moving to the left in the efficiency versus output-power curve. Since the slope of the efficiency curve decreases as the output power decreases, the resulting delta efficiency decreases. This explains the downward trend in the efficiency curve as the peak-to-average ratio continues to increase (Fig. 6).

Although discussion of improved output power is one way to compare system performance, traditional discussions of linearization techniques compare ACP improvements that occur with and without the technique (Fig. 7). For this system, this can be mentioned in two different ways. The first is to look at the improvement in ACP at the failure point for the IS-95 mask. The blue line plots the ACP delta for the first alternate measurement at the point where the system would fail mask specifications without predistortion. In the case of the 11.2-dB peak-to-average signal presented earlier, this would be the ACP delta at +33.25-dBm output power. The red line plots the ACP delta for the first alternate at the maximum delta point. For the 11.2 dB peak-to-average example that is presented, this would occur at +36 dBm, the point where the predistorted ACP is no longer constant, but starts to degrade due to AM-to-AM limitations.

The measured results presented in this article, clearly indicate that baseband digital predistortion can be used to improve the linearity of LDMOS

(continued on p. 164)

Microwave Synthesizer Meets Demanding Communications Requirements

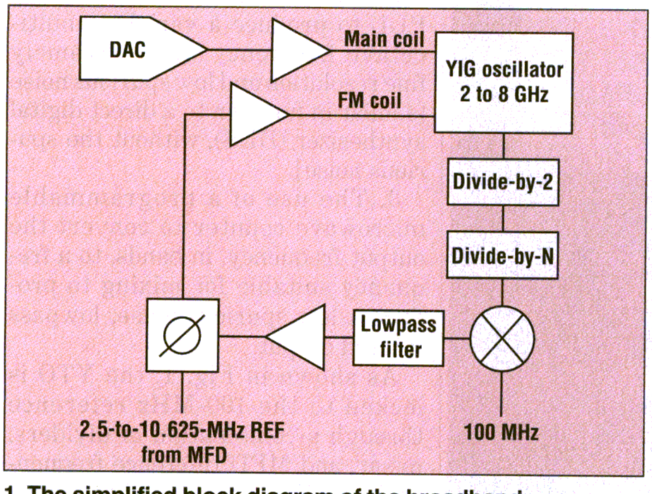
A broadband microwave synthesizer uses advanced technologies to deliver fine resolution and low phase noise for communications systems.

Philip J. Rezin
Principal Engineer
Rockwell Collins, Inc., 855 35th St., Cedar Rapids, IA 52498; (319) 295-4677, FAX: (319) 295-8556, e-mail: pjrezin@collins.rockwell.com.

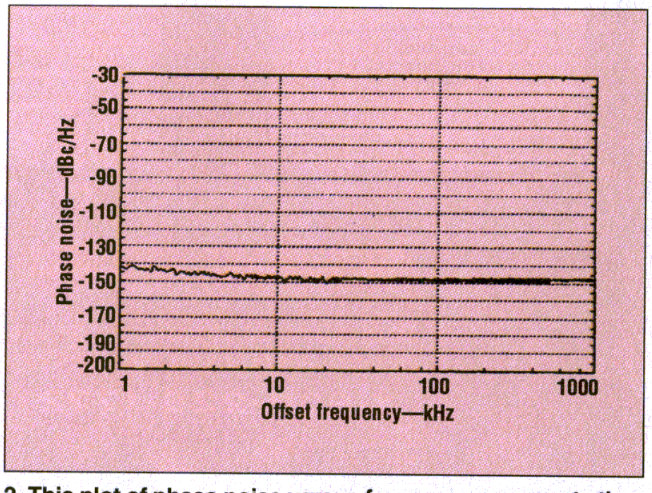
MODERN wireless communications systems depend on the ability to produce high-data-rate analog and digital modulation at high frequencies. Meeting the needs of complex, wideband, high-data-rate systems requires frequency synthesizers with broad tuning and very-low phase noise. With the latest technologies in heterojunction bipolar transistors (HBTs), sigma-delta modulated frequency dividers, and yttrium-iron-garnet (YIG) oscillators, it is possible to produce a broadband microwave synthesizer with fine tuning and low noise to meet the most demanding requirements of today's communications systems.

Synthesizers of high frequencies--in the gigahertz range—frequently use phase-locked loops (PLLs) to stabilize the output-signal frequency. They also include circuitry for downconversion of the output frequency to enable lower-frequency components to lock the output signal to the input reference frequency.[1]

The use of digital dividers to convert the output frequency to the reference frequency suffers from the limitation that excess phase noise is introduced in the output signal by a factor of 20 log (N) where N is the division ratio. Often, it is desired to use a harmonic- or comb-frequency generator and microwave mixer for frequency downconversion.[2] However, downconversion may introduce



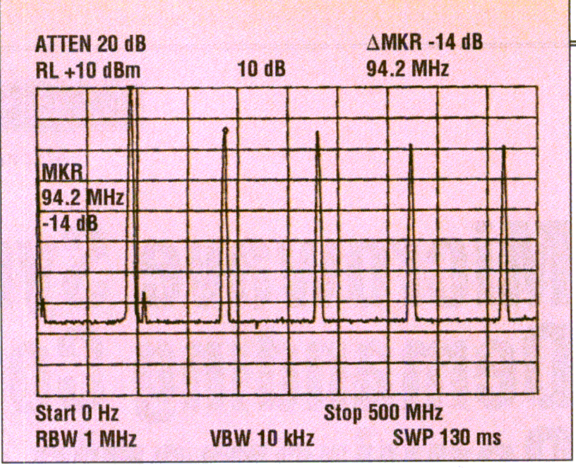
1. The simplified block diagram of the broadband synthesizer shows a new approach to synthesizer design with components such as a modulated fractional divider and a programmable microwave counter.



2. This plot of phase noise versus frequency represents the sum of the noise of two GaAs HBT dividers. At 10-kHz offset, the residual noise is down at –150 dBc/Hz. Three dB must be subtracted to determine the noise of each.

aliasing with difficulty in extracting the desired signal.[3] In addition, it is generally necessary to offset the desired signal from the comb-frequency lines, which requires additional PLLs to provide continuous tuning across comb frequencies.

The key enabling technologies include gallium-arsenide (GaAs) HBT fixed and programmable dividers, a sigma-delta modulated fractional divider (MFD), and a low-noise phase detector. This combination produces an output of 2 to 8 GHz with 1-Hz tuning and integrated phase jitter of less than 0.25-deg. root mean square (RMS), typical (integrated from 100 Hz to 20 MHz).



3. At an output frequency of 93.3 MHz, the programmable divider's even harmonics are relatively well suppressed, giving an output waveshape approximating a square wave.

A block diagram of the proposed synthesizer is shown in Fig. 1. It is comprised of the following components: a YIG-tuned oscillator (YTO) which produces an output frequency in response to a control signal, a microwave programmable divider of limited range, a mixer to produce a feedback signal in response to the output of the programmable counter, a phase detector that outputs the control signal, and a variable reference frequency derived by modulated fractional division of a fixed, high-frequency clock.

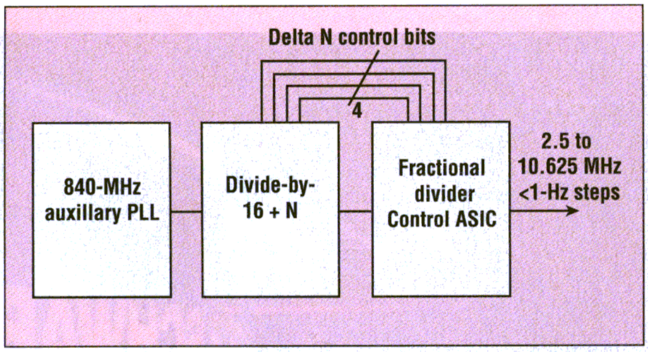
The key features include:

1. Architecture for a microwave synthesizer that contains very few microwave circuits, thus reducing cost and complexity.

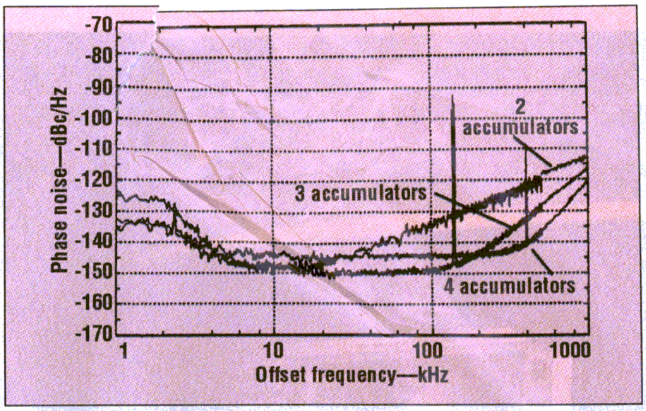
2. The use of an MFD outside of a PLL to produce a variable, multi-octave frequency with extremely fine resolution and low spurious noise [similar in function to a direct digital synthesizer (DDS), without the spurious noise].

3. The use of a programmable microwave counter to convert the output frequency, in bands, to a frequency suitable for mixing to produce a low spurious-noise, lowpass filtered output.

As shown in Fig. 1, the YTO is locked to the 100-MHz reference through a combination of dividers, mixer, and MFD reference frequency. The YTO output is fed to a fixed divide-by-2 GaAs HBT divider to produce an output frequency between 1 and 4 GHz. This signal is



4. The modulated fractional divider (MFD) is configured as the variable reference generator shown here. This divider is located outside of the PLL to obtain finer resolution and low spurious noise.



5. This plot shows the noise on a 10-MHz reference signal from the MFD as a function of various size accumulators.

fed into an HBT programmable divider that divides by 11 to 31 in integer steps to produce an output frequency between 89.375 and 97.5 MHz. The divider output is then mixed with a low-noise 100-MHz standard to produce an intermediate frequency (IF) between 2.5 and 10.625 MHz. This IF is fed to a low-noise phase/frequency detector where its is compared to a fine-tuned reference frequency derived by a modulated fractional division of a high-frequency clock. The output of the phase/frequency detector is fed back to the YTO, completing the PLL.

The relationship between the YTO and the other oscillators is given by:

$$f_{yto}(MHz) = (100 - f_{mfd}(MHz)) \times N \times 2 \qquad (1)$$

Through proper selection of division ratio and MFD reference, it is possible to synthesize all frequencies between 2 and 8 GHz in sub-Hertz steps while maintaining a minimum reference frequency of 2.5 MHz and a maximum division ratio of 82.

Using this approach, the only microwave circuits required are a printed microstrip splitter, a low-cost GaAs HBT monolithic-microwave-integrated-circuit (MMIC) amplifier and the two HBT dividers. After the programmable divider, the PLL is simply an inexpensive very-high-frequency (VHF) loop, which can be produced with repeatable success using surface-mount-technology (SMT) techniques and simple shielding.

As an example of PLL operation, assume that an output frequency of 5.9985 GHz is desired. The YTO is pretuned to 5.9985 GHz ± 50 MHz via microprocessor control of a digital-to-analog-converter (DAC) driver to the YIG main coil. This signal is then divided by 2 to produce 3.0 GHz ± 25 MHz and then applied to the programmable HBT divider which is set to divide by 31 by the microprocessor. The output of the divider is:

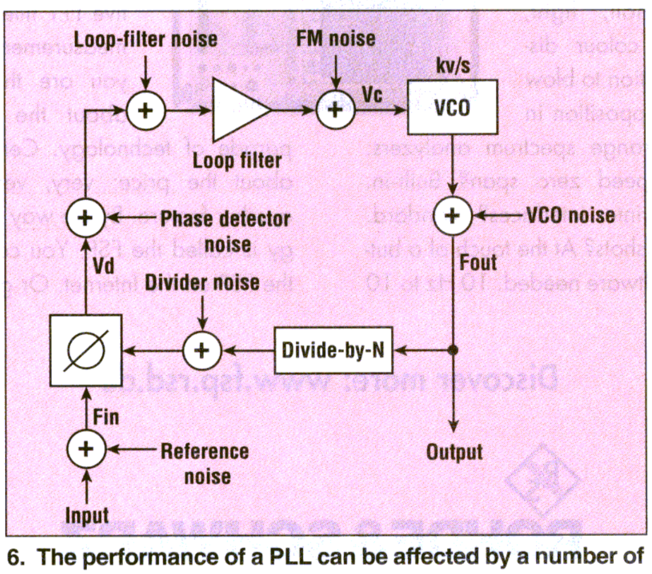
$$f_{div} = f_{yto}/2/31 = 96.75\ MHz \pm \approx 0.8\ MHz \qquad (2)$$

This output is then mixed with 100 MHz to produce 3.25 MHz, which is compared with the low-noise 3.25 MHz from the MFD chain. The phase-detector output is then converted to current to tune the frequency-modulation (FM) coil of the YTO into a locked condition.

HBT DIVIDERS

One enabling technology that makes the architecture achievable is that frequency dividers can operate in the microwave region. Fixed, base-2 dividers at 12 GHz are available from a number of manufacturers. This design uses the new HMMC-3122 from Hewlett-Packard, available in an inexpensive SSOP-8 surface-mount package.[4]

The programmable divider is a custom application-specific IC (ASIC) made by Rockwell Collins using a GaAs HBT process. It divides in integer steps from 4 to 31 and has a maximum input frequency of nearly 5 GHz. The input buffer stage of the divider takes a 0-dBm (nominal) single-ended input signal and converts it to a differential emitter-coupled-logic (ECL) output signal. The input stage has a gain of approximately 3 to increase rise time and improve noise performance. Figure 2 shows a plot of the residual noise for a 2-GHz input and 125-MHz output. The phase noise shown is the sum of two dividers, and 3



6. The performance of a PLL can be affected by a number of different noise sources, all of which can be measured and evaluated.

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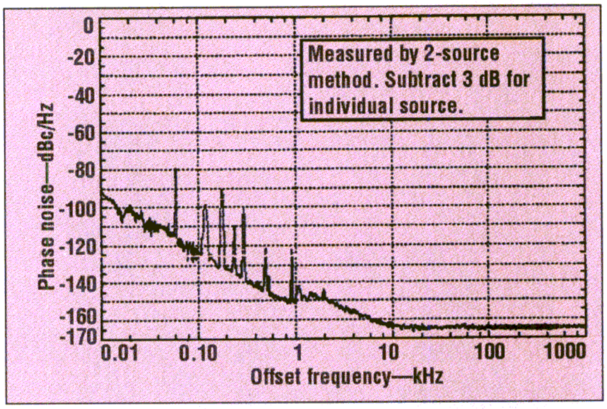
dB must be subtracted to determine the noise of each. As Fig. 2 shows, the residual noise is −150 dBc/Hz at 10-kHz offset.

The second stage is the divider proper, which uses a unique architecture (patent pending) to generate divider ratios other than a power of 2 (i.e., 5, 6, 7, etc.) with small path delays while maintaining a duty ratio of approximately 50 percent. For even-divider ratios, the duty ratio is exact. For odd-divider ratios, the duty cycle is within one count of 50 percent. For example, for a divide by 15, the count of the divider will be high for 8 counts and low for 7 counts. Figure 3 shows the spectrum for an output of 1.4 GHz/15 or 93.3 MHz. The output spectrum shows reasonable suppression of the even harmonics, indicating an output waveform approximating a square wave. The divide output-stage is a high-power buffer designed to drive 50 Ω at a level of 13 dBm.



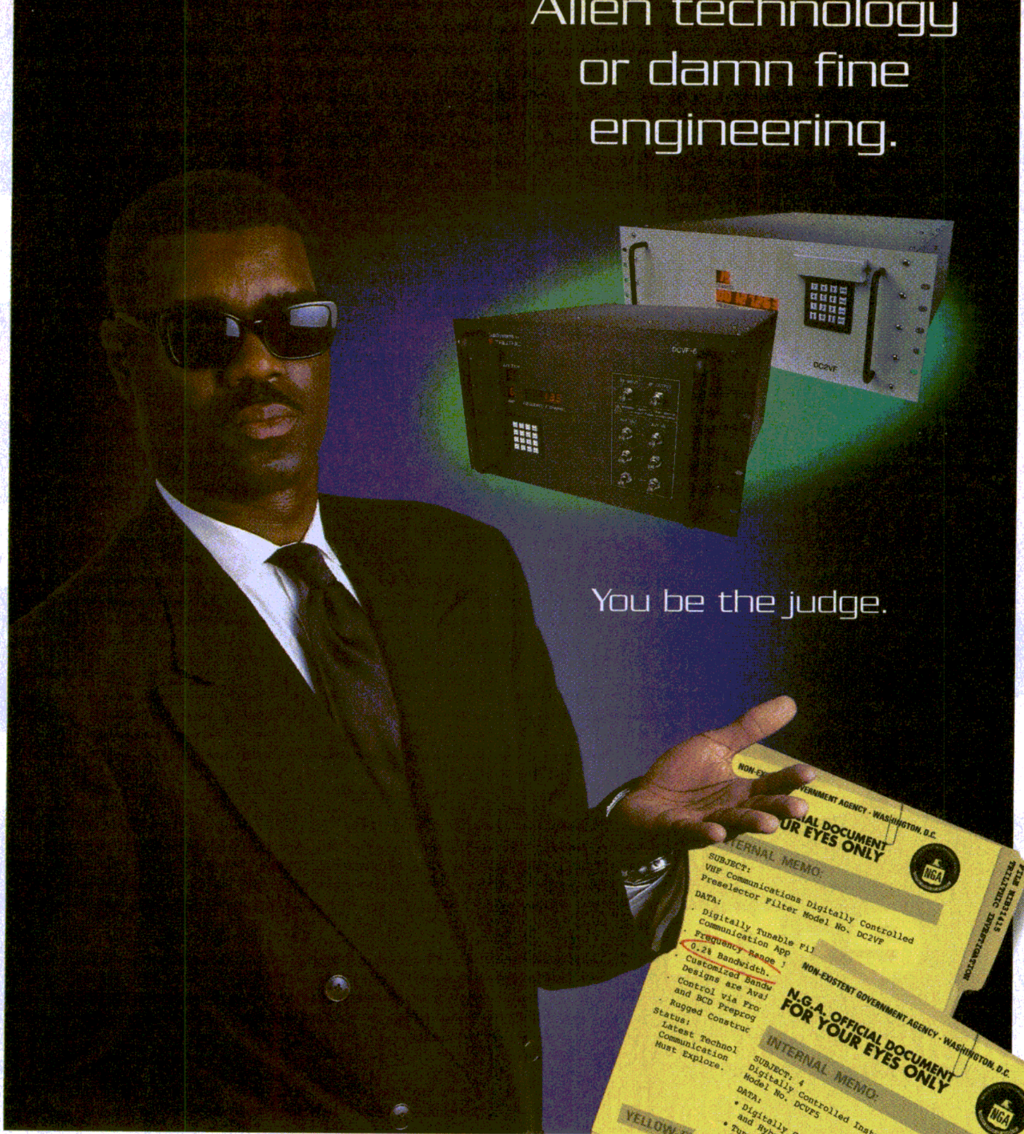
7. As an example of the measurement of a noise source shown in Fig. 6, this plot is of the phase noise measured for the 100-MHz reference.

THE MFD

The second key technology requirement of the synthesizer architecture is the ability to provide very small frequency steps without producing discrete spurious signals. This is accomplished by modulated fractional division of an ultra-high-frequency (UHF) clock.

Spurious levels and phase noise on the reference to the phase detector are increased by 20 log (N) to the synthesizer output, where N equals 82 at 8 GHz. This results in an increase of 38 dB in the noise and spurious noise at the synthesizer output relative to the reference input. Meeting a goal of −60 dBc of spurious noise at the synthesizer output requires the reference to have −98-dBc spurious noise. A modern direct digital synthesizer (DDS) would not meet the requirement across the entire 2.5-to-10.625-MHz reference range. But a sigma-delta modulated divider supports the fine tuning and spurious noise shaping required.

The concept of sigma-delta modulated division is not new. Several methods have been proposed in the literature.[5-10] Most applications center on fractional division within a PLL. In the proposed synthesizer, the MFD is not in the forward path of



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a PLL, but outside the PLL. In this manner, the MFD generates a fine-tuning reference similar to a DDS in other multi-loop synthesizers. Figure 4 shows a block diagram of the variable reference generator.

The idea behind MFD is to eliminate low-frequency phase error by rapidly switching the division ratio.

In this case, a phase-locked auxiliary oscillator at 840 MHz is used in conjunction with multi-modulus divider and a 4-stage sigma-delta modulator with a 24-b accumulator. The divider can take on any value between N −7 and N + 8. The digital correction causes the divide number to vary in a random fashion, producing pure noise. The output frequency of the divider is given by:

$$f_{mfd}(MHz) = (840 \times 10^6) \times 2^{24}/M \qquad (3)$$

where:

M is a 32-b number corresponding to the 8-b integer value and 24-b fractional divider value for the counter.

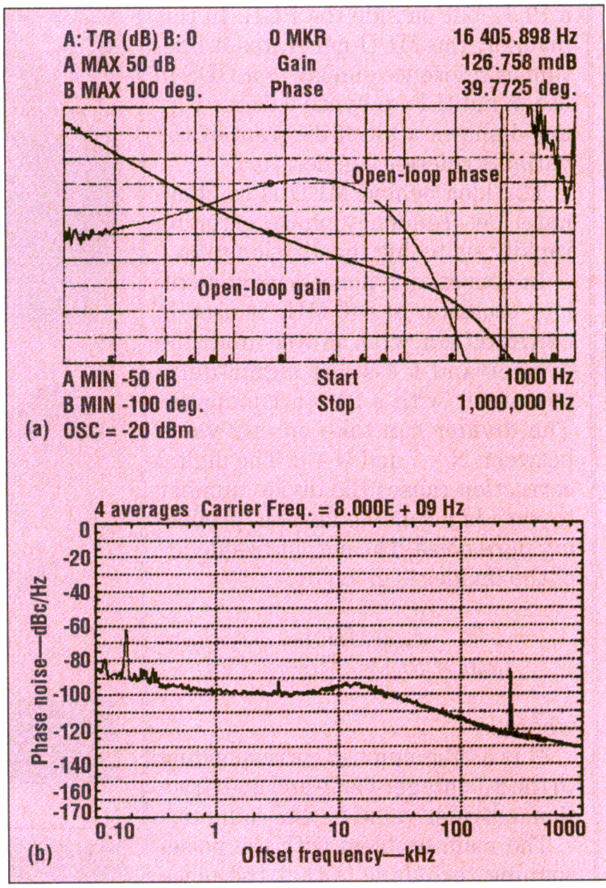
The main problem with the noise-shaping technique is that the noise power rises rapidly with offset frequency. Figure 5 shows a plot of the noise shaping present on a 10-MHz reference generated by the modulated fractional divider with different accumulator sizes. Proper selection of reference frequency and loop-filter bandwidth are required to produce the final desired phase-noise performance.

A well-known PLL noise model (Fig. 6) includes the individual sources of noise within a synthesizer.[12] Each source of noise arises from

Measured phase jitter of the microwave synthesizer

Synthesizer frequency (GHz)	Measured jitter (100 Hz to 20 MHz)
2	0.173 deg. RMS
3	0.177 deg. RMS
4	0.183 deg. RMS
5	0.206 deg. RMS
6	0.223 deg. RMS
7	0.252 deg. RMS
8	0.286 deg. RMS

a different mechanism, and all can be measured and evaluated. Referring to the measured noise of the HBT divider (Fig. 2) and MFD (Fig. 5), and using the measured 100-MHz reference phase-noise data shown in Fig. 7, it is possible to determine the composite phase noise of the reference sources within the loop. Using the phase-noise plot of the YIG oscillator, and knowing the loop-division ratio, phase-detector constant, and YIG-tuning constant, it is then a simple matter of setting the loop-filter values to produce the optimum phase noise.



8. The performance of the entire microwave synthesizer is best represented by measurement of its open-loop frequency response (a) and the phase noise (b). Both measurements were made on a physical prototype of the synthesizer at 8 GHz.

Phase noise was measured with a HP 11740A microwave phase-noise measurement system. Gain/phase measurements were made by using an HP 4194A impedance/gain-phase analyzer and an in-house Bode test fixture. Figures 8a and 8b show the measured open-loop gain/phase as well as the associated phase noise of the synthesizer at 8 GHz. The data agree with predicted values of open-loop gain and phase noise.

The measured jitter for the synthesizer is shown in the table for various tuned frequencies.

The entire synthesizer is used in a single-pitch 6U VME microwave downconverter that tunes from 2 to 8.4 GHz and provides 70- and 140-MHz outputs.

Present design enhancements include an integrated frequency doubler and extended-range YIG to produce output frequencies from 2 to 18 GHz, as well as low-noise ASIC technology development for the programmable dividers and phase detector. ●●

Acknowledgments

The author thanks Max Hawkins for his assistance with the HBT programmable divider and MFD. Thanks are also due to Steven Wilson for his assistance in the synthesizer measurement data.

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Design An Antenna For Pacemaker Communication

Computer optimization helps develop an antenna that is small enough to fit on a pacemaker battery pack.

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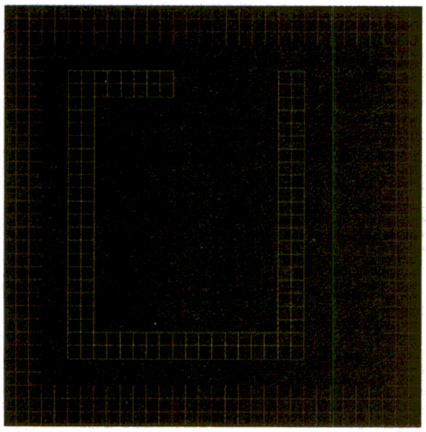
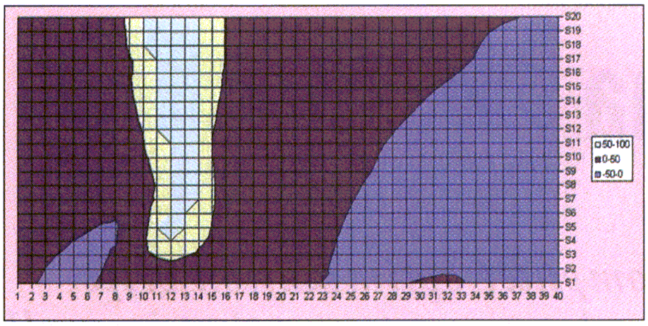
CARDIAC pacemakers are implanted inside the human body to monitor the heart's activity and take control when the heart rate falls below a programmed minimum, typically 60 beats per minute. Pacemakers are programmed with a pulse speed and stimulation waveform, and must be re-programmed to match the heart's changing condition. Pacemakers also collect useful diagnostic information, including the number of times the patient's heart requires excitation. Currently, the only way to re-program the pacemaker and collect diagnostic information from it is to use a large inductive coupler or to operate on the patient and remove the pacemaker from the body. In the latter case, the operation is performed infrequently–once every few years—to minimize trauma to the patient. It would be more desirable to download diagnostic information from the pacemaker and upload improved settings to it on a regular basis with a simple portable-communication device. This could be achieved by fitting a pacemaker with a miniature radio transceiver, allowing it to communicate with similarly equipped diagnostic, monitoring, and programming devices. The challenge is to design an antenna that is small enough to be unobtrusive in the body yet carry radio signals at a frequency that can penetrate body tissue. This article describes the design of a 2-in.^{2}, 433-MHz patch antenna that is small enough to fit on a standard-pacemaker battery pack. The design makes use of electromagnetic (EM) simulation software with an optimization engine.



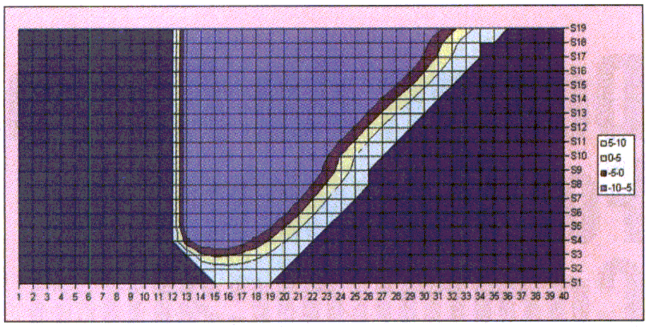
1. This antenna design length = 13 cells, which is an iSIGHT optimal design.

The first step in the design process was to find a frequency where radio signals could easily penetrate body tissue and communicate with external equipment, yet have a wavelength small enough to permit the use of a miniature, unobtrusive antenna, which is in itself small enough to fit on an existing pacemaker. For example, an antenna that is built to operate at an ultra high frequency of 2450 MHz would have a quarter wavelength of approximately 3 cm and could easily be manufactured small and unobtrusive. But at this frequency, radio signals can only penetrate a few centimeters of body tissue. On the other hand, radio signals at a frequency of 433 MHz would penetrate body tissue well enough to communicate with outside equipment, but antennas used at this frequency are normally 5 or 6 in. (12.7 or 15.24 cm) long. This type of antenna that is attached to a pacemaker would protrude into other parts of the body and risk infection or lung punctures. Nonetheless, the authors decided to develop a miniaturized 433-MHz antenna that could fit onto the battery pack of the pacemaker.

The authors chose to pursue a particular type of antenna design known as a microstrip antenna, which



2. Real component of the impedance (note: 50 Ω is good) can be seen.



3. Imaginary component of the impedance (note: close to zero is good) is noted here.

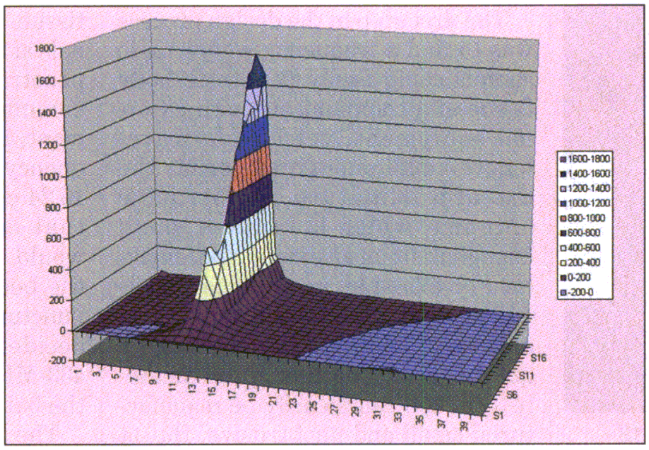
excites an arbitrarily shaped conductor on a dielectric substrate with a backplane conductor. Since there is no analytical solution for designing an insulated and arbitrary-shaped microstrip antenna that is embeded in lossy media, the authors used EM finite-difference, time-domain (FDTD) software known as XFDTD from Remcom, Inc. (State College, PA) to evaluate the performance of specific designs. They selected FDTD software because it is capable of analyzing conductors, lossy dielectrics, magnetics, anisotropic materials, biological tissues, ferrites, as well as many other materials. Furthermore, as problems become electrically complex, the FDTD method quickly becomes more efficient in terms of computer time and memory than other methods since no direct-matrix solution is required. FDTD can provide results for a wide spectrum of frequencies from only one calculation using transient-pulse excitation and Fast Fourier transform (FFT) analysis.

MANUAL SIMULATION

Even with the benefit of powerful simulation tools, the development of microstrip-antenna geometry meeting the stringent requirements of this application is a very difficult task. The authors gridded the microstrip antenna, cardiac pacemaker, and surrounding tissue with a 2-mm cell size. They created and analyzed 109 variations of this base model, primarily by varying the length of the patch, the location of the feed point, and the location of one or more grounding pins. After nine months of analysis, they finally developed two designs that met the requirements of the project—a U-shaped patch antenna and a spiral-shaped antenna.

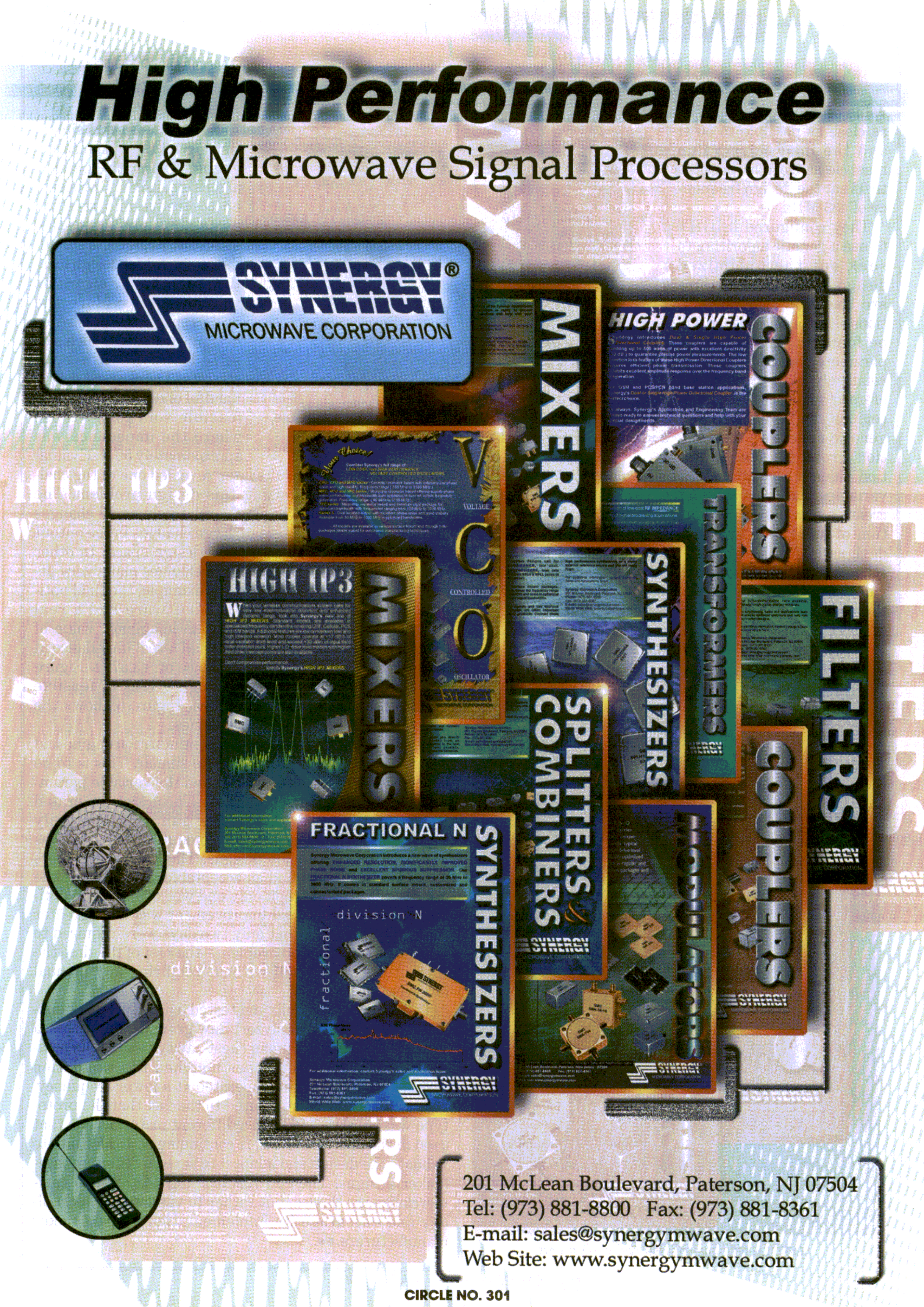
The work demonstrated the viability of the 433-MHz microstrip-antenna concept, but the authors recognized that the design would need to be re-engineered at least a few (and possibly many more) times to meet additional requirements that would arise as the project evolved. The authors did not feel the project could bear the time and expense of multiple iterations of the difficult manual-design process, so they employed an optimization software package called iSIGHT from Engineous Software, Inc. (Morrisville, NC). Essentially, iSIGHT replaces the manual, trial-and-error portion of the traditional design process with an automated, iterative procedure. The software automatically changes the input data, runs the XFDTD-analysis codes, assesses the output, and changes the input again based on instructions from an optimization algorithm chosen for the specific problem. The software optimizes the performance of the overall system while balancing conflicting design requirements and meeting all design constraints. It also provides graphical visualization of how the trade-offs in design parameters affect antenna performance.

In setting up the optimization problem, the authors created an interface between iSIGHT and XFDTD. This process was as simple as selecting the right parameters in the input and output files of XFDTD. The variables that they allowed iSIGHT to control were the length of the spiral antenna and the locations of the source and the ground pins. They set a maximum length of 40 cm and a maximum distance from source to ground of 20 cm. In this case, they fixed the shape of the antenna to a spiral. iSIGHT was given the latitude of selecting a design shape, which at the total length of 40 cm, was a spiral configuration and at minimal length, was an inverted L-shaped configuration. The design that iSIGHT selected was slightly longer than a full U-shaped configuration (Fig. 1).



4. The real feasible values between + or –50 are feasible, while values higher than 50 set to 50.

iSIGHT is also capable of optimizing the shape of the antenna, and the authors plan to use this feature on future designs. The output variables that the program used to evaluate the performance of each design iteration were the real and imaginary values of the antenna impedance. For a matched resonant antenna, the real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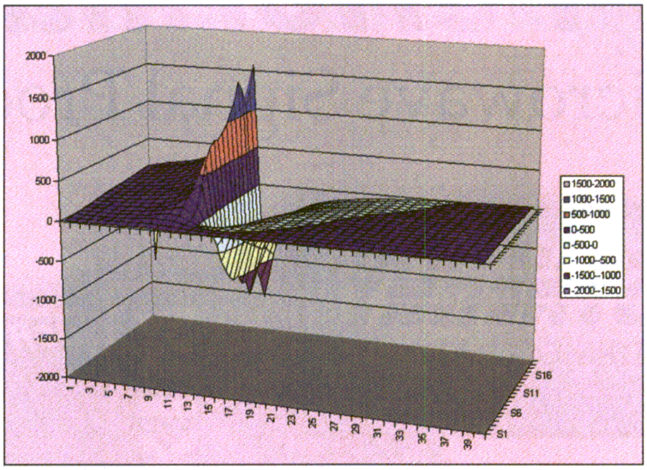
value of impedance must be as close as possible to 50 Ω and the imaginary value must be as close as possible to 0 Ω at the design frequency of 433 MHz.

They configured iSIGHT to use a genetic algorithm to seek out the optimized design. Genetic algorithms are based on two assumptions—the best solutions will be found in the regions of parameter space containing relatively high proportions of good solutions, and these regions can be explored by genetic operators of selection, crossover, and mutation. Genetic algorithms start with a population of design variables that are manipulated with genetic operators to create a new set or generation of designs. Each population of designs is evaluated and a new population of designs is selected based on a survival-of-the-fittest scheme. It is worth noting that the software package provides an optimization advisor that guides the user toward the best of several available techniques for a particular problem.



5. Imaginary feasible values between + or –10 are feasible, and values higher than 10 set to 10.

The next step was simply turning on the optimization engine. Using the genetic algorithm, iSIGHT proceeded to create and analyze approximately 100 to 300 different designs, depending on the initial design location. The genetic algorithm was used to explore the design space until a feasible design was found which satisfied the design constraints. There were only a handful of designs that were feasible solutions—those where the impedance was approximately 50 Ω. Evaluation of the analysis results later revealed one of the challenges of hand-optimizing this design problem. The design space is very "bumpy"—relatively small changes in the design variables create large changes in the output variables in some regions of the design space. In other regions of the design space, large changes make very little difference (Figs. 2 and 3). For example, the final optimized design selected by the software had a real impedance value of 52.99 Ω and an imaginary value of 11.41 Ω. Yet, an adjacent design had a real impedance of 90.38 Ω and an imaginary impedance of −3.57 Ω. Figure 4 is a design-space map showing regions of feasible designs for real impedance values. Figure 5 shows the same for imaginary impedance values.

The solutions developed by the optimization engine and the authors turned out to be quite similar. This is to be expected, since the design goals and the variable parameters were the same in both cases. The design generated by iSIGHT had slightly better gain. But the truly remarkable difference was the amount of time that was required to produce the designs. The human designers took nine months to find an acceptable design, while the optimization engine took only one week to find the optimum design. The actual improvement was even greater considering the fact that only two days were required to set up the optimization problem—the rest of the time the computer ran by itself without requiring any manual intervention.

The antenna was built and tested using a network analyzer and spectrum analyzer. Raw ground beef was used in order to simulate human chest-cavity tissue. The measurement of the prototypes matched the simulation results, except that they operated at a lower frequency. This difference was attributed to the fact that the prototype was constructed from a different material than that used in the analysis. Work is continuing at Utah State to validate the design and adapt it to several different pacemaker designs. The end result is intended not to commercialize the device, but rather to validate the technology and encourage its implementation by pacemaker manufacturers.●●

Reviewing The Basics Of Microstrip Lines

An understanding of the fundamentals of microstrip transmission lines can guide high-frequency designers in the proper application of this venerable circuit technology.

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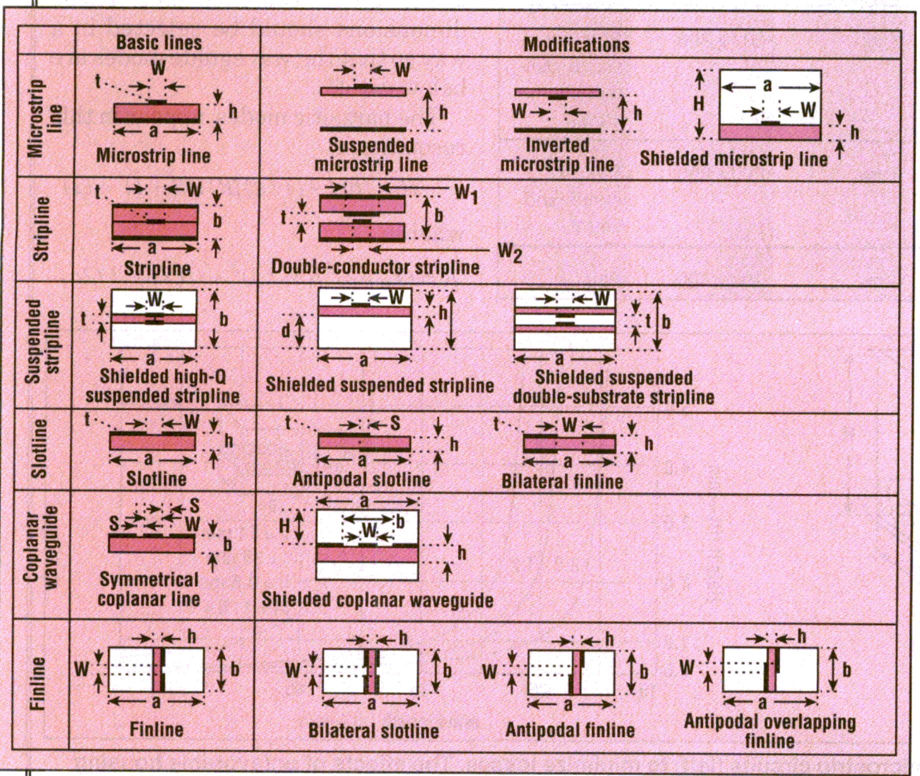
PRINTED transmission lines are widely used, and for good reason. They are broadband in frequency. They provide circuits that are compact and light in weight. They are generally economical to produce since they are readily adaptable to hybrid and monolithic integrated-circuit (IC) fabrication technologies at RF and microwave frequencies. To better appreciate printed transmission lines, and microstrip in particular, some of the basic principles of microstrip lines will be reviewed here.

A number of different transmission lines are generally used for microwave ICs (MICs) as shown in Fig. 1. Each type has its advantages with respect to the others. In Fig. 1, it should be noted that the substrate materials are denoted by the dotted areas and the conductors are indicated by the bold lines.



1. These are commonly used types of printed transmission lines for MICs.

The microstrip line is a transmission-line geometry with a single conductor trace on one side of a dielectric substrate and a single ground plane on the opposite side. Since it is an open structure, microstrip line has a major fabrication advantage over stripline. It also features ease of interconnections and adjustments.

In a microstrip line, the wavelength, Λ, is given by:

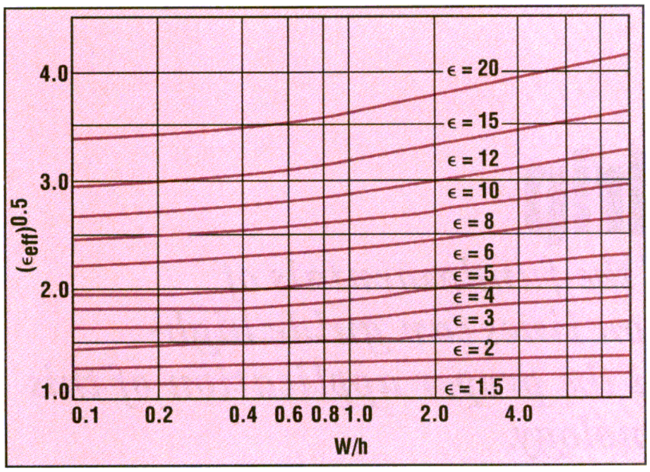
$$\Lambda = \lambda / (\varepsilon_{eff})^{0.5} \qquad (1)$$

where:

ϵ_{eff} = the effective dielectric constant, which depends on the dielectric constant of the substrate material and the physical dimensions of the microstrip line, and

λ = the free-space wavelength.

In a microstrip line, the electromagnetic (EM) fields exist partly in the air above the dielectric substrate and partly within the substrate itself. Intuitively, the effective dielectric constant of the line is expected to be greater than the dielectric constant



2. The values of effective dielectric constant are shown for different substrate relative dielectric constants as a function of W/h.

of air (1) and less than that of the dielectric substrate.[1] Various curves for effective dielectric constant are shown in Fig. 2 as a function of physical dimensions and relative dielectric constant.

Referring again to Fig. 1, it should be apparent that a basic (unshielded) microstrip line is not really a practical structure. It is open to the air and, in reality, it is desirable to have circuits that are covered to protect them from the environment as well as to prevent radiation and EM interference (EMI). Also, the microstrip configurations that have been so far discussed are transversally infinite in extent, which deviates from reality. Covering the basic microstrip configuration with metal top plates on the top and on the sides leads to a more realistic circuit configuration, a shielded microstrip line with a housing (Fig. 1).

The main purposes of the housing or package are to provide mechanical strength, EM shielding, germetization, and heat sinking in the case of high-power applications. Packaging must protect the circuitry from moisture, humidity, dust, salt spray, and other environmental contaminants. In order to protect the circuit, certain methods of sealing can be used: conductive epoxy, solder, gasket materials, and metallization tape.

An MIC mounted into a housing may be looked on as a dielectrically loaded cavity resonator (Fig. 3, left) with the following inner dimensions: a is the width, l is the length, and H is the height of the enclosure. These dimensions should be selected in a way so that the waveguide modes are below cutoff.

The parasitic modes appear in this resonator if:

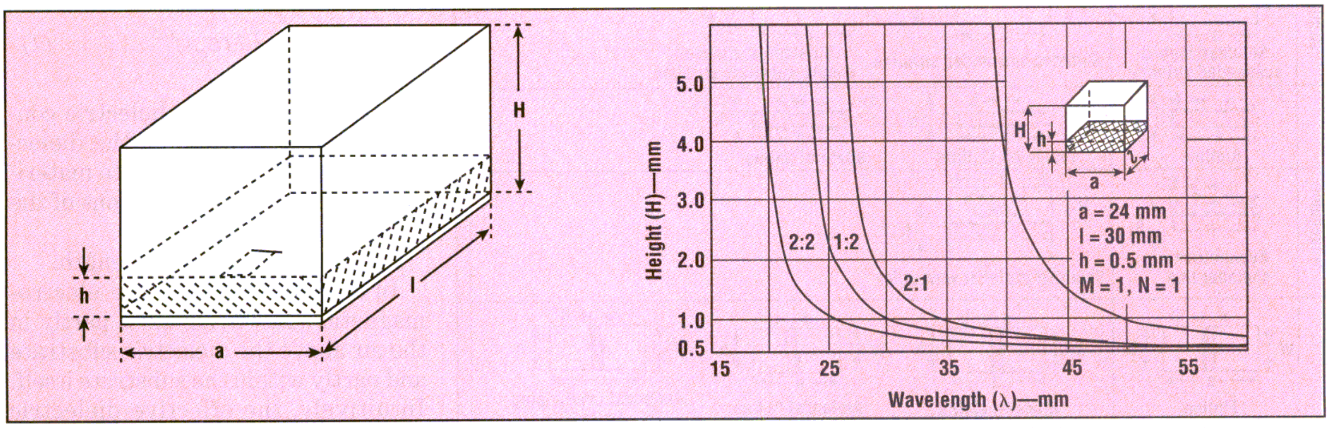
$$H = \{h[1-(1/\varepsilon)]R\}1(R-1) \quad (2)$$

where:

$$R = (\lambda_0/2)^2[(M/l)^2 + (N/a)^2] \quad (2a)$$

A comparison of various transmission-line types

Transmission line	Q factor	Radiaton	Dispersion	Impedance range	Chip mounting
Microstrip (dielectric) (GaAs, Si)	250 100 to 150	Low High	Low	20 to 120	Difficult for shunt, easy for series
Stripline	400	Low	None	35 to 250	Poor
Suspended stripline	500	Low	None	40 to 150	Fair
Slotline	100	Medium	High	60 to 200	Easy for shunt, difficult for series
Coplanar waveguide	150	Medium	Low	20 to 250	Easy for series and shunt
Finline	500	None	Low	10 to 400	Fair



3. Housing dimensions are selected for microstrip circuits (left) to minimize losses. The effects of unfavorable housing height versus wavelength and different parasitic modes is shown (right).

and M and N = positive integers.

From eq. 2, it is possible to obtain the condition of absence of parasitic modes:

$$R - 1 < 0 \,;\, R < 1$$

or

$$\lambda_0^2 < 4 / [(M/1)^2 + (N/a)^2] \quad (3)$$

or

$$\lambda_0 < 2/)[(M/1)^2 + (N/a)^2]^{0.5} \quad (4)$$

Equation 4 is known as the condition for wave propagation in a waveguide with dimensions l × a. In the case of this article, it can also be considered the condition for the absence of parasitic modes in a waveguide of cross-section a × H or l × H. If eq. 4 is not satisfied, parasitic modes can arise, and the height H must be chosen to suppress these modes. Figure 3 (right) illustrates the resulting graphs of unfavorable H versus λ_0 for housing dimensions of a = 24 mm, l = 30 mm, and dielectric substrate with a dielectric constant of 9.8 and THK of 0.5 mm.

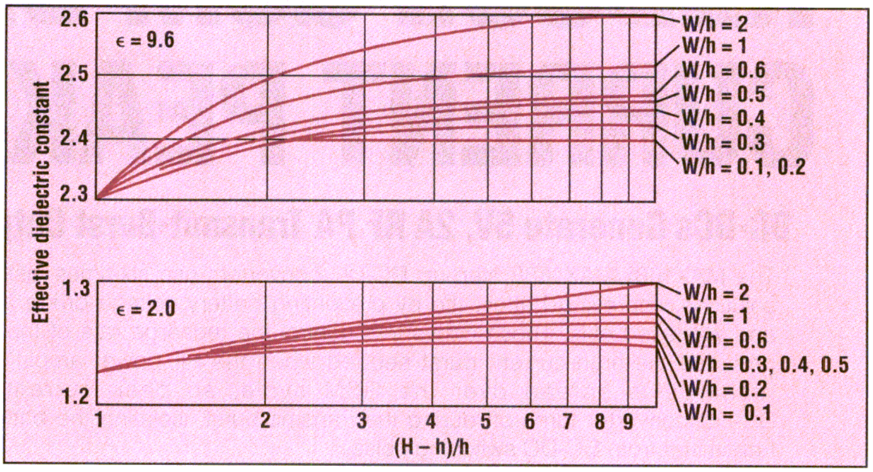
The top and side covers essentially redistribute the field of the more theoretical microstrip and understandably have an influence on the effective dielectric constant.

Figure 4 shows the relationship between the effective dielectric constant and the physical dimensions of the shielded microstrip line for different values of the relative dielectric constant of the substrate material.[2] In these curves, it has been assumed that the side walls are sufficiently spaced so that they only see weak fringing fields and, therefore, have a negligible effect on the effective dielectric constant. The top cover tends to lower the effective dielectric constant (which is consistent with intuition). The top wall enables electric fields in the air above the strip conductor thereby giving the air more influence in determining the propagation characteristics.

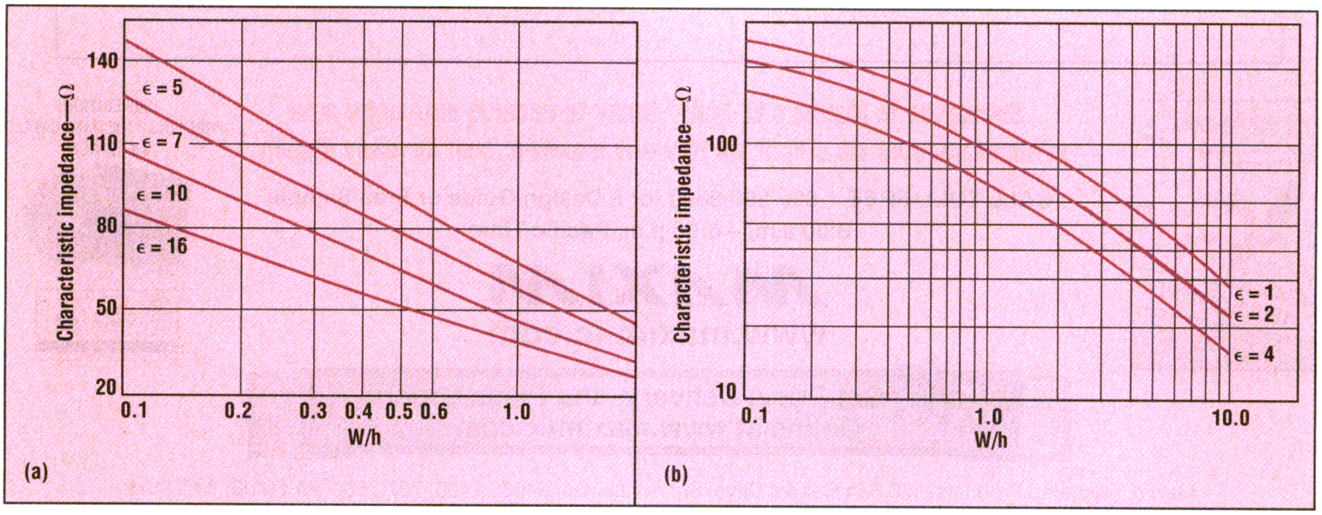
The characteristic impedance of a microstrip line may be approximately calculated by assuming that the EM field in the line has a quasi transverse-EM (TEM) nature. The characteristic impedance of a microstrip line can be calculated using the Wheeler equations.[3,4]

Figure 5 shows the characteristic impedance of microstrip lines for various geometries and substrates of different relative dielectric constants while Fig. 6 illustrates the relationships between characteristic impedance and the physical dimensions of shielded microstrip lines for two examples: substrates with low (2) and high (9.6) relative dielectric constants.[2] The top cover tends to reduce the impedance. When the ratio of the distance from the top cover to the dielectric substrate and the substrate thickness [(H – h)/h] is greater than 10, the enclosure effects can be considered negligible. The characteristic impedance range of a microstrip line is 20 to 120 Ω. The upper limit is set by production tolerances while the lower limit is set by the appearance of higher-order modes.

There are three types of losses that occur in microstrip lines: con-



4. The effective dielectric constant is shown as a function of the relative dielectric constant and physical dimensions for a shielded microstrip line.



5. The characteristic line impedance has been plotted for substrates with high (a) and low (b) dielectric constants.

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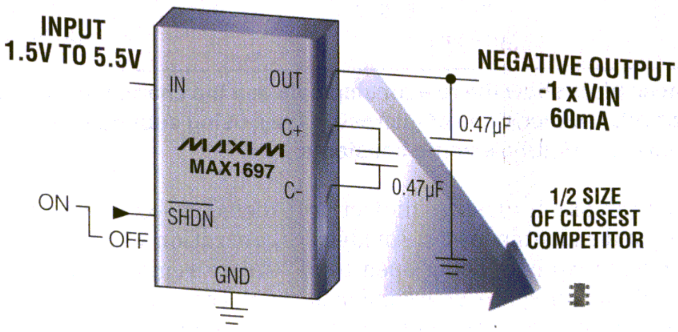

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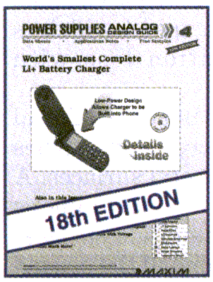

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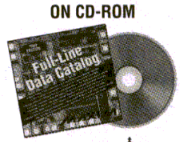

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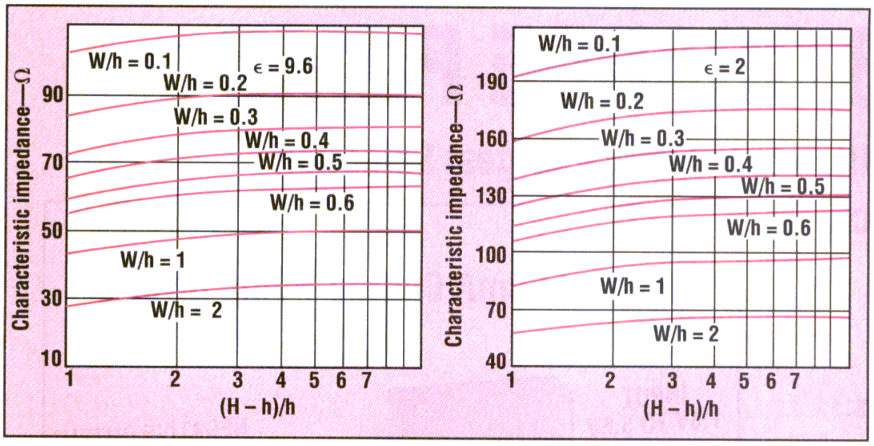


6. These plots show the relationship between the characteristic impedance and the physical dimensions of microstrip lines using substrates with high (9.6, left) and low (2.0, right) dielectric constants.

ductor (or ohmic) losses, dielectric losses, and radiation losses. An idealized microstrip line, being open to a semi-infinite air space, acts similar to an antenna and tends to radiate energy. Substrate materials with low dielectric constants (5 or less) are used when cost reduction is the priority. Similar materials are also used at millimeter-wave frequencies to avoid excessively tight mechanical tolerances. However, the lower the dielectric constant, the less the concentration of energy is in the substrate region and, hence, the more are the radiation losses. Radiation losses depend on the dielectric constant, the substrate thickness, and the circuit geometry.

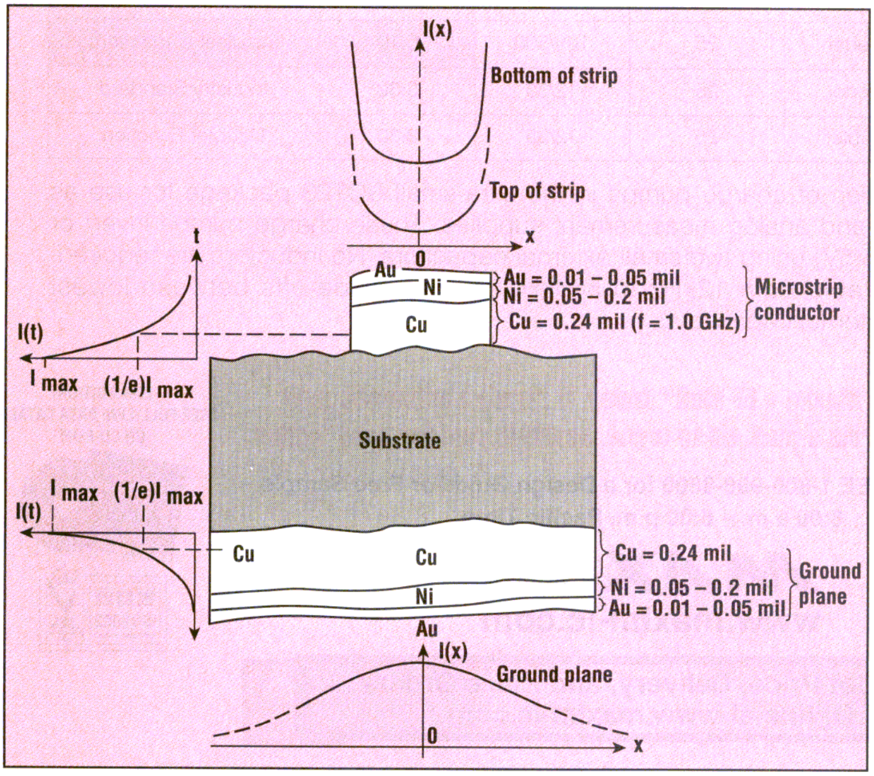
The use of high-dielectric-constant substrate materials reduces radiation losses because most of the EM field is concentrated in the dielectric between the conductive strip and the ground plane. The real benefit in having a higher dielectric constant is that the package size decreases by approximately the square root of the dielectric constant. This is an advantage at lower frequencies but may be a problem at higher frequencies.

In most conventional microstrip designs with high substrate dielectric constant, conductor losses in the strip conductor and the ground plane dominate over dielectric and radiation losses. Conductor losses are a result of several factors related to the metallic material composing the ground plane and walls, among which are conductivity, skin effects, and surface roughness. With finite conductivity, there is a non-uniform current density starting at the surface and exponentially decaying into the bulk of the conductive metal. This is the alleged skin effect and its effects can be visualized by an approximation consisting of a uniform current density flowing in a layer near the surface of the metallic elements to a uniform skin depth, δ. The skin depth of a conductor is defined as the distance to the conductor (Fig. 7) where the current density drops to 1/e from a maximum current density of I_{max}, or 37 percent of its value at the surface of the conductor.

To minimize conductor loss while simultaneously minimizing the amount of metallic material flanking the dielectric, the conductor thickness should be greater than approximately three to five times the skin depth.

In a microstrip line, conductor losses increase with increasing characteristic impedance due to the greater resistance of narrow strips. Conductor losses follow a trend which is opposite to radiation loss with respect to W/h.

The fabrication process of real microstrip devices creates scratches and bumps on the metal surfaces. A cross-section of a microstrip line is shown in Fig. 7. The inside surfaces of the strip conductor and the ground plane facing the substrate repeat the shape of the substrate. The current, concentrated in the metal surface next to the substrate, follows the uneven surface of the substrate and encounters a greater resistance com-



7. This cross-sectional view shows the current distribution across a microstrip conductor and its ground plane.

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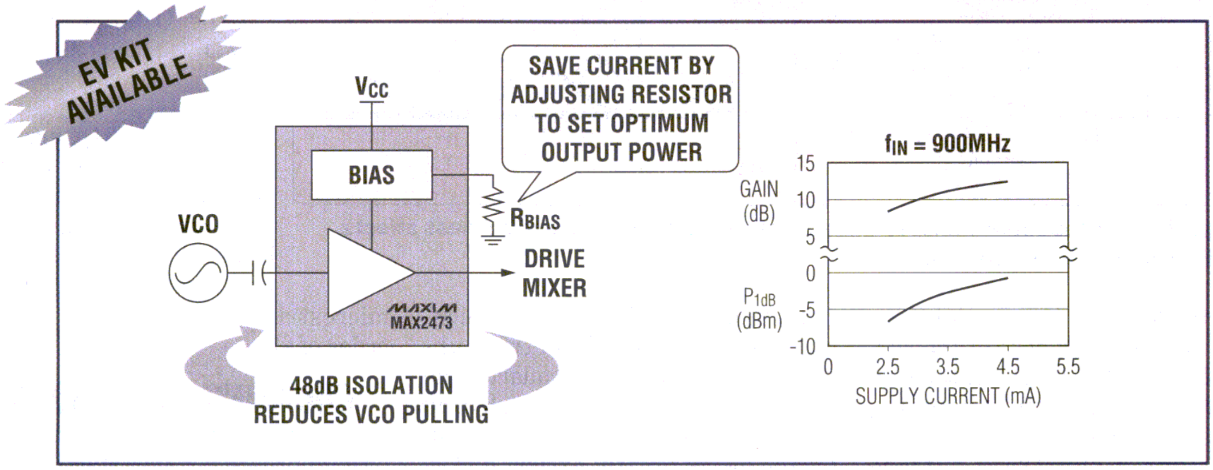

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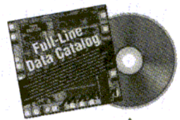

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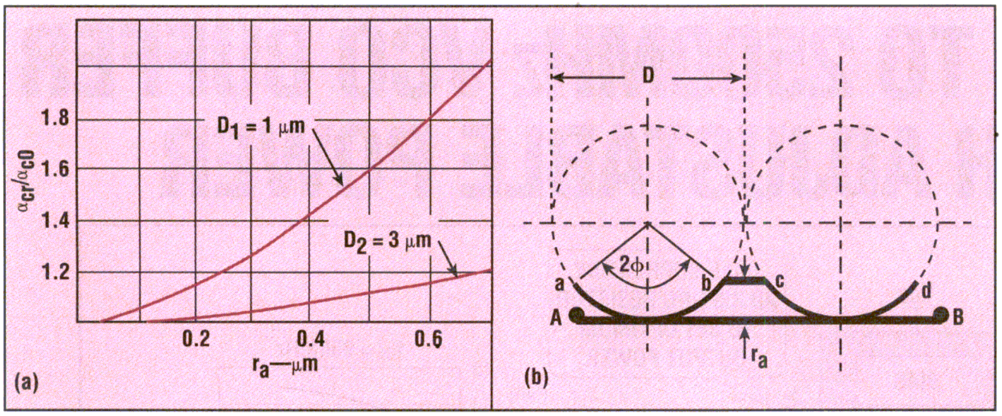


8. The profile of a substrate's uneven surface (a) shows how surface roughness affects normalized conductor losses (b).

pared to the case of a smooth substrate. As the roughness of the surface increases, the length of the current path increases and, therefore, the losses increase.

Consider a substrate surface which, for example, coincides with the shape of the diamond abrasive material that is used to polish the substrate. The path of the current in conductor segment a-d (Fig. 8a) is shown by the line abcd. For an ideally smooth surface, the length of the current path AB is: $I_{AB} = Dn$
where:

n = the number of diamond abrasives within segment AB.

The ratio of conductor losses in the case of an uneven surface, α_{cr}, to losses in the case of a perfectly smooth surface, α_{c0}),[2] is:

$$\alpha_{cr} / \alpha_{co} = 1 + \arccos\,[1-(4r_a/D)] \\ -2\{[(2r_a/D)[1-(2r_a/D)]\}^{0.5} \quad (5)$$

Using eq. 5, α_{cr}/α_{c0} can be plotted as a function of r_a for $D_1 = 1\ \mu m$ and $D_2 = 3\ \mu m$ (Fig. 8b). Analysis of the resulting functions shows that for smaller diameters, conductor losses in the microstrip line are more dependent on the unevenness of the substrate roughness because the extra path length a surface (or skin) current sees is less. For example, consider a copper (Cu) microstrip line with sapphire substrate where typically the roughness is 1 μm.[5] The skin depth at a few gigahertz is 1 μm and the loss is increased approximately 60 percent when the surface roughness is taken into account.

To minimize dielectric losses, high-quality, low-loss dielectric substrates, such as alumina, quartz, and sapphire, are typically used in hybrid ICs. For most microstrip lines, conductor losses greatly exceed dielectric losses. However, in monolithic microwave ICs (MMICs), silicon (Si) or GaAs substrates result in much larger dielectric losses (approximately 0.04 dB/mm).[5]

The preceding sections have considered the individual contributions to losses in microstrip by radiation, ohmic, and dielectric effects. These individual loss components are at most first-order perturbations in the overall EM wave propagation and, consequently, can be combined linearly. To do so, it is convenient to consider the total Q factor, which can be expressed by:

$$1/Q = (1/Q_c) + (1/Q_d) + (1/Q_r)$$

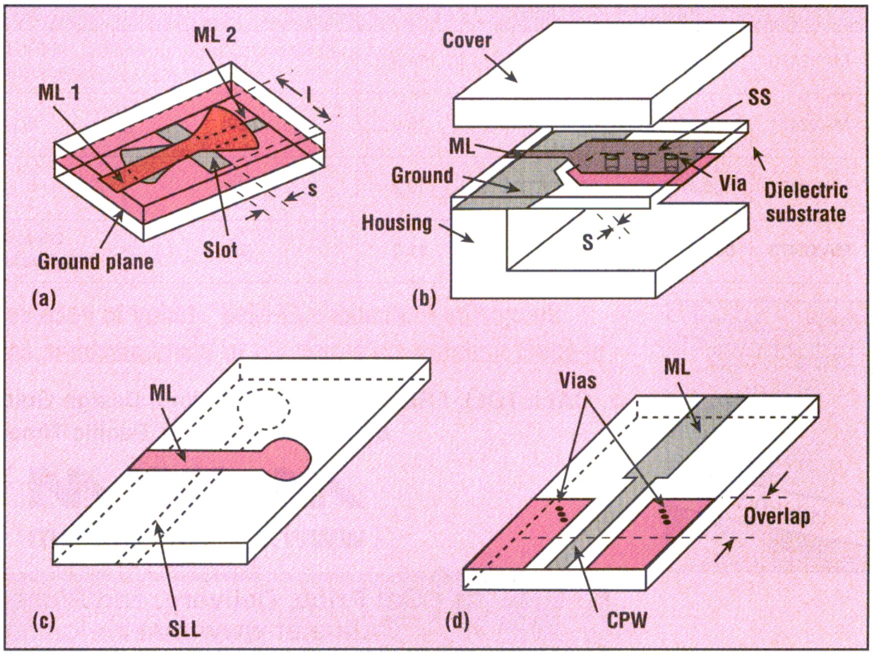
where:

Q_c, Q_d, and Q_r are the quality factors corresponding to the conductor, dielectric, and radiation losses, respectively. The unloaded Q factor of the microstrip line is typically on the order of 250.

CHOOSING DIMENSIONS

For all circuit considerations, a basic approach involves starting with the particular ranges of dimension ratios required to achieve a desired characteristic impedance. Following that, the strip width should be minimized to decrease the overall dimensions, as well as to suppress higher-order modes. It is important to remember, however, that a smaller strip width leads to higher losses.

Factors that affect the choice of substrate thickness are the most contro-



9. Various transitions between microstrip and other circuit structures are possible: microstrip to microstrip (a), microstrip to suspended stripline (b), microstrip to slotted line (c), and microstrip to coplanar waveguide (d).

versial. The positive effects of decreasing substrate thickness are compact circuits, ease of integration, less tendency to launch higher-order modes or radiation, and via holes drilled through the dielectric substrate will contribute smaller parasitic inductances to the overall performance.

However, a decrease in the substrate thickness (h) while maintaining a constant characteristic impedance, Z_0, must be accompanied by a narrowing of the conductor width, W. Narrowing W leads to higher conductor losses along with a lower Q. Also, for smaller W and h, the fabrication tolerances become more severe. Careless handing of thin substrates can cause stress and strain which can modify the performance of the substrate.

Microstrip circuit dimensions decrease with increasing substrate dielectric constant. Losses then usually increase because higher dielectric constant materials usually have higher loss tangents, tan δ, and also because for the same characteristic impedance, reduced conductor line widths have higher ohmic losses. This is a typical conflicting situation between the necessary requirements for small dimensions and low loss. For many applications, lower dielectric constants are preferred since losses are reduced, conductor geometries are larger (and, therefore, more producible), and the cutoff frequency of the circuit increases.

MICROSTRIP TRANSITIONS

The rapid development of high-density modules requires the design of interconnects and transitions, especially for multilayer circuits. Consider useful transitions from microstrip to other printed transition lines. A transition between two microstrip lines (Fig. 9a) can be realized through a slot in the ground plane.

A transition between a microstrip line and a suspended stripline circuit is shown in Fig. 9b.

A transition between a slotline and a microstrip line can be seen in Fig. 9c.[7,8]

An overlay transition between a microstrip line and coplanar waveguide (CPW) is shown (Fig. 9d).[9,10] ••

Acknowledgment

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GaAs MMIC Switch Is Designed Around Low-Capacitance MESFETs

A single-pole, eight-throw (SP8T) GaAs switch uses low capacitance MESFETs to provide low insertion loss and input VSWR which is lower than competitive SP8Ts.

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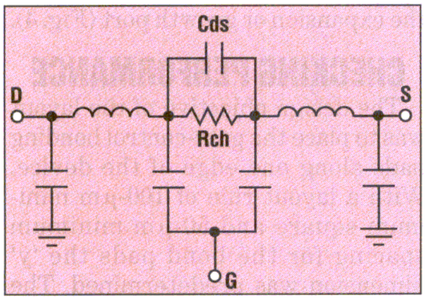


1. The basic MESFET switch contains an equivalent resistance (Rch) and capacitance (Cds) that must be compensated for in order to make an effective switch.

Table 1: MESFET gate length versus Rch, Cds, insertion loss, and isolation

Gate length (µm)	Cds (pF)	Rch (Ω)	Insertion loss at 2 GHz (dB)	Isolation at 2 GHz (dB)
60	0.0048	52	–3.60	–36
100	0.008	31.2	–2.40	–32
200	0.016	15.2	–1.30	–26
400	0.032	7.8	–0.65	–20
600	0.048	5.2	–0.45	–17
800	0.064	3.9	–0.33	–14
1000	0.080	3.12	–0.27	–12
1200	0.096	2.6	–0.22	–11

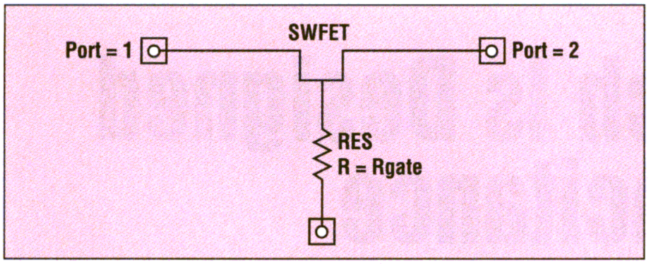
SUITABLE for the commercial market, a unique single-pole, eight-throw (SP8T) gallium-arsenide (GaAs), monolithic-microwave-integrated-circuit (MMIC) switch has been developed. The switch has a common arm match when all ports are biased in the isolation state and a growth port for expansion and integration into a higher-order multi-throw switch. It is packaged in a plastic, QSOP-28, surface-mount package.

The design of a higher-order multi-throw switch requires the addition of a series capacitive element for each throw at the junction. In this case, that element is the drain-to-source capacitance (Cds) of a metal-semiconductor field-effect transistor (MESFET). This capacitance limits the usable frequency response of the device with respect to insertion loss and input VSWR. The challenge in the design is to select a MESFET with sufficiently small gate length in order to not degrade insertion loss and VSWR performance over the target operating band. Simultaneously, the on-arm MESFET channel resistance, Rch, cannot be too large or it will adversely affect insertion loss.

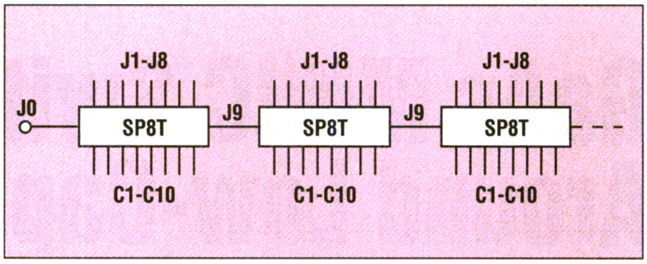
MESFETs

The effects of gate length on Cds and Rch versus insertion loss and isolation over the target frequency of DC to 2 GHz are analyzed through the basic MESFET model (Fig. 1). In Table 1, the gate length varies from 60 to 1200 µm in approximately 200-µm increments. As the gate length increases, Cds increases while Rch decreases. A single-series MESFET circuit model was generated with the Hewlett-Packard Eesof Libra modeling software as shown in Fig. 2. The circuit was simulated to observe the effects of gate length on insertion loss and isolation versus frequency. The results are summarized in Table 1 at a frequency of 2 GHz. Nominal values were chosen for the gate length, except for the 50-Ω condition where the length was extrapolated. A feature of the SP8T is to have the common port matched when all ports are biased off. This is accomplished by adding another throw to the SP8T with the series MESFET connected to ground. This MESFET has a Rch value of approximately 50 Ω at the corresponding gate length of 60 µm.

Switch matrices and arrays can be designed by cascading the SP8Ts as shown in Fig. 3. This can be accomplished by connecting an output port of one switch to the common-arm input of another switch. The output port (J9) in this application is designated the expansion or "growth" port. To balance the arm-to-arm insertion loss of this cascaded net-



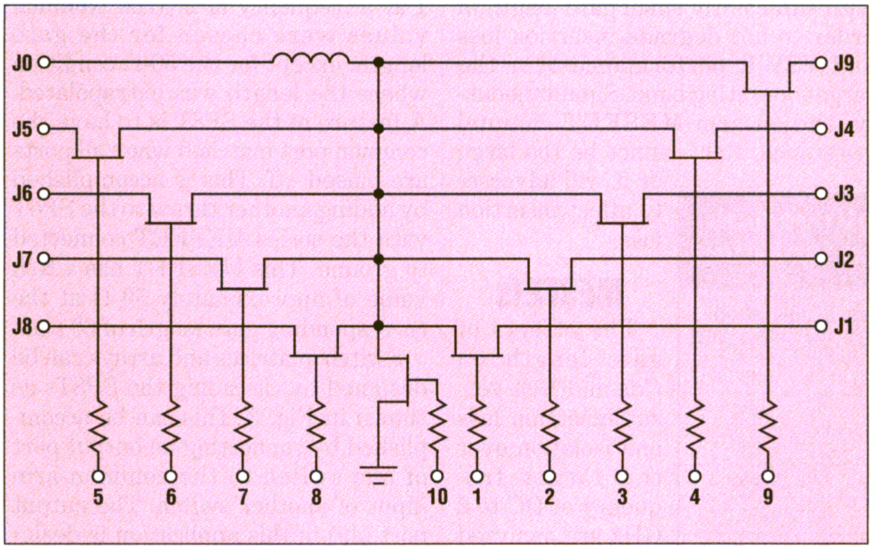
2. A simulation of the switch in Fig. 1 is this Libra model which is used to analyze the effects of gate length on insertion loss and isolation versus frequency.



3. SP8T GaAs MMIC switches can be cascaded by connecting the output port of one switch to the input port of the following switch.

Table 2: Worst-case data for key switch parameters

Parameter	Test conditions	Typical data
Insertion loss	DC to 1.0 GHz	1.8 dB
	1.0 to 2.5 GHz	2.7 dB
Insertion-loss growth port	DC to 1.0 GHz	1.2 dB
	1.0 to 2.5 GHz	2.1 dB
Isolation	DC to 1.0 GHz	29 dB
	1.0 to 2.5 GHz	22 dB
Isolation growth port	DC to 1.0 GHz	23 dB
	1.0 to 2.5 GHz	17 dB
VSWR input	DC to 2.5 GHz	1.4:1
VSWR output	DC to 2.5 GHz	1.5:1
VSWR input terminated	DC to 2.5 GHz	2.0:1
Switching speed	90 percent/10 percent to 10 percent/90 percent RF	10 ns
Input power for 1-dB compression	+20 dBm	+20 dBm
Input IP2	Two tone inputs 5 MHz apart at +5dBm	+60 dBm
Input IP3	Two tone inputs 5 MHz apart at +5dBm	+45 dBm
Gate-control voltage	Insertion-loss	0 VDC
	Isolation	–5 VDC



4. A schematic of the SP8T switch shows the connection between the input port (J0) and the output or growth port (J9) as shown in block diagram form in Fig. 3. The common arm (pin 10) controls the action of all switches on the device.

work, the growth port is designed to have lower insertion loss than the other arms. This can be achieved by selecting a MESFET with a lower Rch and higher Cds. For this application a 400-µm MESFET was selected. With these parameters, an SP8T switch model was constructed.

A 200-µm FET was determined to have the necessary Cds/Rch balance to work at the target frequency of 2 GHz. The 200-µm MESFET was used as the series element for the throws of the switch, a 60-µm MESFET serves for the common arm match and a 400-µm MESFET for the expansion or growth port (Fig. 4).

CHECKING PERFORMANCE

The design objective of the layout was to place the gate-control bonding pads along one edge of the device. With a layout rule of 100-µm minimum square and 50-µm minimum spacing for the bond pads the 'y' dimension was predetermined. The RF ports were placed around the perimeter of the die with the common port near the center. The final size is 0.090 × 0.048 in. (2.25 ×1.20 mm). The difference in distance between the junction and MESFETs was not a factor in the model results or subsequent die testing.

The MMIC is offered to the commercial market assembled in a quad small outline 28-pin plastic package (QSOP-28) for surface mounting. The package was mounted on an FR4 multi-layer board and characterized at the temperatures of −40°C, +25°C, and +85°C. The worst-case data for the key parameters of the SP8T are given in Table 2. ●●

Acknowledgments

The author would like to thank Gerry Cornwell, Chris Weigand, Scott Gatley, Ken McPherson, and Gary Tran for their contributions in developing this product.

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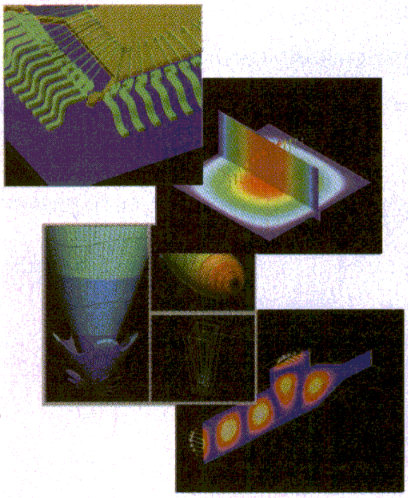

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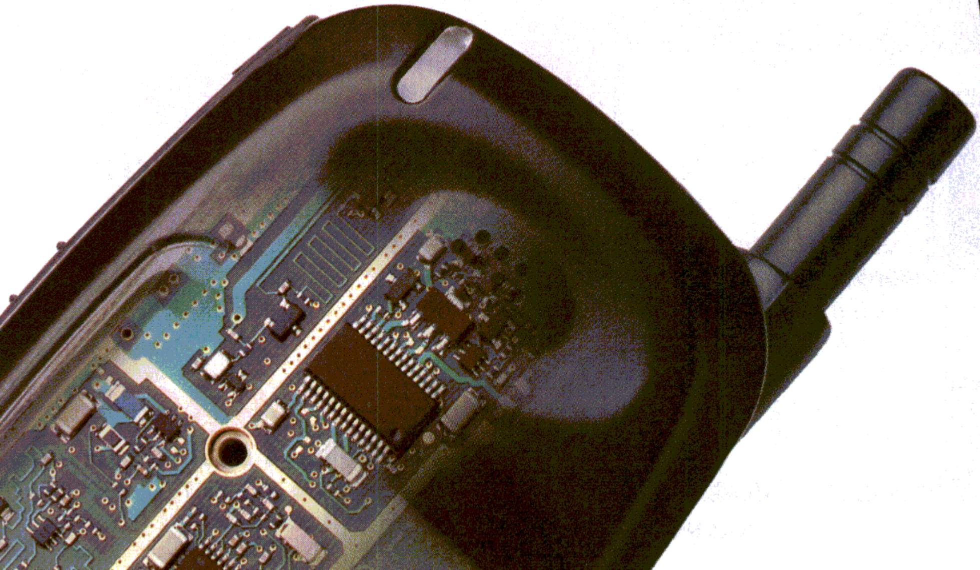

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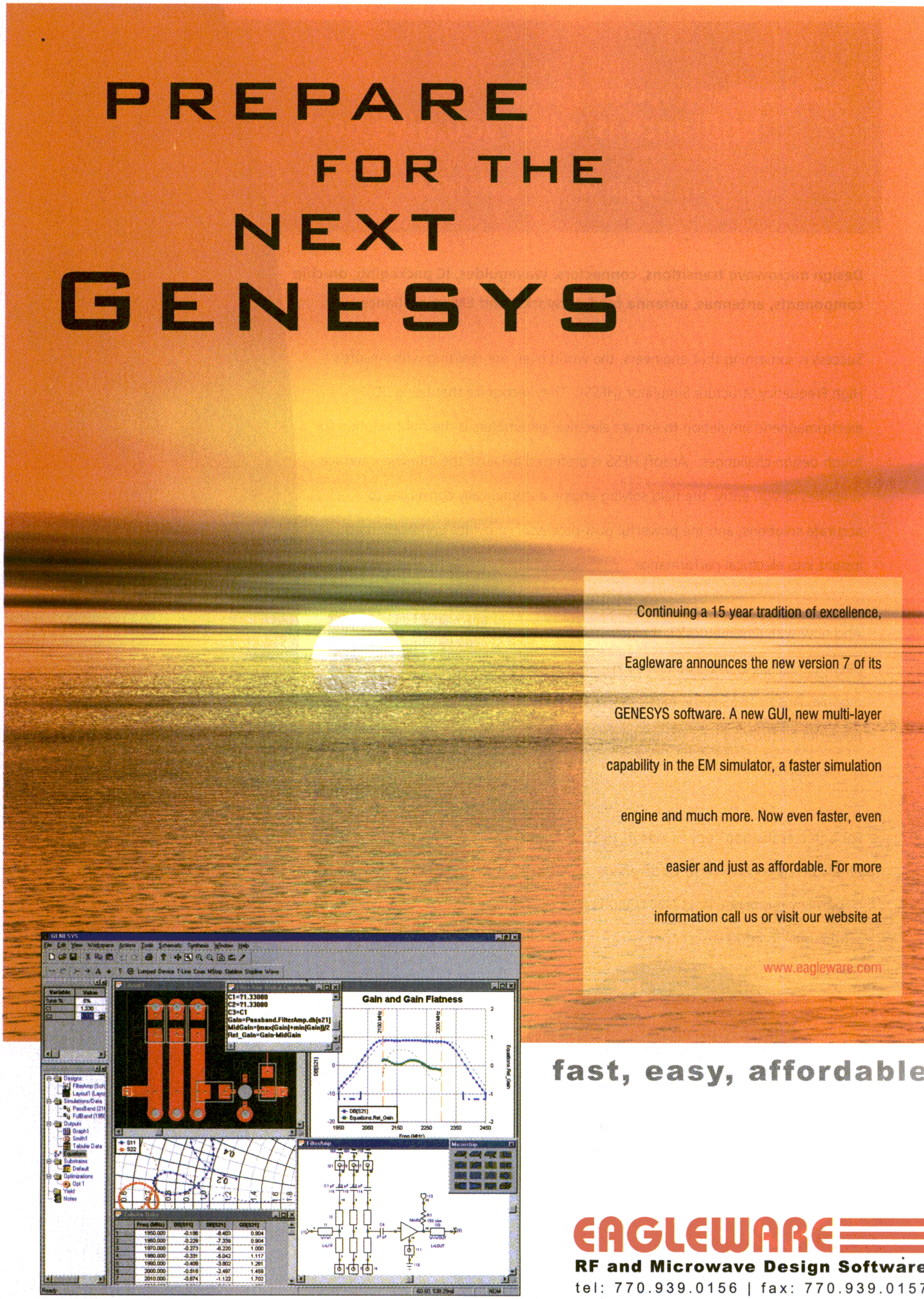

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Extend DDS Bandwidth Above The Nyquist Limit

Extend DDS Bandwidth, Part 2

Several methods can be used to isolate and amplify direct-digital-synthesis (DDS) frequency components above the Nyquist limit.

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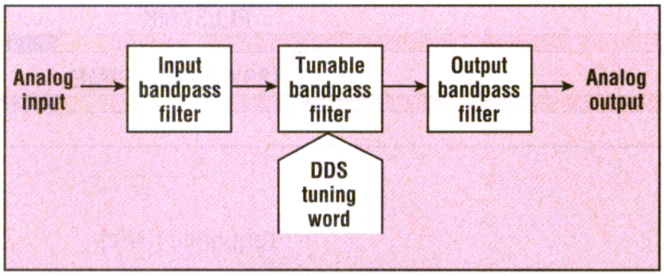
IN Part 1 of this article (*Microwaves & RF,* January 2000, p. 82), two methods were presented to isolate and amplify frequency components in Complete Direct Digital Synthesizer (C-DDS) device output spectrums that are above the Nyquist limit [bandpass and surface-acoustic-wave (SAW) filters]. Part 2 describes the tunable tracking filter, an approach that holds the most promise in preserving DDS advantages at greater than Nyquist frequencies. Good printed-circuit-board (PCB) layout can reduce any parasitic problems encountered. With integration, better transconductors can replace the small-signal buffers in the buffered bandpass filters to enable larger output signal swings.

A tunable tracking filter that uses the DDS input tuning word to tune the filter to isolate and amplify any desired output frequency is shown in Fig. 7.

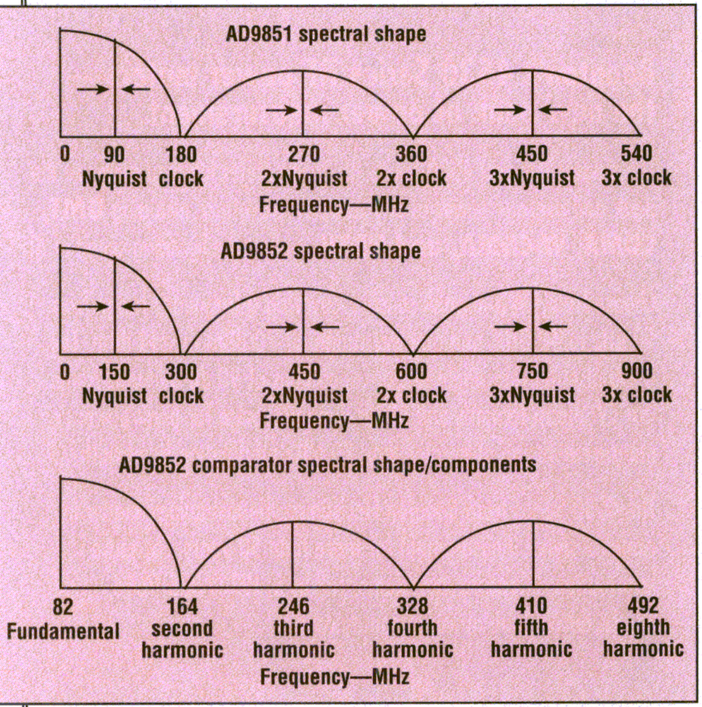
The composite filter uses the filtered-in/filtered-out approach of the previously described SAW filter solution (part 1) and consists of five distinct filters. Four of these filters are fixed, passive, LC lowpass, and highpass filters (a lowpass and highpass filter combined to create a buffered bandpass filter). The lowpass filter is designed to define the maximum frequency that the composite filter is to pass while the highpass is designed to define the composite filter's lowest passband frequency. The fifth filter is an active filter, which uses the DDS tuning word to select frequency bands to be amplified inside the passband defined by the fixed LC filters.

This filter uses the DDS input tuning word to tune the filter to isolate and amplify any desired output frequency.

By properly defining the characteristics of each filter used in the design, the composite filter can be used with any of the C-DDS synthesizer products produced by Analog Devices to pass and amplify any aliased image product or harmonic frequency that is generated by these parts. It is necessary only to define the characteristics of the fixed-frequency filters used in the design, select tuning methods



7. The tunable tracking filter is tuned by the input DDS tuning word which allows it to isolate and amplify any frequency.



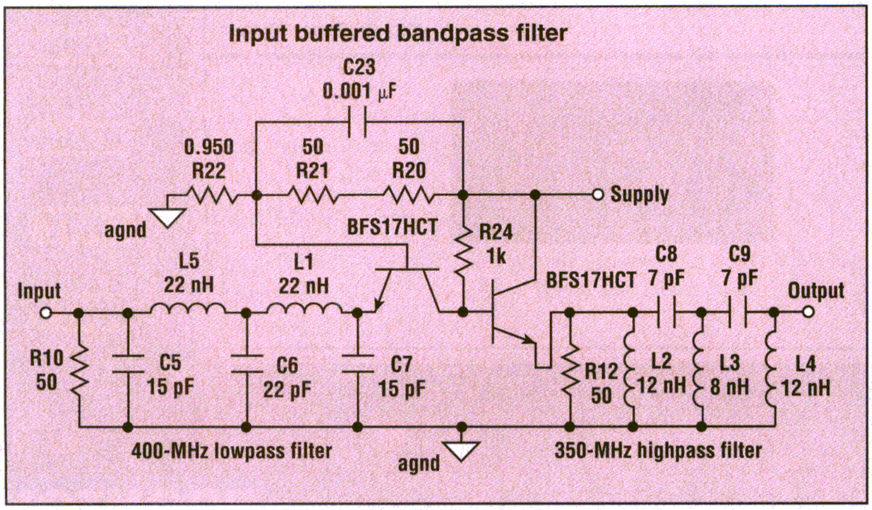
8. These Sin (x)/x characteristics of the AD9851/9852 can be modified by a composite filter to isolate and amplify any alias or harmonic frequency in the band between the Nyquist limit and 450 MHz.

for the filter (the AD9852/54 have an onboard control DAC which can be used to generate a tuning voltage) and select an appropriate inductor to define the tuning range of the central bandpass filter.

FILTER DESIGN

The composite filter can be used to modify the DAC output of any C-DDS device to isolate and amplify any alias or harmonic in the frequency bands between the Nyquist limit and 450 MHz. Figure 8 displays the Sin (x)/(x) characteristics of each of these devices (using the maximum suggested clock rate as a base). It is necessary only to select the clock rate, determine the Nyquist frequency for the selected clock rate, and define the band of frequencies to be amplified.

Once any of the mentioned parameters are selected, the fixed filter characteristics can be determined. The best approach is to consider all frequency bands that do not fall within the confines of the Nyquist rate or its multiple, and the clock rate or its multiple, reject bands. Thus, the Nyquist rate or its multiple and the clock rate or its multiple become the −3-dB breakpoint frequencies of the fixed LC filters. These filters are best designed with



9. The input filter of the bandpass filter uses a common-base amplifier as a buffer between the lowpass and highpass filters.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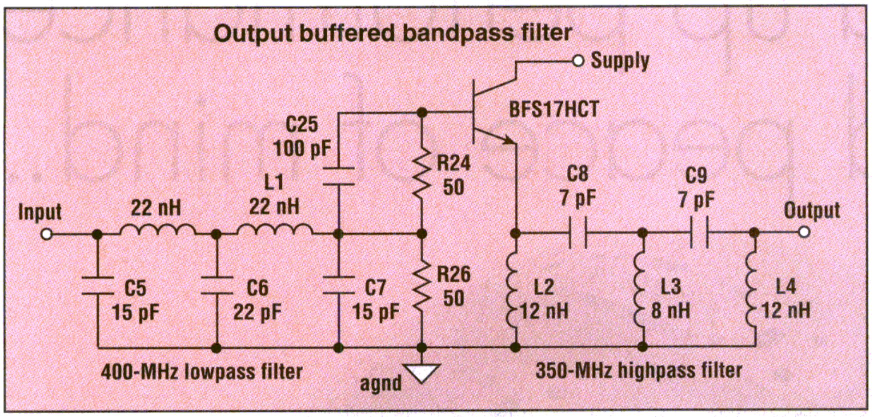


10. At the output of the bandpass filter, an emitter-follower stage is used to buffer the interface between the lowpass and highpass filters.

the aid of MathCAD or some other mathematical software package.

Once these filters are designed, a lowpass and highpass filter are combined to create a bandpass filter. This is to avoid the problems previously mentioned when implementing LC bandpass filters. Buffer amplifiers (Zetex BFS17HCT NPN transistors) are used to buffer the interface between the lowpass and highpass filters.

Due to differing input biasing conditions at the buffered bandpass filter inputs, different buffering schemes must be used for the input and output filters. The input buffered bandpass filter (Fig. 9) uses a common-base amplifier to buffer the interface between the lowpass and highpass filters. The output buffered bandpass filter (Fig. 10) uses an emitter-follower stage as a buffer at the lowpass/highpass interface.

Once the fixed LC filters are designed, the adjustable bandpass filter can be designed. The adjustable bandpass filter (Fig. 11) uses a series of three tuned stages (to take advantage of the bandwidth shrinkage factor) which are gang tuned to produce the adjustable passband of the composite filter.

Bandlimiting in each of these stages is accomplished by LC tank circuits, each using a Zetex varactor diode as a variable tuning capacitance (ZC830). The inductors chosen for these tanks set the tuning range. At resonance, each tank circuit is designed to have a parallel impedance of 25 Ω. This combined with the 75-Ω resistor in the output circuit allows the RF amplifiers to produce maximum gain at resonance.

Referring to amplifiers 1 and 2 in Fig. 11, the tank circuit output is directly fed to an attenuation network, which has a high enough impedance to preserve the tank's Q. The output of each attenuation network is AC-coupled to the next stage, which is the RF amplifier input. These attenuation networks are necessary to prevent the following stage from being overdriven and are designed to be adjustable for producing a variable gain per stage. The gain is set to prevent clipping and distortion in the varactor diodes (i.e., to minimize phase noise), prevent distortion in the RF amplifiers, and adjust for other nonidealities of the circuit layout.

The third and final RF stage is directly coupled to an NPN emitter follower stage which is used to drive current into the output buffered bandpass filter. A diode-connected transistor is used to provide a diode drop in voltage aiding in biasing the emitter follower stage in the output buffered bandpass filter. Equations used in the design are as follows:

The resonant frequency of the individual tank circuits is:

Resonant frequency of individual tank circuits

$$f_0 = \frac{1}{2\pi\sqrt{\left(L + L_{par}\right)\left(\frac{(C_{var})(C_{ser})}{(C_{var}) + (C_{ser})}\right)}} \quad (4)$$

where:

L = the selected inductor,

L_{par} = the parasitic inductor,

C_{var} = the varactor diode capacitance,

C_{ser} = 15-pF capacitance,

f_0 = the resonant frequency, and

r_s = the series resistance of L and L_{par}.

The tank Q is:

$$\text{Tank } Q \qquad Q_0 = \frac{2\pi f_o(L + L_{par})}{r_s} \qquad (5)$$

The bandwidth shrinkage factor at resonance is:

$$B = \frac{\omega_0}{Q_0}\sqrt{\sqrt[N]{2} - 1} \qquad (6)$$

where:

$\omega = 2\pi f_0$, and N = the number of stages

Substituting Eq. 5 for Q_0 provides:

or

$$B = \frac{r_s}{(L + L_{par})}\sqrt{\sqrt[N]{2} - 1} \qquad (7)$$

The gain of the AC circuit at resonance is:

$$A_v = \frac{V_{out}}{V_{in}} = \left(\frac{V_{A1in}}{V_{in}}\right)\left(\frac{V_{A1out}}{V_{A1in}}\right)\left(\frac{V_{A2in}}{V_{A1out}}\right)\left(\frac{V_{A2out}}{V_{A2in}}\right)\left(\frac{V_{A3in}}{V_{A2out}}\right)\left(\frac{V_{out}}{V_{A3in}}\right) \qquad (8)$$

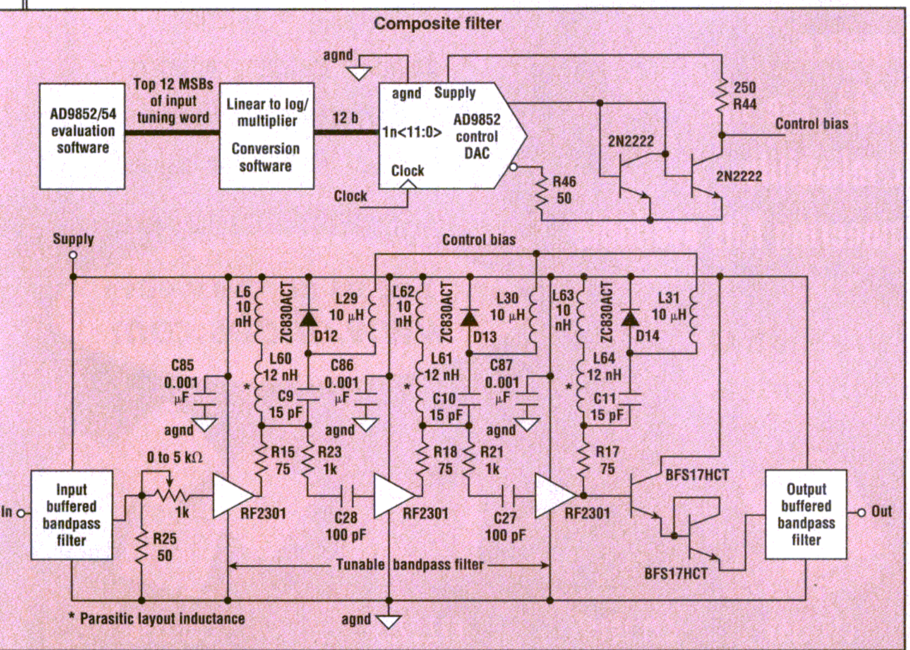
$$A_v = \left(\frac{R_2}{R_1 + R_2}\right)(A_{v1})\left[\left(\frac{R_4}{R_3 + R_4}\right)\left(\frac{R_6}{R_5 + R_6}\right)\right](A_{v2})\left[\left(\frac{R_8}{R_7 + R_8}\right)\left(\frac{R_{10}}{R_9 + R_{10}}\right)\right](A_{v3}) \qquad (9)$$

$$A_{vmin} = (.4)(7)[(.25)(.782)](7)[(.25)(.782)](7) = 5.24 \qquad (10)$$

$$A_{vmax} = (1)(7)[(.25)(.782)](7)[(.25)(.782)](7) = 13.10 \qquad (11)$$

A COMPOSITE FILTER

The circuit in Fig. 11 shows an experimental implementation of the composite filter previously described. It was designed to modify



11. Three tuned stages are gang tuned to produce the adjustable passband of the composite filter. LC tank circuits are used together with a varactor diode which serves as a variable tuning capacitance. The frequency band is 350 to 450 MHz.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the output spectrum of the DAC of the AD9852. The frequency band chosen for this filter is 350 to 450 MHz.

The 350-MHz highpass filter is a fifth-order filter with a with reject band attenuation of −100 dB. The transition band for this filter ranges from 100 to 350 MHz with a 100-dB per decade rolloff.

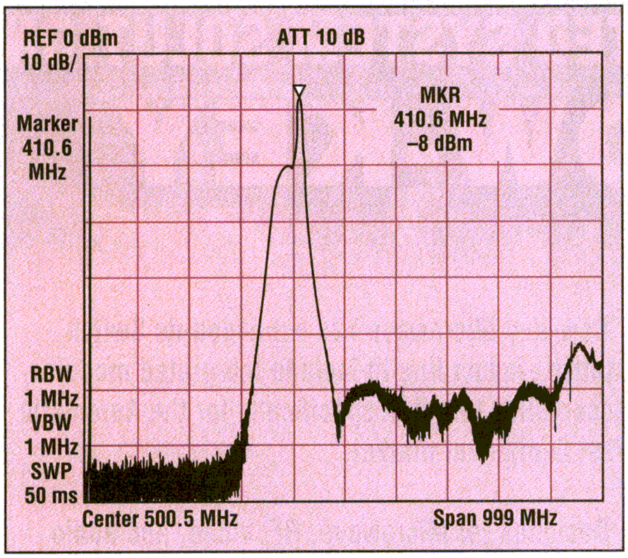
The 400-MHz lowpass filter is also a fifth-order filter with reject band attenuation of −100 dB. The breakpoint of 400 MHz was chosen to provide maximum attenuation of the 450-to-600-MHz image, since it is a negative image and moves in a direction that is opposite that of the positive image (Fig. 8). The transition band for this filter ranges from 400 to 1800 MHz and has a 100 dB per decade rolloff.

In combination, the highpass and lowpass filters produce the filter passband from 350 to 400 MHz. The tunable, central bandpass filter was designed to have a tuning range from 350 to 450 MHz. As is shown in Fig. 11, it consists of three identical, series stages to take advantage of the bandwidth shrinkage factor, cutting each individual stage's bandwidth by half.

The AD9852 DAC or comparator output current is fed directly into the input buffered bandpass filter. The output of the buffered LC bandpass filter is directly coupled to an adjustable attenuation network (consisting of a 5-kΩ variable input resistor and the input impedance (3.6 kΩ of the RF amplifier) and the central bandpass filter's first RF stage. The tunable tank circuit resides on the output of the RF amplifier, and consists of a 10-nH inductor, 15-pF capacitor, and the Zetex varactor diode. The varactor diode capacitance ranges from 15 pF at 0-VDC reverse bias to approximately 8 pF at +3.3-VDC reverse bias. The varactor and 15-pF capacitor combination provides an effective tank capacitance change of 5.0 to 7.5 pF. The inductor selected is smaller than required for the 350-to-450-MHz bands to compensate for approximately 12-nH board inductance present in the implementation of the circuit. The attenuation network consists of a 1-kΩ resistor placed in series with a 100-pF coupling capacitor, which is in series with the RF amplifier input. The output of the final RF stage drives an emitter follower stage which drives current through a diode-connected transistor into the input of the output buffered bandpass filter.

The active, tunable, central bandpass filter is tuned through the top 12 b of the DDS tuning word, which is processed through a linear-to-log conversion/multiplier routine (to compensate for the logarithmic change of the capacitance of the varactor diode and/or multiply the DDS tuning word by a fixed integer factor to track harmonics of the input device). The new processed tuning word is then latched into the control DAC of the AD9852. The AD9852 control DAC output is a current source and its output is mirrored through a 1:1 current mirror (using 2N2222 transistors) to a gain set resistor which develops the tuning voltage for the filter (the DAC complementary output may be used if the negative image is to be selected). The 10-µH inductors in series with the anodes of the varactor diodes are present to provide AC isolation from the circuit signal path.

The filter prototype was constructed on an etched copperclad pressboard and evaluated using a Rhode and Schwarz FSEA 1065.6000 spectrum analyzer, an Advantest Spectrum analyzer with an onboard tracking generator, the AD9852, and its associated hardware and software. A spectral plot of the composite filter transfer function is shown in Fig. 12. The plot was generated using a 2-GHz Advantest Spectrum analyzer using its onboard tracking generator. The output of the tracking generator was −17 dBm and was flat except for the low megahertz ranges. Inspection of the plot reveals four distinct frequency bands. A primary passband region ranges from 350 to 425 MHz and possesses −3-dB attenuation. This is the buffered bandpass filter passband. A second passband region is seen to range from 410 to 426 MHz (16-MHz passband) and possesses +10-dB gain. This corresponds to the tunable filter passband.



12. The spectral plot of the composite filter shows a primary passband region from 350 to 425 MHz and a secondary passband region from 410 to 426 MHz. The reject bands are from 0 to 300 MHz and above 450 MHz.

frequency alias remnants caused by nonideal filtering of the fixed output lowpass filter and PCB layout parasitics.

USING HARMONICS

The AD9852 and AD9854 C-DDS devices contain a comparator that can drive 50-Ω loads. If the input signal to this comparator is properly lowpass filtered, many DDS/DAC nonideal performance results can be reduced. The output of the comparator can be used to drive the Fig. 12 composite filter to suppress the fundamental frequency and amplify harmonics of the output square wave.

A low-frequency reject band region is found to range from 0 to 300 MHz and possesses approximately −75-dB attenuation. This is the highpass filter-reject band. A high-frequency reject band is found to reside above 450 MHz with a worst-case attenuation of −55 dB. This corresponds to the lowpass filter-reject band.

The tuning range of the tunable bandpass filter is 350 to 450 MHz. Overall composite filter gain ranges from 0 to 10 dB. The degradation of the lowpass filter reject band was caused primarily by circuit-board parasitics.

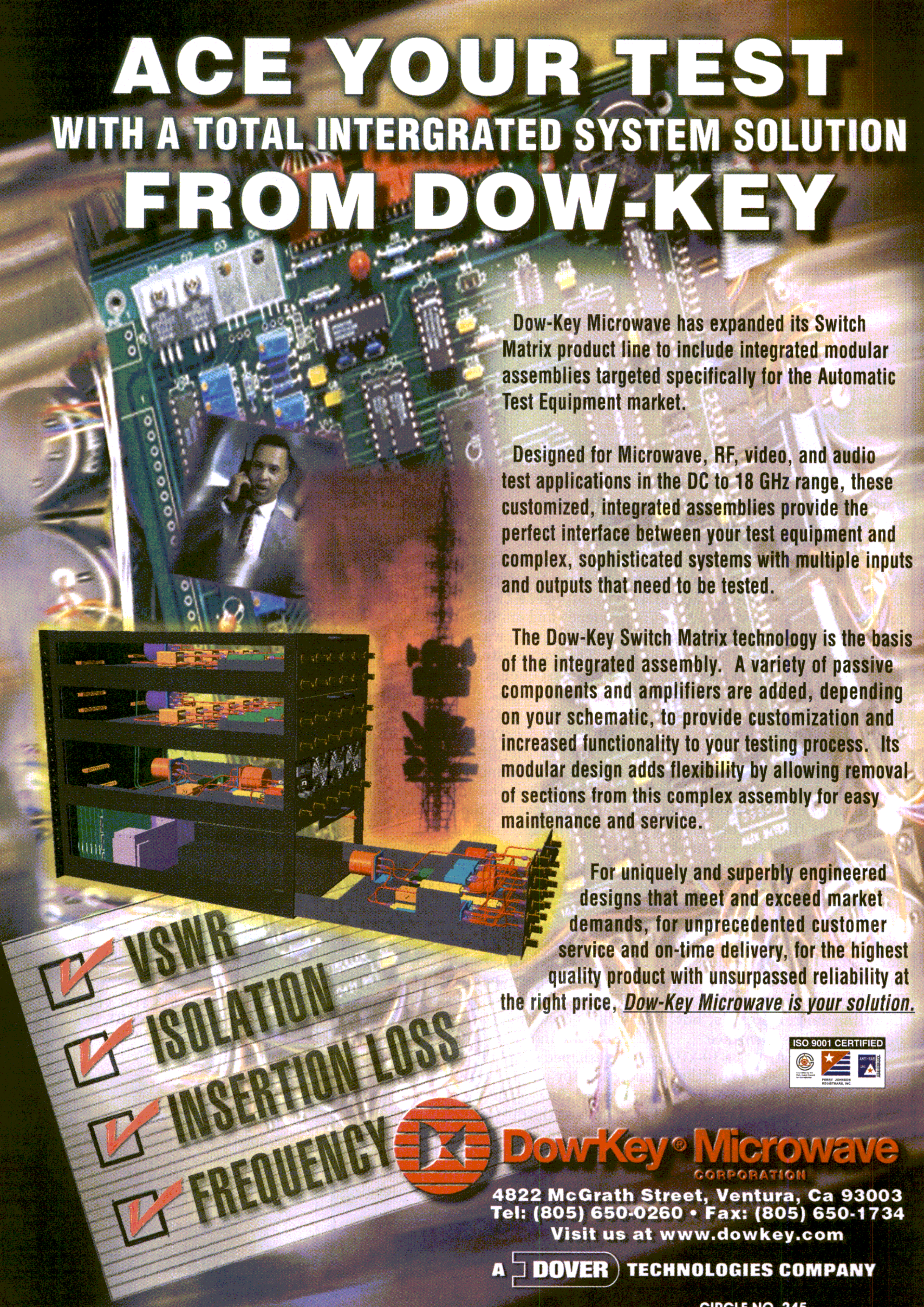
Due to the nature of the small-signal transistors used in the design and layout parasitics, it was decided that the filter would be evaluated under small-signal conditions (less than 25 mV peak-to-peak of input signal). The filter was shown to be capable of handling larger signals, but with degraded AC performance.

Figure 13 illustrates a 10-kHz narrowband frequency plot of the filter. The output shows a 410-MHz alias or image (110-MHz fundamental). The plot reveals a spurious-free dynamic range (SFDR) of approximately 60 dB at the filter output. The SFDR was found to be limited by small-signal input levels, spurious components falling within the passband of the buffered bandpass filters, and high-

The comparator can be used as a driver/frequency generator for the tunable bandpass filter. The differential outputs of the DAC are fed to a differential lowpass filter then presented to the differential inputs of the AD9852 comparator. The differential filter is the same filter shown in Fig. 4 (part 1). Since the DAC outputs are differential currents, voltages across the filter are also differential. These differential voltages cause the comparator to switch at the same point around the midscale value of the DAC output range, thus creating a square-wave output.

If a pure, differential sine source is input, the comparator output spectrum follows a sin (x)/x shaping, with nulls occurring at the even harmonic intervals. This spectrum can be reshaped simply by causing the comparator output to resemble a low or high duty-cycle pulse, moving the nulls as the duty cycle changes. Thus, if the duty cycle is adjusted properly, certain harmonics in the comparator output spectrum can be reduced or cancelled, aiding in the filtering effort.

Since the comparator circuit creates a square waveform, the even



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harmonics are suppressed. This is an advantage in that it eases filter-design requirements, since the frequencies that are to be suppressed are now two times the fundamental frequency away from the harmonic that is to be amplified. In order to use the composite filter of Fig. 12, the output of the comparator must be attenuated to DAC output levels. This improves the noise floor of the comparator output, and prevents the filter amplifiers and tuned circuits from being overdriven. The attenuated comparator signal is fed to the input of the composite filter. The only modification that must be made to the filter is the software that drives the control DAC of the AD9852. The tuning word input to the DAC must not only be converted to a logarithmic function, but must also be multiplied by a factor of the harmonic that is to be isolated and amplified. In this case, if the composite filter of Fig. 12 were used, the multiplication factor would be 5.

With the comparator driving the filter, a 10-kHz SFDR was found to be similar to that in Fig. 13, nearly 60 dB.

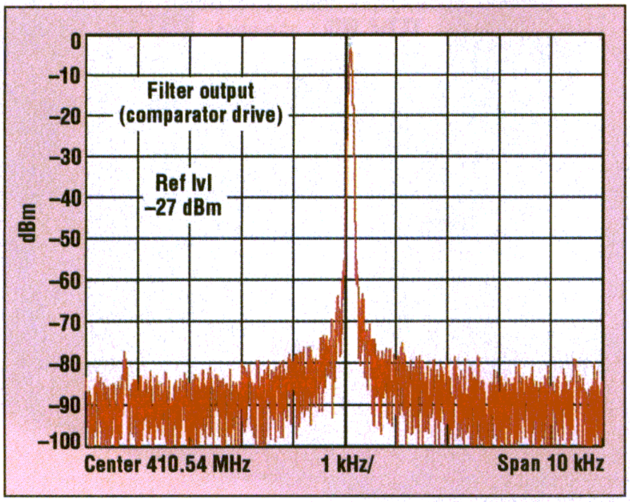
Several methods have been presented that would serve to isolate and amplify frequency components present in C-DDS device output spectrums above the Nyquist limit.

LC filters serve to isolate these frequencies. But they do not provide gain and are hard to construct.

SAW filters provide passband gain, provide excellent isolation from DAC nonidealities, but are extremely passband limited, in that their passbands cannot be adjusted. This non-adjustabiltity of the SAW filter approach is a disadvantage if one wishes to use the DDS frequency-hopping capability.

The tunable tracking filter that is shown preserves the benefits of using C-DDS systems. Discounting the non-ideal performance encountered in the experimental prototype of this circuit (PCB parasitics and the use of small signal devices as buffers in the buffered bandpass filters), this approach holds the most promise in preserving DDS advantages at greater-than-Nyquist frequencies. Better PCB layout and/or integration can reduce the circuit-board parasitic problems encountered. Also, with integration, better transconductors can be used to replace the small-signal buffers in the buffered bandpass filters—to enable larger output signal swings.

Using the AD9852 comparator in a "clock generator" configuration to drive the tunable tracking filter yields similar experimental results as the DAC drive configuration with only slightly reduced SFDR. This solution holds promise in that the comparator output spectrum does not exhibit the spectral-inversion phenomenon seen with the DAC output spectrum, and that the comparator is richer in higher-amplitude frequency components that could be isolated as well as amplified. Again, improved layout of the PCB and/or integration would tend to increase signal amplitudes, drop the noise floor, and increase circuit performance.●●



13. A narrowband frequency plot (10 kHz) of the filter output shows an SFDR of close to 60 dB. SFDR tends to be limited by factors such as non-ideal filtering and board parasitics.

Acknowledgements

Special thanks to Rick Cushing, Applications Engineer at Analog Devices, Inc. for the SAW filter design and data presented in this article.

Special thanks to Ken Gentile, Systems Engineer at Analog Devices, Inc. for the fixed LC filter designs used and presented in this article.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DSP Applications Add Power To Website

This website contains a variety of components and articles developed specifically for digital-signal-processing (DSP) applications.

ALAN ("PETE") CONRAD
Special Projects Editor

Digital-signal-processing (DSP) and digital receiver functions are the basis for many of the high-speed complementary-metal-oxide-semiconductor (CMOS) products developed by Graychip (Palo Alto, CA). Information about these products, and many techniques for applying them in DSP circuits, can be found in the company's website, at http://www.graychip.com. The firm's product lines include all-digital downconverters, upconverters, digital filters, and high-speed quadrature-amplitude-modulation (QAM) modem chip sets.

Included in their family of chip products are a 64-MSamples/s quad digital receiver, a 70-MSamples/s quad digital transmitter, a 70-MSamples/s narrowband receiver, an 80-MSamples/s wideband receiver, a 106-MSamples/s digital filter and 80-MSamples/s digital resampler, and an 80-MSamples/s mixer/carrier remover. Additional products include fast Fourier transform (FFT) receiver chips, pulse-code-modulation (PCM) demultiplexer chips, as well as a 500-MHz multiplexer and differential decoder chip.

The firm's website contains an abundance of application notes in Adobe format *(see figure)*. One of the notes explains how signals sampled at a rate of f and centered at 0 to f/2 can be upconverted to signals centered from f/4 to 3f/4 and sent to a digital-to-analog converter (DAC) at a sample rate of 2f.

This simplifies the analog upconversion of a signal to a higher intermediate frequency (IF), using digital filter chips to split a wide-bandwidth, high-sample-rate signal into two signals with overlapping bandwidths at one-half the input sample rate. The lower band covers from 0 to 0.2 times the input sample rate while the upper band runs from 0.15 to 0.35 times the input sample rate. A third chip can be used to cover the range from 0.3 to 0.5 times the input sample rate.

Another topic includes how to use a single DSP chip to perform the real-to-complex downconversion operation on four signals simultaneously. The input data are split into even and odd samples and then multiplexed into a time-division-multiplexed (TDM) format. The even-sample TDM stream is sent to the A-input of the chip while the odd-sample TDM stream is sent to the chip's B-input. The in-phase (I) portion of the resulting complex output signal will appear at the chip's A output port while the quadrature (Q) portion of the complex output signal will appear at the B output port in the same TDM format as the input signal.

An additional application note shows how to simultaneously filter four or eight ☞

The website at http://www.graychip.com offers extensive information on signal-processing components for digital communications applications.

sampled signals in a single GC2011 digital filter chip. The input data are multiplexed in a TDM format. In the dual path mode, the chip will process two TDM data streams, each containing four signals (for a total of eight signals). In the cascaded mode, the chip will process a single TDM stream containing four signals.

Other notes include an architecture for creating a 5-MB 64QAM modulator for a digital television transmitter, a design for creating a QAM demodulator that supports symbol rates up to 35 MB, and instructions for building high-performance digital modulators for most QAM, quadrature-phase-shift-keying (QPSK), and binary-phase-shift-keying (BPSK) signals.

The site also contains a 42-page application note describing the use of the company's GC2011A general-purpose +3.3-VDC digital filter chip with 32 multiply-add filter cells. The chip operates at rates to 106 MHz. The input data size is 12 b and the coefficient data size is 14 b. The output data size is 8, 10, 12, 14, 16, 20, or 24 b. The 32 multiply-add cells can be arranged as a 32-tap arbitrary phase filter or a 64-tap linear phase filter with even or odd symmetry.

Decimation and interpolation modes double or quadruple the number of taps in the filter. Two input ports allow the 32 filter cells to be shared between two data paths in order to process two signals or to process complex data. Each path becomes a 16-tap arbitrary phase filter, a 32-tap symmetric filter, a 64-tap decimate-by-2 filter or a 128-tap decimate-by-4 filter. Coefficient double buffering and clock synchronization logic permits the user to switch between coefficient sets without causing any undesirable transients in the filter's operation.

Complex coefficients can be handled using an add/subtract cell, which combines the two data paths. A complex coefficient filter requires two chips, one for the I output and one for the Q output. The number of complex taps varies from 16 to 128 depending on the symmetry and desired input/output (I/O) rate. The input data rate can be equal to the clock rate, one-half the clock rate, or one-quarter of the clock rate. The effective number of taps doubles for half-rate data and quadruples for quarter-rate data. The input data rate can be extended to 212 MHz if two chips are used. With two chips the filter size is 32 taps arbitrary phase or 64 taps linear phase. If decimation by two is desired, then only one chip is required and the filter size is 64 taps.

A single chip can be used to convert data between real and complex formats. When converting from real to complex values, the chip mixes the signal down by one-quarter the sampling rate and lowpass filters the results. When making the conversion from complex to real formats, the chip interpolates the signal by two, upconverts it by one-quarter the sampling rate, and sends the real part of the results to the output port.

The two 12-b data paths can be used to process 24-b input data by filtering the upper 12 b in one path and the lower 12 b in the other path. A 12-b shift-and-add circuit merges the results into a 24-b output signal. The chip includes a snapshot memory, which can capture blocks of I or O data. The size of the snapshot can be programmed to be two 128-sample-by-16-b snapshots, two 256-sample-by-8-b snapshots, one 256-sample-by-16-b snapshot, or one 512-sample-by-8-b snapshot. These samples can be read by an external processor and used for adaptive updates of the filter coefficients.

The internal data precision is 32 b, sufficient to preserve the full multiplier products and to prevent overflow in the filter's adder tree. The 32-b results are passed through a gain circuit before they are rounded to 8, 10, 12, 14, or 16 b. The gain circuit can adjust the signal's amplitude over a 96-dB range in 0.5-dB steps.

On-chip diagnostic circuits are provided to simplify system debug and maintenance. The chip receives configuration and control information over a microprocessor compatible bus consisting of a 16-b data I/O port, a 9-b address port, a read/write bit, and a control select strobe. The control registers, coefficient registers, and snapshot memory are memory mapped into the 512-word address space of the control port.

At the time of this review, the site's home page featured preliminary information on the GC4016/GC4017, a multiple-standard quad receiver chip that is suitable for Global System for Mo-

bile Communications (GSM) systems, code-division-multiple-access (CDMA) systems, and mixed-standard base stations. By clicking on the bottom of the advance information section, visitors can receive a full copy of the preliminary data sheet on the four-channel receiver integrated circuits (ICs).

This 75-page document provides a functional description of the quad receiver, including information on the control interface, the input format, synchronization, power-down modes, initialization, clocking, and diagnostic routines. The seven-chapter data sheet provides sections on packaging [for example, the GC4016 is supplied in a thin-quad-flat-pack (TQFP) housing while the GC4017 is supplied in a plastic-ball-grid-array (PBGA) housing] as well as control registers.

The GC4016 provides four independent digital downconversion channels, operating at input rates of 80 MSamples/s. The IC provides enhanced dynamic range, with spurious levels approaching −115 dBc. The out-of-band signal rejection is better than 100 dB while the spurious-free dynamic range (SFDR) is rated better than 115 dB. Output samples can be provided in several word lengths, including 12, 16, 20, and 24 b. Output signals can be in the form of bit serial signals, linking compatible, memory-mapped registers, or, in the case of the GC4017, parallel-port compatible. Both ICs achieve impressive 0.02-Hz tuning resolution.

Each of the four channels can be set for different bandwidths and/or bit rates. The GC4016/GC4017 features a 5-MHz bandwidth mode with 4-times oversampling for Universal Mobile Telecommunications Services (UMTS) applications. The low-power chips are designed for standard industrial operating temperatures of −40 to +85°C. In addition to information about the receiver ICs, the data sheets also provide application notes on using the chips in GSM applications. ●

For more information on these and other DSP-based products, visit the website, at: Graychip, 2185 Park Blvd., Palo Alto, CA 94306; (650) 323-2955, FAX: (650) 323-0206, e-mail: sales@graychip.com.

Internet:
http://www.graychip.com.

Site Offers Powerful Spreadsheet Analysis Tools

This full-featured website contains a host of products for military and wireless applications as well as a series of useful technical notes and design spreadsheets.

ALAN ("PETE") CONRAD
Special Projects Editor

Crystal-oscillator specifiers have long recognized the name Wenzel Associates (Austin, TX). The firm manufactures a wide range of precision, low-phase-noise sources for the communications, navigation, military, and test-equipment markets, in addition to supplying a variety of high-performance instruments and subsystems. Those who are familiar with the quality of the firm's products will not be surprised to find that this quality is reflected in the Wenzel website.

It is a site with generous listings of products, technical articles, application notes, and design tools. In the product area, for example, the company offers its "Blue Tops" RF modules, including amplifiers, multipliers, and phase-locked loops (PLLs). These modules enable RF system designers to configure custom frequency-generation systems to 3 GHz. The modules include integral attenuators, filters, detectors, and other related circuitry to minimize system complexity and reduce cost. Each module consists of circuitry packaged in a small, plated aluminum (Al) case complete with female SMA RF connectors for input/output (I/O), a feedthrough capacitor pin for power-supply connections, and a ground terminal.

The site features a range of application notes and tutorial articles. The notes cover frequency- and time-domain circuits, crystal oscillators, frequency synthesis, and frequency conversion, as well as "how-to" articles on a two-diode odd-order multiplier, phase locking and tuning, and PLLs. Other notes include switching-diode frequency doublers, frequency dividers, frequency multipliers, frequency tripler design, phase noise, amplitude-modulation (AM) noise, and phase-noise measurements. Tutorial articles focus on non-compensated crystal oscillators (XOs), temperature-compensated crystal oscillators (TCXOs), oven-controlled crystal oscillators (OCXOs), and voltage-controlled crystal oscillators (VCXOs). Additional tutorial articles highlight oscillator power-supply circuits, RF amplifier circuits, the effects of shock and vibration on oscillator performance, oscillator types, oscillator terminology, vibration-induced phase noise, as well as vibration isolation systems and techniques for reducing vibration-induced phase noise.

The site also contains a number of magazine-reprint articles covering such topics as an odd-harmonic generator, a low-frequency circulator/isolator constructed without magnetic or ferrite materials, a line of low-noise oscillators, a low-power RF-identification (RFID) transponder, and a constant-reactance VCO.

What really makes this site worth a visit, however, is its unique collection of downloadable Microsoft Excel design spreadsheets. These can be used to calculate the responses of PLLs, calculate oscillator phase noise under different levels of vibration, and to predict Allan Variance. There is a spreadsheet program for calculating the effects of random vibration on an oscillator's performance specification. The first section of the spreadsheet calculates the effect of a single degree-of-freedom vibration isolation system with the natural frequency and the damping factor provided.

An additional spreadsheet program can be used to calculate the bandwidth and damping factor for a type-two PLL, based on entered data for VCO tuning sensitivity, the loop components, and the phase-detector sensitivity. Another spreadsheet calculates the phase noise of a PLL based on the noise of the reference, the VCO, and the loop characteristics. Additional spreadsheets calculate the total root-mean-square (RMS) jitter over a user-specified bandwidth, and the phase noise caused by noise voltage on the electrical tuning line. ●

Wenzel Associates, Inc., 1005 La Posada Dr., Austin, TX 78752; (512) 450-1400, (512) 450-1490, e-mail: wenzel@wenzel.com.

Internet: http://www.wenzel.com.

Website Links FPGA Sources

This website contains links to field-programmable-gate-array (FPGA), complex-programmable-logic-device (CPLD), and simple-programmable-logic-device (SPLD) sources and design tools.

ALAN ("PETE") CONRAD

Special Projects Editor

Logic devices and field-programmable gate arrays (FPGAs) commonly used in communications and wireless systems can be found quickly at a website simply known by its URL: http://www.optimagic.com. The site offers links to companies, product applications, software-design tools, and free software downloads. A simple point-and-click mouse operation provides access to top component suppliers and high-quality tutorial articles from domestic (US) and international websites.

Links are available to leading programmable-logic and FPGA suppliers. This long list of manufacturers includes Altera Corp., Xilinx, Inc. (and its newly acquired portion of Philips Semiconductors), Lattice Semiconductor, Vantis (now part of Lattice Semiconductor), Actel Corp., Lucent Technologies, Cypress Semiconductor, Atmel, and QuickLogic Corp., making the OptiMagic site an effective single source for programmable-logic and FPGA specifiers.

Tutorials include a very-high-speed-integrated-circuit (VHSIC) hardware-description-language (VHDL) article download from Aldec, an FPGA VHDL Synthesis Lab Book by Associated Professional Systems, a general introduction to VHDL, a VHDL methodology paper for FPGAs, a VHDL synthesis tutorial, a coding style guide (from Actel), a Master-Class VHDL Tutorial (from Esperan), a Hardware Engineer's Guide to VHDL, an introductory VHDL tutorial (by Green Mountain Computing Systems), and VHDL help files for Windows.

For example, the OptiMagic site *(see figure)* provides a link to the Computer Science Department of Bucknell University (Lewisburg, PA). There, visitors will find a 30-page handbook on Verilog design by the school's Dr. Daniel C. Hyde. This handbook introduces readers to Verilog VHDL code and its applications.

The OptiMagic site also provides access to the site maintained by Aldec (Henderson, NV) where visitors will find a downloadable VHDL tutorial handbook. The course book guides users through the basic fundamentals for designing architectures with VHDL. The tutorial simplifies the learning curve and includes a series of questions and answers to test a reader's knowledge of the language at the end of each chapter.

Another tutorial, authored by Bob Reese of the Electrical Engineering Department of Mississippi State University, reviews combinational circuit synthesis, sequential circuit synthesis, and system synthesis.

Finally, the OptiMagic site provides access to a synthesis design guide and tutorial developed by Boulder Creek Engineering (Abingdon, MD). The books of this design guide take readers from concepts through VHDL coding, synthesis, routing, and hardware design. The remaining books cover hardware test points, programmable pseudorandom-noise (PN) generators, a matched-filter correlator, and a bit-error tester (BERT). ●

The OptiMagic website at http://www.optimagic.com provides a wide array of articles and design tools for users of FPGAs and programmable-logic devices (PLDs).

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Site Provides Forum For RF Circuit-Board Issues

Visitors to this website can exchange views, news, and comments about all aspects of printed-circuit-board (PCB) advances and applications in the RF and microwave fields.

ALAN ("PETE") CONRAD
Special Projects Editor

Circuit-board design and materials are the key issues at http://www.labtechnical.com. The site offers a variety of links covering over 15 topics, such as material selection, microstrip radiation, gap tolerances, machining, and plating finishes. Upon registering at the site, visitors are provided with a password to limited-access areas that allows them to become involved in technical discussions on the key topics. The technical team at Labtech monitors each forum and contribute useful insights as applicable.

The site includes links to major materials suppliers, including Arlon (http://www.arlonmed.com), Rogers Corp. (http://www.rogers-corp.com), and Taconic Advanced Dielectrics Division (http://www.Taconic-add.com). Free software and application articles are available at these sites for downloading. Links are also included to industry trade publications, such as *Microwaves & RF* (http://www.mwrf.com) and *Wireless Systems Design* (http://www.wsdmag.com), as well as *Microwave Journal* (http://www.mwjournal.com) and *RF Design* (http://www.rfdesign@intertec.com).

The site contains a list of recommended technical books on a wide range of subjects, including advances in microstrip and printed antennas, atmospheric remote sensing, microwave radiometry, broadband patch antennas, computer-aided design (CAD) of microstrip antennas for wireless applications, coherent optical communications systems, computer-aided analysis and modeling, and the design of microwave networks. Other book title include the design of nonplanar microstrip antennas and transmission lines, design of field-effect-transistor (FET) frequency multipliers, as well as the design of RF and microwave amplifiers and oscillators.

The text *Advances in Microstrip and Printed Antennas* reviews the advantages of low-profile printed antennas, their compatibility with integrated-circuit (IC) technology, and their conformability to shaped surfaces. This book considers recent advances in conventional areas of the subject, for example, probe-fed microstrip antennas, aperture-coupled microstrip antennas, and covers new areas of research, such as developments of CAD formulas for the rectangular-patch microstrip antennas made of high-temperature-superconducting (HTS) materials.

The text *Atmospheric Remote Sensing by Microwave Radiometry* presents the state of the art in passive microwave remote sensing. The focus of the book is on the application of various types of microwave observational data for the purpose of determining atmospheric properties. The data enables testing of existing models for the atmosphere's energy balance, depletion of the ozone layer, and climate trends.

The text *Computer-Aided Analysis, Modeling, and Design of Microwave Networks* introduces an alternative to standard microwave CAD—the wave-variable approach to computer-aided analysis, modeling, and design of linear and nonlinear microwave networks. The book includes previously unpublished data on noise analysis, power-sensitivity analysis, and wave-measurement problems. A disk is included, bound in at the back of the book, which performs conventional frequency-domain analysis, noise analysis, and optimization.

The book *The Design of RF and Microwave Amplifiers and Oscillators* provides a practical approach to designing RF and microwave amplifiers and oscillators using the iterative synthesis techniques provided. The book introduces approaches to help readers estimate the 1-dB compression point of Class A and Class B linear circuits, initialize the fundamental component voltages and currents in a harmonic balance simulator, and more easily generate load-pull contours for Class A and Class B transistors. The text helps readers build multistage power amplifiers (PAs), amplifiers with an optimum noise match, and a low-input VSWR and wideband impedance-matching networks. These are only a few of the samples of the recommending reading found at http://www.labtechnical.com.

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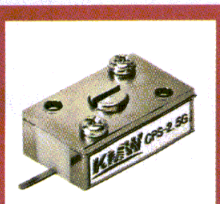
Insertion Loss (Max.)	0.15dB	0.25dB	0.35dB
VSWR (Max.)	1.25:1	1.25:1	1.25:1
Incremental Phase Shift	90 degree min. @ 2GHz		
Electrical Delay	125 psec min.		
Nominal Impedance	50 ohm		
I/O Port Connector	SMA(F) / SMA(F)		
Average Power Handling	20W @ 2GHz		
Temperature Range	-30°C ~ +60°C		
Dimension (inch)	A type : 1.496*1.102*0.457 B type : 1.225*1.102*0.457		



KPH30OSCL000

KPH35OSCL000

Miniature CPS

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

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CIRCLE NO. 374

Specifying coax delay lines for feedforward amplifiers

Coaxial delay lines are used in the correction loops of feedforward amplifiers to create out-of-phase versions of noise signals for noise cancellation. The delay lines must provide precise performance over time and temperature, and over the desired bandwidth of the amplifier. A two-page application note from Micro-Coax, "Coaxial delay lines for feedforward amplifiers, design considerations," details the various requirements needed when specifying coaxial delay lines for feedforward amplifiers.

The application note reports that circular coils provide the most economical and consistent coaxial delay-line configuration. Other configurations, such as oval or square shapes, are expensive to form and require an invasive service loop to compensate for cable physical-length variations.

The note highlights the importance of cable-length tolerances, with electrical-length tolerance recommendations of ±15 to ±20 ps for frequencies between 800 and 2000 MHz. The literature, which includes electrical data for several of the company's coaxial delay lines measured at 820 MHz, covers options for cable types used in the manufacture of coaxial delay lines, packaging considerations, and system-design considerations. Copies of the two-page application note are free, from: **Micro-Coax, 206 Jones Blvd., Pottstown, PA 19464-3465; (800) 223-2629, (610) 495-0110, FAX: (610) 495-6656, Internet: http://www.micro-coax.com.**

CIRCLE NO. 194 or visit www.mwrf.com

Explore multimode and single-mode optical fibers

Local-area networks (LANs) were once served quite adequately with basic copper (Cu) cables. But as data rates increase, other transmission media, such as optical fibers, are needed to provide enough bandwidth without suffering excessive signal losses. A technical article called simply "The Glass Story" from Optical Fiber Corp. (Roanoke, VA) explains how single-mode and multimode optical fibers can effectively improve the performance of LAN systems.

The technical article breaks down the differences between multimode and single-mode fiber-optic cables. The cables share a basic structure that consists of a glass core (which carries the light signals), surrounded by a substrate layer of glass, which does not carry light but adds to the diameter and strength of the glass fiber.

Multimode fibers feature a core diameter that is relatively large compared to the wavelength of light that it carries. Core diameters range from 50 to 1000 μm compared to light-signal wavelengths in the area of 1.0 to 1.6 μm. Older types of multimode fiber are usually step-index fibers, where the index of refraction is the same all across the core of the fiber. Newer multimode fibers are generally graded-index multimode fibers where the index of refraction across the core is gradually changed from a maximum at the center to a minimum near the edges.

Single-mode fibers have a core diameter of approximately 9 μm, which is closer to the wavelength of the light being propagated—about 1.3 μm. This limits the light transmission to a single mode of light, and eliminates multimode effects. For short-to-moderate distance communications, single-mode fibers operating at a wavelength of 1300 nm offer the best combination of low cost and high performance in on-campus and in-building applications. For longer-distance systems, single-mode fibers can also be used at 1550 nm, although the laser sources and detectors are more expensive at this wavelength.

Multimode fibers can be readily differentiated from single-mode fibers by the size of the core and overall glass or cladding. The most popular size of multimode fiber for LAN systems is 62.5/125 fiber, which indicates a core diameter of 62.5 μm and a total glass diameter of 125 μm. Another common multimode fiber is 50/125 fiber for high-bandwidth applications. An older fiber size is 100/140 fiber, which has been used in industrial applications because of its large core size and durability.

The article points out further differences between multimode and single-mode fiber optic cables and how they are used, particularly in Gigabit Ethernet systems. The technical article is contained in the company's latest fiber-optic cable product catalog (Vol. 1). Copies are free, from: **Optical Cable Corp., 5290 Concourse Dr., Roanoke, VA 24019; (540) 265-0690, FAX: (540) 265-0724, Internet: http://www.occfiber.com.**

CIRCLE NO. 195 or visit www.mwrf.com

COVER FEATURE

VNA-Based System Tests Differential Components

By using mixed-mode S-parameters, this test system can evaluate the linear performance of the balanced (differential) devices that are common to high-speed communications.

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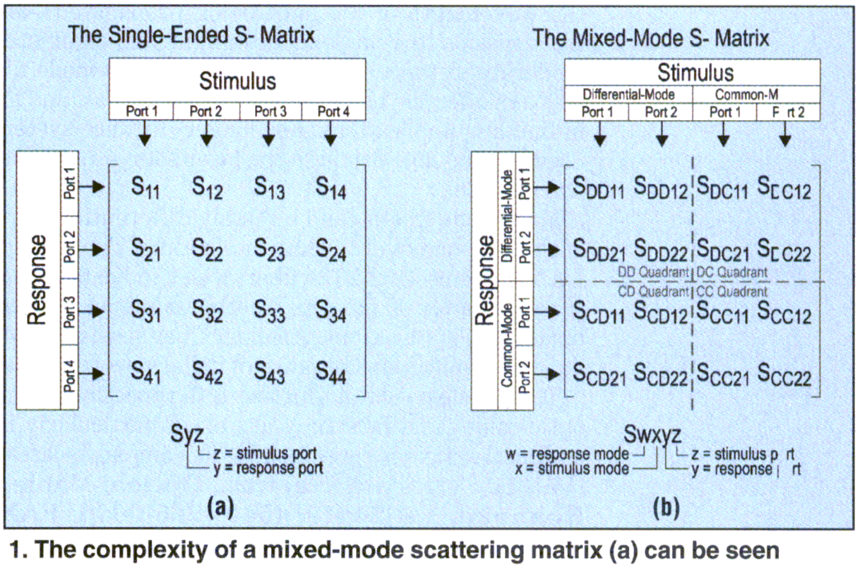
SIGNAL integrity is an issue of growing importance as digital networks increase in speed and bandwidth. Traditionally, RF and digital engineers have had little in common. But in recent years, digital signals are passing that barrier more and more (approximately 1 Gb/s) between being conventional digital signals and having more of an RF/microwave nature. The harmonic content of these signals is many times the frequency of the fundamental tone, calling for greater precision in interconnections and in testing these high-speed digital systems. Fortunately, a system developed by ATN Microwave, Inc. (North Billerica, MA) provides the capability to measure mixed-mode scattering (S)-parameters on the balanced or differential devices common to high-speed digital circuitry. The system essentially brings the precision and accuracy of the vector network analyzer (VNA) to the digital world, for circuits operating to 20 GHz.

Due to the harmonic frequencies found in fast-rise-time digital signals, it is common to think of high-speed interconnections as transmission lines, and to consider the effects of reflected signals. But frequency-domain instruments such as the VNA are not commonly used to address time-domain issues. The differential time-domain reflectometer (TDR) is still the analysis tool of choice for high-speed digital circuits and components.

There are several reasons for this. For one, time-domain analysis of digital



1. The complexity of a mixed-mode scattering matrix (a) can be seen compared to a single-ended scattering matrix (b).

signals seems to make sense—state-to-state transitions need to be preserved and their signal characteristics as a function of time need to be understood. For another, most systems transmitting high-speed data use differential signals, while most network analyzers are designed for single-ended, two-port devices.

TDR is not ideal, however. With increasing data rates, the dynamic range of a very high-speed TDR system is often inadequate for analyzing low-level signals such as crosstalk, or the signal components responsible for generating electromagnetic interference (EMI). Parasitic inductances and capacitances in signal lines and interconnections are generally ignored at lower data rates, but these become significant at higher data rates (frequencies). TDR systems do not correct the systematic sources of error in the measurement equipment, nor do they support de-embedding fixtures or interconnects used with a device under test (DUT).

Frequency-domain measurements with a VNA can address the shortcomings of TDR systems, and provide some additional benefits. VNA systems do not require the application of a large voltage to the DUT, do not require the DUT to have a DC return path, and can provide forward and reverse transmission and reflection data without changing the measurement setup.

Unfortunately, traditional two-port VNA systems are designed for measuring unbalanced devices. However, this system from ATN addresses that issue and provides the ability to measure balanced devices. In doing so, they still do not provide the test data in a format that is meaningful to engineers concerned with digital signal integrity.

Traditional VNAs employ S-parameters. A complete S-parameter matrix accounts for all possible signal paths between any two ports within a device (forward and reverse transmission and reflection). Recently, this concept has been extended from one that describes the linear performance of single-ended devices, to one that can describe the linear performance of balanced (differential) devices. This extension is known as mixed-mode S-parameters.[1]

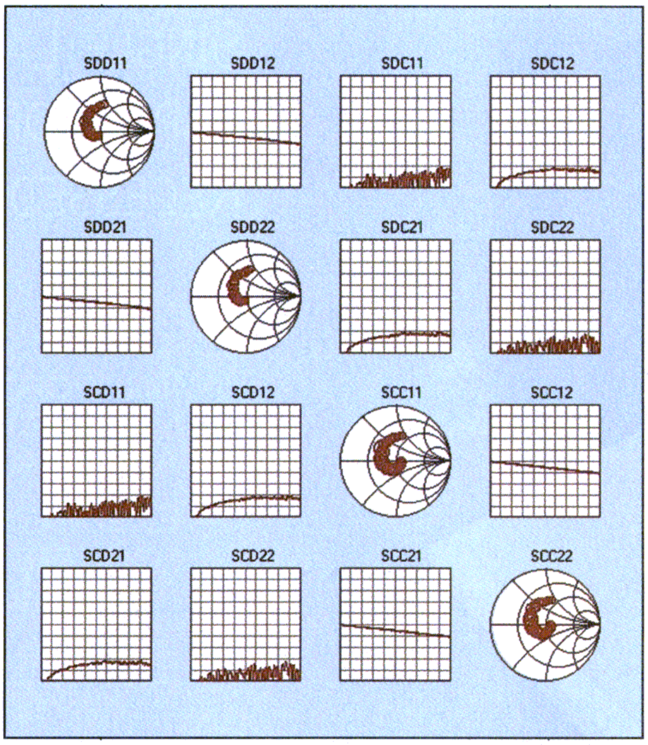
Mixed-mode S-parameters are similar to conventional single-ended S-parameters in that both approaches examine the stimuli and responses between any two ports of a DUT. A single-ended (unbalanced) device has a single terminal and a single mode of operation. A balanced device has two terminals, however, and can support two modes of operation. As a result, mixed-mode S-parameters must consider the stimulus and response mode, in addition to the stimulus and response port.

Single-ended S-parameters are defined in the format S_{yz}, with y and z representing the response and stimulus ports, respectively (Fig. 1). S-parameters for balanced devices

are defined as S_{wxyz}, with w and x representing the additional response and stimulus modes.

A mixed-mode S-matrix can comprehensively characterizes the linear performance of a balanced two-port device. As with single-ended S-parameters, each column represents a different stimulus condition, and each row represents a different response. If the matrix is divided into four quadrants, the device can be considered a two-port device under four specific modes of operation.

To understand the information contained within the mixed-mode S-parameter matrix, it is necessary to examine each quadrant. The upper left-hand quadrant describes the behavior of the DUT when it is stimulated with a differential-mode signal and the differential-mode response is observed. It consists of four parameters—reflection parameters on both balanced ports, and transmission parameters in the forward and reverse directions, all in differential mode. This quadrant describes the fundamental performance characteristics of a differential device.

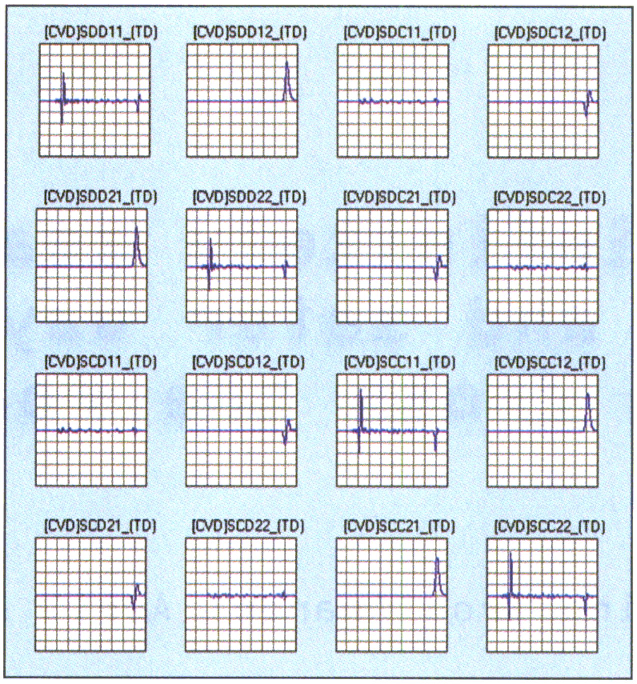


2. These mixed-mode S-parameters were measured for high-frequency transmission lines.

The lower right-hand quadrant describes the behavior of the DUT when it is stimulated with a common-mode signal and the common-mode response is observed. It, too, consists of four parameters—reflection parameters on both ports, and transmission parameters in the forward and reverse directions, all in common mode. If one of the system-level objectives is to

reduce the level of common-mode signals, these parameters can be used to analyze that behavior. Common-mode rejection is often of interest, and it can be calculated by taking the ratio of the differential-mode gain, S_{dd21}, to the common-mode gain, S_{cc21}.

The lower left-hand quadrant describes the behavior of the DUT when it is stimulated with a differential-mode signal and the common-mode response is observed. Therefore, these results show how the DUT converts a differential-mode signal to a common-mode signal. For this mode of operation, it is possible to examine reflections on each port and transmission in the forward and reverse directions. In a perfectly symmetrical balanced device, there is no conversion between modes, and all of these terms will be zero.



3. These time-domain impulse responses were measured for all modes.

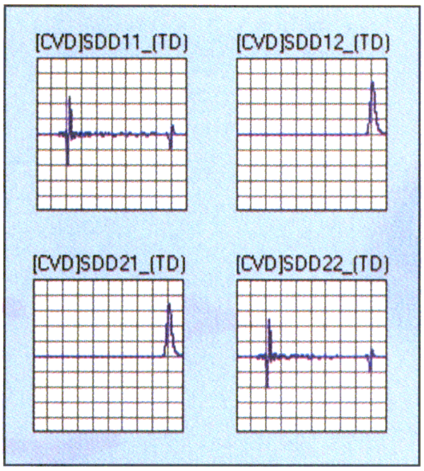
Asymmetry can be seen in a DUT as an imbalance in the amplitude of the signals on each side of the balanced pair, or as a phase skew. The resulting conversion of the signal from a differential-mode to a common-mode will cause signals to appear on ground returns in the system, and therefore, can result in generation of EMI. These grounds return paths could be intentional ground paths, or could be unintentional paths. In devices that handle high-speed digital signals, imbalance will also result in degradation of bit-error rates (BERs).

Finally, the upper right-hand quadrant describes the behavior of the DUT when it is stimulated with a common-mode signal and the differential-mode response is observed. These results show how the DUT converts a common-mode signal to a differential-mode signal. Information in the upper right-hand quadrant shows reflections on each port and transmission in the forward and reverse directions. If a balanced DUT is perfectly symmetrical, there will be no conversion between modes, and all of these terms will be zero.

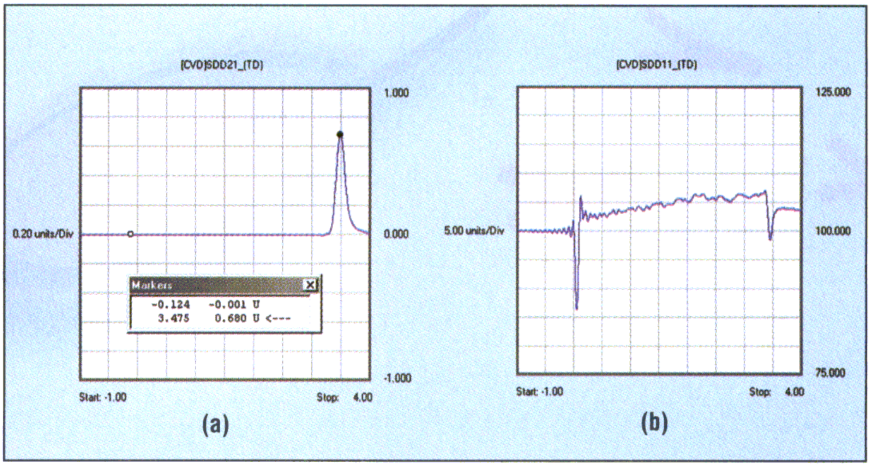
The resulting conversion of the signal from a common-mode to a differential-mode will cause the system to be susceptible to EMI. For example, noise that couples into the system from power supplies, ground connections, or radiated sources is generally introduced as a common-mode signal. If this common-mode noise is converted to the differential-mode, it will superimpose on the intended signal and degrade the signal-to-noise ratio (SNR). Therefore, the mode-conversion behavior has made the system susceptible to these sources of noise and interference.

Clearly, mode-conversion is a phenomenon that must be considered when examining the signal integrity of balanced devices. But since these signals are ideally zero, the dynamic range provided by TDR systems is not sufficient. VNA systems designed to characterize balanced devices excel at this task due to the better than 110-dB dynamic range possible in the time domain.

A VNA is capable of measuring



4. By separating the impulse responses for the differential mode, these are the plots that remain.



5. The plots offer a comparison of a differential TDT impulse response (a) and a differential TDR step response (b) for high-frequency transmission lines.

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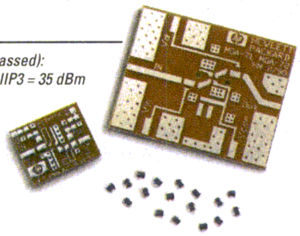
ATF-34143 (output)	4V, 60 mA	0.5	17.5	31.5
ATF-35143 (output)	2V, 15 mA	0.4	18.0	21.0
ATF-38143 (output) *coming soon*	2V, 10 mA	0.5	16.0	22.0

** as a switch (amp bypassed): insertion loss = 2.5 dB, IIP3 = 35 dBm*



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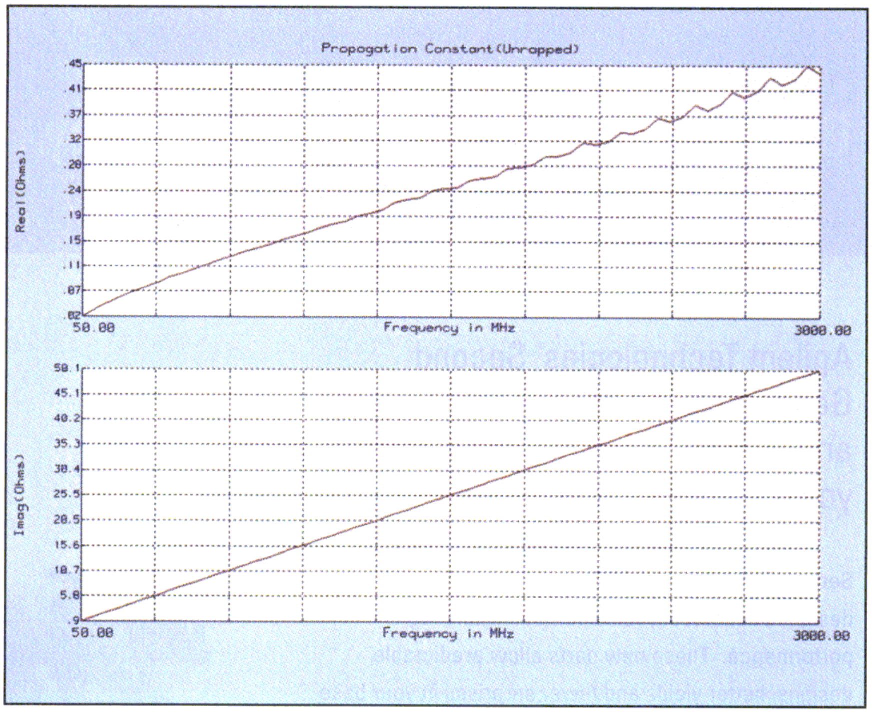
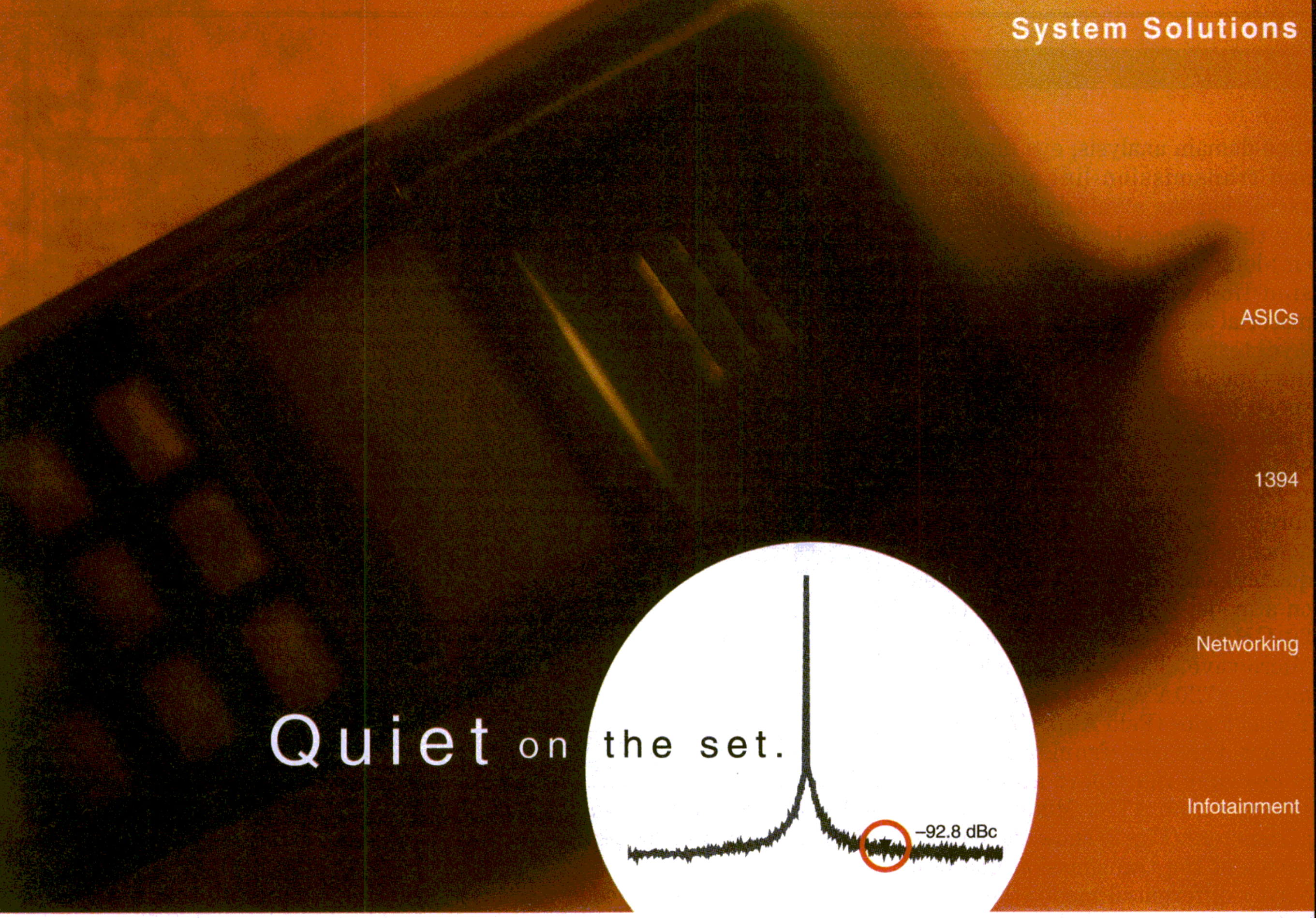
data, characterizing balanced devices, and providing features never before available to engineers concerned with signal integrity in high-speed data systems. However, while it is true that the mixed-mode S-parameter representation of device performance will comprehensively characterize the performance of a balanced device, it is also true that this representation does not directly tell a signal-integrity engineer how his device is performing. This is simply because S-parameters provide a frequency-domain representation, while data signals need to be analyzed in the time domain, and because the S-parameter data cannot be readily used in time-domain circuit simulation tools such as SPICE models. If the S-parameter data provides a comprehensive characterization, though, there is no reason that the data cannot be put into a more-useful format.

The information that is useful for signal-integrity applications includes



6. These measurements show the complex propagation coefficient as a function of frequency for the real and imaginary parts of the transmission line's characteristic impedance.



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Dual PLLs

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MB15F03SL	1.75 GHz 0.6 GHz	2.3 1.2	2.7 2.7
MB15F07SL	1.1 GHz 1.1 GHz	2.5 2.5	2.7 2.7
MB15F08SL	2.5 GHz 1.1 GHz	4.4 2.6	2.7 2.7

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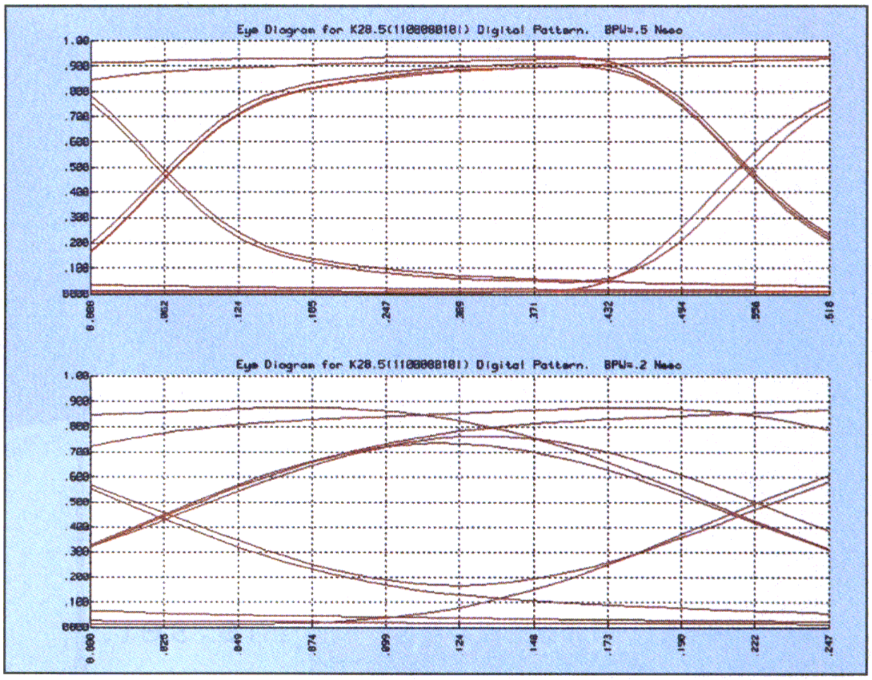
time-domain analysis, eye diagrams, and transmission-line parameter extraction. As a vehicle for examining these, consider a 21-in. (53.34-cm)-long, 100-Ω balanced transmission line printed on Nelco 6000 material as a test vehicle. This multilayer test vehicle is representative of the type of structures used for high-speed backplanes. The lines are fed from an SMA connector and a viahole on the circuit board.

To examine the test circuit, measurements were made for frequency-domain and time-domain data. Frequency-domain data were measured on a model ATN-4002A 50-MHz-to-20-GHz test system from ATN Microwave. The system includes a model 8720A VNA from Agilent Technologies (Palo Alto, CA), a multiport test set from ATN, and microwave multiport application software from ATN. Mixed-mode S-parameters (Fig. 2) show the performance of the transmission line in a purely differential mode (the upper left-hand quadrant of the matrix). Input and output return losses range from 5 to 30 dB over the band of interest, while the insertion loss reaches 12 dB at 6 GHz. Data are referenced to a differential-mode impedance of 100 Ω.

The common-mode performance in the lower-right quadrant (referenced to a common-mode impedance of 25 Ω) shows similar performance, indicating light coupling between the two traces that form the balanced line. The terms in the lower-left and upper-right quadrants show fairly low mode-conversion of only approximately 30 dB in the worst case (which will be examined in greater detail later in this article).



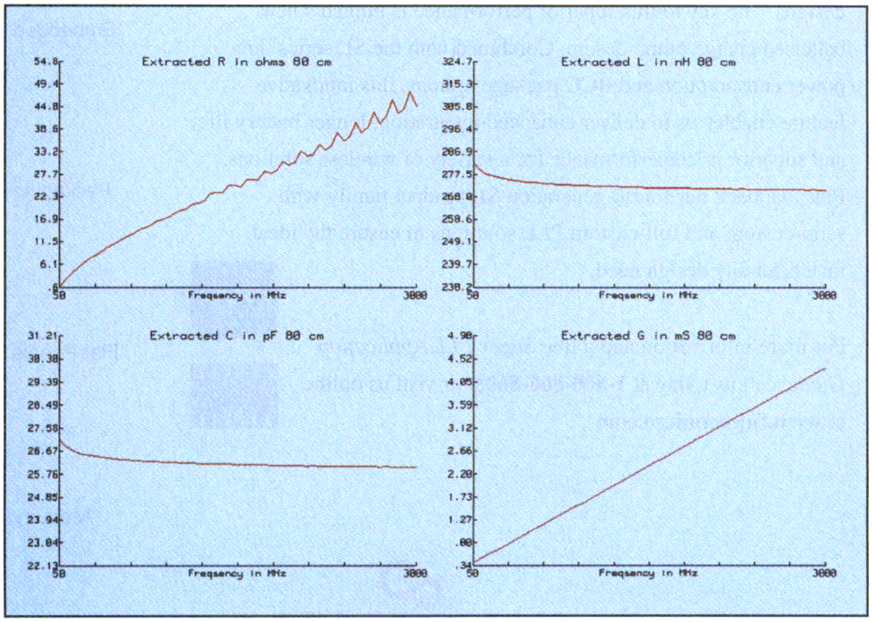
8. The new test system was used to create these eye diagrams for bit widths of 500 ps (top) and 200 ps (bottom).

By performing an inverse Fast Fourier transform (FFT), any of these parameters can also be viewed in the time domain. The time-domain impulse response data (Fig. 3) are organized in the same manner it was for the frequency-domain data. Figure 4 shows only the upper-left quadrant to isolate the differential-mode performance. The parameters on the diagonal show the TDR responses on each port in differential mode, and the off-diagonal parameters show the forward and reverse TDT parameters. The other three quadrants of the full mixed-mode s-matrix show the same type of information for the other modes of operation.

The transmission line has a group delay of 347.5 ps as shown in the differential-mode TDT impulse response (Fig. 5a). The differential-mode TDR step response of the same transmission line (Fig. 5b) clearly reveals the discontinuities of the connectors. The TDR step response can be used to estimate the characteristic impedance of the line, which is slightly higher than 100 Ω. The upward slop of the step response results from the finite loss of the line.

In addition to estimating the characteristic impedance from the step

(continued on p. 162)



7. These extracted telegrapher's parameters were measured for an 80-cm length of transmission line.

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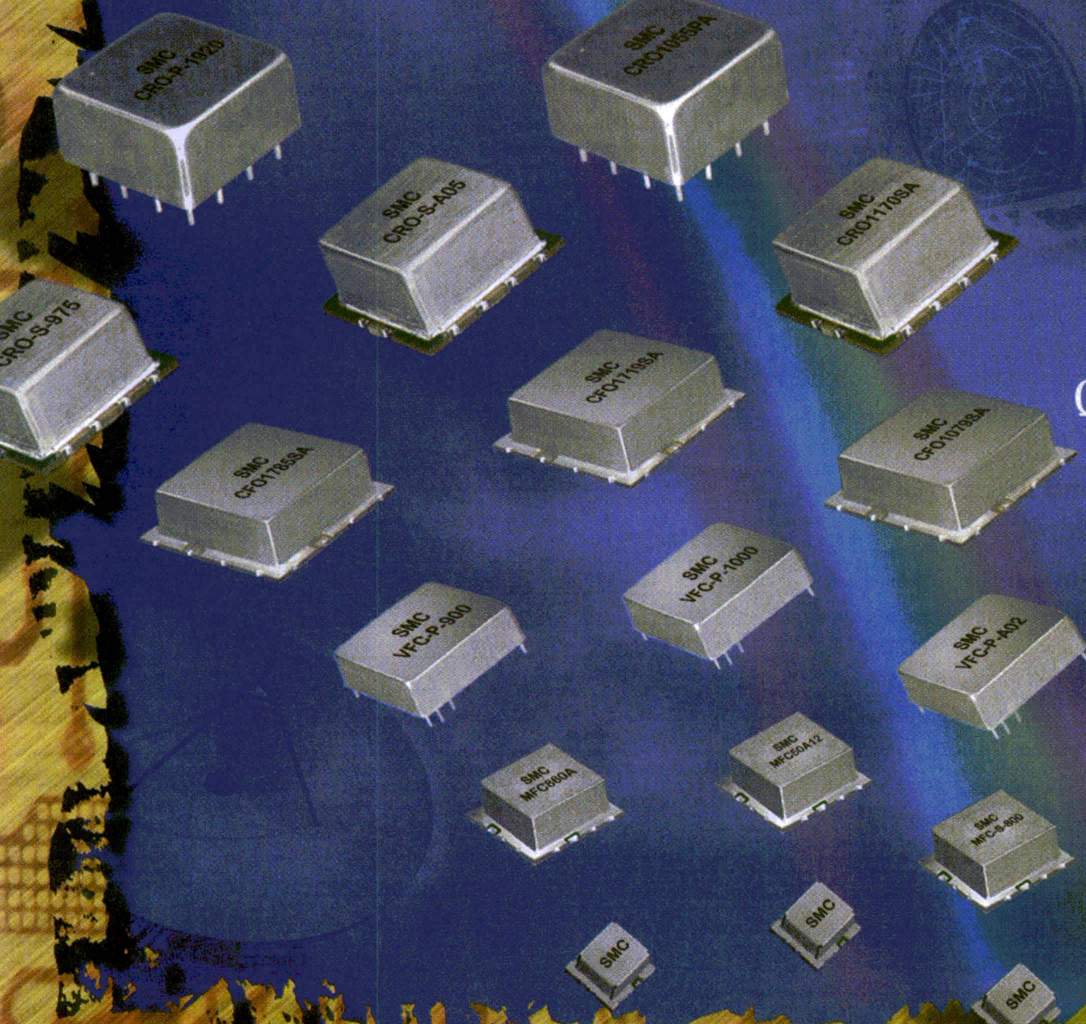

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CIRCLE NO. 264

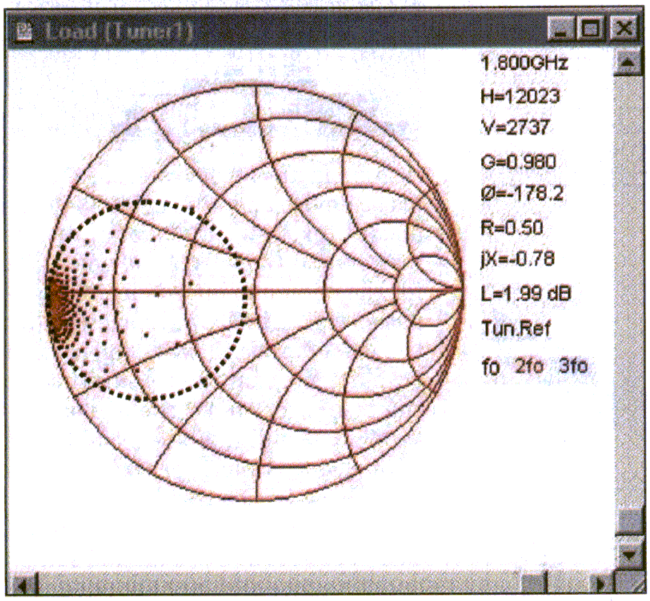
Prematching Tuners Advance Load-Pull Testing

The next generation of load-pull test equipment has arrived in the form of prematching microwave tuners with calibrated SWRs of 150:1 and higher.

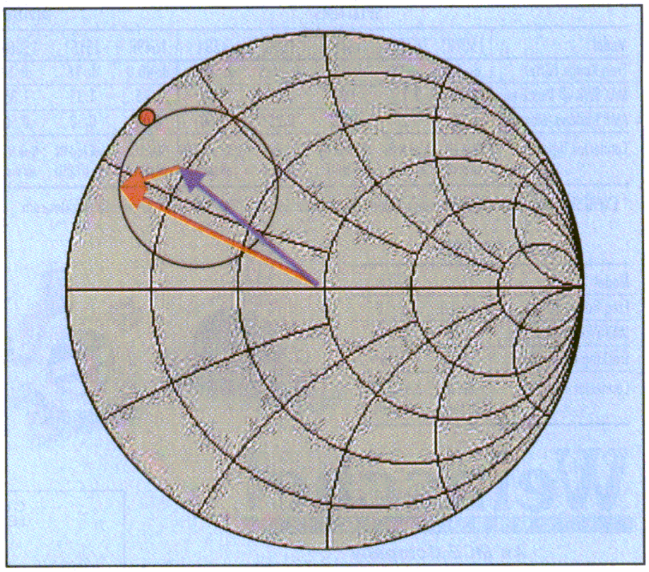
Christos Tsironis
President
Focus Microwaves, Inc., 970 Montee de Lisse, St. Laurent, Quebec, Canada PQ H4T 1W7; (514) 335-6227, FAX: (514) 335-6287, e-mail: christos@focus-microwaves.com, Internet: http://www.focus-microwaves.com.

NONLINEAR characterization of power transistors is critical for progress in commercial markets such as cellular communications. In addition to forcing designers away from pure computer modeling and back to the measurement benches, sophisticated "device engineering" in numerous laboratories employs load-pull data to modify and optimize doping profiles and transistor layout designs for higher efficiency, linearity, and output power. To aid this revolution, a family of programmable impedance tuners based on prematching concepts has been developed by Focus Microwaves (St. Laurent, Quebec, Canada) to improve the measurement accuracy of even the lowest-impedance power transistors, including laterally-diffused-metal-oxide-semiconductor (LDMOS) high-power transistors for cellular base stations.

Power transistors used in cellular and personal-communications-services (PCS) base stations are inherently low-impedance devices that require the synthesis of the appropriate low-impedance matching networks. Ten years ago, transistors with internal impedances of 5 Ω were considered state of the art. Now, high-power transistors exhibit internal impedances below 1 Ω. These transistors are characterized with the aid of adjustable, calibrated load-pull impedance tuners that must provide accurate low-impedance synthesis at high power levels. Automatic testing is also important for timely data acquisition (DAQ), where computer-controlled electromechanical



1. These are the tuner calibration points as seen by a DUT through a quarter-wave transformer.



2. The programmable prematching tuners make it a simple matter to tune around any point on the edge of the Smith Chart.

slide-screw tuners have been used for some time due to their superior power-handling capability, high resolution, and dynamic tuning range, and large bandwidth.

Commercial automatic slide-screw tuners use one or two microwave probes (slugs) to generate high reflection states. The best electromechanical coaxial tuners may accurately generate maximum SWRs of approximately 15.0:1 (corresponding to a real part of the internal resistance of the transistors of approximately 3.3 Ω or a reflection coefficient, Γ, of 0.875). Using two probes rather than one increases the instantaneous operating bandwidth. The probes are independent in vertical direction but move together horizontally. A single probe can provide an instantaneous frequency range to three octaves ($f_{max}/f_{min} = 8$), whereas the combination of two probes may cover more than a decade in bandwidth. Examples of these tuners are available from Focus Microwaves and span the frequency range of 0.2 to 6.0 GHz, 0.8 to 18 GHz, or 2 to 40 GHz.[1-3]

However, using two parallel probes does not raise the reflection factor itself dramatically, even if both probes are mounted close together to increase the capacitive effect. Beyond a reflection level of Γ = 0.875, network-analyzer calibration and impedance-tuner repeatability may cause accuracy and measurement repeatability problems. Many commonly used lossy test fixtures will further reduce the available reflection factor at the reference plane of the device under test (DUT) to unacceptably low values of approximately 10.0:1 for SWR and about 5 Ω for the minimum impedance. The situation can be improved somewhat by the use of quarter-wave microstrip transformers on the test fixture at the test frequency[4], the use of "active" load-pull modules[5] combined with passive tuners, or the use of an entirely active load-pull systems.[6]

Unfortunately, active systems, either complete active load-pull setups or active modules, are not adequate solutions for characterizing packaged high-power transistors due to

MANY COMMONLY USED LOSSY TEST FIXTURES WILL FURTHER REDUCE THE AVAILABLE REFLECTION FACTOR AT THE REFERENCE PLANE OF THE DEVICE UNDER TEST TO UNACCEPTABLY LOW VALUES OF APPROXIMATELY 10.0:1 FOR SWR AND 5 Ω FOR THE MINIMUM IMPEDANCE.

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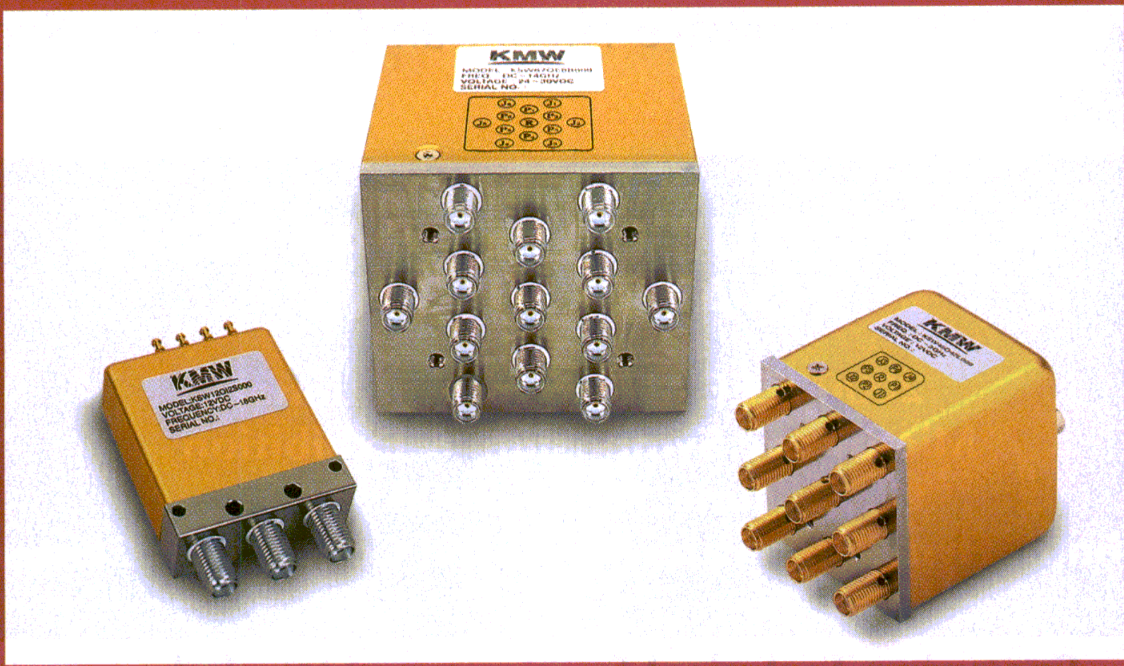

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Operating Mode	TTL Latching with IND.	Latching with IND.	Latching
Actuating Voltage /Current (Max.)	12Vdc ± 10% /240mA (@12Vdc, 25°C)	20 ~ 30Vdc /95mA (@24Vdc, 25°C)	24 ~ 30Vdc /85mA (@26Vdc, 25°C)
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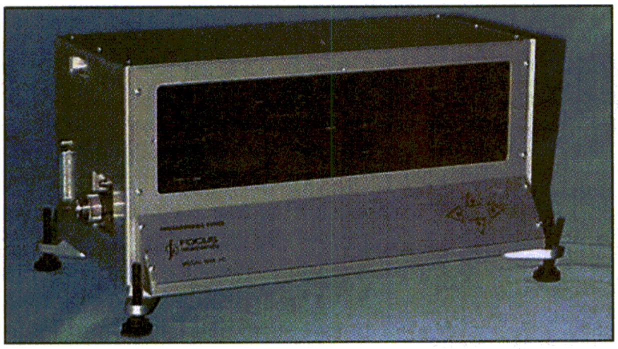


3. **Programmable prematched tuners in the PMT line are currently available for applications to 18 GHz, with models soon available for use to 50 GHz.**

probe raises the reflection factor to a considerably high value. Then a second probe tunes around this level and very close to $|\Gamma| \cong 1$. Since the phase of the first reflection factor vector can be adjusted arbitrarily, it is a simple matter to tune around any point at the edge of the Smith Chart (Fig. 2).

Prematching can also be viewed as a resonance phenomenon where the only limiting factor for obtaining a reflection factor of $|\Gamma| = 1$ is the transmission line losses between the input connector of the tuner and the first (prematching) probe. For probe posi-

high-power limitations and parasitic oscillation problems. The only remaining realistic solution is the quarter-wave transformer method, which is affordable, simple to design, and easy to realize, but also has the shortcomings of being cumbersome, frequency selective, and must be redesigned for each particular DUT. Quarter-wave transformers pretune by nature into a restricted specific area of the Smith Chart, normally (although it may vary from transistor to transistor) the area around a short circuit (where $\Phi = 180$ deg.). One problem is that once the quarter-wave transformers have been fabricated, tuning outside of this prematched area of the Smith Chart is impossible (Fig. 1).

NEW SOLUTION

The new tuner family from Focus Microwaves is based on prematching principles. This means that a first

FOR PROBE POSITIONS OF TUNERS USING GPC-7 CONNECTORS THAT ARE VERY CLOSE TO THE INPUT PORT (PREMATCHING SECTION INITIALIZED HORIZONTALLY), IT HAS BEEN POSSIBLE TO OBTAIN REFLECTION COEFFICIENTS OF BETTER THAN 0.998.

tions of tuners using GPC-7 connectors that are very close to the input port (prematching section initialized horizontally), it has been possible to obtain |Γ| > 0.998. Since this position is not necessarily the one required to tune around Φ = 180 deg. and the first probe must be moved away from the connector, practical maximum values of |Γ| are approximately 0.99 to 0.995, values which are equivalent to SWRs of 200:1 to 400:1 (Figs. 3 and 4).

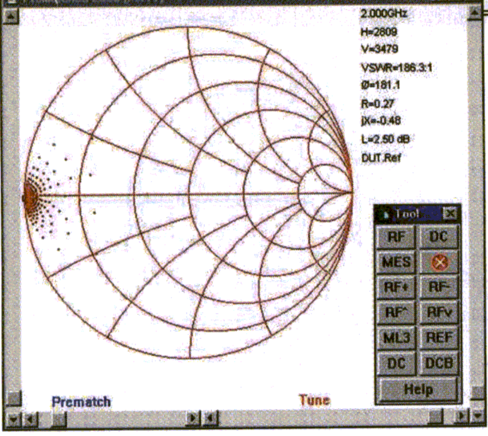
Measurements beyond these levels are limited in measurement accuracy by the choice of vector network analyzer (VNA). Calibration of these high reflections is, obviously, a difficult operation. Focus Microwaves provides a specially designed algorithm for calibrating prematching tuners in a relatively short time (approximately 20 min. per frequency point) enabling the measurement software to synthesize any impedance on the Smith Chart with Γ ≤ 0.99 (or SWR ≤ 200:1) combining the reflections of both microwave probes.

The tuning accuracy of these tuners has been verified at very low impedances. Measurements have been made on deviations between synthesized and measured reflection coefficients of −40 to −55 dB around 0.4 Ω (which is equivalent to Γ ≅ 0.985).[7] The table shows the impedance and reflection factors of a prematched microwave tuner as tuned by the associated software and measured with a calibrated VNA. The difference (ΔΓ) between the measured ($S_{11\,meas}$) and synthesized (S_{11} calc) reflection factors is calculated using the simple formula: $\Delta\Gamma = 20\log|S_{11\,meas} - S_{11\,calc}|$

The first set of data in the table (area 1, tuning around 1 Ω) means that the tuning section of

Tuning accuracy of PMT at very low impedances

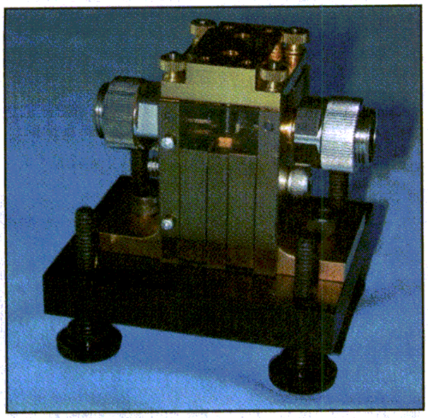
R(Ω)	jX	|Γ|	Φ	Δ(Γ)[dB]
Area 1: tuning around 1.0 Ω				
0.986	0.0918	0.961	179.8	−47.26
0.996	0.2147	0.961	179.5	−49.35
0.949	−0.0245	0.963	180.1	−53.10
0.977	0.3455	0.962	179.2	−45.05
0.845	0.2406	0.967	179.5	−50.32
0.798	−0.0107	0.969	180.0	−47.06
1.093	−0.1666	0.957	180.4	−47.32
1.359	0.4402	0.947	179.0	−55.03
Area 2: tuning around 0.4 Ω				
0.414	−0.1211	0.984	180.3	−47.43
0.409	−0.0106	0.984	180.0	−54.03
0.408	0.0954	0.984	179.8	−51.13
0.406	0.2044	0.984	179.5	−55.49
0.407	0.3217	0.984	179.3	−45.48
0.418	0.4574	0.983	179.0	−51.18
0.428	0.6284	0.983	178.6	−45.02
0.453	0.8637	0.982	178.0	−51.74



4. The maximum tuning range of the PMT tuners approaches an equivalent SWR of 200:1 at 2 GHz.

the prematched microwave tuner (probe 2) moves around vertical zero (initialized), while the prematching section (probe 1) remains set to a high SWR. The second set of data in the table (area 2, tuning around 0.4 Ω) is the final tuning area, when probe 2 (the tuning probe) is close to the central conductor. This is the real operation area of the PMT. The tuning accuracy exhibited by the PMT varies between –40 and –55 dB, which is excellent for this type of operation.

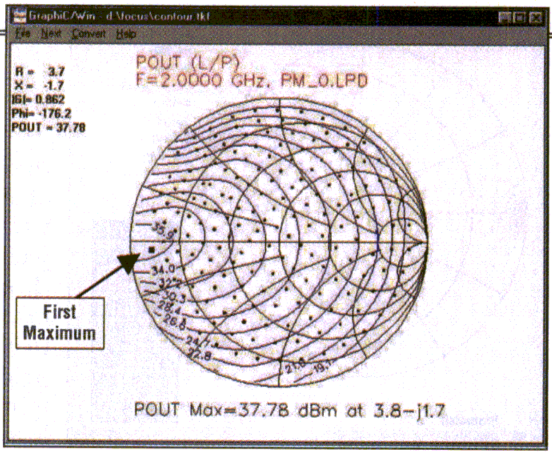
Of course, the advantage of generating high-SWR tuning would be compromised by the use of lossy test fixtures, especially those with low-quality microstrip-transmission materials. To avoid this and take full advantage of the PMT's capabilities, Focus has developed a minimum loss test fixture (MLTF) with an extremely low insertion loss of less than 0.02 dB at 2 GHz (Fig. 5).[8] It is the only commercial test fixture of this kind that is compatible with most power transistor packages. It is available with through-reflect-line (TRL) calibration standards and is simple to use. The MLTF employs proprietary connector clamps to minimize RF and DC contact losses and can be used to maximize the SWR at the device-under-test (DUT) reference plane to values of approximately 100:1 or minimum tunable resistances of 0.4 Ω, without using any transformers, active loads with amplifiers, closed gain loops, or other additional networks.



5. A model MLTF minimum loss test fixture has been developed for use with the PMT tuners. The fixture exhibits less than 0.02-dB loss at 2 GHz.

The prematching tuners offer few shortcomings and many advantages. Using special proprietary techniques developed by Focus Microwaves, the total load-pull system can be calibrated easily, quickly, and very accurately. The tuners themselves are fully characterized at a combination of 400 × 400 impedance points in approximately 20 min. per frequency point around 1 GHz. At higher frequencies, the calibration time is even less. The calibration of prematching tuners is very accurate, despite the extremely high reflections involved. The main reason for this is that the total reflection of the tuners is generated by cascading two medium size reflections with SWR ≅ 12:1, resulting in a total SWR of more than 150:1 at the tuner reference plane. Interpolation routines available in the Focus measurement software support the accurate synthesis of millions of impedances at any point of the Smith Chart, corresponding to reflection factors up to $\Gamma \cong 0.995$. The tuners can handle significantly more power than simple tuners at the same level of SWR, because both probes in the prematching slide-screw tuners stay further away from the central conductor, since each probe needs to generate a lower individual SWR. The tuning area can be pointed at will to any angle of the Smith Chart (not only around 180 deg., as is the case with quarter-wave transformers). The transforming ratio and, consequently, the surface of the tuning area and Γ_{max} can be freely adjusted.

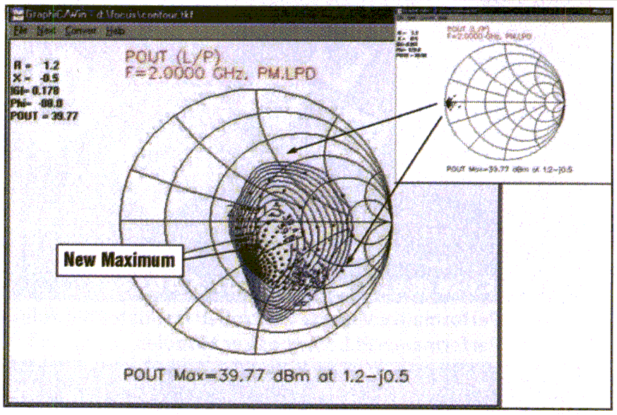
To evaluate the new prematching tuners, high-power transistors were measured with a traditional test setup, then compared to results taken with the new prematching tuners. The comparison is especially telling when the traditional setup cannot reach the low impedance required by



6. These load-pull measurements performed with a normal tuner and microstrip test fixture show that the DUT, a power transistor, has not been properly matched.

THE NEW AUTOMATIC PREMATCHING TUNERS CAN GENERATE HIGH SWRs OF 200:1 IN ANY AREA OF THE SMITH CHART AND AT ANY FREQUENCY FROM 0.4 TO 18 GHz.

the transistor (Fig. 6). In this case, the optimum reflection factor is at the edge of the calibration region of the tuners and test fixture (R_{min} = 3.8 Ω). As can be seen, the contours are not closing around the optimum point in the traditional setup and, by comparison with the contours of Fig. 6, the transistor has not been effectively power matched. Using a combination of the PMT and the MLTF, however, it is possible to envelope the optimum Γ with calibrated points. The contours are closed around the optimum point and the measurement accuracy is increased. In addition, a much higher value is achieved for maximum power (Fig. 7), close to 2 dB more.



7. In contrast to the results of Fig. 6, these normalized load-pull contours, measured with the PMT/MLTF combination, yield +39.8-dBm output power instead of +37.8 dBm with the contours of Fig. 6. The inset shows the non-normalized measurement points.

The new automatic prematching microwave tuners can generate high SWRs of 200:1 in any area of the Smith Chart and at any frequency from 0.4 to 18 GHz. In the near future, additional PMT models will be available with top frequencies that consist of 26.5, 40, and 50 GHz. When used with the firm's low-loss microwave test fixtures, load-pull testing of packaged power transistors is possible at impedances below 0.5 Ω at cellular and PCS frequencies without the use of transformers or active systems. **Focus Microwaves, Inc., 970 Montee de Lisse, St. Laurent, Quebec, Canada PQ H4T 1W7; (514) 335-6227, FAX: (514) 335-6287, e-mail: christos@focus-microwaves.com, Internet: http://www.focus-microwaves.com.**

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Switch Cards Fit Compact VXI Format

These compact switch matrices are ideal for high-frequency and IF front-end signal routing in advanced systems and automatic-test-equipment installations.

JACK BROWNE
Publisher/Editor

COMPACT instrument formats such as VXI make system reconfigurations painless. It is simply a matter of sliding out a VXI card to be replaced and sliding in a new card. This is the ease of installation for two VXI/VME switch-matrix cards from Signal Technology Corp.'s Systems Operation (Webster, MA). The two 4 × 8 switch-matrix modules, which operate from 1 to 32 MHz and 20 to 2000 MHz, offer impressive isolation with fast switching speed and wide dynamic range.

The high-frequency switch matrix, model 6301-004S-0085-7, achieves typical isolation of 50 dB between any two input ports. The isolation from any input port to any output port is typically 60 dB (see table). The signal gain from input to output ports is a nominal 1 dB with amplitude flatness of ± dB.

The model 6301-004S-0085-7 exhibits a noise figure of 10 dB and reaches 1-dB compression with an input signal level of +8 dBm. The second-order output intercept point (IP2) is a lofty +50 dBm while the third-order output intercept point (IP3) is +25 dBm. The 4 × 8 high-frequency switch-matrix card features switching speed of 100 ns and a VSWR of 1.50:1.

The higher-frequency model 6501-004S-008S-7 achieves matrix gain of 3 dB from 20 to 2000 MHz, with gain flatness of ±3 dB. The isolation between different input ports is typically 40 dB while the input-to-output isolation is typically 60 dB. The model 6501-004S-008S-7 offers a noise figure of 12 dB with IP2 of +45 dBm and IP3 of +20 dBm. The maximum VSWR is 2.0:1, with typical performance closer to 1.70:1. This higher-frequency model reaches 1-dB compression at +5 dBm, but shares the 100-ns high-speed switching capability of its HF counterpart.

Both of the switch matrices are rated for operating temperatures from 0 to +50°C. The HF unit consumes 30-W power while the higher-frequency matrix consumes 10-W power during normal operation. Input/output (I/O) RF connectors are SMAs, while the control interface is a standard VXI backplane. Each switch matrix occupies a double-slot C-size VXI rack space.

Both of the switch matrices conform to the VXI plug-and-play specification for ease of installation and use. In fact, switch drivers that are suitable with Windows 95 software are provided with each switch matrix. **Signal Technology Corp., Systems Operation, 37 Sutton Rd., Webster, MA 01570; (508) 943-7440, FAX: (508) 949-1804, Internet: http://www.sigtech.com.**

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The VXI/VME switches at a glance

Specificatons	6301-004S-0085-7	6501-004S-008S-7
Matrix configuration	4 × 8	4 × 8
Frequency range	1 to 32 MHz	20 to 2000 MHz
Matrix gain	1 dB ±1 dB	13 dB ±3 dB
Isolation		
Input/input	50 dB typ.	40 dB typ.
Input/output	60 dB typ.	60 dB typ.
Output/output (diff. input)	50 dB typ.	40 dB typ.
Output/output (same input)	25 dB typ.	20 dB typ.
Noise figure	10 dB	12 dB
Output intercepts		
Second order	+50 dBm	+45 dBm
Third order	+25 dBm	+20 dBm
VSWR	1.50:1	2.00:1 max.
Switching speed	100 ns	100 ns
Input compression (1 dB)	+8 dBm	+5 dBm

Dual Digital Receiver Commands 70-dB Range

A pair of independently controlled SHARC processors gives this receiver the teeth to capture digitally modulated IF signals to 25 MHz.

JACK BROWNE

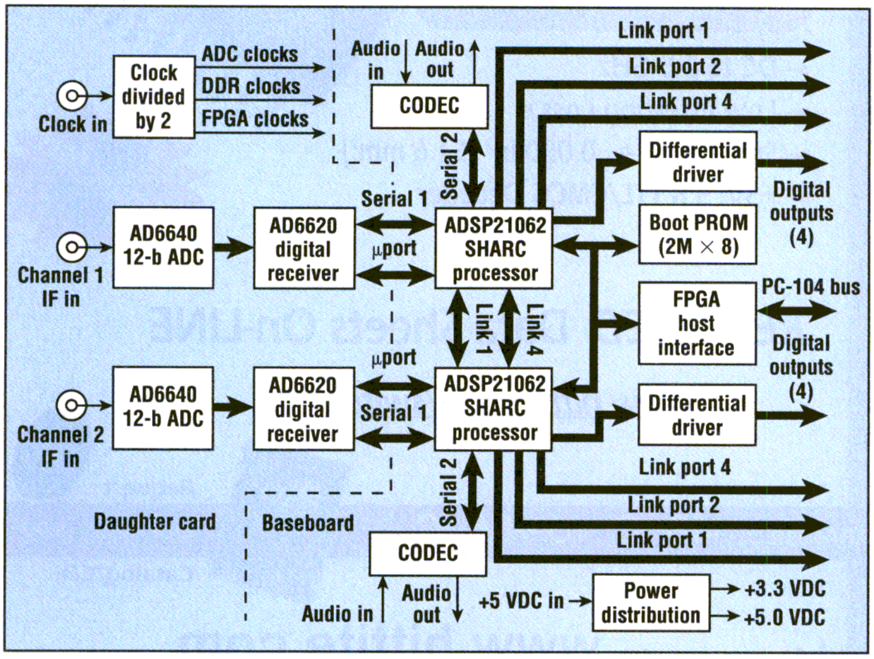
Publisher/Editor

DIGITAL technology has assumed many of the roles that were once played by analog circuitry. The STel-9966 dual digital receiver (DDR) from ITT Industries, Microwave Systems (Lowell, MA) is a good example of digital circuitry in a subsystem that was once perceived as a purely RF architecture. The high-speed receiver employs a pair of fast analog-to-digital converters (ADCs) to digitize input intermediate-frequency (IF) signals to 25 MHz, providing operators with 12-b digital resolution and a spurious-free dynamic range (SFDR) of better than 70 dB.

A key advantage of a digital receiver architecture is its programmability. Rather than relying on IF and baseband filters based on temperature-sensitive ceramic or surface-acoustic-wave (SAW) technologies, the STel-9966 dual digital receiver incorporates digital filters which are defined through software. Due to this, the filters can be readily reconfigured in software to accommodate changing system requirements, such as the addition of channels and/or capacity to a commercial cellular network. The STel-9966 is designed to work with IF signals at 21.4 MHz at input power levels of −65 to 0 dBm. Since any type of baseband filter can be configured under software control, the STel-9966 can effective demodulate almost any form of amplitude- or phase-based modulation scheme, including quadrature-amplitude-modulation (QAM) and quadrature-phase-shift-keying (QPSK) modulation. The STel-9966 is designed for use with external clock rates to 130 MHz at clock input power levels from −5 to +5 dBm.

The STel-9966 dual-channel digital receiver is constructed in a PC/104 format (see figure), with the digital-signal-processing (DSP) functions and digital interfaces on the main circuit board and the equivalent of the digital front end on a connected daughter board. The digital receiver design is based on several integrated circuits (ICs) from Analog Devices (Wilmington, MA), including a pair of AD6640 ADCs, a pair of AD6620 digital receivers, and a pair of ADSP-21062 SHARC DSP chips.

In a typical application, an input clock of 100 MHz is divided by two on the daughter board, resulting in 50-MHz clock signals that are distributed to the digital receiver ICs, the ADC ICs, and the field-programmable-gate-array (FPGA) host



The STel-9966 dual digital receiver is a two-channel receiver with digital filters that can be programmed in software. It can be used as part of a complete software radio solution.

interface. IF input signals at channel 1 or 2 are digitized by an AD6640 ADC and passed along in digital form to the AD6620 digital receiver. Additional signal processing for each channel is provided by a dedicated ADSP-21062 SHARC DSP.

The AD6640 ADC and AD6620 digital receiver chips are designed to work together in a digital receiver. The ADC features 12-b precision with a maximum sampling rate of 65 MSamples/s. The wideband ADC can digitize an entire 25-MHz spectrum at one time for instantaneous coverage of multichannel communications bands. The AD6640 is designed to maintain a SFDR of 80 dB and a signal-to-noise ratio (SNR) of 70 dB over a 25-MHz bandwidth.

The AD6620, which is in effect a set of digital filters, also operates at 65 MSamples/s. Because its filters are readily programmable, it can be quickly tuned to different channels and even different air interface standards. Its four cascaded signal-processing elements are a digital tuner, two fixed-coefficient filters, and a programmable coefficient decimating filter. The AD6620, which is capable of adding 25 to 30 dB of processing gain following the AD6640, maintains a SFDR of 100 dB for the 25-MHz bandwidth.

The ADSP-21062 SHARC DSPs implement a user's preprogrammed signal-processing algorithm for the received signals of interest. These SHARCs are 32-b microprocessors optimizing for the numerical demands of DSP. Fabricated with a high-speed, low-power complementary-metal-oxide-semiconductor (CMOS) process, each ADSP-21062 brings 25-ns instruction cycling time and processing power of 40 million instructions per second (MIPS). Each DSP also includes 2 Mb of on-board static random-access memory (SRAM) for improved operating speed with complex instruction sets.

The STel-9966 even includes an audio codec for each receiver channel, to enable users to demodulate input signals on each channel. The codec deliver analog audio output signals of +1 VDC peak to peak into a load of 600 Ω. The dual digital receiver incorporates four status light-emitting diodes (LEDs) per DSP, one FPGA load-status LED, and one PC/104-to-DSP interrupt per channel. The STel-9966 can be powered via the PC/104 bus in the host mode or through an on-board power connector in a stand-alone mode. It is truly a "next-generation" radio since it can be programmed to meet the needs of today's and tomorrow's air-interface standards: The capacity of the radio can be increased simply by reconfiguring digital filters in software. **ITT Industries, Microwave Systems, 59 Technology Dr., Lowell, MA 01801; (978) 441-0200, FAX: (978) 453-6299, Internet: http://www.stelmsd.com.**

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InGaP Amplifier Targets Wireless Applications

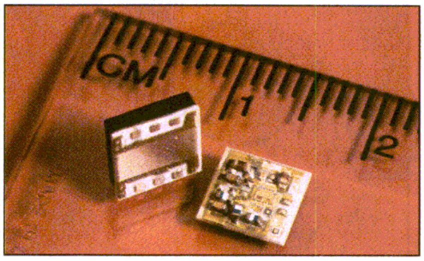
This InGaP HBT amplifier suits wireless communications and is less than half the size of comparable solutions.

PETER STAVENICK
Managing Editor

AMPLIFIERS designed for modern wireless systems must provide the linearity needed to transmit high-capacity voice and data. Small size is essential when reducing circuit-board space. In the case of the CHP1232-PM power module from Celeritek's (Santa Clara, CA) family of TrueTriangle™ +3-VDC power-amplifier (PA) modules, the amplifier offers small size and high linearity. It suits personal communications services (PCS), wireless local loop (WLL), and other high-bandwidth wireless data markets. It has been developed to meet the continuing requirements for small, low-cost, and high-capacity wireless voice and data applications across the 1.85-to-1.91-GHz frequency range.

The indium-gallium-phosphide (InGaP) heterojunction-bipolar-transistor (HBT) amplifier is a 50-Ω-matched linear three-stage PA that meets PCS-1900 or IS-98 requirements (see figure). Operating from a single positive +3.2-VDC supply, the amplifier offers 30-dB typical gain at rated output levels. It also provides 35-percent linear power-added efficiency (PAE) with +28-dBm output power [IS-98 code-division-multiple-access (CDMA) mode]. Although the package only measures 6.0 $mm^2 \times 1.7$ mm, the CHP1232-PM can eliminate up to 14 external elements and has a thermal rating of 20°C/W. With an operating temperature range of −40 to +90°C, the amplifier occupies half the board space of currently available solutions.

The typical input return loss is 10 dB and the noise figure is typically 3 dB. The noise power in the receive band over a 30-kHz bandwidth is −94 dBm. A single positive battery supply covering the range of +3.2 to +4.1 VDC is needed along with a +2.7-VDC fixed, regulated voltage for the bias-control circuitry (V_{ref}). With CDMA modulation at +28-dBm power output and 1.25-MHz offset, the adjacent-channel-power-ratio (ACPR) linearity is −45 dBc in a 30-kHz band. V_{cc} should be applied prior to the reference voltage and before the RF input power. The CHP1232-PM can be operated over a range of supply voltages and bias points through the adjustment of the reference voltage. It is important that the maximum power dissipation of the package be observed at all times and the maximum voltage that is across the device not be exceeded.



These InGaP HBT amplifiers are three-stage linear PAs that meet PCS-1900 or IS-98 systems.

The CHP1232-PM provides on-board bypass decoupling. Inadequate bypass capacitance and inductance around the DC supply lines can compromise the ACPR, reduce power gain, and create oscillations. When biased with the proper V_{cc} and V_{ref}, the CHP1232-PM will achieve the necessary adjacent-channel response for the digital system that is specified. The company tests each product under digital modulation to ensure the correlation to customer applications.

The positive (I_{cc} supply current of the amp is 520 mA at +28 dBm while the quiescent current (I_q) stands at 55 mA. Typical reference current is 1.1 mA with the maximum being 2.0 mA.

Along with the device's small size, the amplifier offers an absolute maximum 5-W power dissipation. Its maximum collector voltage is +5.5 VDC while its collector current is 1.2 A. The storage temperature spans the range of −65 to +150°C while the junction temperature is 150°C.

Specific applications for the CHP1232 include PCS handsets, PCS infrastructure, WLL subscriber units, cdmaOne handsets, and 1X WCDMA 2.5G handsets. Samples of CHP1232-PM are available in bulk and in tape and reel. Quantities are priced at $7.00 per unit. **Celeritek, 3236 Scott Blvd., Santa Clara, CA 95054; (408) 986-5060, FAX: (408) 986-5095, e-mail: truetriangle@celeritek.com, Internet: http://www.celeritek.com.**

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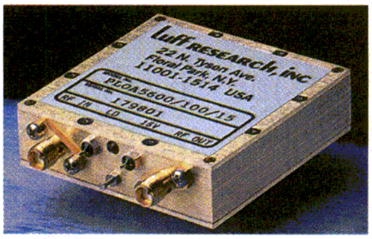

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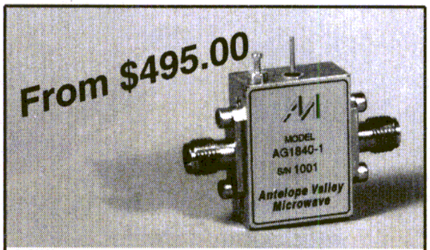

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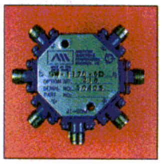

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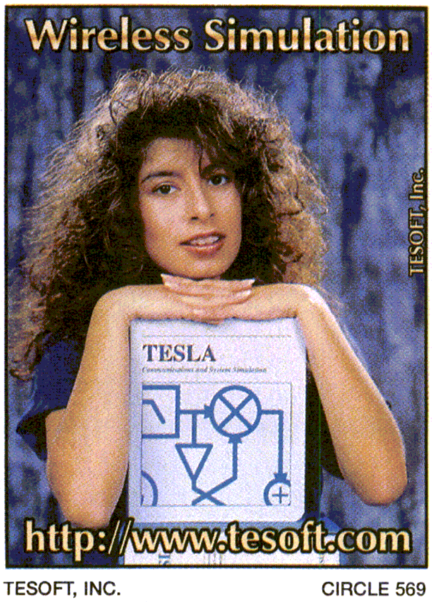

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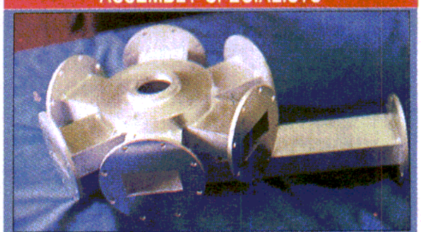

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(continued from p. 136)

response, it is also possible to extract a complete set of transmission line parameters directly from the frequency-domain data. This will give the complex characteristic impedance and propagation coefficient, or equivalently, the telegraphers' parameters (R, L, C, and conductance per unit length).

For the differential mode, measurements of the real part of the characteristic impedance versus frequency shows that the line is very close to 102 Ω from 50 MHz to 3 GHz. The real and imaginary parts of the propagation coefficient are shown in Fig. 6. These are commonly referred to as α (related to the loss) and β (related to the velocity of propagation). Circuit-simulation tools such as SPICE typically model transmission lines using telegraphers' parameters. These are shown versus frequency in Fig. 7 for a unit length of 53 cm. The capacitance and inductance are fairly constant with frequency, the resistance is proportional to the square root of frequency (due to the skin effect), and the conductance is directly proportional to frequency (due to dielectric losses).

In characterizing devices for data communication applications, they should add minimal distortion to waveforms passing through them. Such devices can be characterized by comparing output waveforms to input waveforms through an eye diagram. A test pattern, based on a 10-b idle pattern, was used to evaluate the test transmission line. With a bit width of 500 ps, the eye pattern is clearly distinguishable (Fig. 8, top). When the bit width is shorted to 200 ps (Fig. 8, bottom), the eye closes, indicating difficulty in distinguishing between high and low logic states (and leading to higher BERs).

There are many advantages to applying frequency-domain measurements to the analysis of signal integrity. Frequency-domain measurement systems such as ATN's model ATN-4002A can provide much higher accuracy than TDRs due to correction of systematic sources of measurement error. They can also provide a dynamic range of more than 110 dB in the time domain, which is useful for isolating crosstalk and identifying EMI problems. Also, with a 20-GHz frequency-domain system, impulse and rise-time measurements of 35 ps can be made. The system can also be used to analyze the effects of phase skew on balanced transmission lines. In addition, such systems allow the operator to quickly evaluate components at different characteristic impedances. **ATN Microwave, Inc., 101 Billerica Ave., Building 4, North Billerica, MA 01862; (877) ATN-TOOLS, (978) 667-4200 ext. 8029, FAX: (978) 667-8548, Internet: http://www.atnmicrowave.com.**

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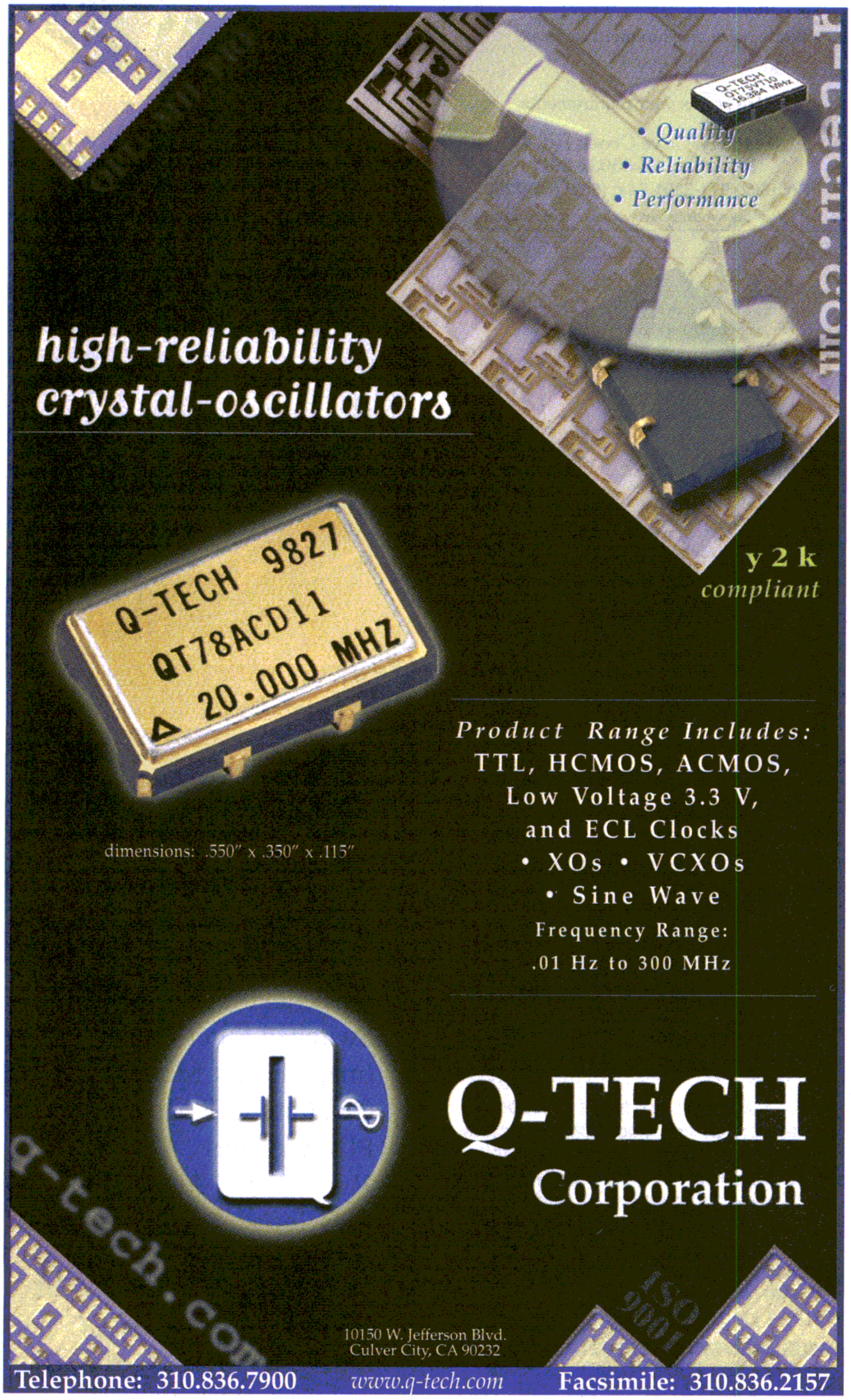

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TOIP (dBm)	34.0	36.0	34.0
P1dB (dBm)	20.0	20.0	20.0
N.F. (dB)	3.9	3.8	2.9
Supply Voltage (Vdc)	4.2	5.0	5.2
Supply Current (mA)	75	80	75

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CIRCLE NO. 268

(continued from p.61)
RF PAs for IS-95 applications. The results also show the limitations of the technique due to the amplifier saturation. A better than 6-dB improvement in ACP is obtained for the composite LDMOS PA tested.

In a full implementation of digital predistortion in a production environment, a number of key issues not addressed in this article have to be considered. The most important is the change in bandwidth requirements in the transmission path. Since digital predistortion results in a baseband signal that exhibits a much wider bandwidth than the original signal, all components from the digital predistorter to the input of the PA must pass the wider band signal without any further distortion. In the experimental analysis on CDMA signals, it was observed that three times the original bandwidth is needed in order to maintain the predistorted signal integrity. This bandwidth requirement will certainly increase the requirements on the DAC. Additionally, based on the transmitter architecture (i.e., homodyne versus heterodyne), substantial filter redesign may be required.

The other important issue is determination of DSP million instructions per second (MIPS) required to achieve the necessary predistortion. This will be the critical factor in determining whether added cost and power consumption of baseband predistortion is justifiable for a particular application. The implementation of the lookup table also plays an important role. Specifically, the effectiveness of the predistortion algorithm will be constrained by the word length of the coefficients in the lookup table as well as the size of the lookup table. The finite word length and table size are manifested as quantization noise in the system.[10] ●●

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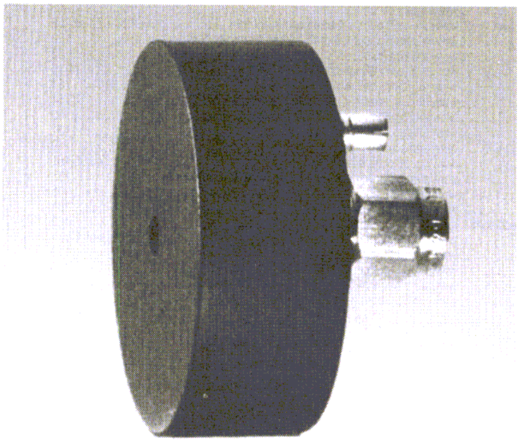

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Microwaves & RF April Editorial Preview

Issue Theme: Communications

News

Communications is the hottest market area for high-frequency electronics at present, offering growth opportunities much like those of the computer industry a decade earlier. What are the hottest markets in communications, and what technologies are serving them? Don't miss this Special News Report for the answers, only in the April issue of *Microwaves & RF*.

Design Features

Technical articles in April will offer various design and test techniques for improving communications-system performance. For example, an author from Agilent Technologies will provide clear guidelines for making electromagnetic-interference (EMI) radiated-emissions measurements, while another author from Wavetek Wandel Goltermann will describe some of the tests necessary for production-line evaluation of CDMAone mobile telephones.

Product Technology

April's Product Technology section will feature a host of new products from the recent Wireless Symposium/Portable By Design Conference & Exhibition held in February. For example, a complete fiber-optic communications link handles data rates in excess of 10 Gb/s, while a tunable filter helps sort cable-television (CATV) test signals from spurious distortion.

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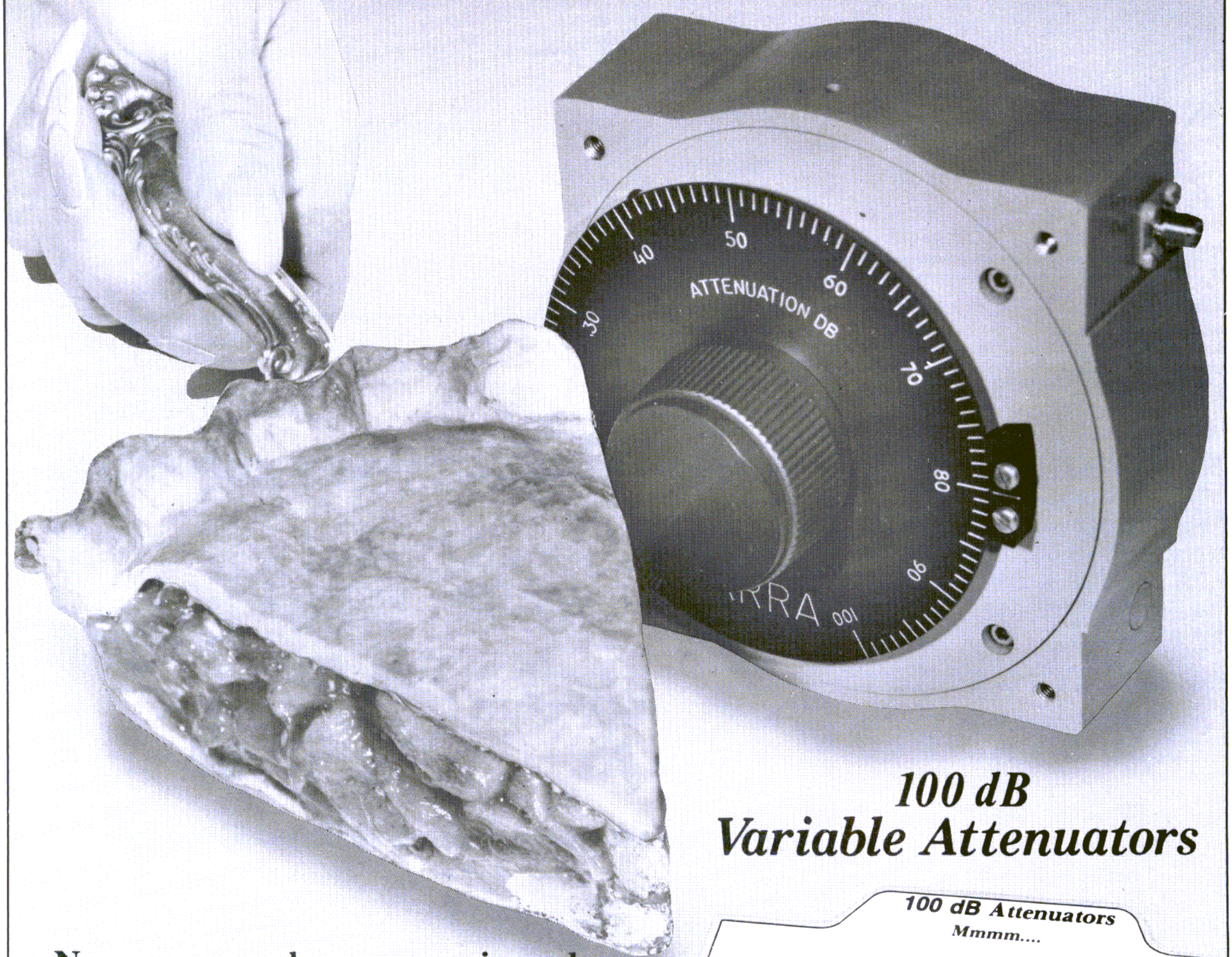

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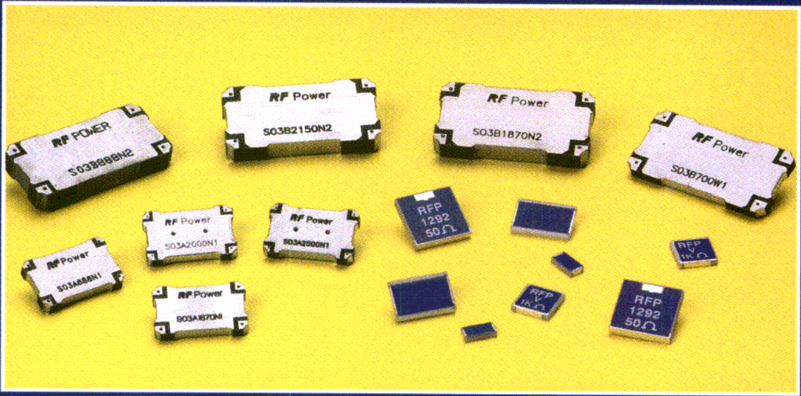

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Custom Devices

Custom devices and assemblies

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S03A888N1	815-960Mhz	100W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.25dB
S03B888N2	815-960Mhz	200W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.20dB
S03A1870N1	1750-1990Mhz	100W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.25dB
S03B1870N2	1750-1990Mhz	200W	+/-0.30dB	+/-1.5	20dB	1.25:1	0.20dB
S03A1960N1	1930-1990Mhz	100W	+/-0.20dB	+/-1.5	20dB	1.25:1	0.25dB
S03B1960N2	1930-1990Mhz	200W	+/-0.10dB	+/-1.5	20dB	1.25:1	0.20dB
S03A2000N1	1500-2500Mhz	100W	+/-0.30dB	+/-2	20dB	1.20:1	0.25dB
S03B2150N2	2000-2300Mhz	200W	+/-0.20dB	+/-2	20dB	1.25:1	0.20dB
S03A2250N1	2000-2500Mhz	100W	+/-0.30dB	+/-2	20dB	1.20:1	0.25dB
S03A2500N1	2000-3000Mhz	100W	+/-0.35dB	+/-2	20dB	1.20:1	0.30dB
S03D3500NR5	3000-4000Mhz	50W	+/-0.30dB	+/-2	18dB	1.30:1	0.30dB

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RFP-250250-6Z50-2	16	1.25:1	3 GHz
RFP-250375-4Z50-2	25	1.20:1	2 GHz
RFP-375375-6Z50-2	30	1.25:1	3 GHz

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