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June 2001

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The broadband game: Is SCM the strong player?

The unexpected benefits of polyphase filters

The rules of non-linearity in RF components

The key design details for broadband tuners

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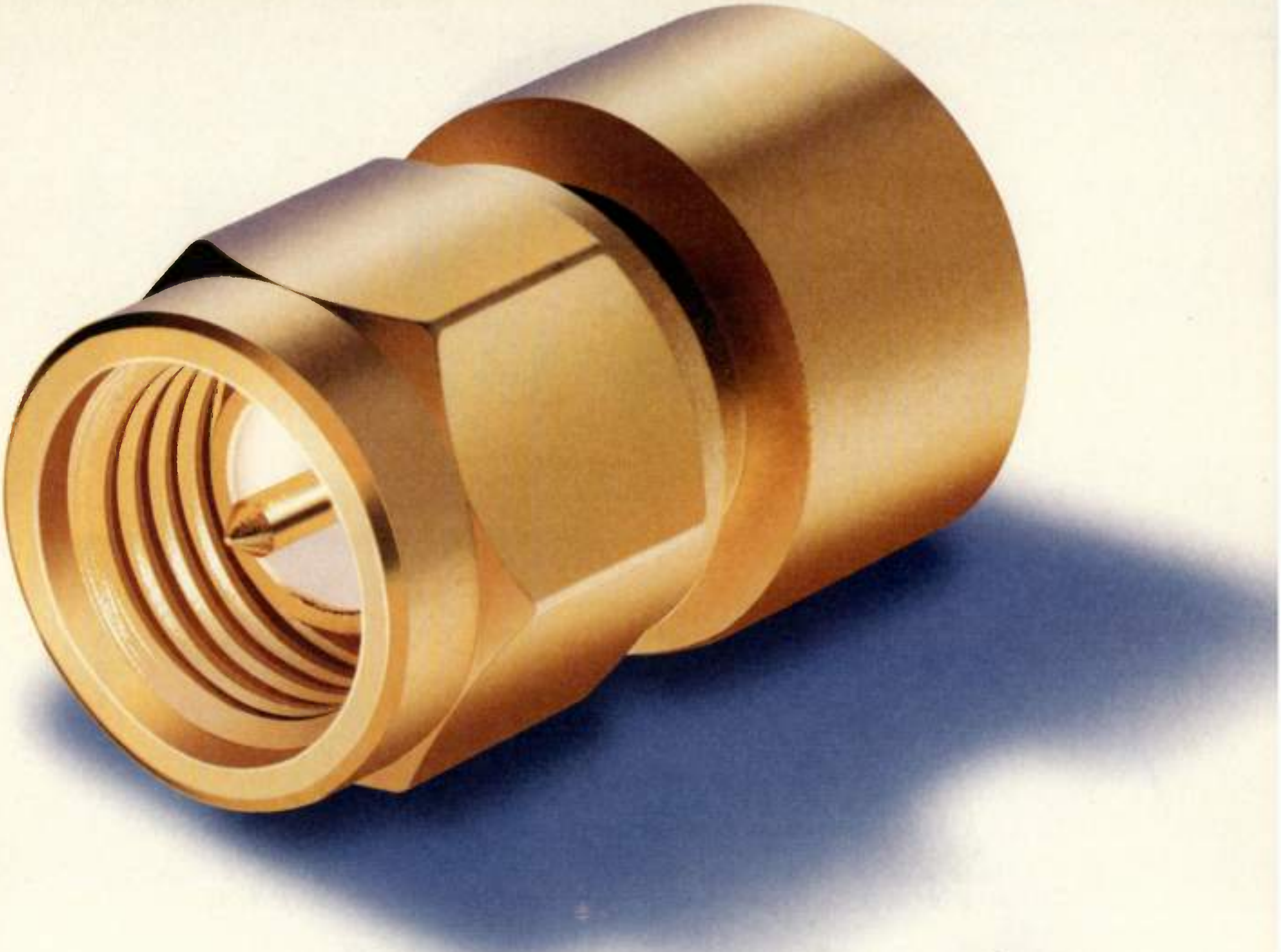

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DC to 4	35
DC to 6	28

ROSE-50
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Freq. Range (GHz)	Return Loss (dB, Typ)
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DC to 4	35
DC to 6	28

ANNE-50
SMA Male Connector
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Freq. Range (GHz)	Return Loss (dB, Typ)
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4 to 10	30
10 to 20	20

ANNE-50CN
with 3 1/2" Chain & Cap
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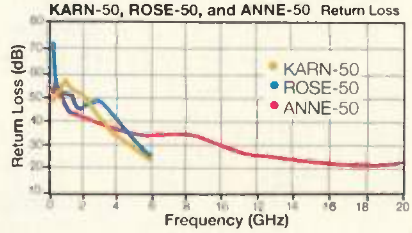

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Actual Size

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GAL-1	DC-8000	12.7	11.8	±0.5	12.2	4.5	27	108	40	3.4	.99
GAL-21	DC-8000	14.3	13.1	±0.6	12.6	4.0	27	128	40	3.5	.99
GAL-2	DC-8000	16.2	14.8	±0.7	12.9	4.6	27	101	40	3.5	.99
GAL-33	DC-4000	19.3	17.5	±0.9	13.4	3.9	28	110	40	4.3	.99
GAL-3	DC-3000	22.4	19.1	±1.7	12.5	3.5	25	127	35	3.3	.99
GAL-6	DC-4000	12.2	11.8	±0.3	18.2	4.5	36	93	70	5.2	1.49
GAL-4	DC-4000	14.4	13.5	±0.5	17.5	4.0	34	93	65	4.6	1.49
GAL-51	DC-4000	18.1	16.1	±1.0	18.0	3.5	35	78	65	4.5	1.49
GAL-5	DC-4000	20.6	17.5	±1.6	18.0	3.5	35	103	65	4.4	1.49

•Low freq. cutoff determined by external coupling capacitors. ▲Models tested at 2GHz except GAL-4, -5, -6, -51 at 1GHz
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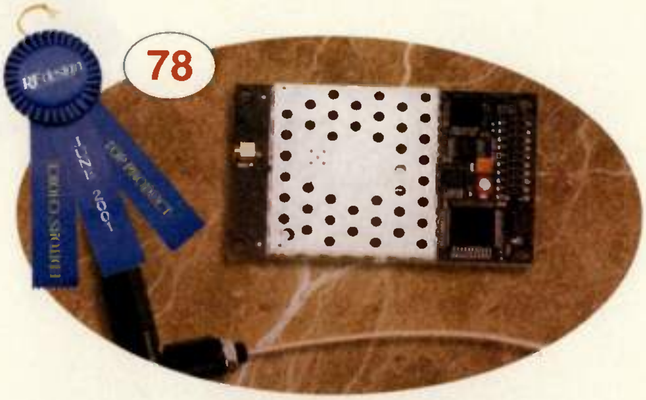

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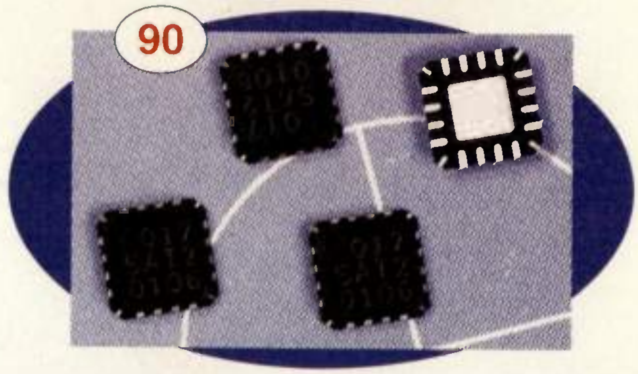

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RF editorial

What? Me worry?

By Roger Lesser
Editor
rlesser@intertec.com

I'm a science-fiction buff. Have been since I was 13 and read Edgar Rice Burroughs' "At the Earth's Core." I think that is one reason I'm a real technology buff. Now, I read other books as well, especially history books. And every now and then you can even find me reading everything from news magazines to, dare I say it, *Mad* magazine.

While I'll admit it has been sometime since I've actually picked up a copy of *Mad*, I had a friend pass one along to me recently. It was an older issue that offered a parody of Star Wars. The *Mad* folks did their usual good pickling of the Star Wars universe. But, as I thumbed through it, I couldn't help but look back at the cover with Alfred E. "What? Me worry?" Neuman.

Which brings me to my column last month. You may recall that I am not jumping on the demise of telecom bandwagon. After writing the column, I saw Warren Buffet, the legendary investor, revel in his glory as tech stocks plummeted. Once again he was all about the "blue-chip" stocks. But as I watched him it struck me that I was watching the investment world's Alfred E. Neuman. It also struck me that his lack of faith in tech stocks probably means he never read a science-fiction book.

Tech stocks vs. blue-chip stocks

"Alfred E. Buffet," I think, will still be proven to be a false idol. Who can argue against the blue-chip "What? Me worry?" stocks. But to say the tech stocks should be avoided (which he again advocated) is absurd. Why do I, "Roger E. Neuman" not worry? Because technology is the cornerstone of our future. It is our communications future. It is our defense future. It will make the future.

You may ask, and rightfully so, just how I can adopt such an Alfred E. Neuman attitude. Even with the layoffs and slowdowns that are facing the industry, there are still rays of light. (Fiber optic?) A report from the Associated Press notes that DSP manufacturers are not seeing a downturn. Semiconductor stocks have been upgraded in the eyes of investors. The projections for a rebound, which initially had the upturn taking place next year, are now more positive. Some analysts are now seeing the upturn beginning as early as late summer.

And the beat goes on

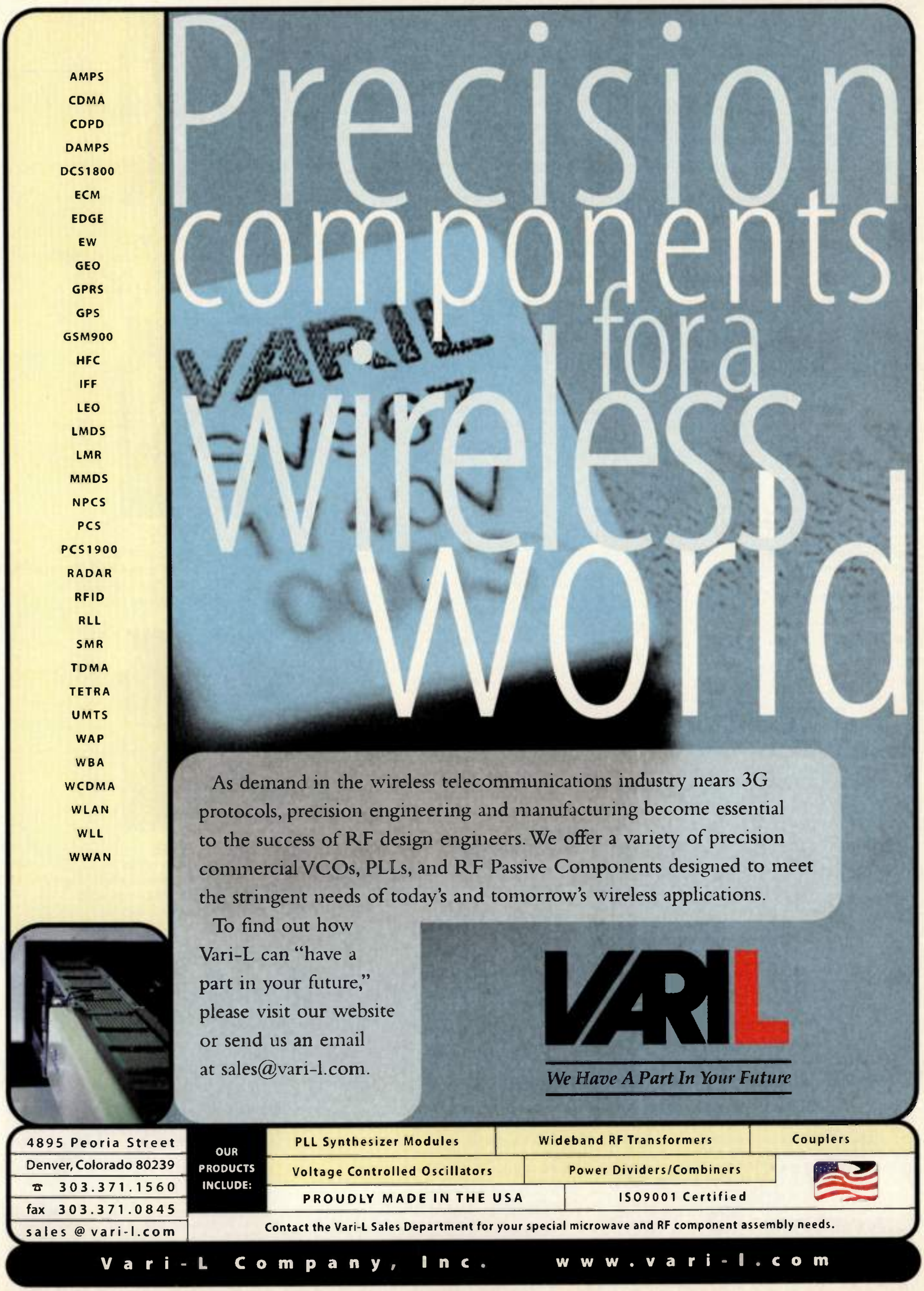
While I'm beating the drum for a more positive approach to the marketplace, I'm also beating the drum of technological advancements. I again point to the number of new products on the market. While these were designed before the downturn, companies are not entrenching or postponing production. Also, from the IEEE MTT-S IMS product announcements I have seen, there is not exactly a slowdown.

And what about drumming up support for new applications? They're coming. All one has to do is watch TV to see the spot ads for new apps.

And what about defense spending? The Bush administration will, I believe, correct the under spending of the previous administration. This will mean more opportunity for telecommunications technology and applications. Remember former Defense Secretary William Perry's commercial-off-the-shelf (COTS) mandate. Well, it is alive and well. The Department of Defense will turn to COTS devices and capability to augment its current command and control (C^2). How do I know this? First, enhanced communications capability is always a top priority. Second, while defense spending has been down over the past eight years, the one area that has maintained growth has been C^2. Third, 21 years dealing with military electronics did not leave me without a few trustworthy contacts. Enough said.

Bottomline — Stay the course and if somebody asks how you feel, smile and say, "What? Me worry?"

Roger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RF calendar

JUNE

24–27 **WCA Annual Conference** – *Boston* – Information: Wireless Communication Alliance (WCA). Tel. 202.452.7823.

24–29 **The Oxford Programme** – *Oxford* – Information: RTT Programmes Limited. Tel. +44 (0) 20. 8844.1811. Web site: **www.sttonline.com**

JULY

9–12 **Embedded Systems Conference** – *Chicago* – Information: Web site: **www.embedded.com/esc**

AUGUST

13–17 **IEEE-EMC Symposium** – *Montreal* – Information: Web site: **www.ieee.org.**

21–23 **ITCom and Opticomm 2001** – *Denver* – Information: Web site: **spie.org/info/itcom**

SEPTEMBER

4–7 **Embedded Systems Conference** – *Boston* – Information: Web site: **www.embedded.com/esc**

24–26 **EDA: Front-To-Back** – *Santa Clara* – Information: Penton Media. Tel. 1.888.947.3734.

OCTOBER

1–3 **34th Annual Connector and Interconnection Technology Symposium** – *Anaheim* – Information: Web site: **www.ec-central.org**

1–4 **Communications Design Conference** – *San Jose* – Information: Web site: **www.CommDesignConference.com**

2–4 **Sensors Expo Fall** – Philadelphia – Information: Web site: **www.sensorsexpo.com**

23–25 **Cleveland 2001 Advanced Productivity Exhibition** – *Cleveland* – Information: SME Customer Service. Tel. 800.733.4763. Web site: **www.sme.org/cleveland**

DECEMBER

3–6 **Internet World Wireless West 2001** – *San Jose* – Information: Web site: **www.ccievents.com**

RF courses

AGILENT TECHNOLOGIES – *RF and Microwave Fundamentals* – Aug. 29–31, Dec. 4–6; *Network Analysis Measurements* – Oct. 16–17; *Spectrum Analysis Measurements* – Oct. 18–19. Information: Tracey Bull, Eskdale Rd., Winnersh Triangle, Wokingham, UK; Tel. +44.118.927.6741; Fax: +44.118.927.6862; e-mail: *tracey_bull@agilent.com*

UNIVERSITY OF MISSOURI-ROLLA – *Grounding and Shielding Electronic Systems* – June 19–20, Boston; August 8–9, Toronto; *Circuit Board Layout to Reduce Noise Emission and Susceptibility* – June 21, Boston; August 10, Toronto; September 19, Denver. Information: Web site: **www.umr.edu/-conted>**

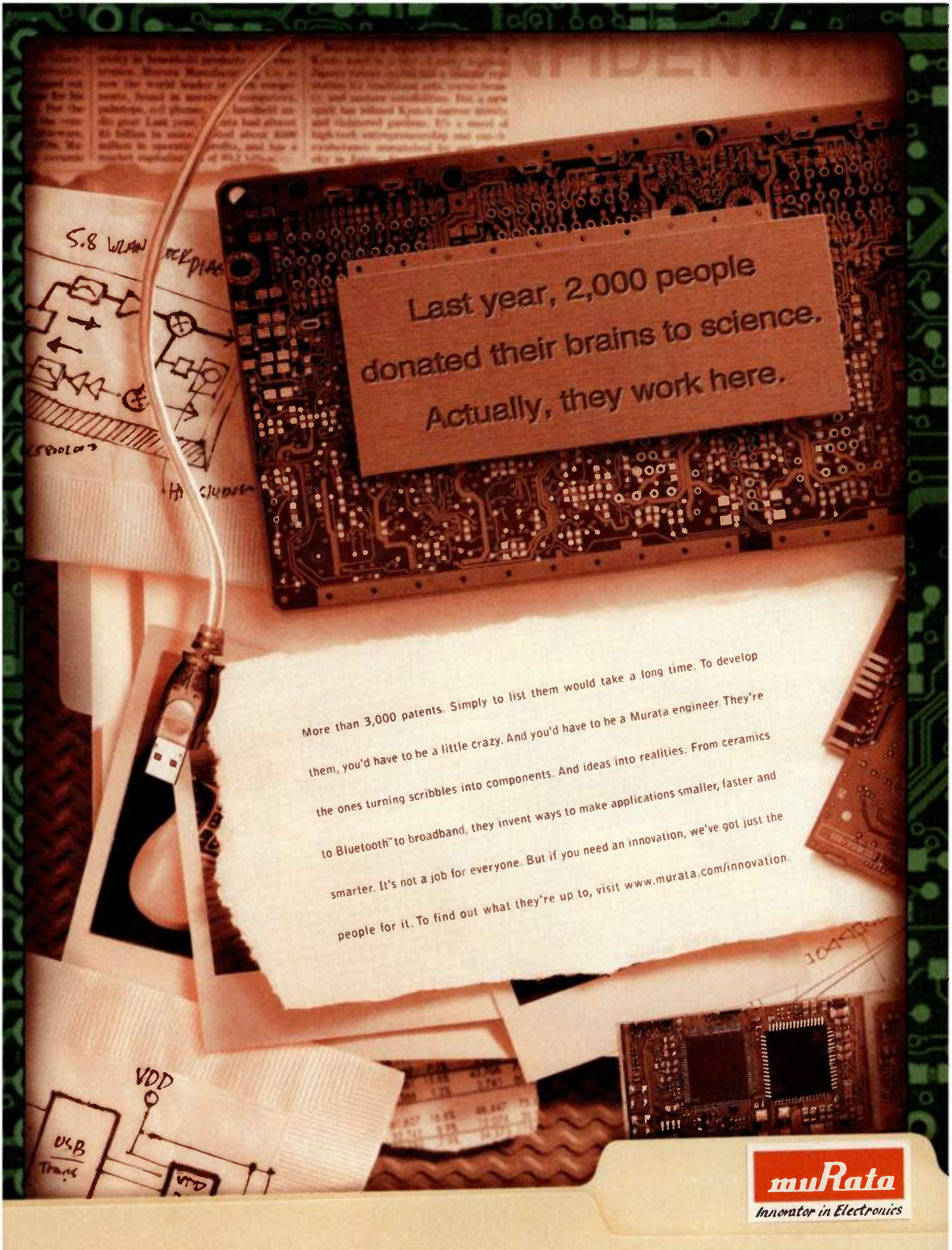
ALEXANDER RESOURCES – *Making Money in the U.S. Wireless Internet Market* – July 10–11, Chicago. Information: Jeff Stone, Alexander Resources,15851 N. Dallas Pkwy, Addison,TX 75001; Tel. 972.818.8225; Fax: 972.818.6366; e-mail: *jstone@alexanderresources.com.*

BESSER ASSOCIATES – *RF and Wireless made Simple* – July 12–13, San Francisco; Aug. 7–8, San Diego. Information: Besser Associates, 201 San Antonio Circle Building E, Suite 280, Mountain View, CA 94040; Tel. 650-949-3300; Fax: 650-949-4400; e-mail: *info@bessercourse.com*; Web site: **www.bessercourse.com**

UCLA – *Biometric Identification: Theory, Algorithms and Applications* – July 30–Aug. 1; *Digital Signal Processing: Theory, Algorithms, and Implementation* – Aug. 13–17; *Bluetooth: Technology, Applications, and Performance* – Aug. 20–22, Los Angeles, Information: Information Systems and Technical Management Short Courses. Tel. 310.825.3344; e-mail: *mhenness@unex.ucla.edu*; Web site: **www.uclaextenstion.org/shortcourses**

R.A. WOOD ASSOCIATES – *Introductory RF and Microwaves* – Sept. 20–21; *RF and Microwave Receiver Design* – Sept. 24–26; *RF Power Amplifiers, Classes A–S: How Circuits Operate, How to Design Them, and When to Use Each* – Sept. 27–28, Lake George, NY. Information: R.A. Wood Associates, 1001 Broad St. Ste. 450, Utica, NY 13501; Tel. 315.735.4217; Fax 315.735.4328; e-mail: *RAWood@rawood.com;* Web site: **www.rawood.com**

MASSACHUSETTS INSTITUTE OF TECHNOLOGY – *Design of Analog Integrated Circuits* – June 18–22; *Fundamentals of Lasers, Fiber optics and Their unique Applications* – July 9–13; *Principles of Optical Sensors* – Aug. 6–10, Cambridge, MA. Information: Tel. 617.253.2101; Fax 617.253.8042; e-mail: *professional-institute@mit.edu* Web site: **www.web.mit.edu/professional/summer**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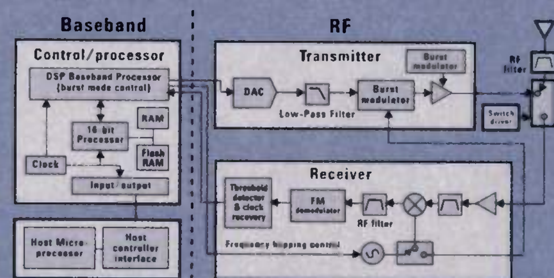
Something for the RF experts, too. If you are approaching *Bluetooth* wireless technology from an RF background, we can offer advice on the most-efficient test procedures and toolsets to solve a wide range of *Bluetooth* measurements.

The *Bluetooth* big picture. Most of the *Bluetooth* work we're seeing today involves the integration of a *Bluetooth* module into a new product design:

- Evaluating module performance and characterizing interoperability
- Understanding host-module integration issues
- Designing and debugging the host module interface
- Conducting pre-qualification RF testing
- Getting *Bluetooth* Qualification
- Manufacturing quality products

Some of the more interesting problems show up in the second stage, as you bring the RF transceiver into your host products.



Watch out for some interesting interoperability problems when you integrate a Bluetooth module into your host device

Baseband signal integration. Challenges here include verifying transmission and receipt of data packets, viewing the actual data values transmitted, quantifying system bottlenecks, identifying logic errors, and resolving DSP and mixed-signal issues.

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Bluetooth

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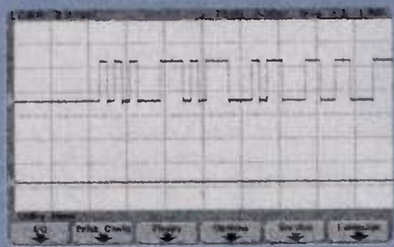
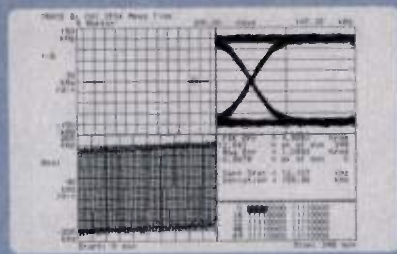
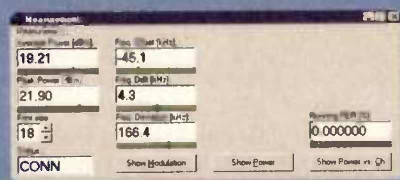
For instance, once you've found the preamble, you can identify the entire bit stream, including the access code, header and payload. Learn more in our free *Bluetooth* baseband application note.



The first two pulses in this idealized transmit signal correspond to the 0101 pattern of the preamble; the access code follows immediately after.



Bluetooth measurement tools range from powerful design analysis to fast, automated tests for the production line. Above, a modulation characteristics test verifies proper performance of the modulation circuitry to ensure reliable data transfer over the Bluetooth communication link.



Above, an automated test combines pass/fail indications with numerical readouts.

RF receiver tests. RF receiver performance is key to both *Bluetooth* qualification and overall product performance. For example, a sensitive radio that is immune to interference will reduce file transfer times and therefore increase battery life. You need to make sure the RF receiver will not be adversely impacted by the harmonics of high-frequency digital signals or other noise sources likely to be present in your system.

Receiver performance is tested in a number of ways for qualification, including carrier/interference and blocking tests. You probably won't need to run all the tests if you're integrating someone else's module, but they can be complicated so clear information and simplified procedures are important.

RF transmitter tests. The *Bluetooth* specification covers a wide range of transmitter tests, some to insure interoperability between *Bluetooth* devices (e.g., modulation characteristics) and others to meet regulatory limits (e.g., spurious emissions). Given the concerns about interference with other wireless systems, output spectrum tests are also important.

Integrating a module can create problems that affect transmitter performance, sometimes in unexpected ways. For example, power supply ripple coupled through your system can degrade the modulation characteristics.

You must be able to show that your device stays within both *Bluetooth* and regulatory limits, and the more of this work you can do on your design bench, the better. Some of the tests are complex and potentially time consuming to understand and perform. Our free online application resources can help you look for and fix problems quickly.

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Agilent Technologies

Location, location, location

By Nikki Chandler
senior associate editor
nchandler@intertec.com

Last summer, I spent the day at a public safety answering point (or dispatch center) observing the management of radio traffic and listening in on the dispatchers' 9-1-1 calls. Most calls that came in during that long afternoon were from cellphones.

The dispatchers explained that Good Samaritans often call when someone is in trouble–an accident occurs or a car breaks down–but many of the callers cannot give their location. At first that was difficult for me to comprehend. Then I remembered the last time I tried to find my way home from downtown Kansas City and ended up in a less-than-safe part of town.

Absence of location capability jeopardizes public safety. Many people purchase cellphones for security, not realizing that if an emergency does happen, they may not be found if they cannot identify or describe their location.

Automatic location identification (ALI) was designed for landline phones. It doesn't work yet with all cellphones, and most of the public doesn't realize that. In 1996, under pressure from the public safety community, the Federal Communications Commission (FCC) adopted enhanced 9-1-1 (E9-1-1) location rules and set Oct. 1, 2001, as the deadline for cellular ALI capability.

But will these companies meet this deadline? AT&T Wireless and Nextel have already filed for extensions. Nextel even told the FCC that it would give $25 million to the Association of Public-Safety Communications-International (APCO) for dispatch centers (like a 16th-century indulgence) if the commission would just give it more time.

Technical obstacles to creating E9-1-1 services remain, but the technology to implement the service exists. If the FCC hadn't set some kind of deadline, the service would be delayed indefinitely, and more lives would be lost.

The FCC has a scarcity of engineers, and the commission often issues orders and passes rules without appreciating the technical feasibility of compliance. But the officials from APCO and the National Emergency Number Association, who have called for this mandate, contend that the technology exists.

By forcing this issue on carriers, the FCC has the public interest at the forefront. As technology advances, carriers can update equipment and networks for more cost-effective and accurate results. They will also find ways to exploit this mandatory capability for additional applications–for profit.

In the meantime, if someone carjacks my vehicle with me in it, and if I can dial 9-1-1 without his knowing it, I'd like the dispatcher to know where we are going. We may be heading into downtown Kansas City.

Nikki

FCC recommends increased enforcement

Federal Communications Commission (FCC) Chairman Michael Powell has recommended that Congress increase the forfeiture level imposed on common carriers violating local competition provisions of the Telecommunications Act of 1996 from the current statutory limit of $1.2 million per violation to at least $10 million per violation.

"In my discussions with competitive local exchange carriers," said Powell, "they site enforcement as the key area for increased regulatory effort."

In a letter sent to leaders of the Senate and House Commerce and Appropriations Committees in May, Powell also recommended a longer statute of limitations than the current one-year period for FCC investigations of local competition violations. He also asked for Commission authority to award punitive damages in formal complaint cases.

The FCC must "vigorously enforce the local competition provisions of the 1996 Act," Powell said. "I believe there is more that we can do with the help of Congress."

OADM market to reach $4 billion in 2006

The market for optical add-drop multiplexers (OADM) will grow from $338 million in 2000 to more than $4 billion in 2006, representing a 53% compound average annual growth rate, according to a new report by Allied Business intelligence (ABI), Oyster Bay, NY.

This growth will be driven by both growth in dense wavelength division multiplexing (DWDM) systems and the increasing use of more expensive reconfigurable OADMs.

ABI's new report, "*Optical Add-Drop Multiplexers: World markets and Opportunities for Fixed and Reconfigurable OADMs in the Long-Haul and Metro Networks*," provides an overview of the markets for fixed and remotely configurable OADMs. It forecasts shipments to submarine, long-distance terrestrial and metro segments of the world.

DWDM is still in the early stages of development. Its deployment in the terrestrial long-haul applications is well-advanced in North America and has occurred over the past two years

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ADE-12H	500-1200	+17	28	1.1	6.7	8.95
• MBA-591L	4950-5900	+4	15	1.1	7.0	6.95
SYM-25DLHW	40-2500	+10	22	1.2	6.3	7.95
SYM-25DMHW	40-2500	+13	26	1.3	6.6	8.95
SYM-24DH	1400-2400	+17	29	1.2	7.0	9.95
SYM-25DHW	80-2500	+17	30	1.3	6.4	9.95
SYM-22H	1500-2200	+17	30	1.3	5.6	9.95
SYM-20DH	1700-2000	+17	32	1.5	6.7	9.95
SYM-18H	5-1800	+17	30	1.3	5.75	9.95
SYM-14H	100-1370	+17	30	1.3	6.5	9.95
SYM-10DH	800-1000	+17	31	1.4	7.6	9.95

*E Factor = [IP3 (dBm) - LO Power (dBm)] ÷ 10. See web site for E Factor application note.
ADE models protected by U.S. patent 6,133,525
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BUSINESS BRIEFS

PDF Solutions, Cadence join to develop software — PDF Solutions, San Jose, and Cadence Design Systems, San Jose, have announced plans to jointly develop a complete yield-analysis and improvement software for analog and mixed-signal integrated circuit (IC) designers. The software will be designed to improve the profits and competitiveness of both fabless and merchant IC companies through improved design yields and performance.

TRW creates new company — TRW, Redondo Beach, CA, announces the creation of a new company that will develop and sell TRW's high-speed indium phosphide (InP) and gallium arsenide (GaAs) telecommunications products for digital and analog applications. The company is called Velocium and is located in Manhattan Beach, CA.

NEC to spin off Microwave and Optical Semiconductor Division — NEC, Santa Clara, CA, announces its intention to spin off its Compound Semiconductor Device Division (CSDD) into a new company, effective Oct. 2001. CSDD designs and manufactures silicon and GaAs RF and microwave semiconductor devices, as well as optical semiconductor devices.

LSI Logic licenses Newlogic Bluetooth technology — LSI Logic, Milpitas, CA, announces an agreement with Newlogic, Lustenau, Australia, in which the company will license Newlogic's Boost Core and Boost Software IP products. The agreement allows LSI Logic to add Bluetooth functionality to its portfolio of standard products and offer the Boost Core as an element in its CoreWare library of intellectual property.

Emblaze Systems, Sonofon sign trial agreement — Emblaze Systems (formerly GEO Interactive Media Group), Las Vegas, NV, announces plans to conduct multiphase trials of its video streaming solutions over wireless networks with Sonofon, Denmark. The trial began in March and is projected to last through Oct. 2001.

Zucotto Wireless, Plazmic partner — Zucotto Wireless, San Diego, and Plazmic, Tokyo, announce a partnership whereby Plazmic will port its Media Engine to Zucotto's Xpresso Java technology-based processor. The partnership is intended to make it easier for content developers to create content for devices that use the Xpresso processor.

StratEdge to design, produce, assemble packages for TriQuint — TriQuint Semiconductor, Dallas, has contracted StratEdge, San Diego, to assemble and test TriQuint OC-192 broadband amplifiers using patented, high-performance StratEdge packages. StratEdge has adapted its standard 580286 leaded ceramic power amplifier package for TriQuint's application.

Siliconix to transfer LDMOS IP, business to Linear Integrated Systems — Siliconix, Santa Clara, announces that it is transferring all of its lines of lateral DMOS switch transistors to Linear Integrated Systems, Fremont, CA. Under the agreement, Siliconix will transfer to Linear Systems' masks, technical expertise, and marketing information for the covered products, which include the SD5000, SD5400, SST211, and SST215. In exchange, Linear Systems will support the market for these products for a minimum period of four years, and will also pay Siliconix royalties.

in Europe. DWDM is just now appearing in long-haul applications in other parts of the world and in metro networks. The metro market is forecasted to grow rapidly, beginning in North America and Europe.

Fixed OADMs have been used in terrestrial long-haul applications for years. However, remotely reconfigurable OADMs offer advantages, particularly in metro networks, where there will be frequent reprovisioning of the systems as carriers gain and lose customers.

Remotely reconfigurable OADMs are more expensive, and their use will depend on the market. Carriers outside of Western Europe and North America, where economies are less developed, might avoid the extra cost.

Consortium researching 3G phones for the deaf

A consortium of researchers from Germany, Sweden, Spain and the U.K. is working together on a project to incorporate sign language recognition technology into third-generation (3G) cellular phones, according to a May 11 CNN.com article by Rick Perera. The project is called Wireless Information Services for Deaf People on the Move (Wisdom).

With this technology, hearing-impaired users will be able to call up news, weather and sports information in sign language from a video server via 3G phones, give commands to their phones in sign language, and access a real-time interpretation service to help them communicate with hearing people.

In laboratory conditions, the technology can already understand sentences having as many as nine signs in them with a recognition rate of 90%.

The European Union (EU) is giving Wisdom 6 million euros (U.S. $5.3 million) in funding over the next three years, with sights set on helping 0.2% of the European public that communicates primarily in sign language.

Bluetooth market to shine: 955M units in 2005

Despite delays, economic slowdown and a recent slew of negative reports, the emerging Bluetooth market will eventually shine, according to Cahners

In-Stat Group, Scottsdale, AZ.

The firm forecasts that demands for Bluetooth-enabled devices will provide substantial opportunities for the technology, with Bluetooth-enabled equipment shipments soaring to 955 million units in 2005, a 360% five-year compound annual growth rate (CAGR). The semiconductor opportunity in this area will also be substantial as Bluetooth radio and baseband silicon will rise to $4.4 billion in 2005.

The first "hot spot" projects have already appeared in hotels, shopping malls, golf courses, and airports. And more are expected to come to fruition by the end of the year. Aside from hardware, there is activity happening in application development, both on the client side and the server/services side. In-Stat expects that this activity will increase.

In-Stat has also found that adapters and cards will rule the lion's share of the market in the near term, then relinquish the market to embedded Bluetooth solutions; however, the adapter/card market will represent a significant market even in 2005.

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PACKAGING

VARIETY

Using polyphase filters as image attenuators

Using active polyphase filters in receivers with image rejection mixers not only reduces component count, but also provides some unexpected benefits.

By Tom Hornak

A large market exists for low-cost, gigahertz-band radio receivers used for data communications. Low cost demands single-chip implementation with minimum off-chip components. An important goal is to replace any external intermediate frequency (IF) filters with on-chip IF filters while maintaining sufficient image rejection. One possible solution for this challenge is a direct conversion approach—for example, using a zero IF frequency. But direct conversion has many well-known drawbacks, such as DC offset, 1/f noise and local oscillator leak-through. A solution that avoids these problems uses a low-megahertz IF frequency, where on-chip filters can be built within the high-frequency limitations of IC processes[1].

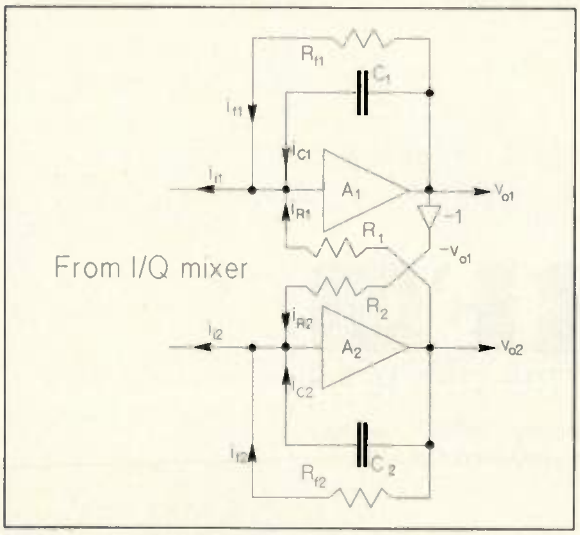

Figure 1. An active polyphase filter stage.

However, a low IF keeps the image frequency so close to the target frequency that suppressing the image in front of the mixer would require an impossibly high Q of the filter preceding the mixer. The solution here is to use an image-rejecting I/Q mixer that delivers two outputs in quadrature to two IF filters. The target signal is then separated from the image signal by the unique phase difference between the two mixer outputs. If the first output's phase lags behind the second output's phase by 90° for the target signal, then the first output's phase leads the second output's phase by 90° for the image signal. The two mixer outputs are usually then filtered by two separate matched IF filters that do not discriminate between the target signal and the image signal. Image rejection is then achieved by an image rejector that shifts the phase of one filter's output by an additional 90° and adds it with the second filter's output. When choosing the proper setup, the two filter outputs from the target signal enter the rejector in phase and add up, while the two filter outputs of the image signal enter the rejector in opposite phase and get subtracted.

Enter the polyphase filter

A better possibility is to replace the two separate filters with one polyphase filter[1]. This technique has three advantages. First, the frequency response of a polyphase filter depends on the phase difference between its two input signals. So, contrary to two separate filters, it has a passband response for the target signal and an attenuating response for the image signal. In low-IF data receivers, the data bandwidth is a significant fraction of the IF filter's center frequency, i.e. the IF filters must have low Q. The frequency response of conventional low-Q bandpass filters is not symmetrical around the passband's center frequency. That distorts the received data's eye diagram.

The second advantage of the polyphase filter is that its bandpass response is symmetrical around the passband's center frequency, independent of its Q. (The polyphase filter keeps the data's eye diagram intact.)

The third advantage of polyphase filters is that, for the same degree of image suppression, the matching of their components is less stringent than the required matching in two separate IF filters and in the subsequent image rejector. This is because of the polyphase filters' close cross-coupling.

Polyphase filters can be entirely passive, built of only resistors and capacitors (see Figure 6 of [1]). An implementation more suitable for monolithic integration is the active polyphase filter [2,3]. The operation of an active polyphase filter is not obvious from its circuit topology. But analyzing the filter's voltage and current phasors leads to a clearer understanding of the filter's useful properties. The filter's bandpass response to the target, its attenuation of the image and its sensitivity to component mismatch will be examined. All voltage and current symbols appearing in the following sections represent magnitudes. The phase relations are described in the diagrams.

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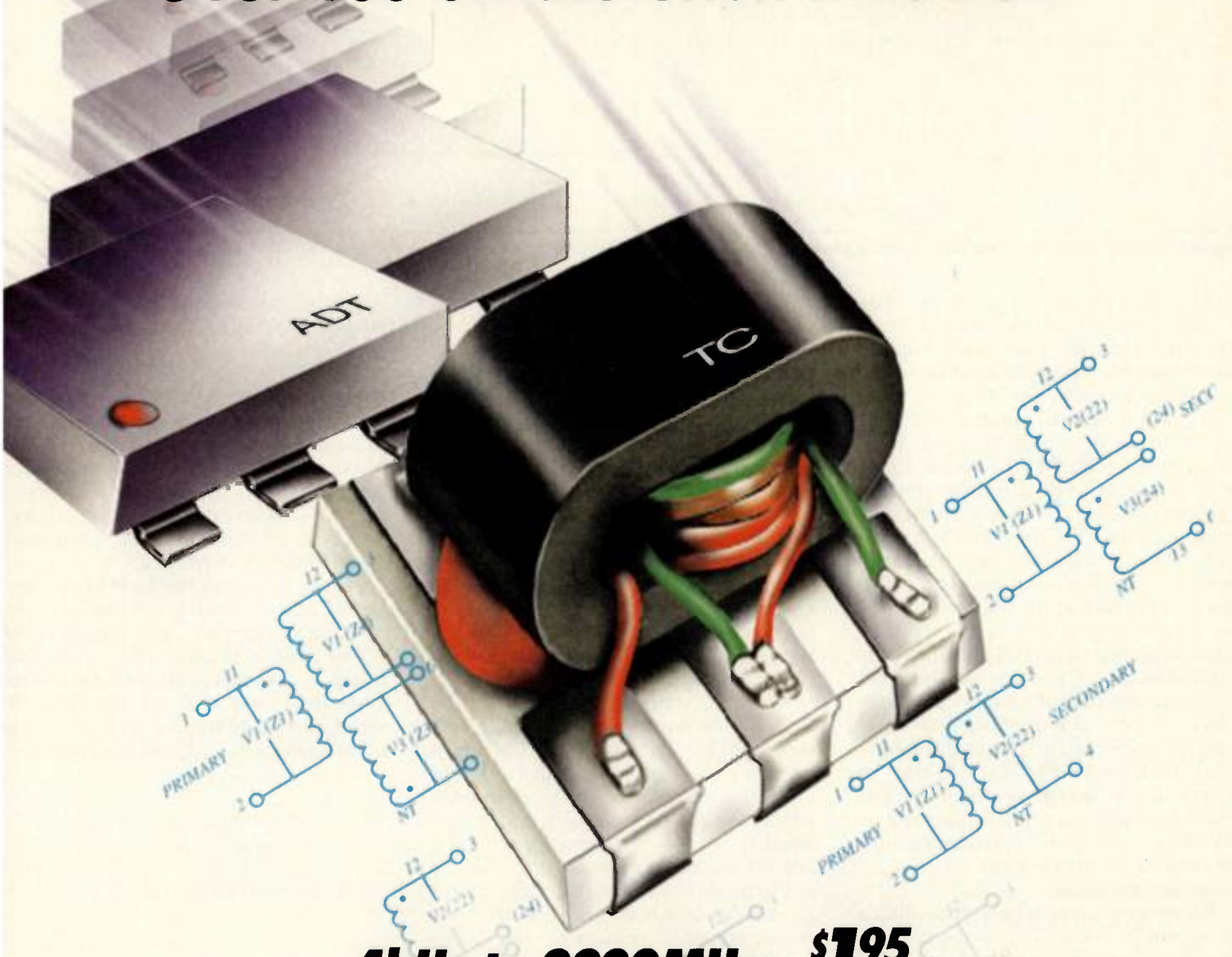

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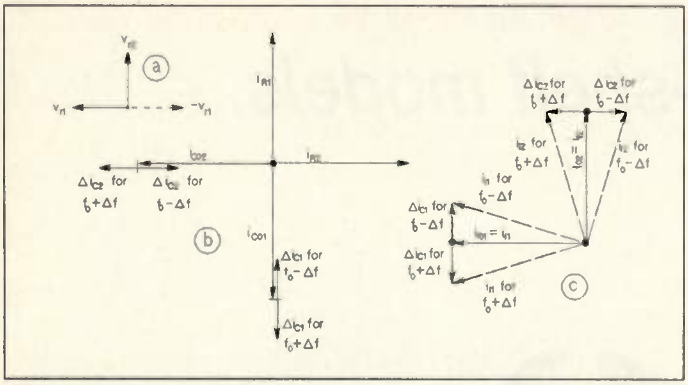


Figure 2. Phasors of the polyphase filter receiving a target signal

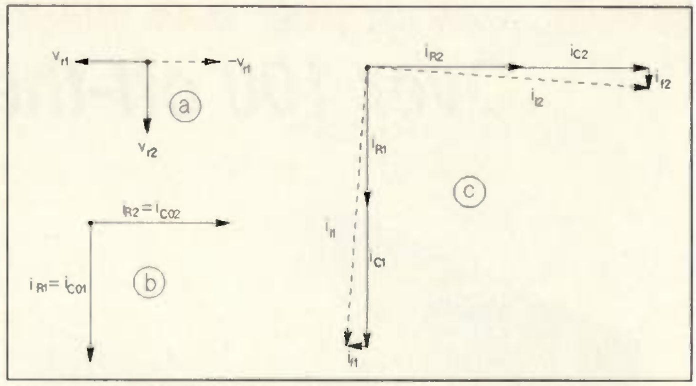


Figure 3. Phasors of the polyphase filter receiving an image signal.

The polyphase filter stage (see Figure1) includes two damped integrators (A, C, R_f), two cross-coupling resistors (R) and one unity-gain inverting amplifier (-1). In an ideal polyphase filter, all components with Suffix 1 exactly match the same component with Suffix 2. In an ideal polyphase filter, the operational amplifiers (A) also have sufficient open-loop gain at the filter's operating frequency band to keep the amplifiers' input voltages within very small fractions of the amplifiers' respective output voltages. Then, the voltage across all resistors and the integrating capacitors (C) is essentially equal to the filter's output voltage. In monolithic IC form, it is preferable to build the filter in full differential mode (and not as shown for clarity in Figure 1). In that case, the filter has four input signals phase-staggered by 90°. The unity-gain inverting amplifier (-1) is implemented simply by crossing two leads.

When measuring a polyphase filter's frequency response, one would normally test the filter's transimpedance Z(f) = v_o/i_i (note how the filter's output voltages v_o vary with frequency f when the input currents i_i are kept fixed). However, understanding this filter's operation is easier when investigating the filter's transadmittance Y(f) = i_i/v_o. Thus, follow how the filter's input currents i_i must be varied with frequency f to keep the filter's output voltages v_o fixed and equal to preset reference voltages v_{r1} and v_{r2}, respectively.

Bandpass property

Figure 2a shows the integrators' fixed output voltage v_{r1} and v_{r2} and the inverted output voltage $-v_{r1}$. Phasor v_{r1} leads v_{r2}. Figure 2b shows current i_R in resistors R and i_{C0} in capacitors C when the filter's input is at its center frequency f_0 . As follows from Figure 1, current i_{R1} is in phase with voltage v_{r2}, and current i_{R2} is in phase with $-v_{r1}$. Capacitor current i_{C01} leads voltage v_{r1} and is in opposite phase with current i_{R1}. Current i_{C02} leads voltage v_{r2} and is in opposite phase with current i_{R2}.

To tune the filter's center frequency to f_0, resistors R are set to $R = 1/(2\pi f_0 C)$. The filter's bandwidth f_b is set by the feedback resistors $Rf = 1/\pi f_b C$. The filter's Q, or the ratio f_0/f_b, is $Q = R_f/(2R)$. Following Figure 1, for any frequency f, the current in resistors R is $i_R = v_r/R$, the current in resistors R_f is $i_f = v_r/R_f$, and the current in the integrating capacitors C is $i_C = 2\pi f C v_r$. At center frequency f_0, the capacitor currents are $i_{C0} = 2\pi f_0 C v_r$, and because $R = 1/(2\pi f_0 C)$, $i_R = i_{C0}$. The filter's input currents i_{i0} are the sum of i_f , i_R and i_{C0}. Thus, at f_0 with i_R and i_{C0} of opposite phase, the filters' input currents i_{i0} are equal to the filter's feedback currents i_f and the output voltages are $v_r = i_{i0}R_f$.

Figure 2c shows the feedback current i_{f1} and i_{f2}, again in phase with their respective drive voltage v_{r1} and v_{r2}, as follows from Figure 1. As stated above, at center frequency, f_0 input currents i_i match the feedback currents i_f. Therefore, in Figure 2c, phasors i_{i0} are identical with phasors i_f.

When $f = f_0$, currents i_{C0} in Figure 2b are equal to i_R and cancel one another. However, with $f \neq f_0$, currents i_C in the integrating capacitors (C) are different from i_{C0}. Thus, with a fixed capacitor voltage $v_c = v_r$, the capacitors demand difference currents $\Delta i_C = i_C - i_{C0} = 2\pi(f - f_0)Cv_r$. When $f = f_0 + \Delta f$ ($\Delta f > 0$), the reactance of capacitors (C) is smaller, thus currents i_C must increase, and currents Δi_C for $f = f_0 + \Delta f$ are in phase with current i_{C0}. On the contrary, when $f = f_0 - \Delta f$, the reactance of capacitor (C) is larger, thus currents Δi_C for $f = f_0 - \Delta f$ are of opposite phase than currents i_{C0}. However, with v_o fixed at v_r, the currents $i_R = v_r/R$ and feedback currents $i_f = v_r/R_f$ are also fixed, independent of frequency f and equal to their magnitude at f_0. Therefore, difference currents Δi_C can come neither from R nor from R_f, but can come only from the filter inputs. Any change Δi_C in capacitor currents i_C due to a deviation from the center frequency f_0 will add an equal component Δi_C to the input currents i_i.

Figure 2c shows the total input currents i_{i1} and i_{i2}, each consisting of a vector sum of current i_f and the currents Δi_C for $f = f_0 + \Delta f$ and $f = f_0 - \Delta f$, respectively. Figure 2c confirms that currents i_i are the lowest at $f = f_0$. Thus, the filter's transimpedance for a target signal is the highest at $f = f_0$ and is equal to:

$$Z_{0t} = \frac{v_r}{i_{i0}} = \frac{v_r}{i_f} = R_f$$

The total input current is:

$$i_i = \sqrt{i_f^2 + \Delta i_c^2} = v_r\sqrt{\left(\frac{1}{R_f^2}\right) + (2\pi C)^2 (f - f_0)^2}$$

With input current i_i being a quadratic function of the frequency difference $f - f_0$, the filter's transadmittance Y(f) = i_i/v_r is symmetrical around center frequency f_0. The same applies to the filter's transimpedance Z(f) = 1/Y(f). As stated in the introduction, this is one of the benefits of polyphase bandpass filters when applied to low IF data receivers.

Image suppressing property

Phasor v_{r1} was chosen, leading v_{r2} in Figure 2, to achieve a bandpass response of the filter. Thus, to ensure an attenuating response, a phasor v_{r2} will now be chosen, leading v_{r1}.

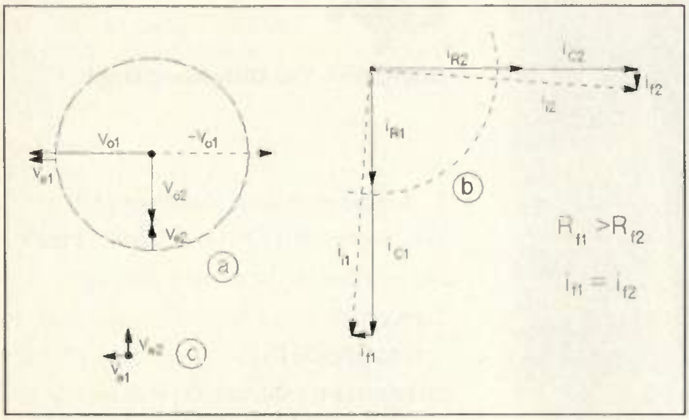


Figure 4. Mismatch of feedback resistors R_f.S.

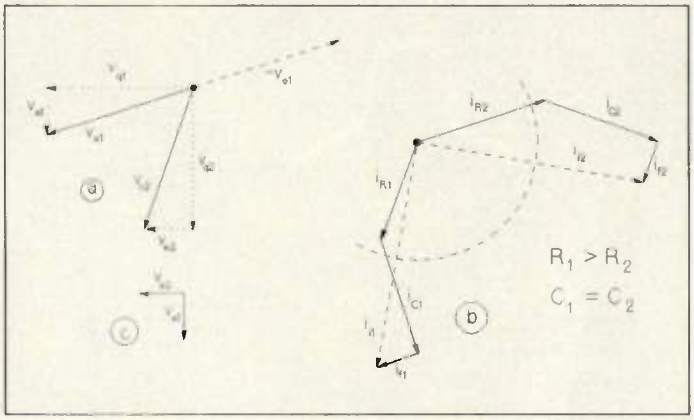


Figure 5. Mismatch of cross-coupling resistors R_1 and R_2.

The situation at frequency f_0 will be followed, which is also the center frequency of the image signal. Figure 3a shows output voltage v_{r1} and v_{r2} and the inverted output voltage $-v_{r1}$. Figure 3b shows the current i_R in resistors (R) and current i_{C0} in capacitors (C). As follows from Figure 1, current i_{R1} is in-phase with voltage v_{r2}, and current i_{R2} is in-phase with $-v_{r1}$. Capacitor currents i_{C01} and i_{C02} lead the respective output voltages v_{r1} and v_{r2}.

Following Figure 1, the current in resistors (R) is $i_R = v_r/\mathrm{R}$, and substituting for (R), $i_R = 2\pi f_0 C v_r$. The current in the integrating capacitors (C) is $i_{C0} = 2\pi f_0 C v_r = i_R$. In Figure 2b, the phase of capacitor currents i_{C0} was opposite the phase of currents i_R, therefore they canceled and made no contribution to input current i_{i0}. In Figure 3b, the phase of capacitor current i_C is the same as the phase of current i_R. Therefore, as shown in Figure 3c, their contribution to input current i_{i0} is their sum.

Figure 3c displays feedback current i_{f1} and i_{f2}, again, in phase with their respective drive voltages v_{r1} and v_{r2}. At frequency f_0, the filter's input currents i_{i0} will be the vector sum of feedback currents i_f, i_R and i_{C0}, and with $i_{C0} = i_R$, the vector sum of i_f and $2i_R$. With $i_f = v_r/\mathrm{R}_f$ and $i_R = v_r/\mathrm{R}$, input currents i_{i0} at frequency f_0 will be:

$$i_{i0} = v_r\sqrt{\left(\frac{1}{R_f^2}\right)+\left(\frac{4}{R^2}\right)}$$

The transimpedance for the image will be:

$$Z_{0i} = \frac{v_r}{i_{i0}} = \frac{1}{\sqrt{\left(\frac{1}{R_f^2}\right)+\left(\frac{4}{R^2}\right)}}$$

Comparing Figs. 2c and 3c, it can be seen that the filter's bandpass response occurs when input current i_{i1} leads input current i_{i2}, while attenuating response occurs when input current i_{i2} leads input current i_{i1}.

The degree of image suppression S_0 at frequency f_0 is equal to the ratio of the filter's transimpedance Z_{0i} for the image signal and the transimpedance Z_{0t} for the target signal. Substituting for $Z_{0t} = R_f$ yields:

$$\frac{Z_{0i}}{Z_{0t}} = S_0 = \frac{1}{\sqrt{1+4\left(\frac{R_f}{R}\right)^2}} = \frac{1}{\sqrt{1+16Q^2}} \approx \frac{1}{4}Q$$

Component mismatch sensitivity

In the previous two sections, it was assumed that the polyphase filter's components with Suffix 1 exactly match the same component with Suffix 2. Now analyze the effect of a mismatch between components in Figure1 when the filter input is an image signal, i.e. i_{i2} leading i_{i1}.

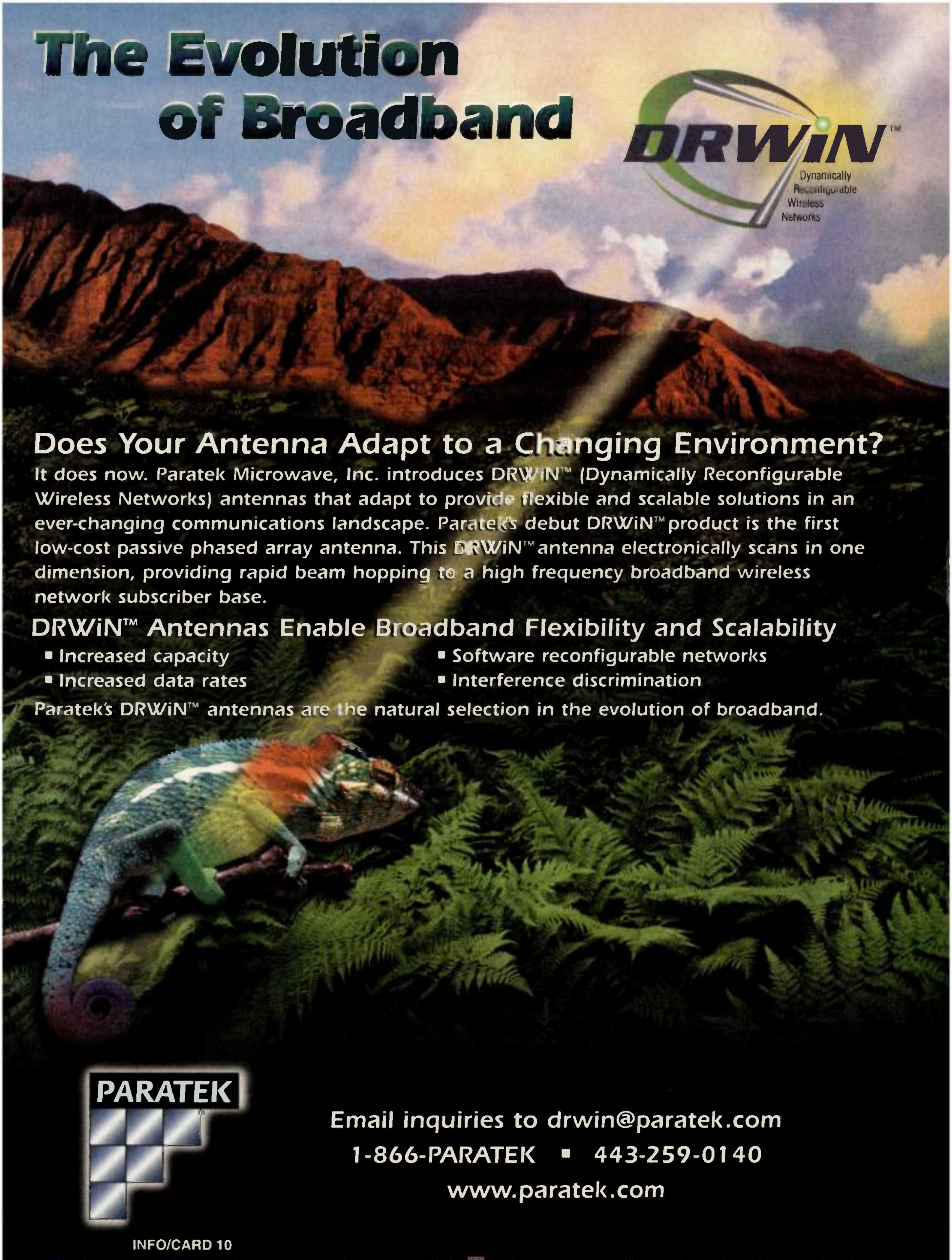
In an ideal polyphase filter, the target and image signal differ at the filter output by the unique phase difference of output voltages v_{o1} and v_{o2}, as is the case for the filter's input currents i_i. In the circuit of Figure 1, a target signal will result in a phasor v_{o1}, leading phasor v_{o2} according to Figure 2a. However, the image signal will cause v_{o2} leading v_{o1} according to Figure 3a. Subsequent polyphase filter stages or an image rejector stage will pass any target signal, but will suppress the image further according to their image attenuation.

Always assume that any component mismatch is evenly distributed between the two members of the mismatched pair, i.e. if the fractional mismatch of a pair is p (e.g. p = 1%), one of the components is off by $p/2$, the other by $-p/2$. For clarity, the mismatches shown in the following figures will be much larger than any occurring in a real integrated circuit.

The phasor diagram of a polyphase filter with mismatched feedback resistors R_f is shown in Figure 4. The output voltages v_{o1} and v_{o2} are in quadrature, however, their magnitudes differ by error components v_{e1} and v_{e2}. Current phasors i_R and i_C in Figure 4b depend again on voltages v_o according to the circuit of Figure 1 and are proportionally mismatched as well. However, the sums $i_{R1} + i_{C1}$ and $i_{R2} + i_{C2}$ and feedback currents i_f are not influenced by the mismatch. When combining error components v_{e1} and v_{e2} (in Figure 4c), it can be seen that v_{e1} leads v_{e2}, i.e. a configuration corresponding to a target signal.

The phasor diagram of a polyphase filter with mismatched cross-coupled resistors $R_1 > R_2$ is shown in Figure 5. The effect of the mismatch is that the phase difference between v_{o1} and v_{o2} is less than 90°. However, their magnitudes remain balanced (see Figure 5a). Current phasors i_R, i_C, and i_f in Figure 5b depend again on voltages v_o according to the circuit of Figure 1. As in Figure 3, the vector sums of i_R, i_C, and i_f are equal to the input currents i_i. Because $R_1 > R_2$, current i_{R1} is smaller than current i_{R2}.

The output phasors v_{o1} and v_{o2}, resulting from mismatch of resistors (R), can be decomposed into quadrature components v_{q1} and v_{q2} and respective error components v_{e1} and v_{e2}, as shown in Figure 5a. The quadrature components represent an image signal (vq2 leads vq1) further attenuated by subsequent image rejecting circuits, if any. However, when error components ve1 and ve2 are again separately joined in Figure 5c, it can be seen that ve1 is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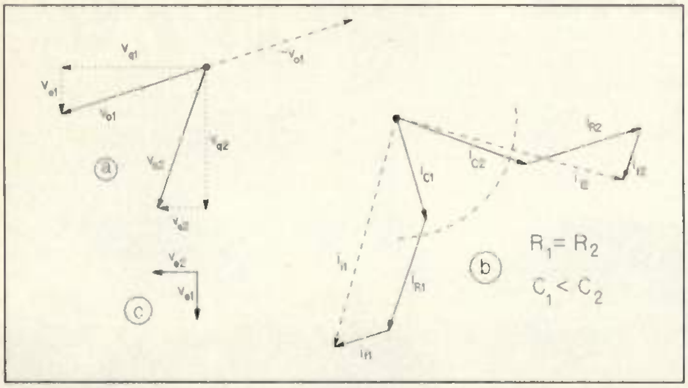


Figure 6. Mismatch of cross-coupling capacitors C_1 and C_2.

leading ve2, again creating a configuration corresponding to a target signal.

When the mismatch is between the integrating capacitors and $C_1 < C_2$, the phase difference between the filter's output voltages is also less than 90° (see Figure 6). When the resistors (R) or the capacitors (C) are mismatched in the opposite sense as described above, the phase difference between the filter's output voltages becomes more than 90° (see Figs. 7 and 8).

Any component mismatch in the polyphase filter results in some error components v_e. Thus, a part of the image signal appears at the filter output as an "image leak" that mimics a target signal. Understandably, any subsequent image-rejecting circuit will pass that leak signal with no attenuation because it cannot distinguish it from a genuine target signal. It is therefore important to develop a quantitative relation between component mismatch and image leak to avoid unpleasant surprises or, on the contrary, to avoid excessive component matching that costs chip area and power dissipation.

To find the relation between mismatch and leak, one must first assume that the filter is not mismatched and that it is driven by an image signal. Its outputs will be $v_{r1} = v_{r2}$, as in Figure 3. Next, mismatch a component pair, but keep the filter output fixed to v_{r1} and v_{r2} and the filter input voltage at zero. The mismatch will cause an error current of i_{e1} in one, and of i_{e2} in the other member of the pair. Finally, assume that a mismatch-free replica of the analyzed filter has error currents i_{e1} and i_{e2} as its input currents i_{i1} and i_{i2}. The magnitude of the replica filter's output will be a good approximation of error components v_{e1} and v_{e2} in the mismatched filters of Figs. 4 and 5.

To calculate error currents i_e when the mismatched pair is the feedback resistors, use the following formulas:

$R_{f1} = R_f(1 + p/2)$

and

$R_{f2} = R_f(1 - p/2)$.

With p << 1:

$$i_{e1} = v_{r1}\left(\frac{1}{R_{f1}} - \frac{1}{R_f}\right) \approx -\frac{v_{r1}p}{2R_f}$$

and

$$i_{e2} = v_{r2}\left(\frac{1}{R_{f2}} - \frac{1}{R_f}\right) \approx +\frac{v_{r2}p}{2R_f}$$

It can be seen that i_{e1} is of opposite phase to v_{r1}, while i_{e2} is in phase with v_{r2}. Because phasors v_{r1} and v_{r2} represent an image signal (see Figure 3), the constellation of currents i_{e1} and i_{e2} must be that of a target signal. It has been found that the transimpedance for a target signal is $Z_{0t} = R_f$. So, when driving the replica filter with i_{e1} and i_{e2}, its outputs will be $v_e = i_e R_f = v_r p/2$. When p << 1 and the mismatched filter's input is an image signal, its feedback current i_f is in close quadrature with input current i_i, similar to Figure 3. Input current i_i is then close to $i_i = i_R + i_C$ and is almost equally divided between i_R and i_C.

Thus, with good approximation, one can write:

$v_r = i_i R/2$.

And, furthermore, for a mismatch p in R_f:

$v_{ef} = i_i R p/4$.

When the same procedure is applied with the mismatched pair using the cross-coupled resistors, the result is:

$v_{eR} = i_i R f p/4$.

Finally, the same applies when the mismatched pair is the integrating capacitors C_1 and C_2.

To assess the significance of the image leak v_e, the filter's transimpedance Z_{0t} must be compared for a target signal with its transimpedance Z_{0p} for an image signal with a mismatch in R_f, R or C, respectively. It is known that $Z_{0t} = R_f$. From this:

$Z_{pf} = v_{ef}/i_i = Rp/4$ for a mismatch of feedback resistors R_f, and:

$Z_{pR} = v_{eR}/i_i = Rfp/4$ for a mismatch of cross-coupled resistors (R) or capacitors (C).

One important ratio is $Z_{ipf}/Z_{ipR} = R/R_f = 1/2Q$. It means that the matching of

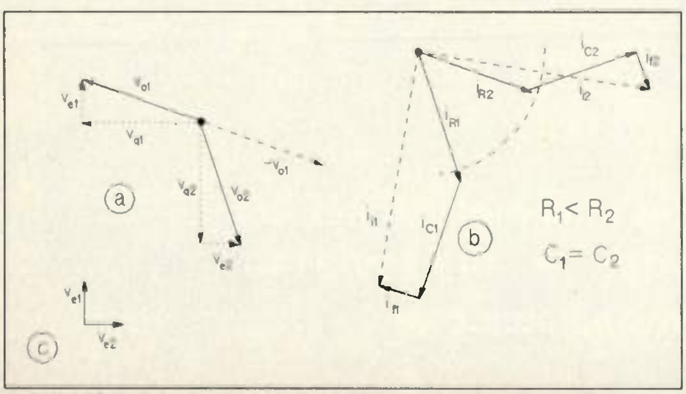


Figure 7. Mismatch of cross-coupling resistors opposite to Figure. 5.

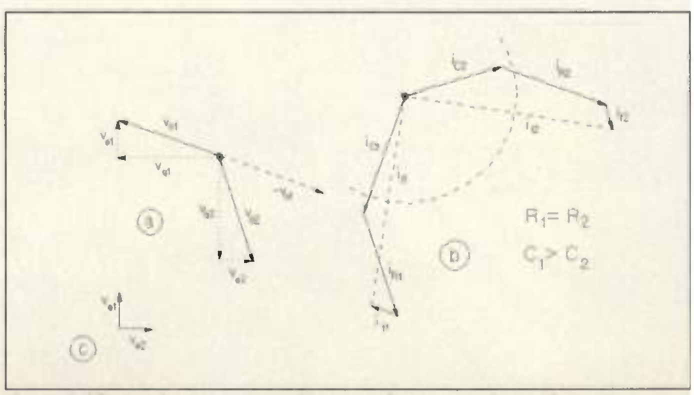
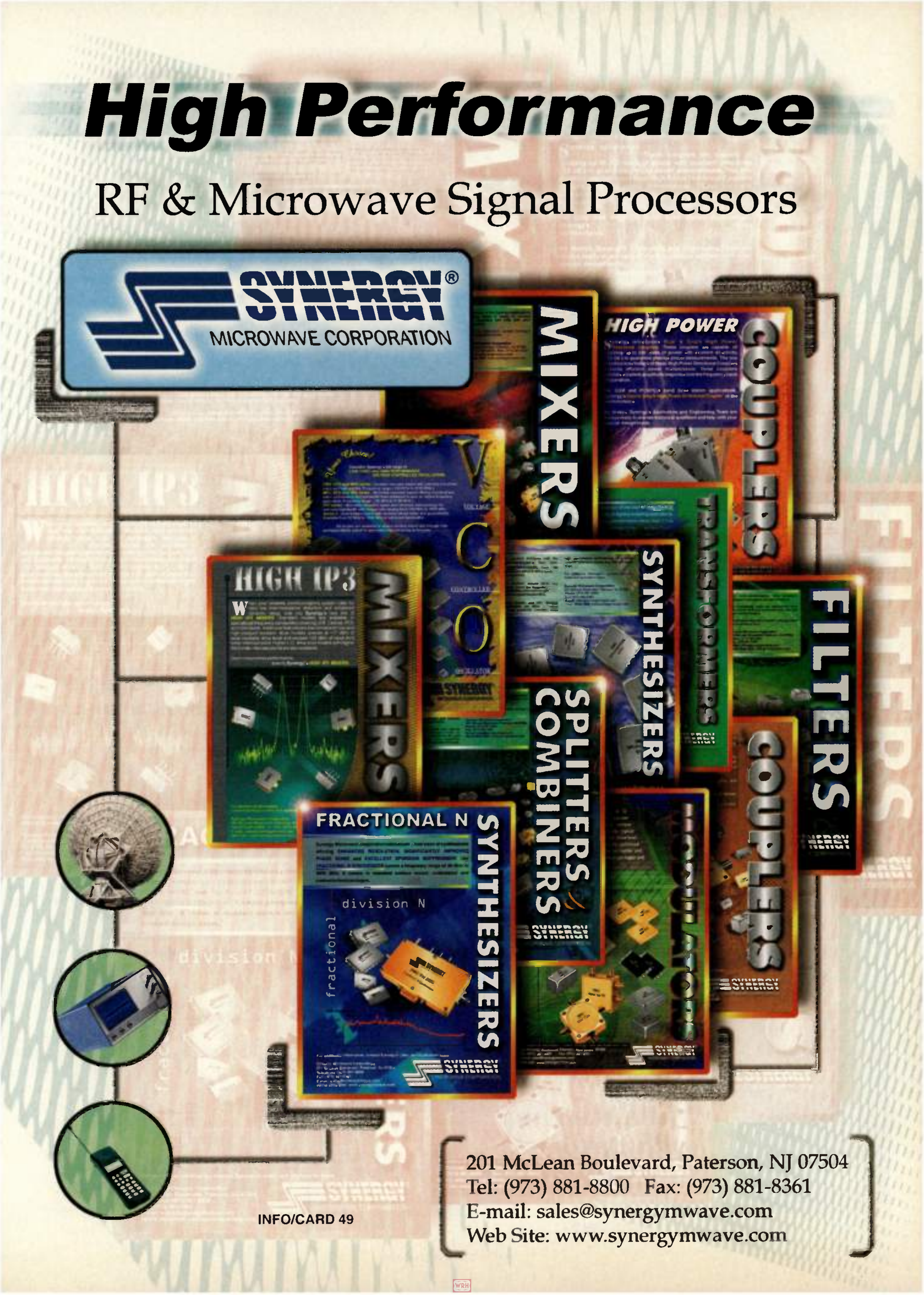


Figure 8. Mismatch of cross-coupling capacitors opposite to Figure 6.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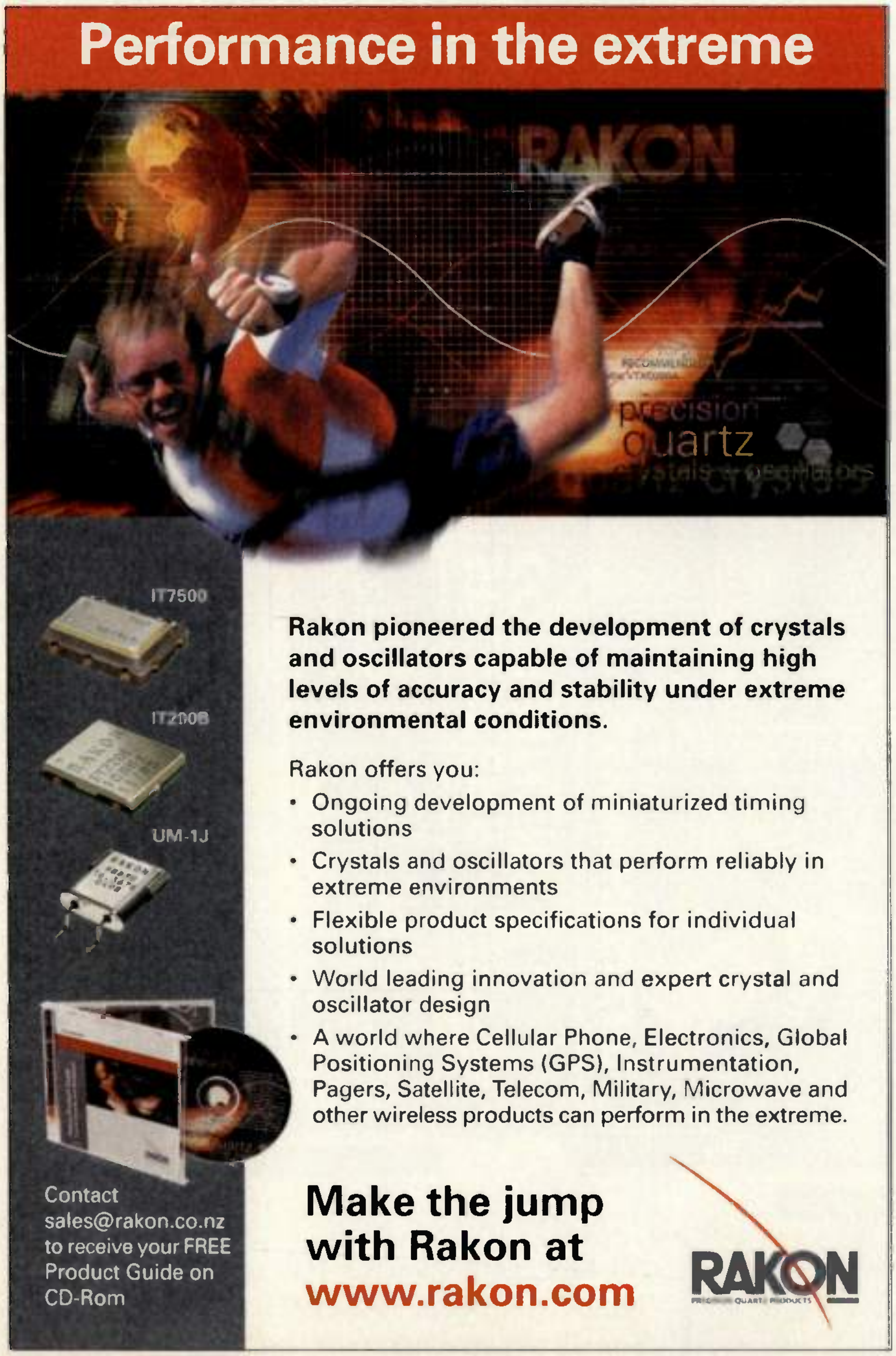
R_f can be relaxed 2Q-times over the matching of R or C to cause the same image leak. Another important ratio is $Z_{pR}/Z_{0t} = p/4$. This says that if, for example, the ratio of image signal to target signal is 1000:1 (60 dB), to keep the image leak smaller than the target signal, the mismatch p of (R) or (C) can be as high as 0.4%. This is one advantage of the polyphase filter over two separate IF filters, where, in the same case, a mismatch of no worse than 0.1% is required.

The assessment

The operation of an active polyphase filter when used for image attenuation in low IF data receivers has been clearly visualized by a phasor analysis of the filter's voltages and currents. The influence of the filter's component mismatch on its image suppression performance has been quantitatively analyzed. It has been shown that the image-attenuating performance of a polyphase filter is superior to two separate IF filters.

RF



References:

[1] Jan Crols, Michiel Steyaert "*A Single Chip 900 MHz CMOS Receiver Front-End with a High Performance Low-IF Topology*," IEEE Journal of Solid State Circuits, Vol. 30, No.12, December 1995, pp. 1483 - 1492.

[2] Johannes O. Voorman "*Asymmetric Polyphase Filter*," US Patent No. 4,914,408.

[3] Jan Crols, Michiel Steyaert "*An Analog Integrated Polyphase Filter for a High Performance Low-IF Receiver*," Proceedings of the VLSI Circuit Symposium Kyoto, June 1995, pp. 87-88.

About the author

Tom Hornak received his M.S.E.E. degree from the Bratislava Slovak Technical University and a Ph.D. degree in electrical engineering from the Czech Technical University, Prague. From 1947 to 1968, he worked at the Tesla Corporation's Radio Research Laboratory. He then worked at the Computer Research Institute in Prague. He emigrated from Czechoslovakia in 1968 and joined Hewlett-Packard's Corporate Research Laboratories (HP Labs) in Palo Alto, California. In his years in HP Labs, his interests were high-speed analog-digital converters, high-speed fiber-optic communication systems, electronic instruments and ICs for wireless communication. He retired form HP Labs in 1999. Hornak was lecturer at the Czech Technical University in Prague and has published more than 50 papers. He holds more than 60 U.S. and foreign patents. He became an IEEE Fellow in 1985. Hornak can be reached by e-mail at: *Thornak68@aol.com*

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The Design Engineers Search Engine

USER'S GUIDE

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TM

The Design Engineers Search Engine

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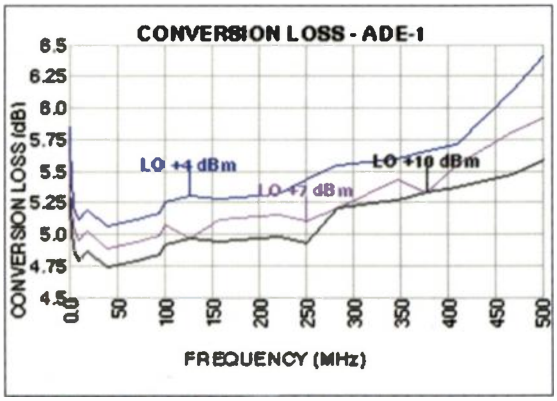

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Contents

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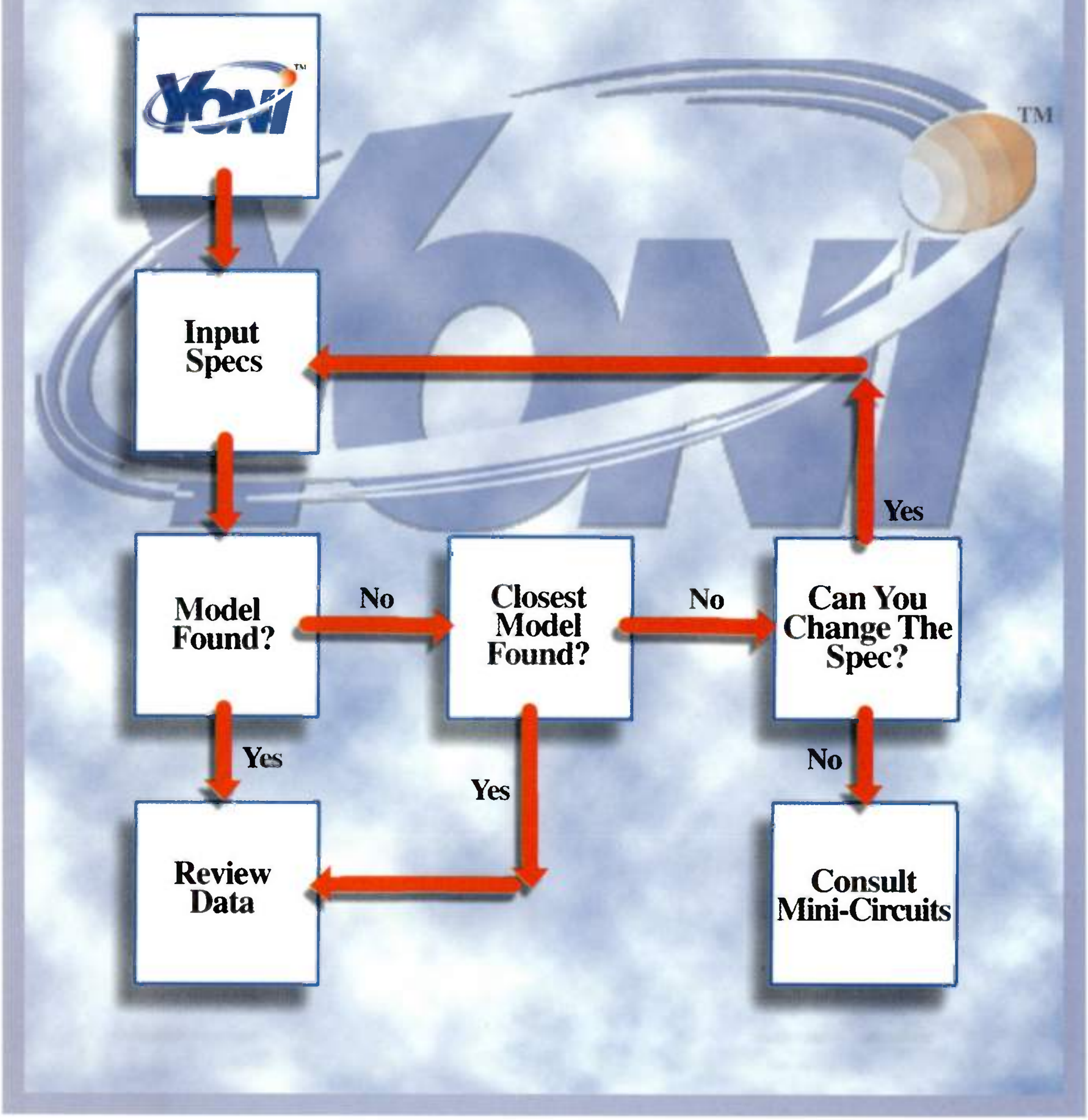
Flow Diagram Of Yoni™ Operation.



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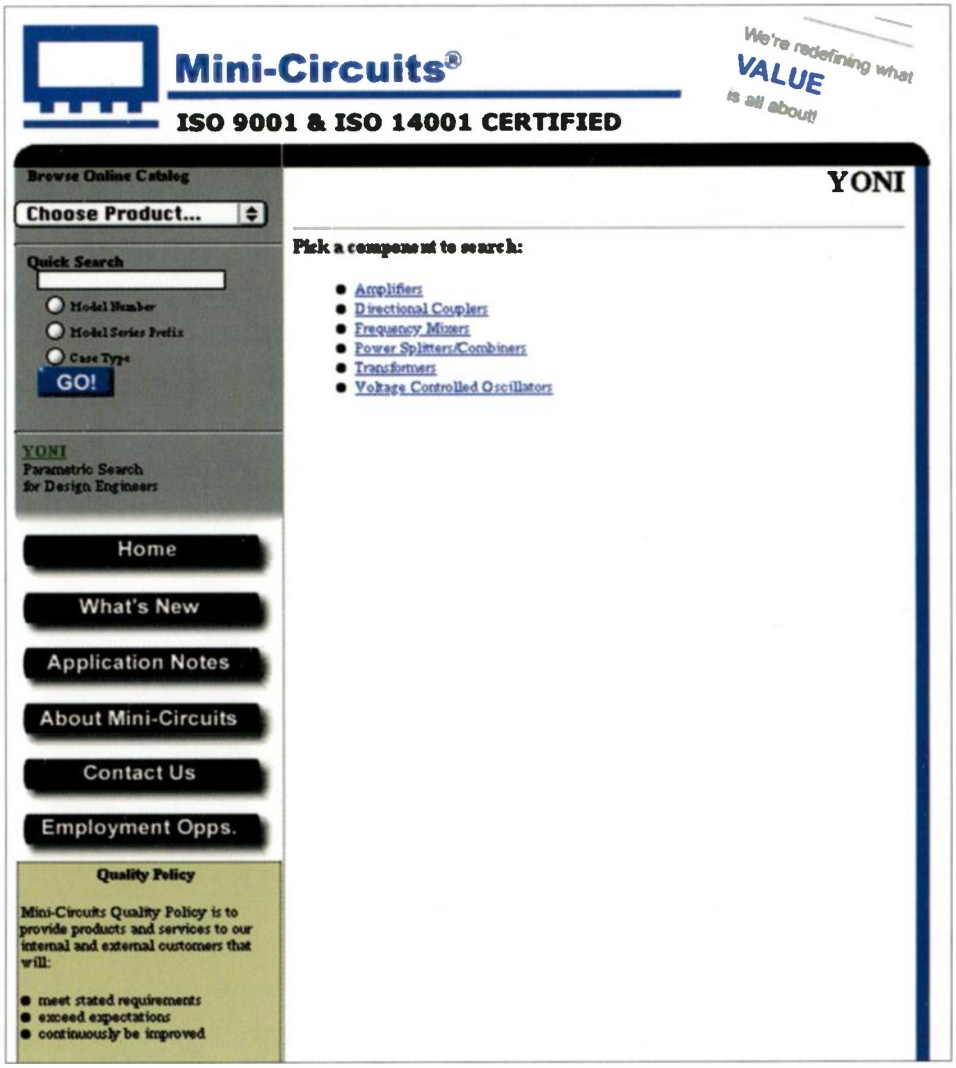
New products, performance data, specification revisions, and new Case Style Drawings and Outline Dimensions are added to YONI on a daily basis as they become available. This provides a jump start to use new product developments and assures a timely search.



Pick a product category to search in YONI.

YONI Component Categories

Upon entering the YONI Search Engine, you will be asked to "Pick a component to search" from the available categories. Select a component category with a single click.

YONI searches are available for categories listed on the screen. Additional search categories will be added in the future.

For the purposes of this User's Guide, we will use the category of "Frequency Mixers" for demonstration.

TIP:

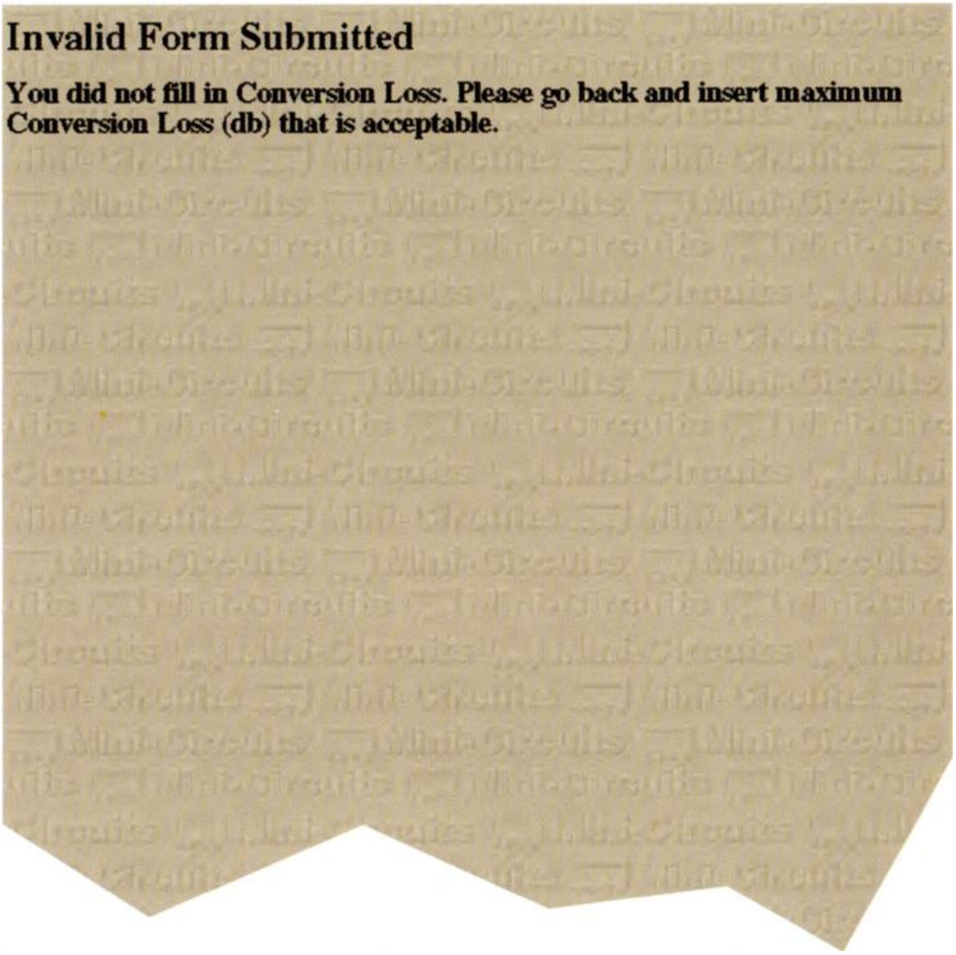
If your Component Type does not appear on the YONI search menu, browse our Online Catalog, and also check the "What's New" feature.

Completion of all minimum data fields (red) are required to proceed with a search.

YONI Data Entry Fields

Minimum Required Data Entry Fields

After picking a component family from the choices given, a "Data Entry Information Page" will be displayed. The completion of all the Minimum Data Requirement Fields on this page are mandatory for YONI to proceed with a successful search. Mandatory data fields are highlighted in red.



Invalid Form Submitted

You did not fill in Conversion Loss. Please go back and insert maximum Conversion Loss (db) that is acceptable.

If all required fields are not completed, YONI will issue a friendly reminder.

YONI Will Remind You To Complete ALL Required Fields

If a search is attempted with any REQUIRED field left blank, YONI will display a page titled "Invalid Form Submitted". At this point, simply click the "Back" button in your browser to return to the "Data Entry Information Page" and enter the additional REQUIRED specification(s).

Typically, more optional fields completed = tighter component match. Fewer optional fields completed = broader component match.

Optional Data Entry Fields

On the same Data Entry Information Page, Optional Data Entry Fields are headed in black. Each "Optional Data Entry Field" that is completed narrows YONI's search to display a subgroup of all components that meet or exceed the entered specs. Hence, to obtain the tightest component match to your specific performance requirements, complete as many Optional Data Entry Fields as possible.

After all the REQUIRED fields and as many OPTIONAL fields as desired are completed, click Search located at the bottom of the page to launch the YONI search engine, or click Clear to clear all entered data in Data Entry Fields, and start over.

Frequency Mixers
LO Power Level 7 dBm

Click on the reference number to see Performance Data and Graph.

Reference No.	Frequency MHz		Conversion Loss dB		LO-RF Isolation dB			LO-IF Isolation dB			Case Style	Conn-ection	Notes
	LO/RF	IF	Mid-Band	Total Range	L	M	U	L	M	U			
LRMS-1J	0.500-500	DC-500	7.0	8.5	50	25	20	45	23	19	QQQ569	w	
RMS-1	0.500-500	DC-500	7.0	8.5	50	25	20	45	23	19	TT240	w	
ADE-1	0.500-500	DC-500	6.5	7.8	50	35	30	45	25	20	CD636	ht	*,I,4
ADE-1LH	0.500-500	DC-500	6.5	8.2	50	35	26	40	22	20	CD636	ht	
ADE-ED8019/1	0.100-1112	5-1050									CD542	ht	U
ADE-ED7941/1	0.100-2000	10-1000									CD542	hu	U
ADE-ED7879/1	0.400-953	0-520									CD542	ht	U
ADE-ED7693/1	0.400-900	0-520									CD636	ht	U
LRMS-ED6552/1	0.500-965	5-530									QQQ569	w	U
ADE-ED6529/1	0.000-1540	100-1145									CD542	hu	U

L=low range(f_L to $10f_L$) M=mid range($10f_L$ to $f_U/2$) U=upper range($f_U/2$ to f_U)

Pin Connections - see case style outline drawing

Port	LO	RF	IF	Gnd Ext.	Case Gnd	Not Used
w	1	4	5	2,3,6	-	-
ht	6	3	2	1,4,5	-	-
hu	6	2	3	1,4,5	-	-

Notes:

1 - +7dBm LO, up to +1 dBm RF

Absolute maximum power, voltage and current ratings:

4 - a. RF power, 50mW

b. Peak IF current, 40mA

U - Non-catalog model. Please consult factory for price and delivery.

* - General Quality Control Procedures and Environmental Specifications are given in Mini-Circuits Guarantees Qual
Hi-Rel, MIL description are given in Hi-Rel and MIL.

* - Prices and Specifications subjects to change without notice.

Back

YONI will display a list of models (Reference No.) that fall within your requirements, complete with case style number.

YONI Search Results

Display Of Models With Specifications And Case Styles

After you click "Search", YONI will instantly display a list of Mini-Circuits standard catalog models and/or Engineering Design (ED) references from our internal development database. Typically, the *TOTAL* frequency range for these units will be shown along with guaranteed performance specifications. *YONI selected these units because they meet or exceed the performance criteria you entered within the frequency range you specified.*

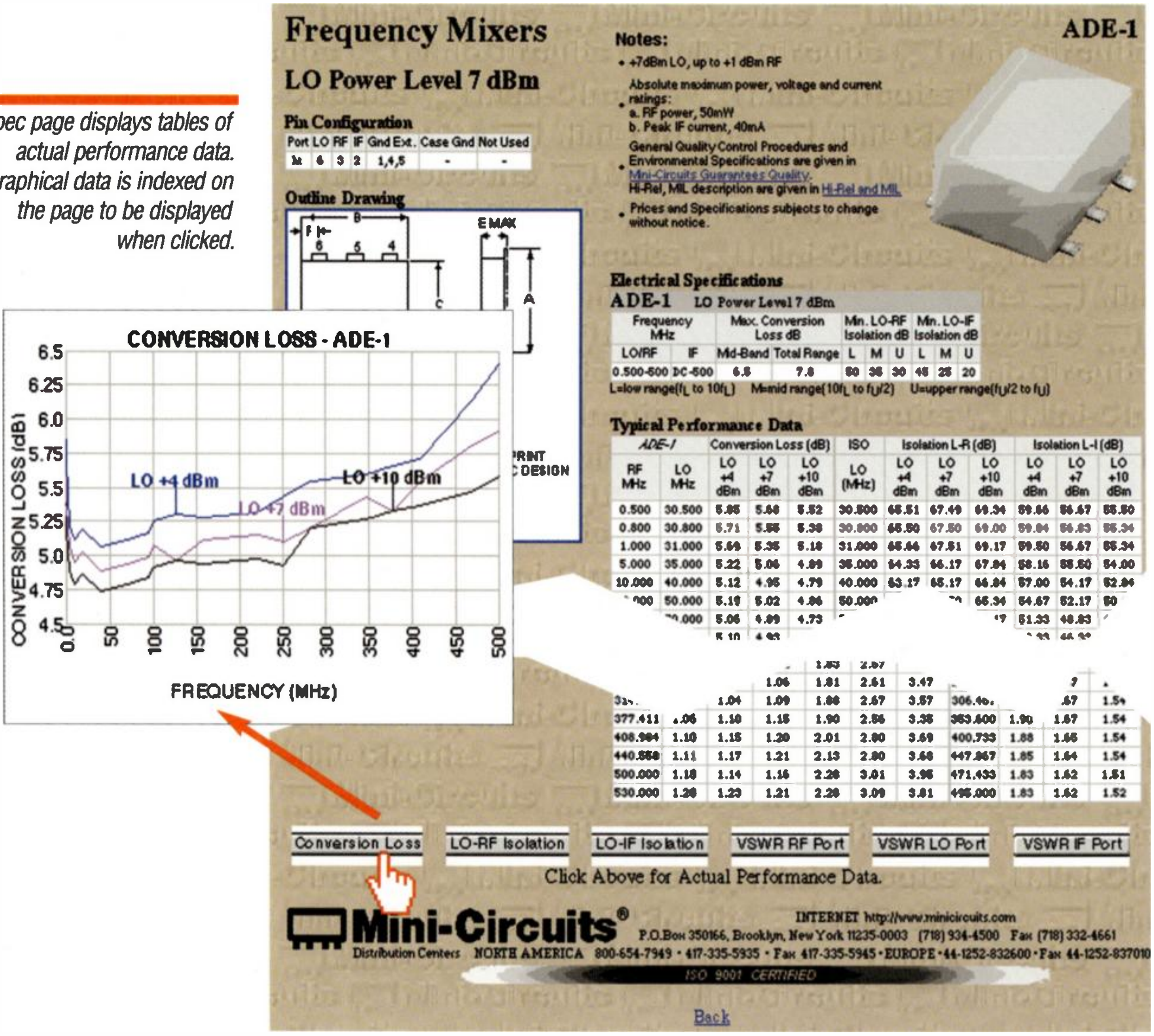
How To Find Models With The Lowest Price

After preliminary evaluation of all models shown, check the price for each model you selected by using our "RF/IF Designers Guide" (see page 23), or the "Browse Online Catalog" feature on the Mini-Circuits home page (see page 22). This feature accesses a "Catalog Model Index" from the "Choose Product..." pull-down menu. Locate your component category in the index, then click on a Model Number to view prices. By using YONI in conjunction with our catalog, you can get relative pricing for the models you selected. For high quantity or "ED" model pricing, contact factory or complete a "Price Quotes" form available by following the "Contact Us" link on the Mini-Circuits home page.

TIP:

To aid your evaluation, click the Case Style number corresponding to the Reference No. to display complete case style information.

Spec page displays tables of actual performance data. Graphical data is indexed on the page to be displayed when clicked.



Viewing Actual Performance Data And Case Styles

Viewing Tables And Graphs

Start with the lowest priced part that meets your requirements and click on the reference number to display a specification page. The resulting page shows *actual performance data* in tabular form and indexes actual graphical performance data available for up to 7 important electrical specifications.
Click on the bar for the performance specification you wish to view graphically and a Java Applet Window will appear with corresponding actual performance curves.
To view another performance curve, simply click another performance specification bar.

Frequency Mixers

LO Power Level 7 dBm

ADE-1

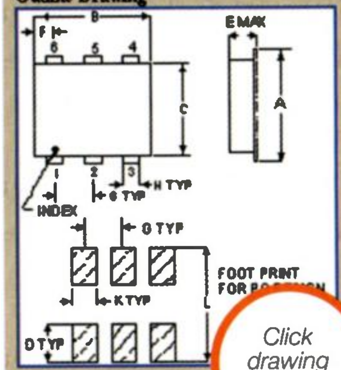
Pin Configuration

Port	LO	RF	IF	Gnd Ext.	Case Gnd	Not Used
M	6	3	2	1,4,5	-	-

Notes:

- +7dBm LO, up to +1 dBm RF
- Absolute maximum power, voltage and current ratings:
  a. RF power, 50mW
  b. Peak IF current, 40mA
- General Quality Control Procedures and Environmental Specifications are given in Mini-Circuits Guarantees Quality. Hi-Rel, MIL description are given in Hi-Rel and MIL.
- Prices and Specifications subjects to change without notice.

Outline Drawing



Click drawing to enlarge

Electrical Specifications

ADE-1 LO Power Level 7 dBm

Frequency MHz		Max. Conversion Loss dB		Min. LO-RF Isolation dB			Min. LO-IF Isolation dB		
LO/RF	IF	Mid-Band	Total Range	L	M	U	L	M	U
0.500-500	DC-500	6.5	7.8	50	35	30	45	25	20

L=low range(f_L to $10f_L$) M=mid range($10f_L$ to $f_U/2$) U=upper range($f_U/2$ to f_U)

Typical Performance Data

ADE-1		Conversion Loss (dB)			ISO	Isolation L-R (dB)			Isolation L-I (dB)		
RF MHz	LO MHz	LO +4 dBm	LO +7 dBm	LO +10 dBm	LO (MHz)	LO +4 dBm	LO +7 dBm	LO +10 dBm	LO +4 dBm	LO +7 dBm	LO +10 dBm
0.500	30.500	5.86	5.60	5.52	30.500	65.51	67.49	69.34	59.66	56.67	55.50
0.800	30.800	5.71	5.55	5.38	30.800	66.50	67.80	69.00	59.84	56.83	55.34
1.000	31.000	5.69	5.35	5.18	31.000	65.66	67.51	69.17	59.50	56.67	55.34
5.000	35.000	5.22	5.06	4.89							
10.0[illegible]	[illegible]40.000	5.12	4.95	4.79							
20.000	50.0[illegible]	5.19	5.02	4.86							
40.000	70.000	5.06	[illegible]	4.73							
63.247	93.247	5.10	4.93	[illegible]							
94.821	124.821	5.16	4.99	4.83							
100.000	130.000	5.25	5.08	4.91							
126.395	156.395	5.30	4.96	4.96							
157.968	187.968	5.20	5.11	4.94							
221.116	251.116	5.32	5.15	4.98							
250.000	280.000	5.43	5.10	4.93							
284.263	314.263	5.54	5.20	5.20							
347.411	377.411	5.60	5.43	5.27							
378.984	408.984	5.66	5.33	5.33							
410.558	440.558	5.71	5.54	5.37							
470.000	500.000	6.14	5.81	5.47	500.000	41.16	44.67	44.50	33.50	28.16	25.83
500.000	530.000	6.41	5.91	5.58	530.000	41.00	43.82	43.00	33.67	28.50	25.80

RF/LO	YSWR RF port			YSWR LO port			IF	YSWR IF port		
FREQ. (MHz)	LO +4 dBm	LO +7 dBm	LO +10 dBm	LO +4 dBm	LO +7 dBm	LO +10 dBm	FREQ. (MHz)	LO +4 dBm	LO +7 dBm	LO +10 dBm
30.500	1.50	1.46	1.45	1.69	2.51	3.46	0.100	2.17	1.74	1.49
30.800	1.38	1.32	1.30	1.67	2.56	3.57	0.300	2.13	1.74	1.49
31.000	1.34	1.29	1.26	1.71	2.56	3.57	0.500	2.20	1.78	1.52
35.000	1.29	1.19	1.12	1.67	2.51	3.46	0.800	2.20	1.85	1.55

Case Style - CD636 (inch,mm)
grams.

A	B	C	D	E	F	G	H	J
.280 7.112	.310 7.874	.220 5.588	.100 2.540	.155 3.937	.055 1.397	.100 2.540	.030 0.762	
K	**L**	**M**	**N**	**P**	**Q**	**R**	**S**	**T**
.065 1.651	.300 7.620							

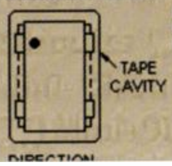
Tolerance: .x± .1 .xx± .03 .xxx± .015 inch.

Material and Finish:

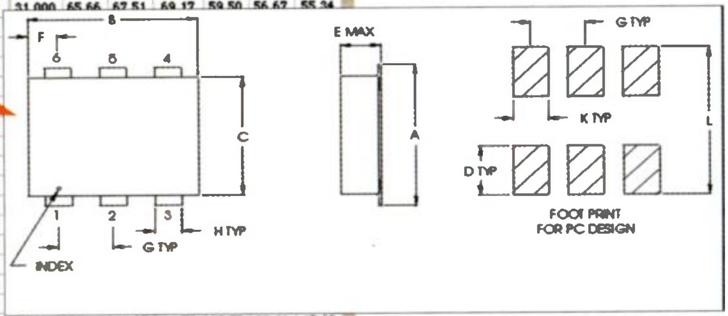
Case material: plastic. Lead finish: solder plate.

Packaging:

UNIT ORIENTATION



Packaging information:
Tape Width(mm): 16
Reel Size(inches): 13
Device Cavity Pitch(mm): 12
Devices Per Reel: 1000



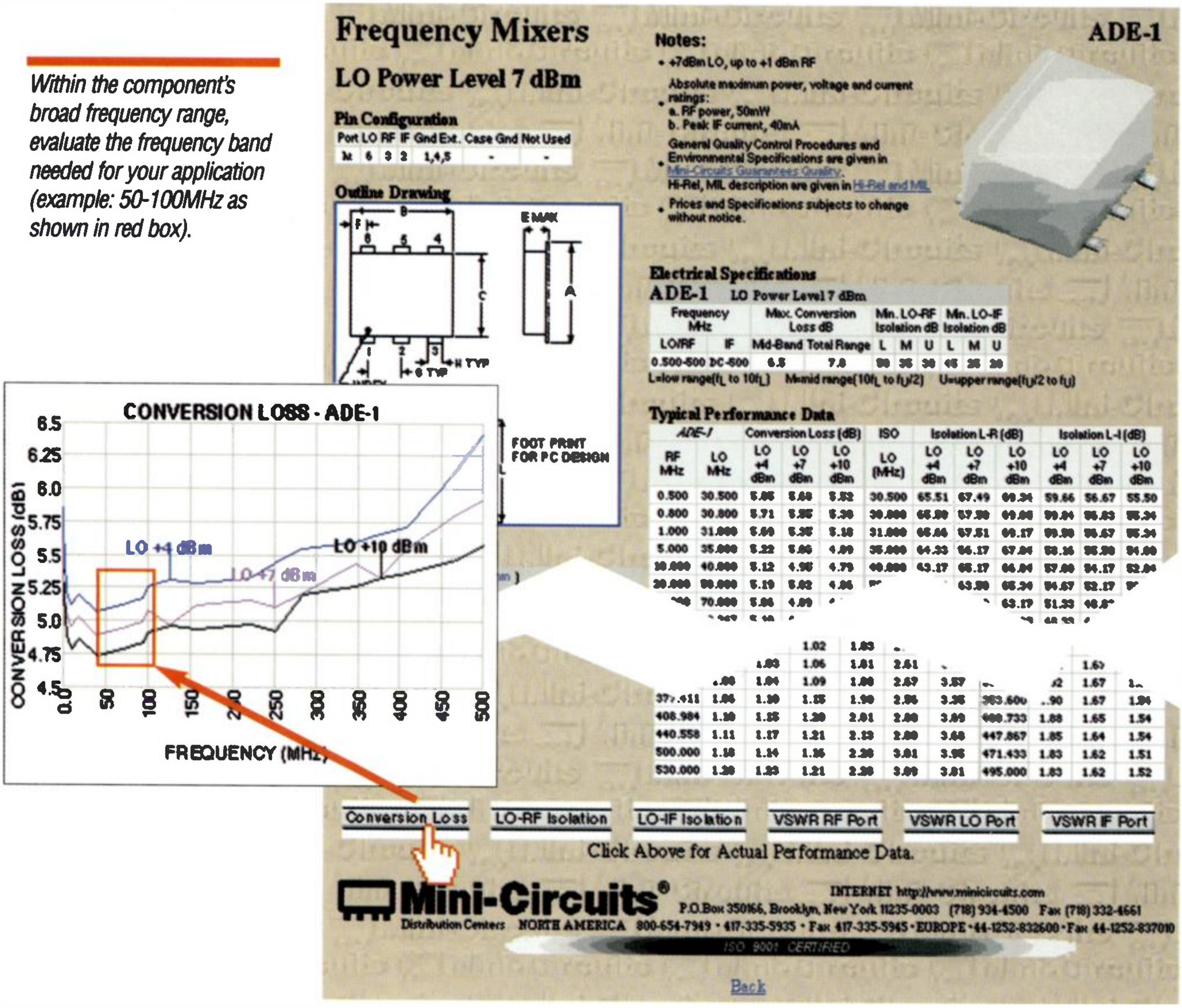
YONI provides complete Case Style information.

Viewing Case Style Information

YONI displays complete case style information including:

- *Drawings*
- *Outline Dimensions, Tolerances*
- *Weight*
- *Plus Material & Finish, and Mounting information.*

Within the component's broad frequency range, evaluate the frequency band needed for your application (example: 50-100MHz as shown in red box).



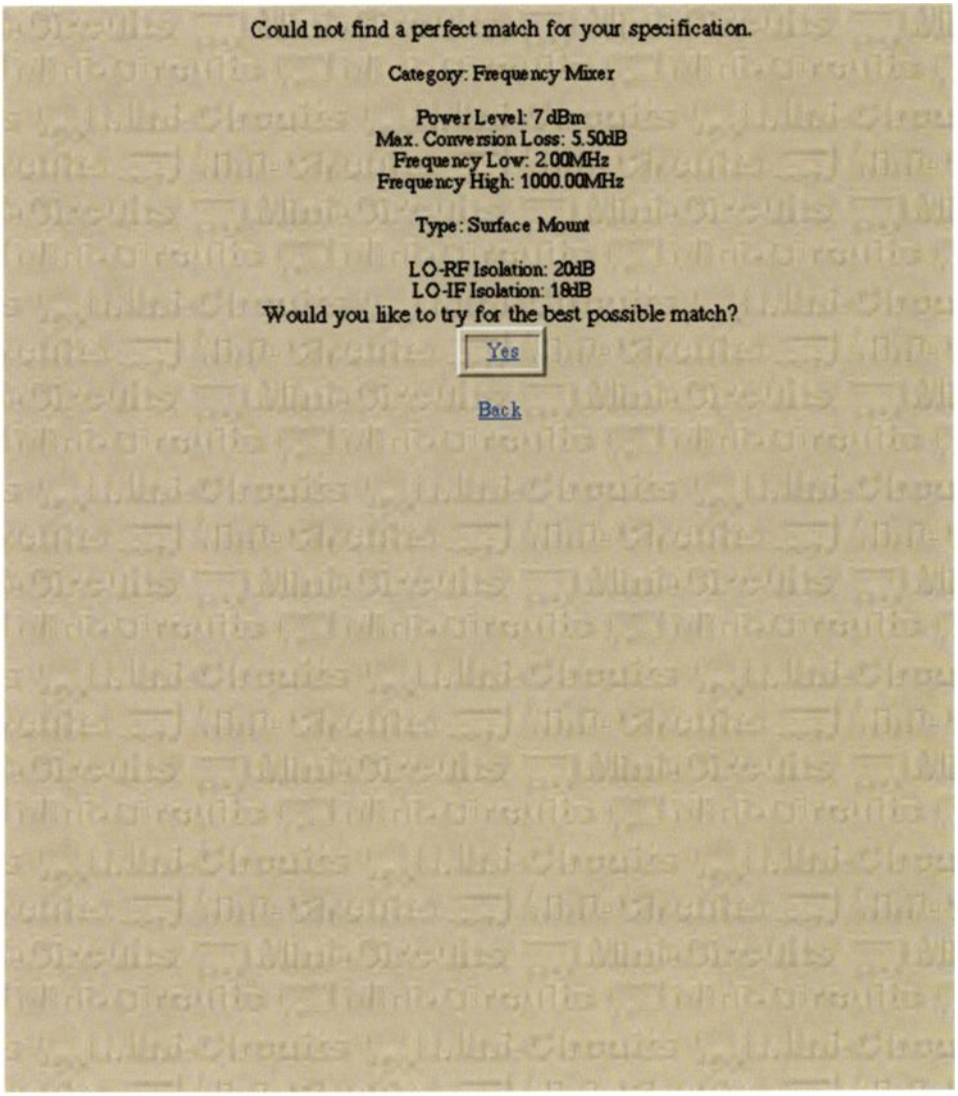
How To Evaluate, Select, And Order Your Parts

How To Evaluate And Select Your Parts

Within the Tables and Curves, locate the frequency range you specified when you initiated the search and evaluate the corresponding electrical specifications within that frequency band. Evaluate the package style as well.
Repeat this procedure with each model you selected to obtain the optimum price/performance solution for your application.

How To Order Your Parts

To order your parts on the internet, contact Mini-Circuits sales by clicking the "Contact Us" bar at the Home Page and following the "Sales" link, or email: sales@minicircuits.com .
Be sure to advise that you selected your model using YONI so we will be confident you selected the optimum model for your application.



If your criteria are not matched, YONI asks, "Would you like to try for the best possible match?".

What Happens If YONI "Draws A Blank"?

SCREEN #1: ***Obtaining A Best Possible Match***
If YONI could not find an exact match for your specification criteria, best possible component matches can be found. In this situation, YONI will display a page listing your stated criteria and will ask: "Would you like to try for the best possible match?".
To view a list of components displaying best possible matches, Click Yes .

Frequency Mixers
LO Power Level 7 dBm

Click on the reference number to see Performance Data and Graph.

Reference No.	Matched Criteria		Frequency MHz		Conversion Loss dB		LO-RF Isolation dB			LO-IF Isolation dB			Case Style	Conn-ection	Notes
	CL,LR,LI	VSWR RF,LO,IF	LO/RF	IF	Mid-Band	Total Range	L	M	U	L	M	U			
SYM-2	N,Y,Y	, ,	2.00-1000	DC-1000	7.2	9.5	45	30	25	40	24	20	TTT166	x	*,1,4
SY[illegible]11	N,Y,Y	, ,	1.000-2500	10-600	9.0	10.5	40	24	20	40	20	15	TTT167	x	*,1,4
SY[illegible]	N,Y,Y	, ,	1.000-2500	DC-500	8.5	9.8	50	25	20	45	10	8	TTT167	x	*,1,4
TUF-5XSM	N,Y,Y	, ,	1.000-1500	1-1000	7.0	9.0	40	20	17	45	25	20	NNN150	gm	*,1,4
ADE-2ASK	N,Y,Y	, ,	1.000-1000	DC-1000	6.8	9.5	45	30	20	40	22	12	CD542	hr	*,1,4
SYM-11LH	N,Y,Y	, ,	1.000-2000	10-600	8.3	9.8	40	25	25	40	20	20	TTT167	x	
SYM-11J	N,Y,Y	, ,	1.000-2500	10-500	8.0	9.8	40	24	20	40	20	15	CG581	ka	*,1,4
ADE-ED8019/1	N,Y,Y	, ,	0.100-1112	5-1050									CD542	hr	U
ADE-ED7941/1	N,Y,Y	, ,	0.100-2000	10-1000									CD542	hu	U
ADE-ED7907/1	N,Y,Y	, ,	2.00-1577	0-1490									CD542	hr	U
ADE-ED7878/1	N,Y,Y	, ,	0.500-1050	0-1000									CD542	hr	U
LRMS-ED6326/2	N,Y,Y	, ,	1.000-1970	3-900									QQQ569	w	U
ADE-ED6529/1	N,Y,Y	, ,	0.000-1540	100-1145									CD542	hu	U

L=low range(f_L to $10f_L$) M=mid range($10f_L$ to $f_U/2$) U=upper range($f_U/2$ to f_U)

Pin Connections - see case style outline drawing

Port	LO	RF	IF	Gnd Ext.	Case Gnd	Not Used
w	1	4	5	2,3,6	-	-
x	2	1	3	4,5,6	-	-
gm	4	2	1	3	3	-
hr	6	3	2	1,4,5	-	-
hu	6	2	3	1,4,5	-	-
ka	11	5	2	1,3,4,6,7,8,9,10,12	-	-

Notes:
· +7dBm LO, up to +1 dBm RF
...olute maximum power, vo...
...wer, 50mW

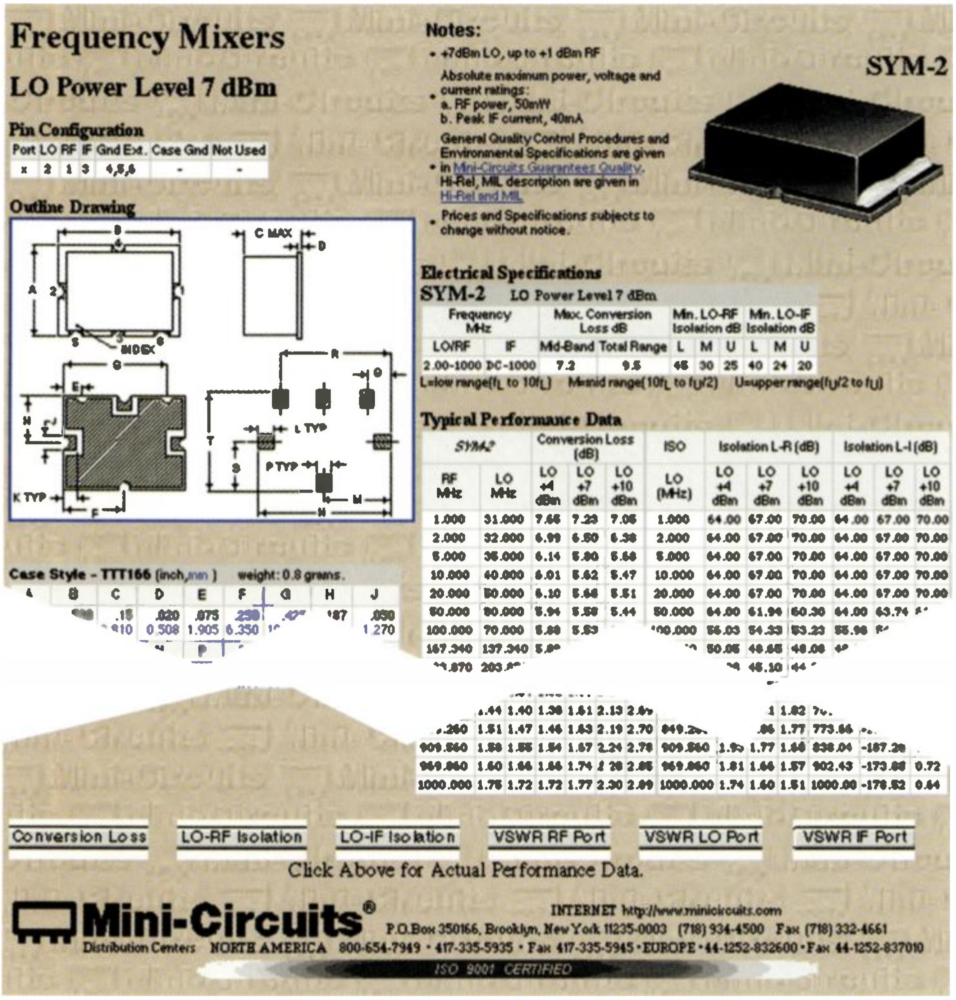
If YONI could not match your performance criteria, you can select from alternate models.

YONI Suggests Alternate Models

After clicking [1m], YONI will display a page listing Models and Engineering Design Reference Numbers with specifications and case styles of components that closely match your entered criteria. A column headed "Matched Criteria" will be displayed. This column is keyed to your specific search criteria and will indicate which of your specifications were met (Y) and which were not (N).

Y=Search Criteria Meets Spec.
N=Search Criteria Does Not Meet Spec.
Blank=Criteria was not defined.

By viewing the specifications of suggested alternate components, and knowing at a glance which specifications meet your criteria and which do not (Y or N), you can determine if alternate units are acceptable for your consideration.



Frequency Mixers

LO Power Level 7 dBm

SYM-2

Pin Configuration

Port	LO	RF	IF	Gnd Ext.	Case Gnd	Not Used
x	2	1	3	4,5,6	-	-

Outline Drawing

Case Style - TTT166 (inch,mm) weight: 0.8 grams.

A	B	C	D	E	F	G	H	J
[illegible]	[illegible]	.15	.020	.075	.250	.42	.187	.050
		3.810	0.508	1.905	6.350	[illegible]		1.270

Notes:

- +7dBm LO, up to +1 dBm RF
- Absolute maximum power, voltage and current ratings:
  a. RF power, 50mW
  b. Peak IF current, 40mA
- General Quality Control Procedures and Environmental Specifications are given in Mini-Circuits Guarantees Quality. Hi-Rel, MIL description are given in Hi-Rel and MIL.
- Prices and Specifications subjects to change without notice.

Electrical Specifications

SYM-2 LO Power Level 7 dBm

Frequency MHz		Max. Conversion Loss dB		Min. LO-RF Isolation dB			Min. LO-IF Isolation dB		
LO/RF	IF	Mid-Band	Total Range	L	M	U	L	M	U
2.00-1000	DC-1000	7.2	9.5	45	30	25	40	24	20

L=low range(f_L to $10f_L$) M=mid range($10f_L$ to $f_U/2$) U=upper range($f_U/2$ to f_U)

Typical Performance Data

SYM-2		Conversion Loss (dB)			ISO	Isolation L-R (dB)			Isolation L-I (dB)		
RF MHz	LO MHz	LO +4 dBm	LO +7 dBm	LO +10 dBm	LO (MHz)	LO +4 dBm	LO +7 dBm	LO +10 dBm	LO +4 dBm	LO +7 dBm	LO +10 dBm
1.000	31.000	7.65	7.23	7.05	1.000	64.00	67.00	70.00	64.00	67.00	70.00
2.000	32.000	6.99	6.50	6.38	2.000	64.00	67.00	70.00	64.00	67.00	70.00
5.000	35.000	6.14	5.80	5.68	5.000	64.00	67.00	70.00	64.00	67.00	70.00
10.000	40.000	6.01	5.62	5.47	10.000	64.00	67.00	70.00	64.00	67.00	70.00
20.000	50.000	6.10	5.66	5.51	20.000	64.00	67.00	70.00	64.00	67.00	70.00
50.000	80.000	5.94	5.58	5.44	50.000	64.00	61.94	60.30	64.00	63.74	[illegible]
100.000	70.000	5.88	5.53	[illegible]	[illegible]	56.03	54.33	53.23	55.98	[illegible]	[illegible]
167.340	137.340	[illegible]	[illegible]	[illegible]	[illegible]	50.05	48.65	48.08	[illegible]	[illegible]	[illegible]
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	46.10	[illegible]	[illegible]	[illegible]	[illegible]

[illegible]	[illegible]	1.40	1.38	1.61	2.13	[illegible]	[illegible]	[illegible]	[illegible]	1.82	[illegible]	[illegible]	[illegible]
[illegible]	1.51	1.47	1.46	1.63	2.19	2.70	[illegible]	[illegible]	[illegible]	1.77	773.66	[illegible]	[illegible]
909.560	1.58	1.55	1.54	1.67	2.24	2.78	909.560	[illegible]	1.77	1.66	838.04	-167.28	[illegible]
969.560	1.60	1.66	1.66	1.74	[illegible]	2.85	969.560	1.81	1.66	1.57	902.43	-173.88	0.72
1000.000	1.75	1.72	1.72	1.77	2.30	2.89	1000.000	1.74	1.60	1.51	1000.00	-178.52	0.64

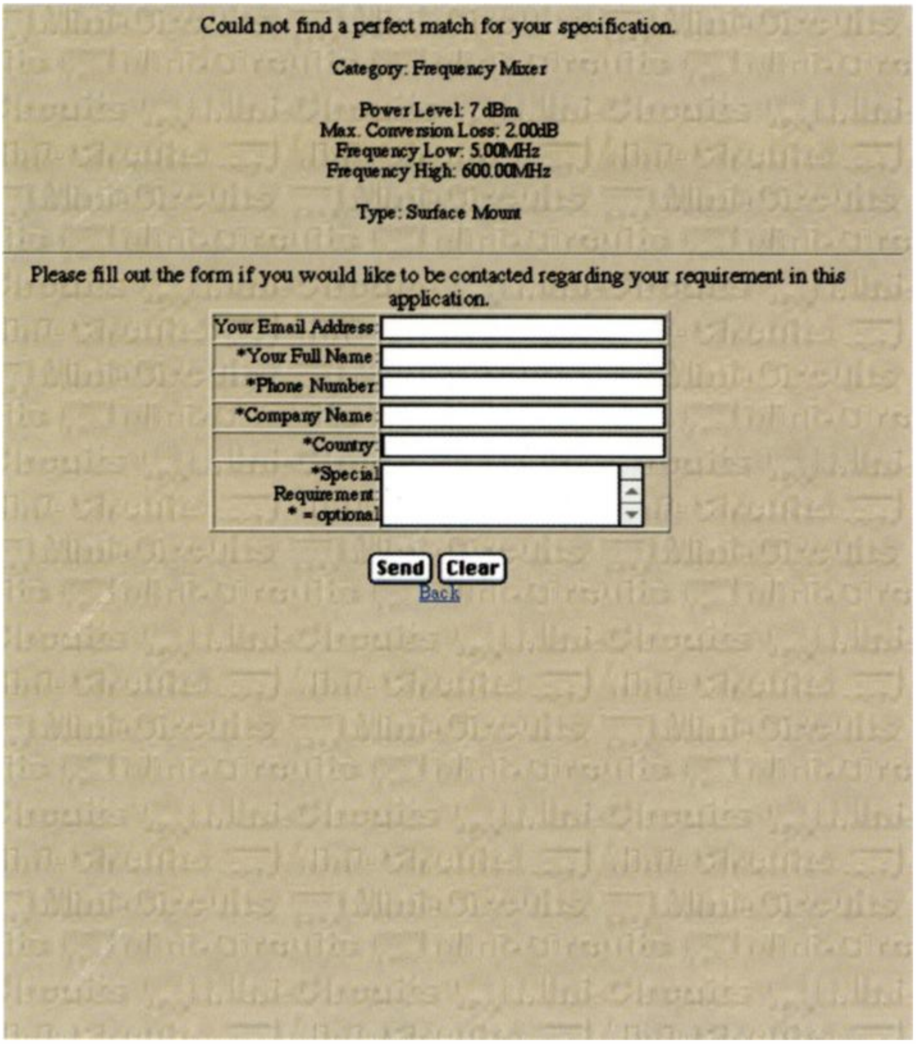
Conversion Loss | LO-RF Isolation | LO-IF Isolation | VSWR RF Port | VSWR LO Port | VSWR IF Port

Click Above for Actual Performance Data.

Evaluate the alternate component suggested using Tabular and Graphical displays.

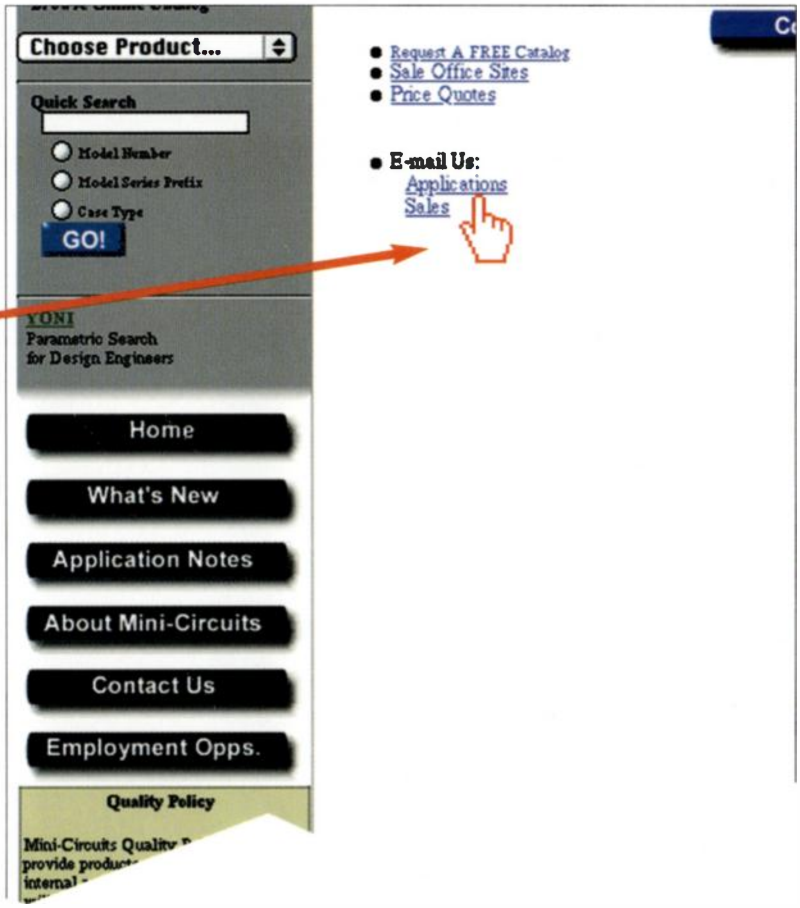
You can review each alternate component's specification page by clicking on the Model or Engineering Design Reference Number. The resulting page will display actual measured performance data in the form of tables, and indexes performance curves.

If YONI could not provide a part to meet your requirements, contact Mini-Circuits for a fast response.



SCREEN #2: *Contacting An Applications Engineer*
If a component match is not found and YONI could not suggest alternate models, a form will be provided to communicate your requirements to Mini-Circuits (see screen above).

Our team of dedicated applications Engineers are prepared to respond courteously, knowledgeably, and promptly to help find the Mini-Circuits part for your applications.



To obtain help finding your part in The YONI Search Engine, click the "Contact Us" bar on the Mini-Circuits home page, then click the "Applications" link.

Limitations Of The YONI Search Engine

YONI may not locate your part on the search engine if the required frequency range you entered is too narrow.
If YONI does not suggest a part, return to the Data Entry Fields (see page 8), broaden the frequency range, and proceed with a new search.

If The YONI Search Engine still cannot find an acceptable part for your requirements, you may contact Mini-Circuits through the "Contact Us" link on the home page. Detail your requirements in the form provided, and an Applications Engineer will contact you within 48 hours.

TIP:

The YONI "Quick Search" option can display spec pages by entering a complete model number or series prefix. It can also provide a detailed case style drawing.

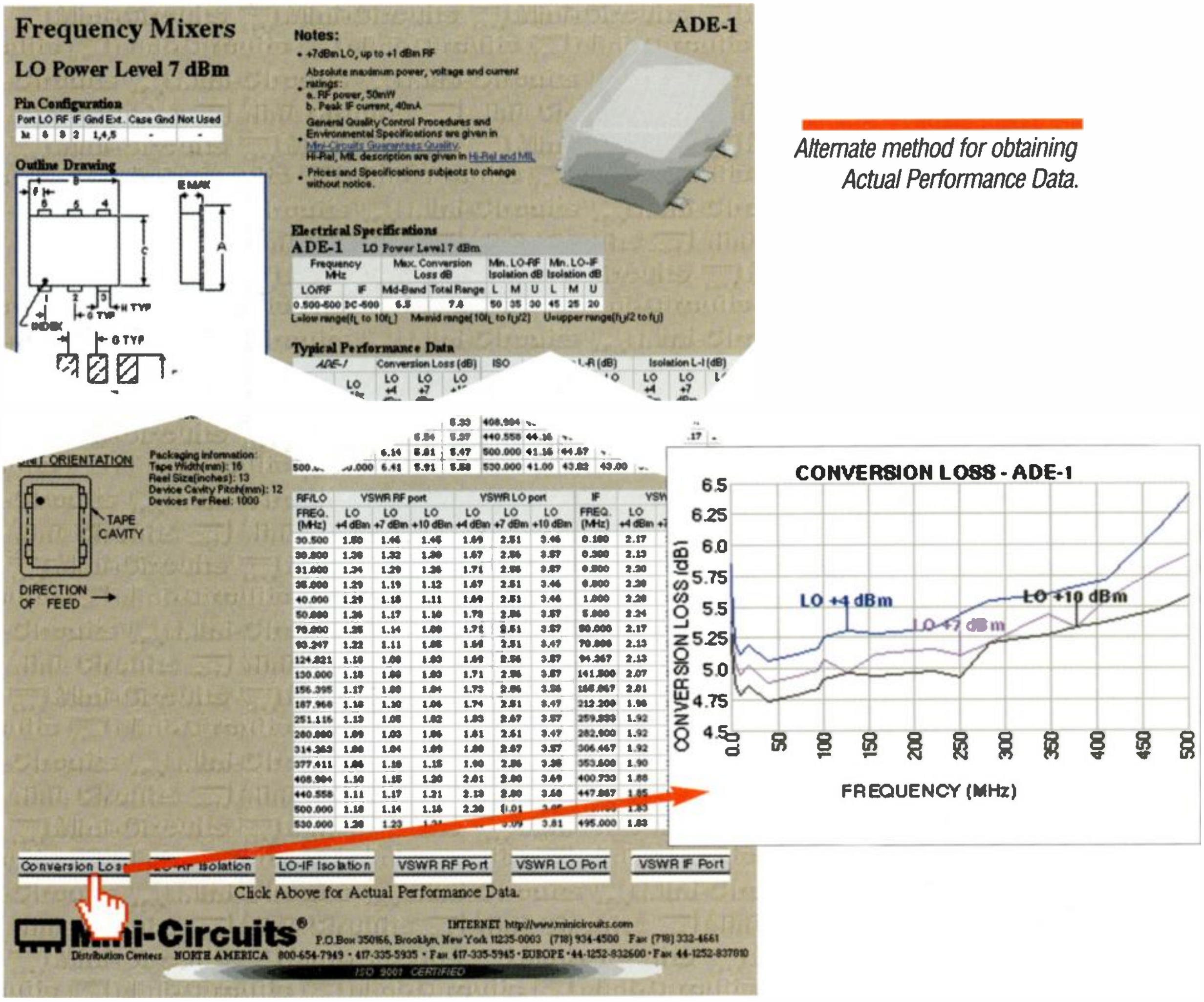
How To Obtain Actual Performance Data Using The Component Model Number

Actual performance data may be obtained via "The YONI Search Engine" using a component model number as well.

Return to Mini-Circuits Home Page at:
www.minicircuits.com
and enter the model number in the "Quick Search" field. Select the "Model Number" circle under the field, then click "GO".

This is an alternate method for accessing the YONI Search Engine to obtain actual performance data, tables, curves, and case style information for component categories that are indexed (see page 7).

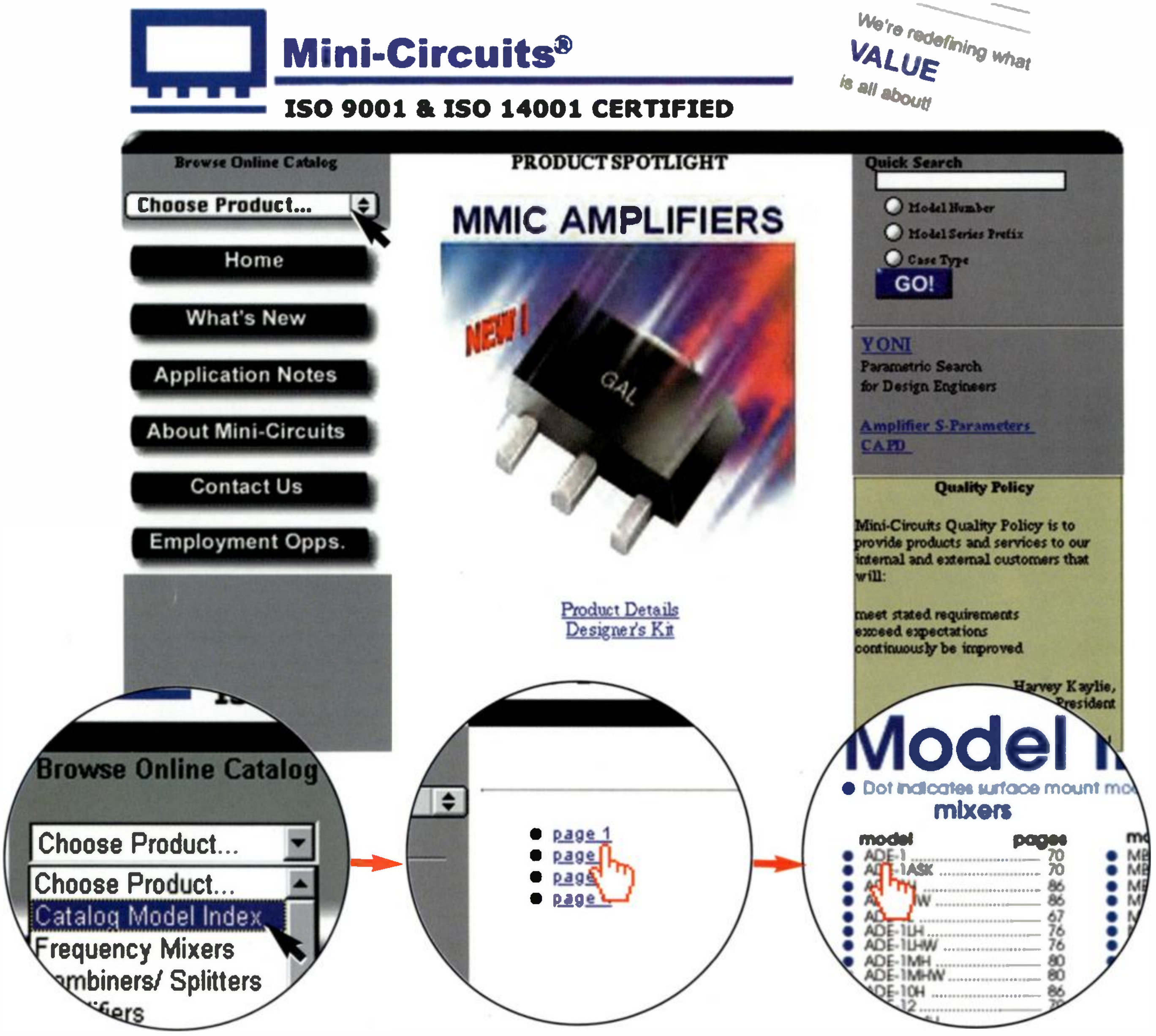
Note: This feature works with Internet Explorer version 4.0 or higher or Netscape 4.0 or higher.



Alternate method for obtaining Actual Performance Data.

If the part entered is indexed in YONI, the resulting page allows you to view actual performance data for the model entered. Tables will be displayed and graphical data may be viewed by clicking the appropriate specification bar.

For YONI navigating instructions, see pages 12 to 14.



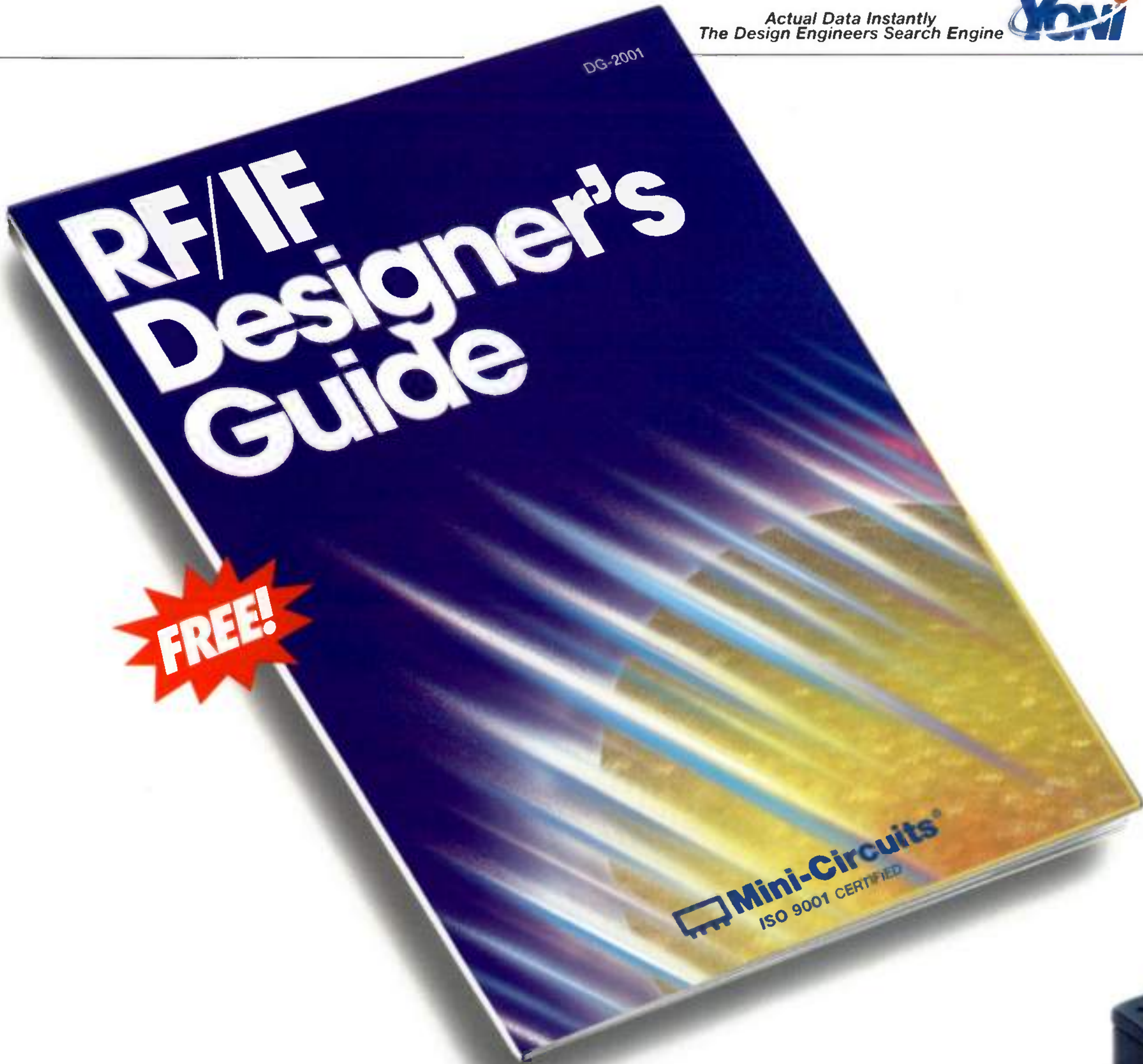
Select "Catalog Model Index" from the "Browse Online Catalog" drop-down menu.

Browse pages for your component category.

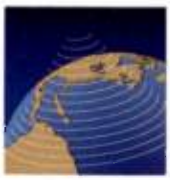
Click the model number to view specs and prices.

If the Model Number you entered is not indexed in YONI, return to the Mini-Circuits Home Page and locate your Model Number in "Catalog Model Index" from the drop-down menu located in the "Browse Online Catalog" feature.

Locate your component category in the index, then Click on a Model Number to view information from our Online Catalog.



AVAILABLE WORLDWIDE...NOW!



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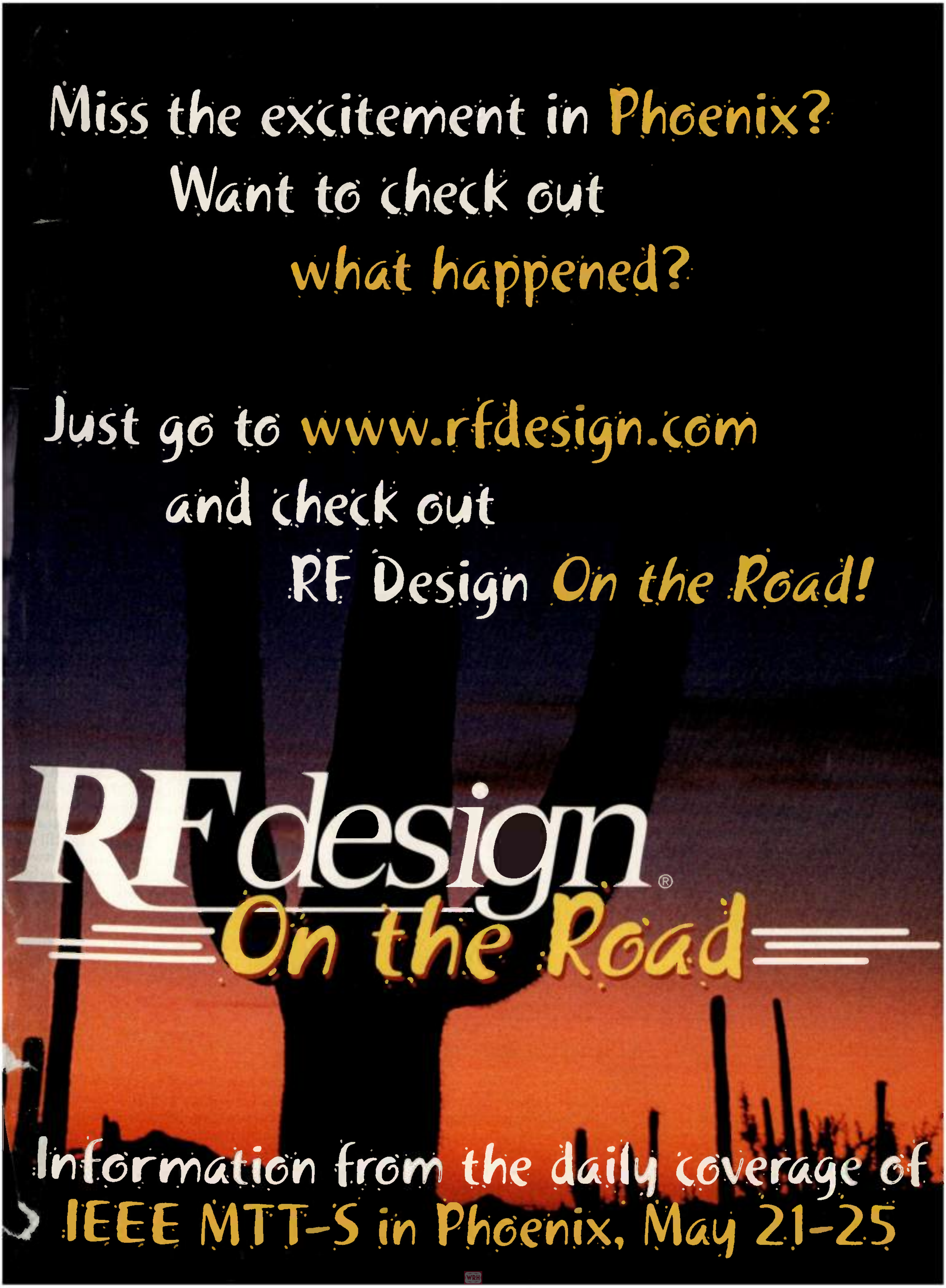
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Two-tone IMD measurement techniques

Seven rules to ensure the best characterization of non-linear RF components

By Keith Barkley

Two-tone testing for intermodulation distortion (IMD) has a long and venerable history. Since the dawn of RF engineering, it has been used to characterize the non-linearity of RF components, both active and passive.

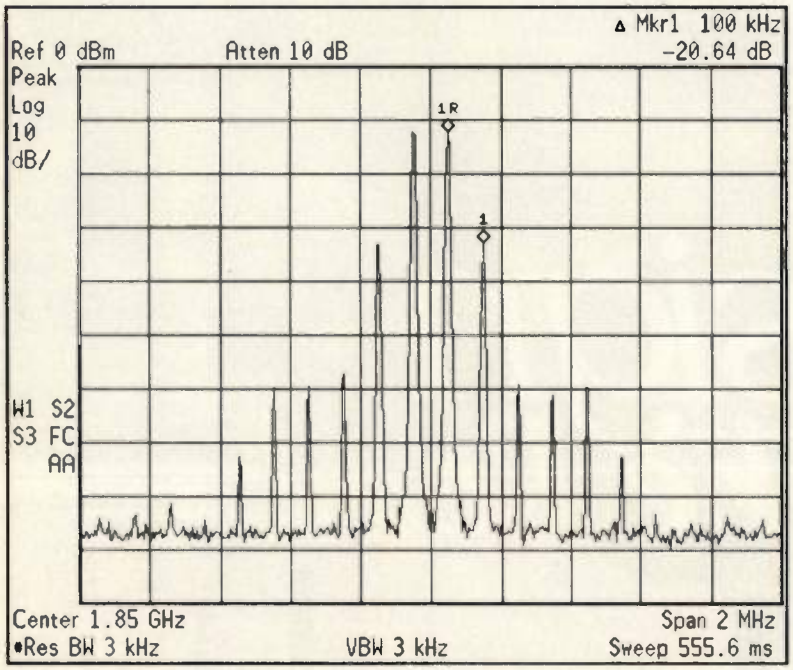

Figure 1. An example of a device with poor IMD response.

Measurement methods, pitfalls and tidbits of information have been gathered over the years by myriad sources, here and elsewhere. A few simple rules can help in setting up a test system that can accurately measure two-tone IMD for RF power transistors. For this discussion, f_1 and f_2 will denote the two signals' input into the device and Δf will be the frequency separation between them.

Where it comes from

While the causes and intricacies of the generation of IMD are beyond the scope of this discussion[1], intermodulation distortion is a result of a non-linear transfer function. Everything, with the possible exception of a signal in free space, generates IMD. *Rule #0: Everything generates IMD.*

A device that adds IMD to its output signal will have unwanted frequency components generated at specific frequencies. The third-order products will occur at a frequency of $(2 \bullet f_1 - f_2)$ and $(2 \bullet f_2 - f_1)$, showing up as extra frequency components Δf above and below the two input frequencies. Fifth-order IMD will show up as extra frequency components above and below the third-order distortion, exactly Δf apart. The other odd-order IMD products follow suit. The sample spectrum, as seen on a spectrum analyzer, of a device exhibiting poor IMD characteristics is shown in Figure 1.

Some may consider it an oversight that most literature touches only on the odd-order IMD products, leading one to believe that second-order distortion products have gone the way of TV channel 1 in the United States. In fact, second-order products are generated, they just tend to be out-of-band for most RF applications. The second-order products fall at the sum and difference of the two tones, or at $(f_2 - f_1)$ and $(f_2 + f_1)$.

Designers of multi-octave amplifiers must pay attention to second-order effects, as well. The easiest way to lessen second-order distortion is to use a push-pull amplifier. The second-order products generally cancel, greatly reducing them.

GSM, CDMA, PCN, 8VSB, π/4DPQSK, NAMPS, 64QAM and EDGE are just part of the alphabet soup that is proposed or current digital modulation schemes. How does old-fashioned two-tone testing fit into all of these standards? The jury is still out. It is highly desired to find some correlation between the performance of a component under two-tone test conditions and its behavior subjected to these complicated digital signals[2].

How to set up an IMD test station

A generic two-tone IMD station is shown in Figure 2. It consists of two parallel paths with two signal generators, two amplifiers and four circulators summed in a two-way combiner routed through an (optional) adjustable attenuator and then fed into the device under test (DUT). A spectrum analyzer is used to perform the actual measurement.

Unless one is cost-conscious, it is best to use [illegible] synthesizer as opposed to an RF sweeper. Whi[illegible] using two sweepers is possible, at small Δf the f[illegible] quency will need continuous readjustment to k[illegible] the tone spacing relatively constant. At a tone s[illegible] ing of less than 50 kHz, it is not worth the tr[illegible] especially in production testing.

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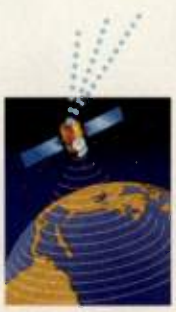

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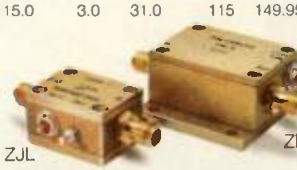
ZJL-7G	20-7000	10.0	±1.0	8.0	5.0	24.0	50	99.95
ZJL-4G	20-4000	12.4	±0.25	13.5	5.5	30.5	75	129.95
ZJL-6G	20-6000	13.0	±1.6	9.0	4.5	24.0	50	114.95
ZJL-4HG	20-4000	17.0	±1.5	15.0	4.5	30.5	75	129.95
ZJL-3G	20-3000	19.0	±2.2	8.0	3.8	22.0	45	114.95
ZKL-2R7	10-2700	24.0	±0.7	13.0	5.0	30.0	120	149.95
ZKL-2R5	10-2500	30.0	±1.5	15.0	5.0	31.0	120	149.95
ZKL-2	10-2000	33.5	±1.0	15.0	4.0	31.0	120	149.95
ZKL-1R5	10-1500	40.0	±1.2	15.0	3.0	31.0	115	149.95

NOTES:
1. Typical at 1dB compression.
2. ZKL dynamic range specified at 1GHz.
3. All units at 12V DC.



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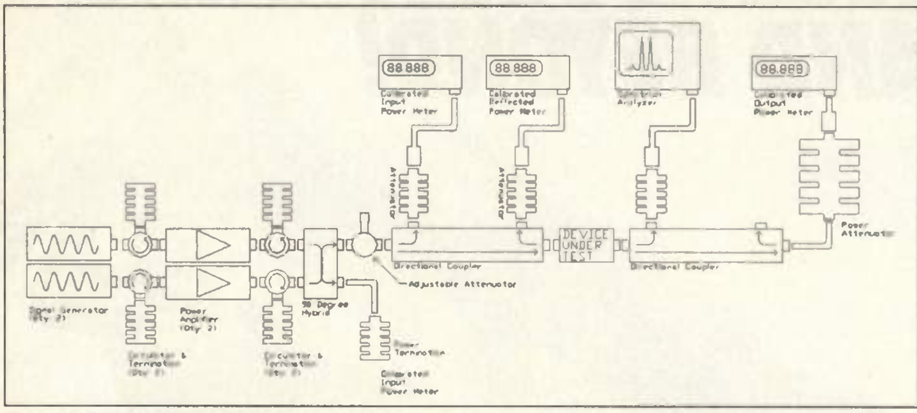

Figure 2. Typical IMD measurement system.

els of the RF driver amplifiers constant, while allowing for an adjustable RF power level at the DUT. This will keep the source IMD (SIMD) from changing as the power is adjusted, and it will allow the adjustment of both tone levels simultaneously. The only disadvantage is that the attenuator must be able to handle the average power and, if it is a non-linear device, generates IMD of its own (Rule #0).

It is important to measure the residual, or SIMD of the test station to ensure that it will not interfere with the measurement. At the same time, also look for spurious responses coming from the system that may contribute to SIMD from unlikely or unnoticed sources. Synthesizers use frequency references at nice round frequencies that may or may not coincide with the required Δf. To measure the SIMD, simply remove the DUT and connect the two-tone source to your load and measure the IMD of the source over all required power levels and frequencies. (The formulas for calculating SIMD are given at the end of the article.)

Fine tuning

It appears that locking the synthesizers and spectrum analyzer together with a common frequency reference would be a good thing to do, and in general, probably is, especially at narrow tone spacing. If the spacing is above 10 kHz, there is probably no benefit, and it may exacerbate any problems due to spurious responses from the synthesizer. One other benefit of locking them together is that one may be able to use the test station to achieve a single frequency power of twice the individual amplifier power. The synthesizers can be set to exactly the same frequency, and the phase of one of them can be adjusted to maximize the available power.

The other reason to look closely at the output of the two-tone generator is to identify any spurious responses that may confuse things if seen while testing a component.

Reverse IMD test and check

A slight modification of the two-tone measurement system is a test of reverse IMD, that is, the IMD of a component when the second tone is injected into the output port of the DUT. This test simulates a situation

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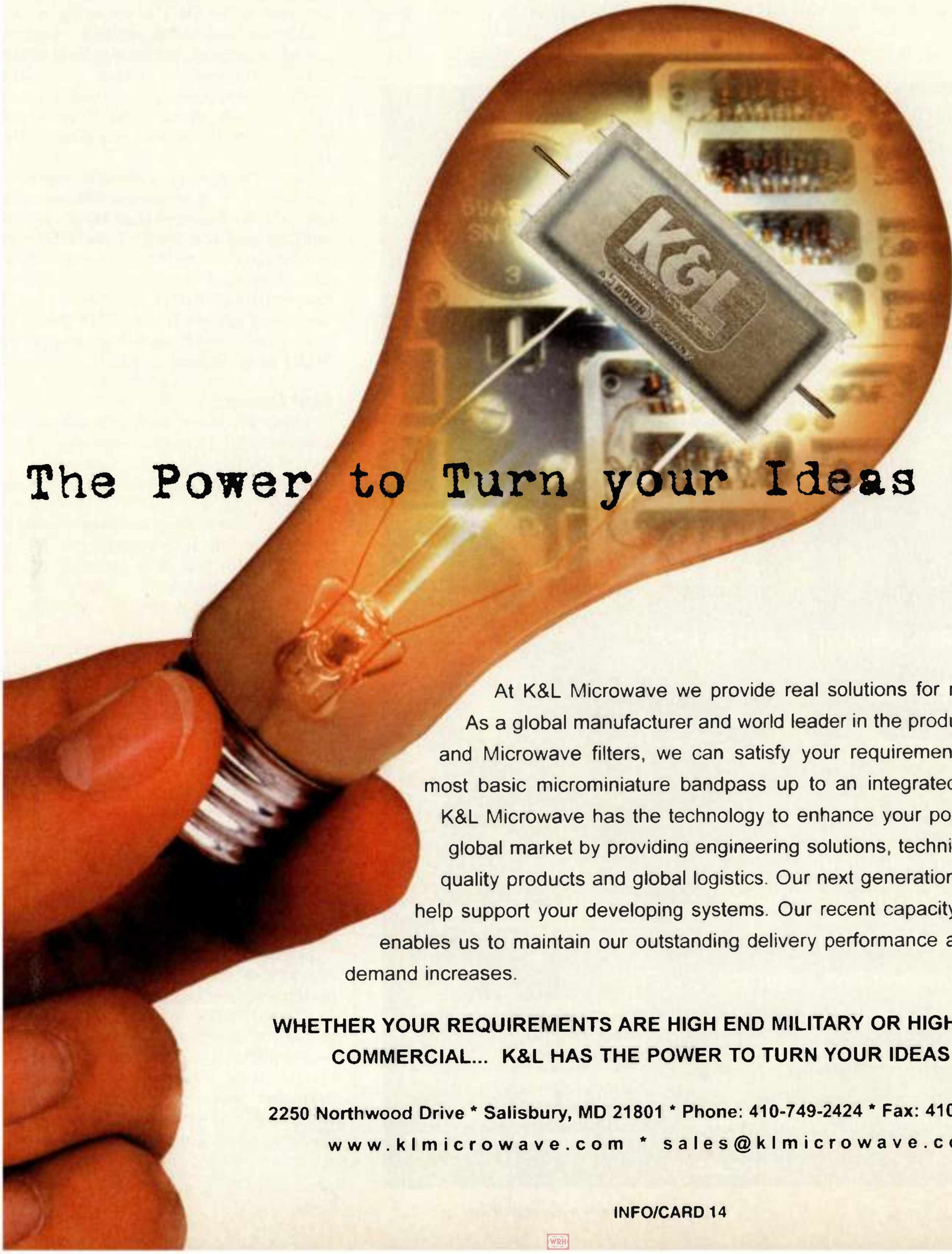

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common in communications equipment where an interfering signal from a co-located transmitter is conducted back through the system and into the amplifier. The equipment to perform this test is shown in Figure 3.

A directional coupler is used to inject the signal back into the DUT. The circulator is needed to prevent the signal source from generating IMD products of its own and from interfering with the measurement.

Correlation of the test system must be done without the DUT in place. A power meter (with attenuation, if necessary) is connected at the output port of the DUT looking toward the load. The signal is applied to the directional coupler. Typically, this involves injecting the signal into what is usually thought of as the port that couples out the signal from the main line going toward the load. Most directional couplers have internal terminations that cannot handle much power (check the data sheet carefully. The signal source is adjusted so that the level at the output port of the DUT is set at the specified level. Because directional couplers are symmetrical, the power level of the reverse tone at the signal generator will be attenuated by the coupling factor, so it may need to be 20 to 30 dB higher than the power required at the DUT.

Once the signal level is calibrated, the DUT can be replaced and set into operation. The reverse tone can be applied and the level of the IMD can be checked as stated in the specification. For example, the specification may call out for a 10 dBm reverse tone, and all resulting IMD products need to be –80 dBc or better, compared to the main tone at 40 dBm.

Fast forward

There are three secrets to setting up a successful two-tone station: isolation, isolation and more isolation.

Rule #1: Get as much isolation as possible.

It may seem that too many circulators are in the test system, but a single circulator has only about 20 dB of isolation. It may be necessary to cascade two to double the isolation between stages.

In fact, some signal coupling may not come through the RF cables at all, but through radiation or even conduction through the AC power line. If big problems exist with excessive IMD in a test system, try moving the equipment around. Take the generators out of the 19-inch rack and pile them around the bench with some separation between them. Plug the two halves of the system into separate AC circuits. At the signal levels involved with –80 dBc IMDs, it does not take much signal to cause a problem.

Rule #2: Check your test system.

It does seem that a lot of added expense exists in using two amplifiers instead of one larger Class-A amplifier. It will take a large amplifier to generate acceptably small levels of source IMD at the DUT. A good rule of thumb is to assume that the driver amplifier will have –30 dBc of third-order IMD at its rated power level. (With the test tones set at one-fourth


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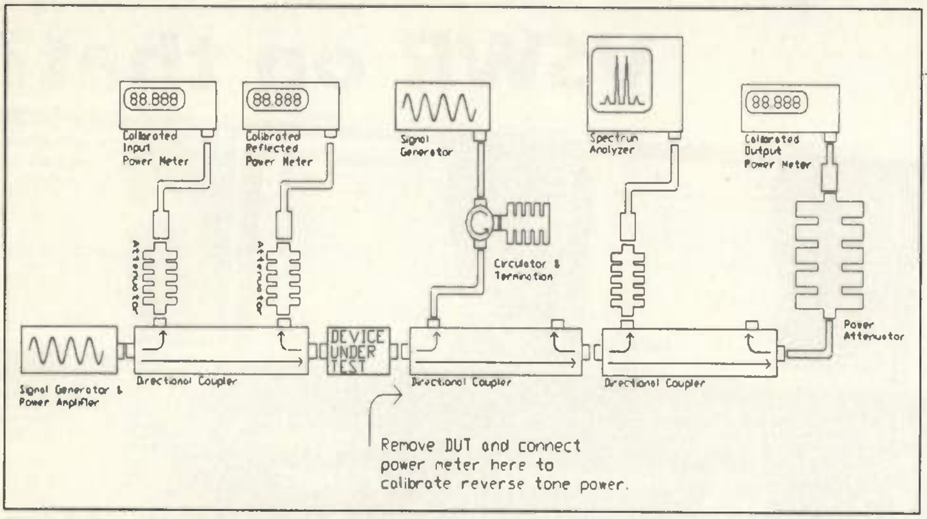

Figure 3. A reverse IMD measurement system.

the rated power per tone.) 2 dB of third-order IMD improvement will be seen for every 1 dB the power is reduced. For example, if a 100 W Class-A driver amplifier is used at 10 W (2.5 W per tone), one can expect a source IMD improvement of 20 dB, or –50 dBc of third-order source IMD.

Pitfalls of IMD testing

When specifying and making two-tone IMD measurements, one must be precise about the terminology used to state the amplitudes of the tones involved. There are three ways to specify the power of the tones:

- Per-tone—The absolute power of the individual tone. This is most accurately and easily seen on a spectrum analyzer. If a data sheet says that the part is tested at 1 W per tone, it means to apply the two signals such that each signal has an amplitude of 1 W. It may be referred to the input or the output of the device, though output is most convenient for measurement if there is only one spectrum analyzer.
- Average—The average value of the two tones. This is most accurately and easily measured with a true-RMS power meter. The value is simply the sum of the two powers of each tone. So, if there is 1 W per tone, the average is 2 W.
- Peak—Otherwise known as peak envelope power (PEP), it is the maximum instantaneous power of the combination of the two signals. A two-tone signal looks similar to an AM modulated tone. The envelope of the RF signal varies as a sinusoidal with a frequency of Δf. When the voltages of f_1 and f_2 are out of phase, they cancel and the envelope is at a null. When the two voltages are in-phase, they add, and the voltage at the instantaneous peak is twice that of either tone. Because power is equal to the square of the voltage divided by the resistance, when the voltage increases by a factor of two, the power increases by a factor of 2^2, or four. Therefore, the PEP is four times the per-tone power and twice the average. In this example, the per-tone power is 1 W per tone, the average power is 2 W and the PEP is 4 W.

The common industry practice uses an average reading power meter and calculates the PEP as twice the average power reading. This is not technically correct because the PEP is always less than twice the average power in the presence of harmonic and intermodulation distortion. However, at the typical IMD specification levels, around –30 dBc, the error is negligible. If a peak reading power meter is used to measure the PEP directly, that fact should be noted with the results so that measurements can be compared with another system that uses average reading meters.

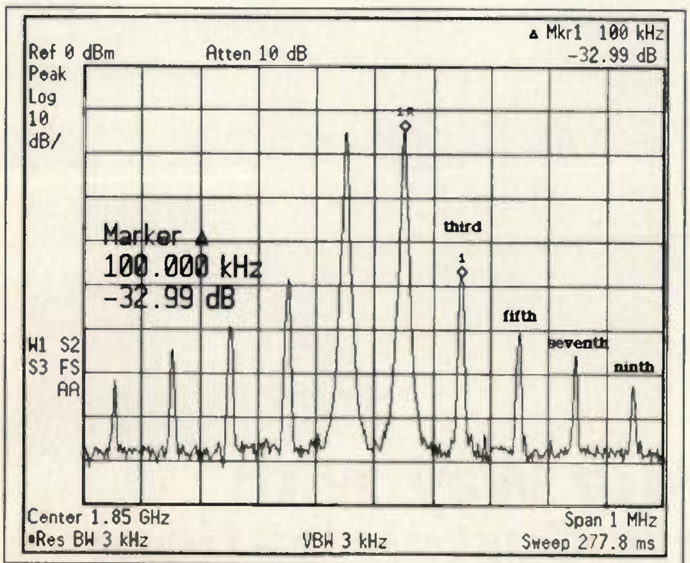

Figure 4. Example IMD measurement.

It is sometimes easy to get confused, so I suggest the next rule: *Rule #3: Stick to one method of power measurement.*

Clearing the throat

Now that the terms are more clearly defined, conventions can be specified more clearly. Depending on the manufacturer, company data sheets may specify either the per-tone or the PEP of the signal. It may be referred to as the input or the output. However, the IMD in dBc is always referred to as the output, and the reference level of the carrier (the "c" in dBc) is the per-tone level. In other words, the IMD is measured from the highest of the two input tones to the highest of the IMD product being measured (third, fifth...).

The reason to belabor these points is that one must read manufacturers' data sheets carefully. It is easy to manipulate the numbers to make one manufacturer's device seem better than its competitor's.

One manufacturer of power RF semiconductors uses the PEP as the reference for the carrier level, giving it a 6 dB boost in third-order IMD performance over the industry. Unfortunately, there may be no way to tell by looking at a data sheet where the carrier reference is set because the data sheet usually just says "dBc."

Figure 4 shows a real-world example. This spectrum analyzer is attached to a coupler at the output of the DUT. This particular device is operated at its data sheet conditions of 26 V, and an ICQ of 25 mA. The output power is set at the specified 2 W PEP. To make a measurement, both tones are turned on and the adjustable attenuator is set to achieve 1 W on the true RMS reading power meter, for the 2 W PEP. In this case, because the PEP is set through the power meter, the spectrum analyzer is being used as a relative measuring device only. Therefore, it is not necessary to calibrate the spectrum analyzer to the absolute power level.

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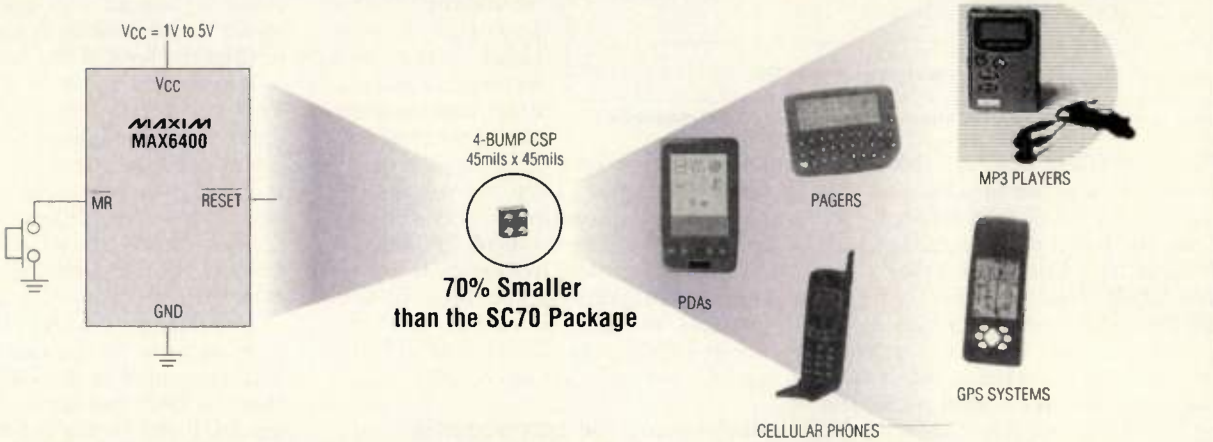

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MAX6403/MAX6409	3.30 to 4.63	✔			MAX6403 Only	MAX6409 Only
MAX6404/MAX6410	3.30 to 4.63		✔		MAX6404 Only	MAX6410 Only
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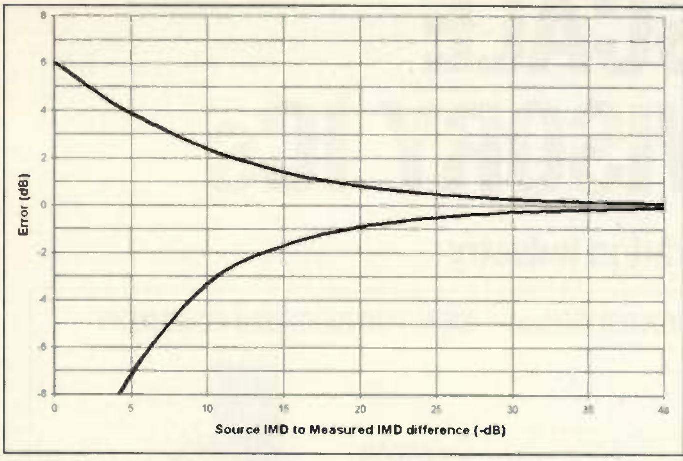

Figure 5. Maximum and minimum error vs. SIMD/measured IMD difference.

The levels of the tones are checked on the spectrum analyzer to make sure they are within 0.2 dB of each other. If they are not, the levels of the signal generators must be adjusted to bring them within 0.2 dB. (Note: Observe the behavior of the delta marker function on the spectrum analyzer. On some analyzers, when the delta reference is set, it stays at an absolute power level no matter what the signal is doing "underneath." More friendly analyzers will have the delta reference follow the level of the signal. If an analyzer is of the first type, make sure to adjust the level of the tone where the delta marker is located, not the tone where the delta reference is set.)

When the tones are set to equal amplitudes, the IMD products are measured by comparing the level of the highest of the two tones to the highest of the desired IMD product. Some spectrum analyzers have a feature that automatically shows the absolute or relative level of all the peaks on the display. With this function, measuring the levels of all the IMD products can be accomplished with just a glance at the analyzer.

Measurements at other PEP levels can be easily accomplished by adjusting the power to a new level with the adjustable attenuator, and making the measurement. (Note: If the tone spacing is wide, 1 MHz or greater, then the amplitude flatness over frequency may change, requiring that the levels of the tones be correlated at each frequency to maintain the proper relationship.) In this example, the third-order IMD products are at –33dBc, the fifths are at –47 dBc and the sevenths are at –53 dBc.

Understand the components

A power meter may provide an inaccurate reading of average power in two situations. The first is when something other than a true RMS power meter is used. Thermocouples, thermistors and diode detectors operated in the square-law region are true RMS readings and give an accurate indication of the actual average power. Power meters that use diodes in the non-square-law region must be watched closely to get accurate results. In fact, it could be suggested that one only use true RMS power meters for two-tone IMD measurements.

The second way to get misleading average power readings is if the signal does not look like a two-tone signal. Any frequency components besides the two tones (including IMD products) add to the average power. Luckily, the problem is linear and well-behaved. If the third-order distortion is –30 dBc, then the error is 0.1 %. If the thirds, fifths and sevenths are all –15 dBc, the average power meter reading is about 10% higher than the level of the two main tones.

Harmonics must also be watched. The power in harmonics will add linearly to the measurement and cause error. A peak-type diode power sensor is extremely sensitive to harmonics. When using one of these meters, a harmonic filter is almost always required for power devices that generate significant harmonic powers.

Any IMD generated in the test system is going to be at the exact frequency of IMD generated by the DUT. This means that the IMD measured at the output of the DUT will be the result of the vector sum of the SIMD and the DUT IMD. This leads to more error than you might expect. It is not a simple linear addition of the two powers, but an addition of the voltages, much like the PEP discussion. *Rule #4: Power meters can lie.*

Doing the math

Here are the formulas for calculating the worst-case error due to SIMD:

The maximum positive error (the voltages are in phase) is:

$$(err+) = 20\log\left(1 + 10^{\frac{SIMD - MIMD}{20}}\right)$$

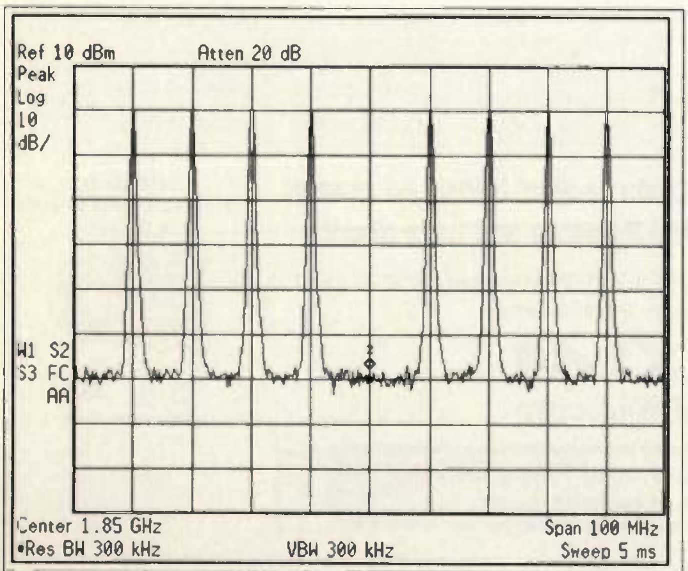

Figure 6. Multi-tone test signal.

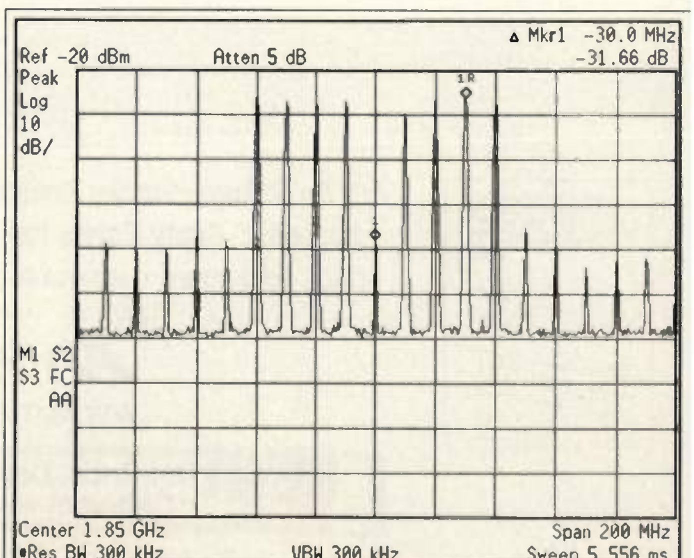

Figure 7. Device response to multi-tone test signal.

Number of tones	average power (W)	PEP (W)	Peak to average ratio
2	2	4	2
3	3	9	3
4	4	16	4
5	5	25	5
6	6	36	6
7	7	49	7
8	8	64	8

Table 1: Comparison of peak and average power for various numbers of tones.

The maximum negative error (the voltages are out of phase) is:

$$(err-) = 20\log\left(1-10^{\frac{SIMD-MIMD}{20}}\right)$$

Where SIMD is the absolute value of the IMD at the input of the DUT, and MIMD is the IMD measured at the output of the DUT.

For example, what is the possible error if a –30 dBc third-order IMD is measured at the output of the device and the residual SIMD was –40 dBc? The difference between the measurement and the source is –10 dB. Plugging the numbers in the formula yields:

$$(err+) = 20\log\left(1+10^{\frac{-40-(-30)}{20}}\right)$$

$$= 20\log\left(1+10^{-0.50}\right) = 2.4$$

and

$$(err-) = 20\log\left(1-10^{\frac{-40-(-30)}{20}}\right)$$

$$= 20\log\left(1-10^{-0.50}\right) = -3.3$$

The somewhat surprising results are that if the SIMD is 10 dB below the IMD measured at the DUT, the error is between +2.4 and –3.3 dB. A plot of the maximum and minimum error vs. the IMD difference is shown in Figure 5. The true IMD of the DUT lies somewhere between the two curves, depending on the magnitude and phase of the SIMD and IMD generated by the DUT. This leads to the suggestion that the source should have IMDs at least 30 dB below where it is desired to take accurate measurements. *Rule #5: Keep Source IMD to a minimum.*

The importance of the SA

The spectrum analyzer (SA) is the heart of the IMD measurement system. Generally, setting the controls on the spectrum analyzer for measuring IMD is simple[3]. The eye can be a good judge. Be sure that the power going into the mixer is less than the power that the manufacturer uses in the IMD specification of the analyzer.

For example, if the specifications of the analyzer say that third-order intermodulation distortion is less than –80 dBc at –20 dBm input to the mixer, make sure that the per-tone power is –23 dBm, or lower, to prevent the analyzer from influencing the measurement. If an analyzer has a minimum distortion/minimum noise switch, use the minimum distortion setting.

Span, sweep speed, resolution and video bandwidths can be set to get a readable trace at whatever conditions yield the best dynamic range and show all the required IMD products on the screen. If necessary, the span can be lowered to show only a single component and the center frequency adjusted to the proper frequency. This is usually the best way to programmatically read the IMD. (Here is a situation where having the reference frequencies linked to a master clock can really help.)

Make sure that the level of the lowest IMD product to measure is above the noise floor. A signal only 3 dB higher than the noise level is still accurately measured[4]. With a slow enough scan and averaging, almost any IMD component can be resolved from the noise.

Be careful when adjusting the reference level to get a lower noise. This may lower the internal attenuation level and overload the mixer. It is better to narrow down the span, reduce the filter bandwidths and zero in on the desired IMD product. Verify that the delta marker function does what you expect in these circumstances.

Also, be sure to set the spectrum analyzer so the mixer is not being overloaded on one hand and not lost in the noise on the other. For typical two-tone measurements on power devices, these conditions are met easily.

Because almost every spectrum analyzer is different, and there are often neat features (as discussed above) hidden inside, another rule can be stated: *Rule #6: Get to know your spectrum analyzer.*

Accuracy and expectations

It can be difficult to be truly accurate because of all the variables involved. However, several of the errors can be named, and most can be accounted for with careful attention to detail.

- Power level — For every 0.1 dB of error in measuring the device's PEP, the third-order IMD will be off by 0.2 dB.
- Frequency response — (This is more of a problem for wide Δf.) The frequency response of the measurement system, especially any directional couplers, should be factored in. With narrow tone spacing, (up to 100 kHz) this can usually be ignored.
- The log response of the spectrum analyzer — After the signal is detected, the video signal is sent to an amplifier that has a logarithmic response vs. signal amplitude. The output of this log-amp is then displayed on the screen. The log-amp error of a typical HP 71200A spectrum analyzer from the early 1980s is specified at 0.5 dB from 0 to 90 dB at IF bandwidths between 30 and 100 kHz. Newer analyzers digitize the detected video output (or even the IF) and use DSP to perform the log functions. A Rohde and Schwarz FSEB has both digital and analog specifications. The analog specification at resolution bandwidths of greater than 1 kHz is less than 0.3 dB over a range of 0 to 50 dB and less than 0.5 dB from 50 to 70 dB. In digital mode, the error is less than 0.3 dB over a 0 to 100 dB range at resolution bandwidths less than or equal to 1 kHz. So the error due to the logarithmic transformation can be estimated to be from 0.3 to 0.5 dB.

The end analysis

Putting all these errors together, a rule of thumb is (when comparing two systems) do not trust differences of IMD measurements less than 1 dB. In absolute terms, a test system can be accurate to 0.75 to 1 dB. Two IMD measurements can be compared, if they are taken on the same system, at the same conditions, to within 0.5 dB.

Rule # 7: The termination on the combiner must be able to handle half the average power input to the combiner.

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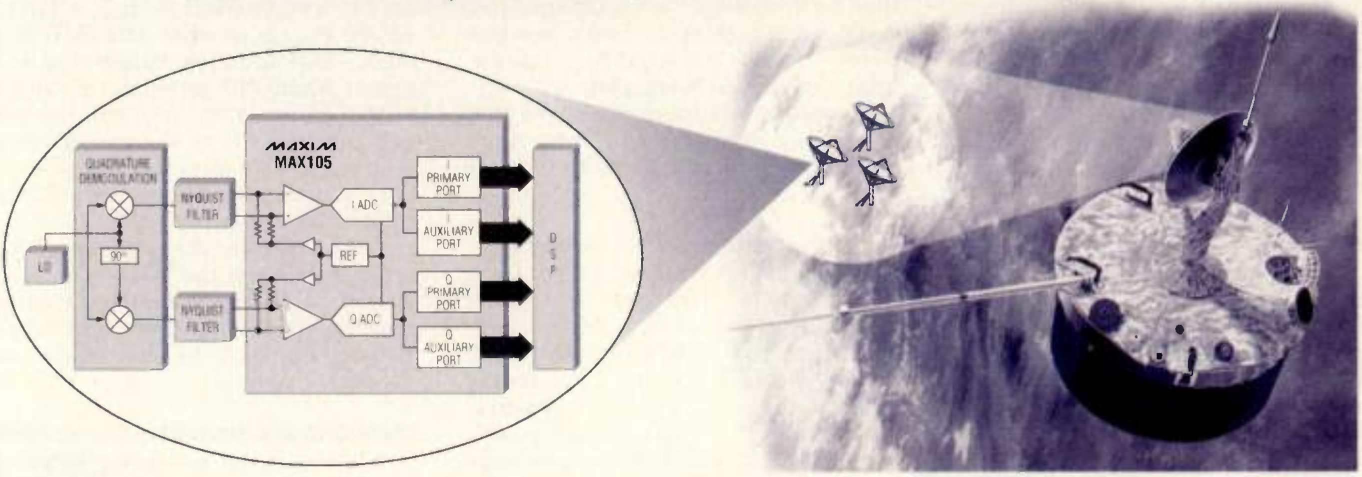

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Say that the system is set up, as in Figure 2, with two 10 W amplifiers to boost the level of the signal generators. Connect the system to a power attenuator and power meter and prepare to verify the IMD levels of the system as this note suggests. Turn on all equipment, set signal generators to drive amplifiers to 10 W each, crank the adjustable attenuator to the minimum attenuation and look on the power meter expecting to see 20 W average, instead getting about 5 W average.

When it doesn't add up

What happened? All the components are taken to the network analyzer and about 2 dB of loss can be accounted for. Where is the other 4 dB going? About 1 dB is because of the fact that a circulator can have higher loss in power conditions than at small signal. (This loss is variable and may never come up.)

By far, the largest loss, 3 dB, is dissipated in the isolation port of the combiner. (If using a Wilkinson two-way combiner, a "hidden" resistor is inside the unit that is dissipating the extra power. Check the data sheet to make sure that Rule #7 is being followed.) Most combiners are only "lossless" when the two input signals are at the same frequency and a particular phase relationship. All that extra power is being dumped into the 3.5 mm termination that was borrowed "only for a second" from the network analyzer calibration kit.

There are expensive combiners that are lossless with differing frequencies, but they are narrowband, difficult to find, and expensive.

Measuring third-order intercept

A common measurement for Class A amplifiers is third-order intercept (TOI). In a well-behaved non-linear amplifier, the third order IMD will increase at a 3:1 ratio. This means that for every 1 dB the output is increased, the IMD (in dBc) will increase by 2 dB. If the curves are drawn for the output power and the third-order IMD power (not the ratio in dBc), the intersection of the two lines is the TOI. (In practice, one can never measure where they intersect because the amplifier will usually saturate long before the TOI point. The lines are extrapolated from the linear region.) It is easy to calculate the TOI. If one knows the IMD at a particular tone power, the TOI is (P_{tone} + IMD/2) where P_{tone} is in dBm and IMD is in –dBc. TOI must be calculated in a region where the measured distortion of the device is linearly well-behaved and the errors from the measurement system are at a minimum. In practice, it is a good idea to calculate TOI at IMD levels around –40 dBc.

That being said, it never makes sense to measure the TOI of a class AB amplifier. Most Class AB amplifiers have narrow areas where distortion is well-behaved and the IMD is often not even monotonic.

Multitone testers and notched noise

While it is not commonly specified another option for linearity testing is multitone testing where more than two tones are used. Because the PEP (instantaneous peak) increases as the

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CLV0815E	806	824	0.5-4.5	11	-113	-35	5.0	11
CLV0950E	865	1035	1-10	27	-114	-11	5.0	24
CLV0915A	902	928	0-4	17	-108	-30	3.0	10
CLV1085E	1050	1086	0.5-4.5	21	-112	-20	5.0	20
CLV1385E	1370	1400	0.5-4.5	18	-110	-20	5.0	20
CLV1550E	1500	1600	0.5-5.0	44	-106	-35	5.0	22
CLV2465E	2436	2496	1-4	26	-107	-20	5.0	25

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SMV0162A	125	200	0.7-8.3	12	-100	-6	5.0	36
SMV1570L	1540	1600	0.5-2.5	128	-90	-15	2.7	9
SMV2165A	2118	2218	0-3	148	-91	-10	3.3	16
SMV2390L	2290	2485	0-4	116	-90	-11	5.0	16
SMV2660L	2620	2700	0.5-4.5	90	-91	-17	5.0	21

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USSP2330	2300	2360	0.5-2.5	0±3	-83	-15	2.7	8

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PLL0210A	200	230	100	0.50	-105	3.5±2.5	+5	25
PLL0930A	900	960	100	0.75	-101	3±2	+5	40
PLL1260A	1230	1290	1000	0.75	-102	1±2	+5	40
PLL1456A	1420	1490	1000	0.75	-103	1±2	+5	40
PLL2710A	2670	2740	1000	1.25	-98	1±4	+5	30

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square of the number of carriers, signals can be generated with high peak-to-average ratios rapidly. Table 1 shows how quickly the peak power of a number of 1W tones increases (this table should give one an appreciation for the linearity requirements of a CATV line amplifier, which has to handle 70–80 channels). Multitone testing is an effective test of an amplifier's linearity, especially for multichannel amplifiers.

Generators are on the market that use separate oscillators to generate the multitone signal. These oscillators can be phase-locked to provide random phase relationships, or adjusted to provide worst-case phase or best-case phase relationships. (While it is beyond the scope of this note, as the number of tones increases, the amount of time spent at the PEP level decreases. The time spent near the peak is a function of the phase of the individual tones. By adjusting the phase of the tones, the time spent at the PEP of the signal can be changed to near zero, which changes the peak value of the signal without changing the average power level.)

How to use a multitone generator

With more tones, the peak level increases greatly while the average increases slowly. A multitone generator can output eight tones at about 2 mW per tone[5]. With eight tones, the PEP is 128 mW, which may be sufficient for testing medium-power RF transistors or high-gain devices like modules. If more power-per-tone is needed, the high peak-to-average value of the signal almost requires one to use separate amplifiers for each tone and then sum them in an eight-way combiner. The spectrum at the output of the eight-tone generator is shown in Figure 6. And after amplification by a power transistor, the signal is shown in Figure 7.

In practice, the eight tones are evenly spaced, possibly with a missing one as shown in the figures. When measuring the DUT, the distortion products at the missing tones, either in band or outside the band, are measured and compared to the level of the main tones. The device used as a source for this discussion has a multitone performance of –23 dBc.

Next-generation signal generators are now available and use an arbitrary waveform generator with an IQ modulator to produce almost any modulation on the carrier signal. It is possible to create an IQ waveform that will generate a multitone signal. This approach works fine with the following caveats:

- Generating the proper IQ signal — Software is available to generate the proper IQ signals (Rohde and Schwarz supplies software that specifically does this), but a new IQ signal must be loaded for every variation in tone spacing, tone phase or tone amplitude. No aspect of the signal can be adjusted, except level of all tones and center frequency, "on-the-fly."
- The generator is limited to a PEP level that is equal to the maximum that the generator can deliver, usually less than 100 mW — because the individual tones are not available for amplification, either the level from the generator or a "monster amp" must be used to get the tones to the level desired with little source IMD.


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One other linearity test is to generate a band-limited noise signal and use a notch filter to remove a portion of the noise.

When passed through the DUT, the IMD products "pile up" in the notch and the distortion is read as the ratio of the level at the top of the noise to the noise in the notch. This represents a severe test of amplifier linearity and probably is the technique that represents the real world the best. This is the easiest to extend to higher powers because the noise is generated and amplified to the desired level. A high-power notch filter is used to remove the noise from the measurement band. Any distortion products added by the driver amplifier are removed by the notch filter.

A spectrum analyzer with an RMS detector makes measuring the noise level easier. Remember that the interest is in simply comparing two noise levels, not measuring absolute noise. It is probably not necessary to use the noise marker to measure noise-per-root-Hz, or make any elaborate corrections usually required for noise measurements with spectrum analyzers.

The IQ signal generators mentioned above can also generate these notched-band-limited noise signals. However, the PEP limitation also applies to these signals[6].

RF

Summing up — remember the Rules:

#0: Everything generates IMD. Corollary to #0: Everything generates IMD in unexpected ways.

#1: Get as much isolation as possible.

#2: Check the test system.

#3: Stick to one method of power measurement.

#4: Power meters can lie.

#5: Keep source IMD to a minimum.

#6: Get to know your spectrum analyzer.

 # 7: The termination on the combiner must be able to handle half the average power input to the combiner.

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SCM — analog modulation for 21st century digital content

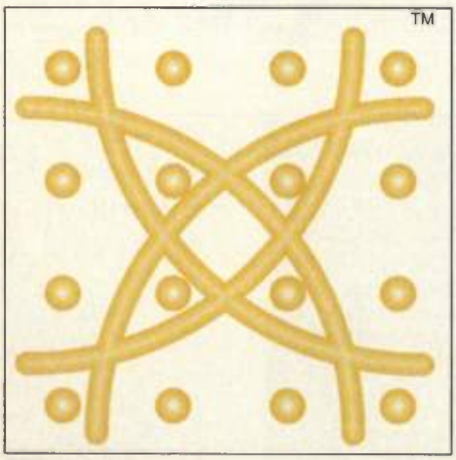
Contrary to popular belief, analog modulation isn't going the way of the dinosaur. Signal code modulation, a hybrid digital/analog scheme, may offer a viable alternative.

By Eli Pasternak

The industry is in the midst of a digital revolution, where virtually every form of information is either digital — or if it is not already digital (e.g. speech, music and video), it is digitized, stored in digital media, or transmitted digitally over communications channels. Is analog modulation a relic of the past, relegated to a time when AM and FM broadcasts were introduced? Should analog modulation ever be considered for state-of-the-art communications? Are there any clear benefits over digital modulation? The answer is yes and signal code modulation (SCM), a hybrid analog/ digital scheme with unique advantages, is one strong player.

Introduction to signal code modulation

Signal code modulation is a method for transmitting analog information over a noisy channel. SCM provides an analog pipe through which any band-limited waveform can pass, including truly analog information or the output of a digital modem. The operations that SCM performs on the payload signal are simple, as illustrated in Figure 1.



SCM constellation diagram.

The waveform is sampled and quantized, just like a typical pulse code modulation (PCM) transmission, and the digital signal is then transmitted over the noisy channel using any digital technique, such as quadrature amplitude modulation (QAM). The digital signals are denoted by the symbol *D*. However, unlike PCM, SCM does not discard the quantization error. This error signal is extracted and transmitted over the noisy channel as an analog symbol, *A*.

The SCM transmission and reception processes are depicted in Figure 2. The transmission channel is divided into two channels. Channel 1 is analog, and channel 2 is digital. In a process essentially identical to PCM, the original analog signal at the system input is sampled at the appropriate rate, based on the sampling theorem, and converted to digital values. The resulting *D* symbols are transmitted via channel 2 using a digital transmission technique optimized for the channel. Those *D* symbols represent *N* bits per analog input sample. To produce the quantization error *A*, the PCM data is converted back to analog and subtracted from the original input. This *A* symbol is amplified by a gain of 2^N or any gain that will optimize the voltage swing of the *A* symbol with that of channel 1.

The SCM receiver performs the opposite operation, combining the *A* and *D* symbols into an analog stream replica of the original analog signal. This replica is not a precise copy of the original signal, because noise in the channels could vary the *A* symbols or cause bit errors in the *D* symbols. However, the 2^N amplitude gain in channel 1 has provided noise power immunity of 2^{2N} to the *A* symbols. This is one of the key benefits of SCM.

Its application

This signal processing method is a straightforward approach in implementing a real wireless application, as shown by the following example. A radio channel of bandwidth *B* is splitting *B* between channel 1 and channel 2 (see Figure 2). The *D* symbols are transmitted as digital QAM symbols, and the *A* symbols are combined in pairs and transmitted as analog QAM. The RF channel bandwidth is divided by time division into the *A* and *D* symbols. The actual transmission may consist of a stream of ADADAD... symbols. If a *D* symbol contains eight bits and the RF channel is suitable for four bits per symbol, the transmitted signal may be arranged as ADDADDADD..., wherein *D* is a 16-QAM symbol. If, on the other hand, *D* is a two-bit symbol and the channel is suitable for four-bit symbols, the transmission will be AADAADAAD.... Each particular application will determine the optimum SCM mode in terms of choice of number of bits per *D* symbol.

The *D* symbols may be aggregated and encoded using forward-error correction (FEC) techniques, using typical framing and scrambling techniques. The SCM modem design can become a modification of a conventional all-digital QAM modem, using the synchronization and channel equalization techniques with slight modifications; thus SCM is suitable for implementation by existing digital modem techniques. Furthermore, even the *A* symbols can be processed digitally using digital signal processing (DSP) techniques. The presence of *D* symbols next to the *A* symbols simplifies the design of such a modem by having the receiver rely on the digital symbols to calibrate the signal gain and perform

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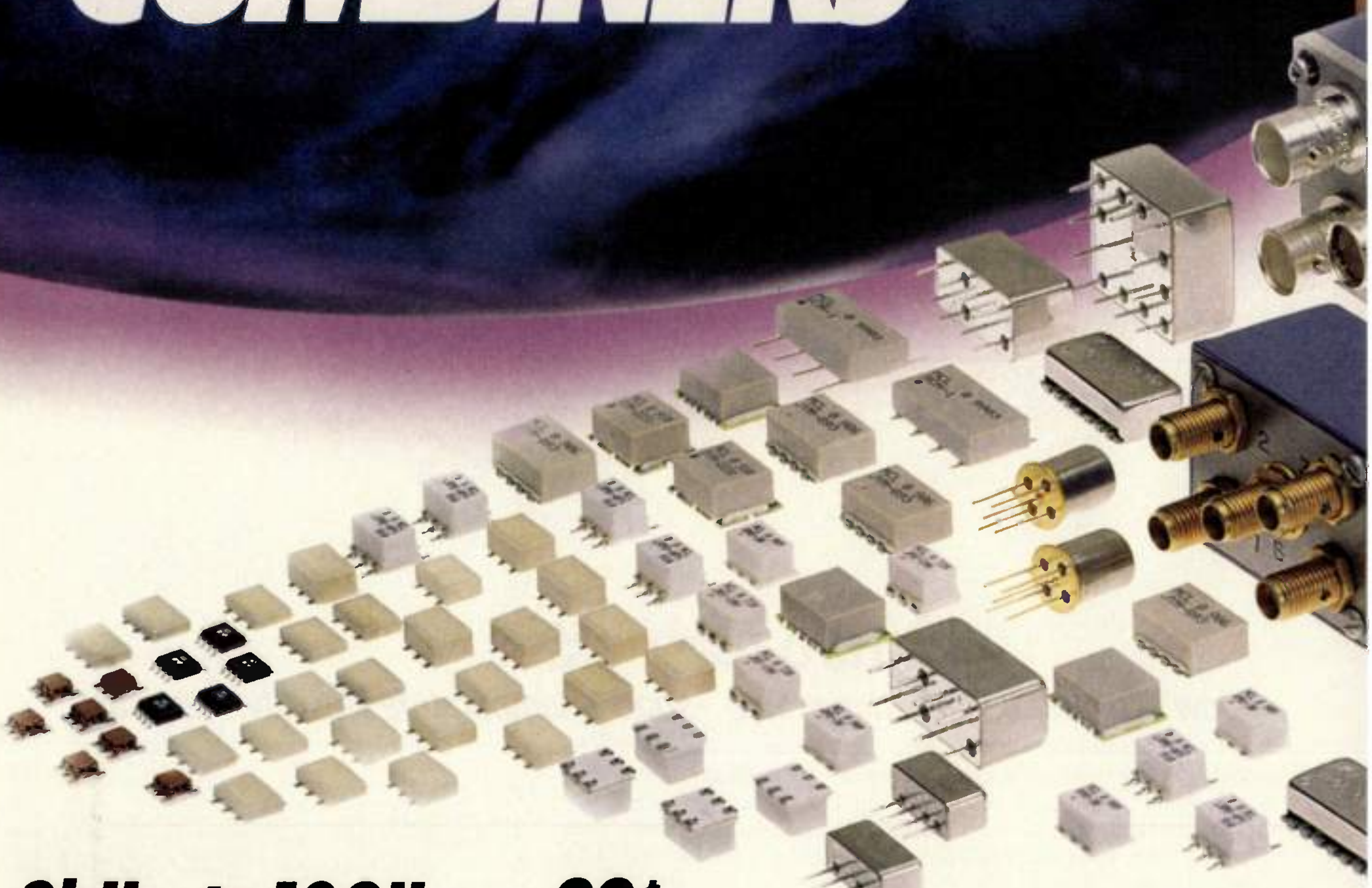

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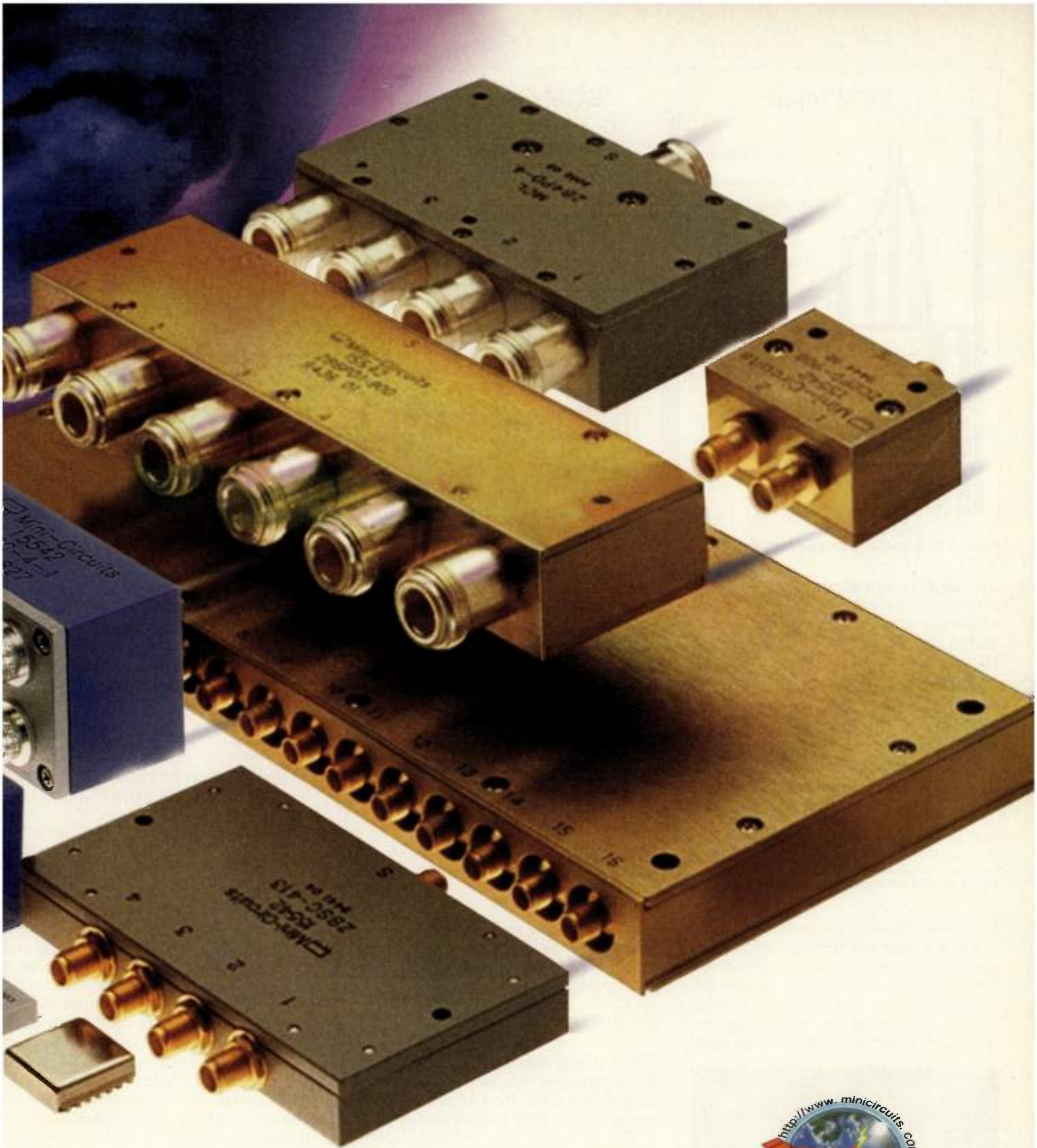

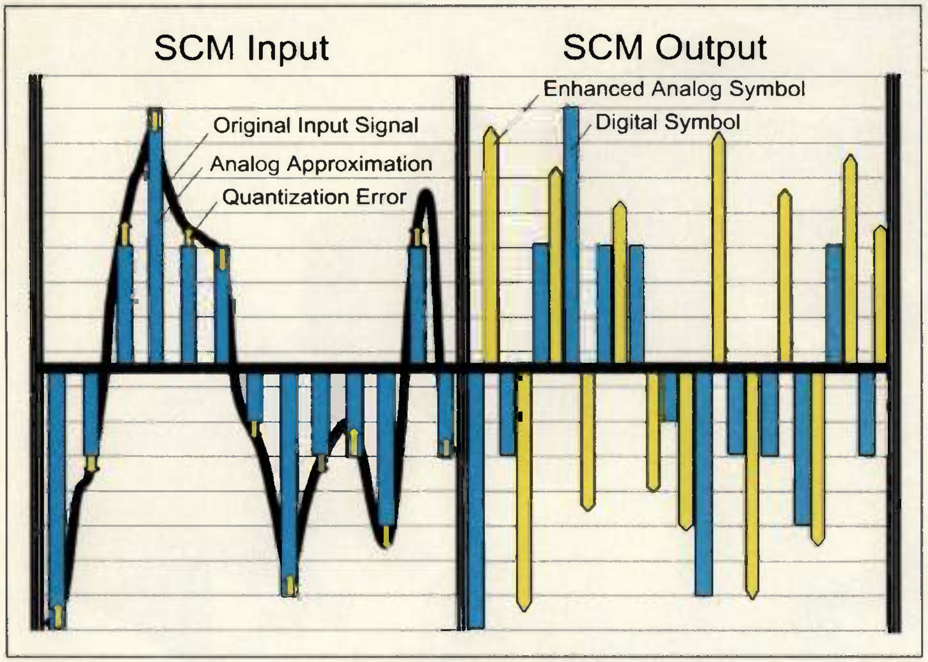

Figure 1. SCM operation on a payload waveform.

This is because FM suffers from a threshold phenomenon in which the performance degrades drastically with channel SNR, several decibels (dB) above the PCM[1]. Shannon, who studied this subject, introduced rate distortion theory, from which the performance of an ideal communications system could be derived. Such system performance will depend on the bandwidth expansion factor b, which is the ratio between channel bandwidth and source information bandwidth.

The numbers

Shannon has derived the following equation[2]:

$$SNRd = \left(1 + \frac{SNRc}{b}\right)^{b} - 1$$

PCM can meet this SNR curve in one point, but the quantization noise remains unchanged as channel SNR exceeds the threshold value. A similar expression can be derived for SCM, as shown in Appendix A. SCM performance is as follows:

$$SNRd = \frac{SNRc}{b}\left(1 + \frac{SNRc}{b}\right)^{b-1}$$

This expression is plotted for $b=2$ and $b=4$ in Figure 4 for both SCM and the Shannon Bound.

In the threshold corner of SCM, the channel SNR is a fraction of a decibel from the ideal Shannon Bound. If the SCM curve is to follow the Shannon Bound for every channel SNR, SCM must adapt its bit rate and error-coding scheme for each SNR. A practical implementation of SCM is more likely to be optimized for one threshold point only: for example, 23 dB output SNR in Figure 4. The straight line "single point SCM" depicts the resulting performance. The advantage of SCM is now becoming apparent; while optimum PCM performs as well at the threshold channel SNR, SCM continues to improve as channel SNR improves. Any practical communications link operates most of the time at a significant margin above threshold. PCM would remain in threshold performance regardless of SNR, while SCM improves.

Because SCM is essentially an ideal modulation scheme for analog signals, it is difficult to come up with a significantly better scheme. Therefore, SCM is likely to remain useful for the foreseeable future, making it a preferred

the adaptive equalization. An example of a transmission constellation in which the original signal is an unmodulated carrier and the D symbols are 16-QAM is shown in Figure 3. (Note: An explanation of Figure 3 appears in Appendix B.)

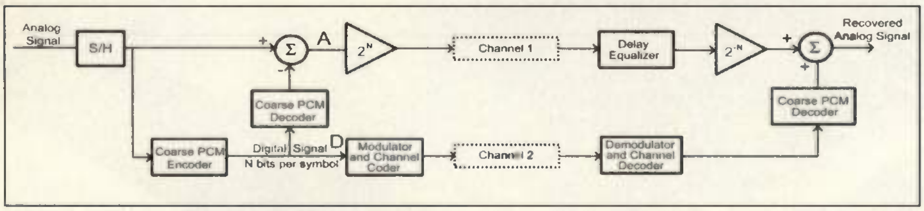

Figure 2: SCM transmission and reception processes.

Performance comparison

Before explaining why SCM is a nearly ideal analog communications method, it is necessary to define the ideal reference and compare it with existing alternatives. A communications link designer faced with an additive white Gaussian noise channel of bandwidth B and a limited signal-to-noise ratio (SNR) might choose a digital link as a first choice. The analog samples are converted to digital with a resolution of M bits per symbol. It is well-known that, by using an ideal digital error-correction coding technique, the channel can pass the information error-free at a bit rate that is called the channel capacity C, where:

$$C = B \log_2 (1+\text{SNR})$$

If the analog signal-sampling rate is R, the number of bits per symbol cannot exceed $M = C \bullet R$; thus M is limited and quantization noise is unavoidable.

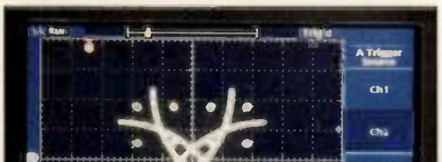

Figure 3. SCM constellation with unmodulated carrier input. (An explanation of this figure appears in Appendix B.)

The designer may consider analog modulation, such as frequency modulation FM. Frequency modulation is known to improve the output signal-to-noise ratio, referred to below as the signal-to-noise ratio destination (SNRd), compared with the signal-to-noise ratio channel (SNRc). FM accomplishes this advantage at the expense of bandwidth increase. The designer will soon find that FM is inferior to PCM at the minimum-channel SNR.

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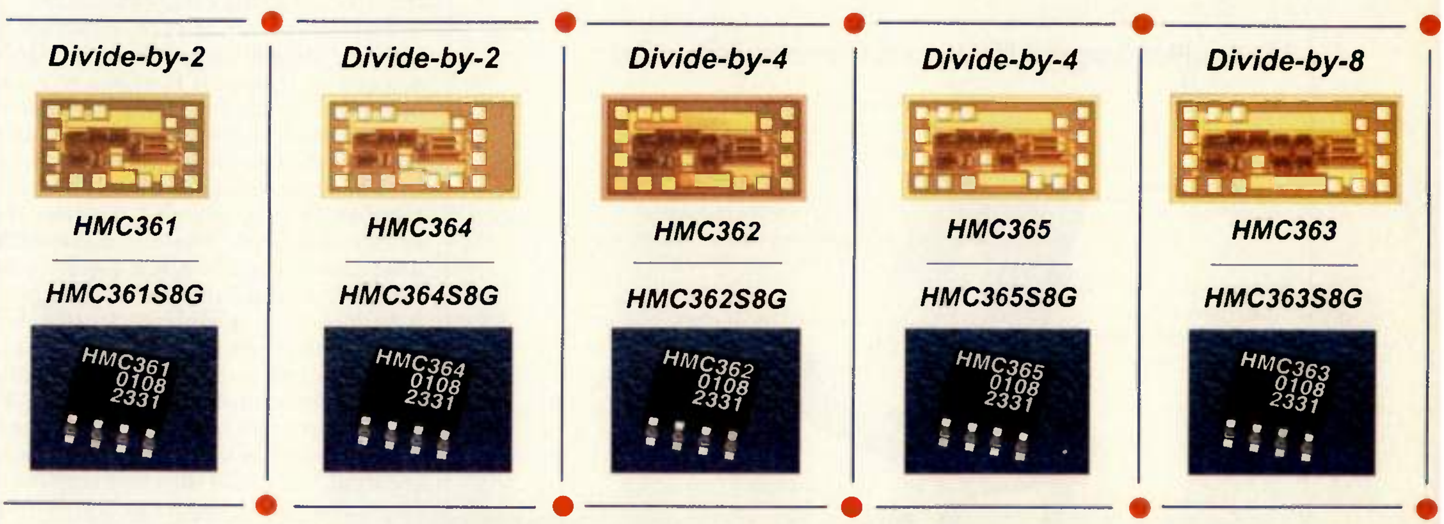
		DC - 12.5	-145 dBc/Hz	HMC364S8G
÷ 4	High Efficiency Med. Output Power	DC - 12.0	-149 dBc/Hz	HMC362
		DC - 12.0	-149 dBc/Hz	HMC362S8G
÷ 4	High Frequency High Output Power	DC - 13.0	-151 dBc/Hz	HMC365
		DC - 12.5	-151 dBc/Hz	HMC365S8G
÷ 8	High Efficiency Med. Output Power	DC - 12.0	-153 dBc/Hz	HMC363
		DC - 12.0	-153 dBc/Hz	HMC363S8G



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choice for emerging applications. The obvious question is: Who needs an analog modulation scheme? Surprisingly, most SCM applications are related to digital communications, and this paradox will be resolved in the next section.

Wide range of applications

An SCM-based communications link is basically a transparent, band-limited analog pipe with near-ideal performance in noisy channels. Every analog signal could potentially use SCM because it can outperform other existing modulation schemes. However, SCM has a compelling advantage for digital communications applications as well. For example, SCM can pass digital information by acting as a repeater of a digital channel. This application provides a wireless extension of cable modem digital information.

As illustrated in Figure 5, a cable modem termination system (CMTS) transmits a 42 Mb/s 256-QAM signal in a 6 MHz cable channel shared among the cable modems located at the subscribers' premises. The return upstream path from the cable modems is a 10 Mb/s 16-SQAM signal in a 3.2 MHz cable channel. The signals are carried by a combination of fiber and coax referred to as a hybrid fiber/coax (HFC) network. The fiber delivers a large amount of bandwidth over long distances with strong noise immunity. Coax cables distribute the signal between the fiber and each subscriber.

To reach a business park located beyond the reach of the existing HFC network, the cable operator installs an SCM-based point-to-multipoint wireless access system at any point on the HFC network that has line-of-sight to the business park. All customers located at a particular site share the SCM radio located at that site. The subscribers simply use low-cost cable modems that connect to the SCM radio via a shared coax cable. The wireless subscribers can even share the same cable channels with purely wired subscribers because the wireless link is transparent to the cable equipment.

The significance of SCM in this application is its ability to take a 256-QAM signal and transport it over a wireless link suitable only for a lower modulation scheme, such as 16-QAM. SCM provides significant additional noise immunity as is depicted in Figure 4 because it uses bandwidth expansion to improve the destination SNR. There is a non-SCM alternative: the 256-QAM signal could first be demodulated back to the original data bits, then modulated as 16-QAM transmitted over the wireless link demodulated at the destination, and finally remodulated using 256-QAM. This alternative would be much more costly, given the amount of processing required. It would also add significant latency to the information transported because an efficient channel must perform the error correction of the original signal before transmitting it over the wireless link. Furthermore, because SCM provides a transparent link that is not sensitive to protocol evolution or variations, it is more future-proof and versatile than specific digital standards.

A second example is an application from another discipline: digital audio

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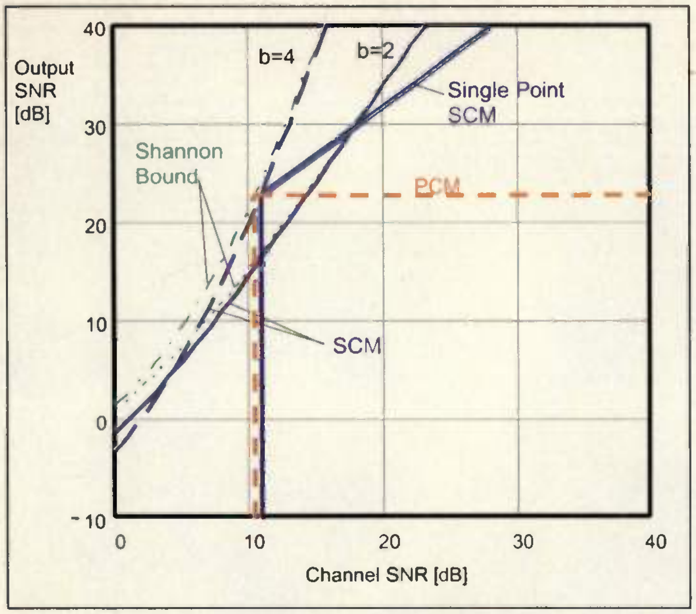


Figure 4. SNR performance of an ideal analog communications channel, PCM and SCM.

recording. A new-generation audio CD could include a digital track identical to and compatible with existing CD tracks, and in addition, have an analog track to provide the enhanced quantization error. Such an analog track would provide audio performance that depends on the quality of the recording and of the disc player. The most discriminating audiophile could use the more sophisticated player for true analog reproduction, while the less discriminating users would enjoy the low-cost CD technology in its current format.

The call

SCM is a versatile, hybrid analog-digital modulation scheme for transmitting analog signals to provide a repeater function for digital signals. The transparent nature of SCM allows it to relay signals without depending on transmission protocols and modulation formats. Such transparent links provide significant cost reduction because they use mass-produced end equipment such as cable modems. The near-ideal performance of SCM makes it a good choice for low-cost transparent links.

References

1. B. P. Lathi, *Modern Digital and Analog Communication Systems*, Third Edition, 1998, page 720.
2. Ibid., *eqn. 15.70a*, page 718.

Appendix A: SNR performance of an ideal SCM channel

Without loss of generality, it can be assumed that the source signal is sampled at the Nyquist rate and is transmitted as a discrete symbol. If b=1, the modulation is simply pulse amplitude modulation (PAM). It is well-known that this link meets the theoretical limit of performance set by the Shannon capacity theorem[i]:

$$SNRd = \left(1 + \frac{SNRc}{b}\right)^b - 1 \qquad \text{(Eqn. 1)}$$

SNRd is the output (destination) SNR and *SNRc* is the channel SNR. Clearly for *b=1*, there is no gain, and thus PAM meets the ideal performance. For *b>1* there is a significant performance gain but how can it be achieved?

Suppose that a band-limitec Gaussian channel is used for binary transmission and that by channel cod- ing, the channel capacity is achievec digitally. Such coding is not practical but today's codes closely approach the theoretical limit. The capacity of a Gaussian channel is:

$$C=B\log_2(1+\text{SNRc}) \qquad \text{(Eqn. 2)}$$

C is the capacity in bits per second o error-free transmission over a channe of width *B*.

Next, consider the following mixec link:

Total bandwidth is *bB*.

It has an analog portion of band width *B*.

It has a digital portion of bandwidtl *(b-1)B*.

It maintains the same peak power a the original PAM signal. Thus, with th bandwidth increase, the channel SNR i decreased by the factor *b* (i.e., noise powe increases by the factor *b*). The availabl capacity of the digital channel is:

$$C = (b-1)B\log_2\left(1 + \frac{SNRc}{b}\right) \qquad \text{(Eqn. 3}$$

These bits are used for qualifying th analog symbols in the analog portion. A there are *2B* symbols/sec and *C* bits/sec there are $M = C/2B$ bits per analo symbol. Now the analog signal in th range [-a, a] is not transmitted in full Instead, it is divided into 2^M equal seg ments. We assume that M is an intege however, we treat M as a continuou variable in the following analysis. Fo each symbol, only one of the segment contains the analog information. Thi segment is magnified to the range [-a,a i.e., it is amplified by a factor of 2^M an transmitted with PAM modulation. Th M bits associated with it are transmit ted in the coded digital channel an recovered. The receiver will then tak the analog signal, shrink it back by 2 and translate it to the original level. Th signal-to-noise increase is the square o the magnification, thus it equals 2^{2M}

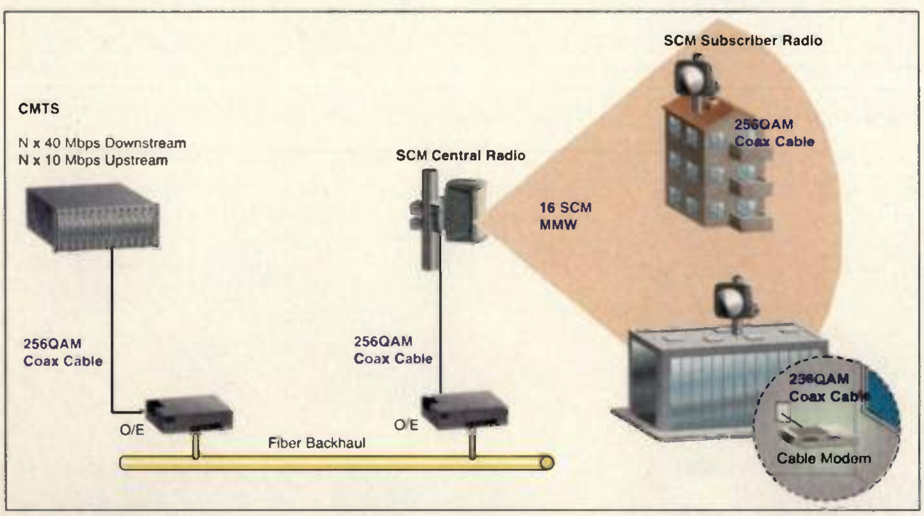
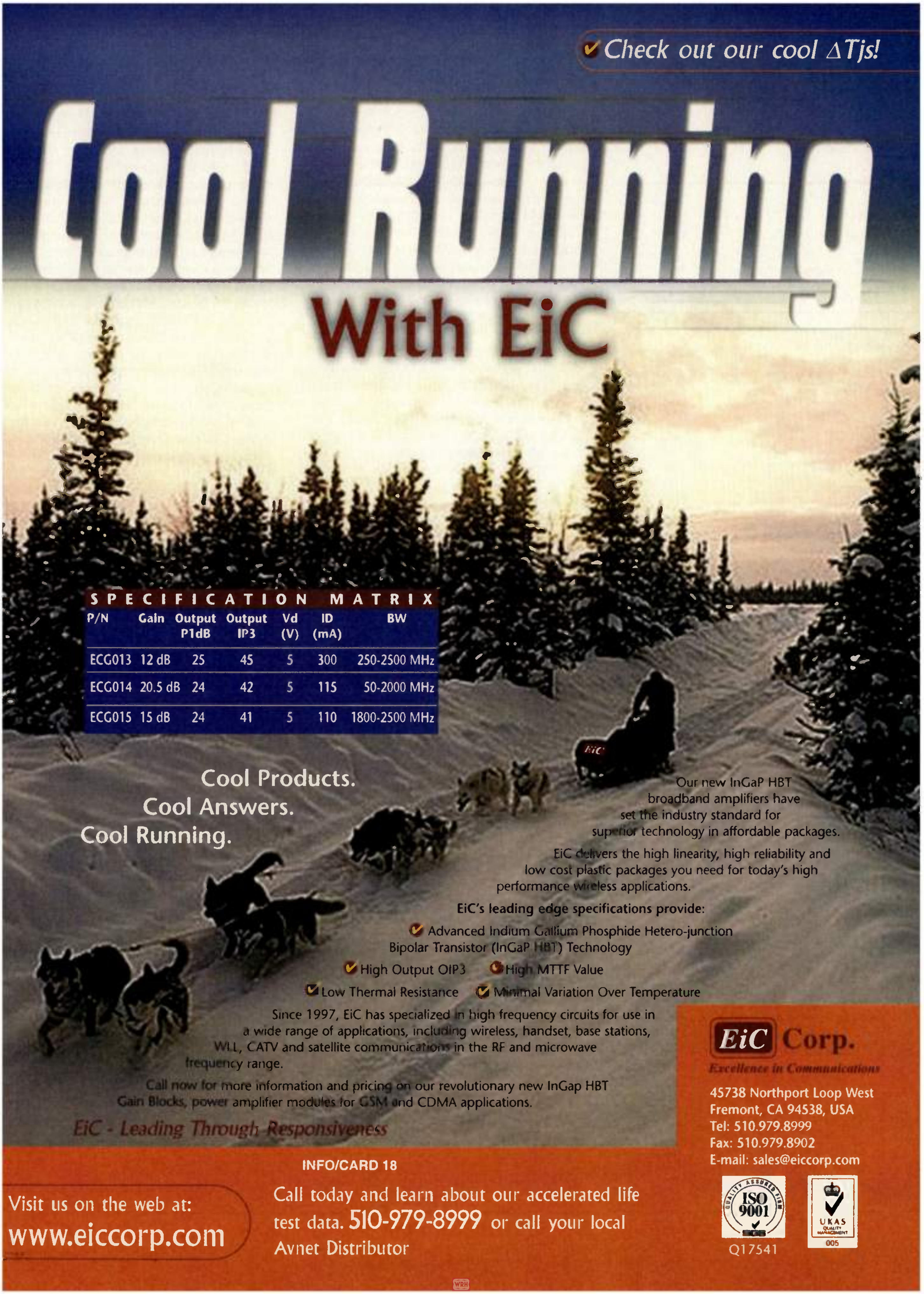


Figure 5.



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$2^{C/B}$. Therefore:

$$SNRd = \left(\frac{SNRc}{b}\right)2^{\frac{C}{B}} \quad \text{(Eqn.4)}$$

Substituting Eqn. 3 for C in Eqn. 4:

$$SNRd = \left(\frac{SNRc}{b}\right)2^{(b-1)B\log_2\left(1+\frac{SNRc}{b}\right)}$$

And simplifying:

$$SNRd = \frac{SNRc}{b}\left(1+\frac{SNRc}{b}\right)^{b-1}$$

Reference

1. B. P. Lathi, p. 711.

Appendix B: SCM constellation explanation

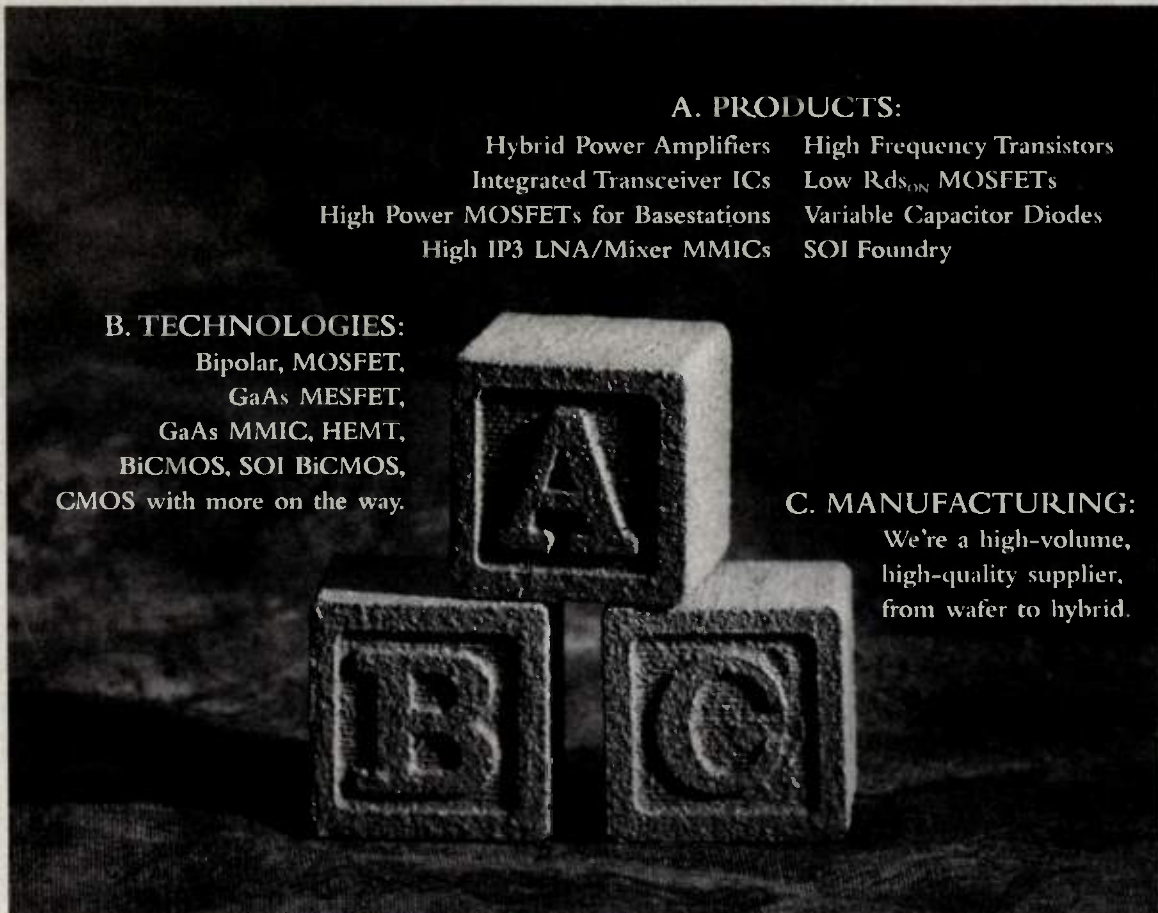
How does SCM cause this unusual constellation diagram on page 54?

This diagram represents a specific SCM mode, AAD, wherein the bandwidth expansion is b=1.5, and the input is an unmodulated carrier within the system pass band, although not necessarily at center frequency. I is a 16-QAM symbol carrying four bits, and the SCM conversion uses only a single bit per A dimension. The image is a superposition of many sampling points, each representing a single A or D point. The collection of all D points creates the familiar 16-QAM constellation. When depicted as I vs. Q, the original analog carrier would produce a circle, a familiar Lissageau waveform. However, SCM, in this example of N=1 bits per dimension, divides such a circle into four quadrants. The analog difference signal is a quarter of a circle. The four quarters are superimposed in the image, thus forming the symmetrical waveform. By changing the amplitude of the carrier, the quarter circles change their radii, creating different images.

RF

About the author

Eli Pasternak is co-founder, senior vice president and chief technology officer of BridgeWave Communications. Prior to BridgeWave, he served as CTO and chief scientist at Netro Corporation, which he co-founded in 1994. In his role at Netro, he was responsible for most of the IP related to both AirMAN and AirStar products and wrote four fundamental core technology patents in the area of wireless ATM and its associated MAC layer. Prior to Netro, Pasternak was a co-founder of Telestream and a technology consultant in a broad range of technical disciplines, including medical electronics, wireless, microwave products, communications products, system architecture, and food processing. He also served as a consultant in two related areas, as the director of engineering during the first year of P-COM's operations, and as a system architect for Plantronics' wireless infrared headset products. He received B.S.E.E. and M.S.E.E. degrees from the Technion in Israel. He can be contacted at: **www.bridgewave.com.** A demonstration of BridgeWave's technology can be viewed at this site as well.



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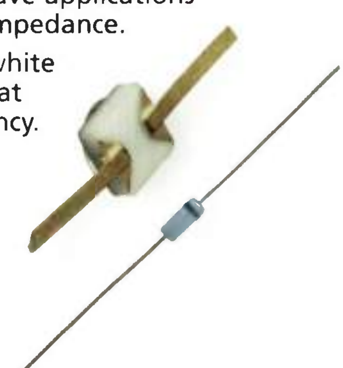

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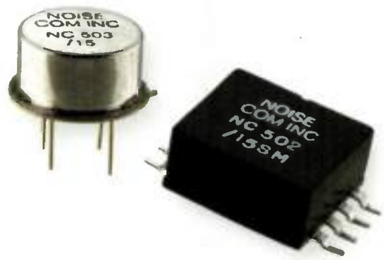

NC500 SERIES Standard Models		
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NC502/15	0.2 MHz – 1 GHz	31 dB
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NC506/15	0.2 MHz – 5 GHz	31 dB
NC513/15	0.2 MHz – 2 GHz	51 dB

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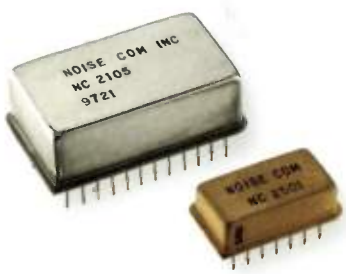

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UFX7218	2 GHz – 18 GHz	-20 dBm
UFX7911	5 MHz – 1 GHz	+30 dBm

Broadband tuners for modern systems

Designing tight performance parameters for today's high-performance broadband tuners requires careful attention to key design details.

By John Norsworthy

More than 300 million broadband tuners are produced globally each year for operation in the TV band. These tuners are integrated into a variety of consumer electronics, ranging from familiar household standards such as televisions and VCRs to newer, more complex devices that include cable-set top boxes, cable modems, cable telephony systems, WebTVs, PC/TVs and the various implementations of digital television.

Functioning as the RF broadband "gateway," the basic function of a tuner in these devices is to receive all available channels in the input bandwidth. It is also required to select the desired channel and reject all others, and to translate the desired channel to a standard intermediate frequency (IF). These traditional tuners operate over a frequency range of 54 to 862 MHz, taking into consideration those frequencies used by broadcast television and cable operators.

Tuners that are enabling products to offer PC, television, and Internet functionality have different performance requirements than the traditional TV tuner. As applications become more sophisticated, tuners with higher performance are required. New concerns brought on by the latest tuner applications include smaller form factors, high reliability, compliance with standards (data over cable system interface specification [DOCSIS], OpenCable, PacketCable), and ease of manufacture.

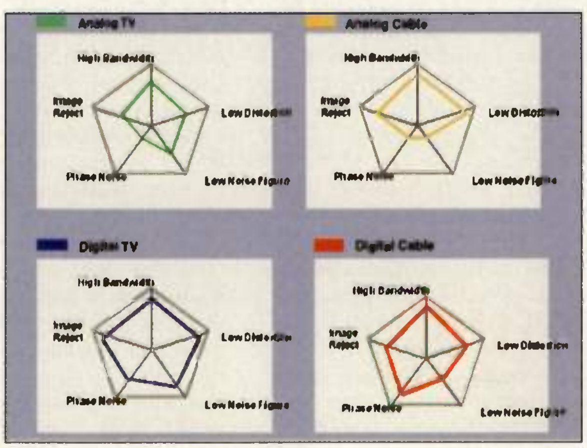


Figure 1. Tuner performance parameters across multiple applications.

For tuners operating in the TV band, performance can be summarized by five parameters: dynamic range, phase noise, noise figure, spurious responses and image rejection. The relative importance of each parameter and its specification range vary by application. For the designer of the latest communications solutions, understanding these tuner performance parameters and the available configuration options is essential to successful, high-performance, low-cost design.

Dynamic range

Dynamic range defines a tuner's ability to handle signals of varying strengths. For example, a TV tuner must be able to receive a weak signal from a distant transmitter while simultaneously handling a strong signal if the transmitter is close. Thus, a tuner must incorporate amplification that varies from strong amplification to strong attenuation. This degree of gain variation is the dynamic range of the receiver.

Different applications require different degrees of gain variation. For example, a tuner for digital terrestrial applications may be required to receive a signal as weak as –83 dBm, and simultaneously be able to handle a signal as strong as –15 dBm. The difference between these two signal levels is 68 dB, or 6.8 orders of magnitude, and that is the dynamic range of the receiver.

In cable applications, a much higher degree of uniformity exists among the channels because they all originate from a common head end and are transmitted at similar power levels. There is, however, tilt in the spectrum, which derives from frequency-dependent losses in the cabling and the methods used to compensate for it. Also, depending on where a home connects to a cable, with respect to the most recent signal amplification, the signal may be either relatively strong or relatively weak. In general, the cable network is regulated so that the minimum analog video carrier signal strength seen is 0 dB/mV and the strongest signal seen is +20 dB/mV. For digitally modulated signals, DOCSIS provides for signal levels that are –15 to +15 dB/mV. Typically, the dynamic range for a cable modem would be 30 dB for an analog cable receiver, and 20 dB and 35 dB for a tuner designed to receive both analog and digital signals.

In the above discussion, two sets of units are used for signal levels: dBm, popular in the terrestrial community; and dB/mV, popular in the cable community. These units are related by 48.75, such that power in dB/mV = power in dBm + 49 (approximately). Also, it is typical in a receiver to divide the gain or amplification over two stages: intermediate frequency and RF. The RF amplifier is in the front end of the tuner, while the IF amplifier is in the IF stage immediately following the tuner. Common practice is to operate the RF amplifier at full gain until the desired input signal level is 0 dB/mV, followed by reducing the gain on a decibel-for-decibel basis as the signal level increases. In conjunction, the IF amplifier operates at full gain for the weakest signals and is

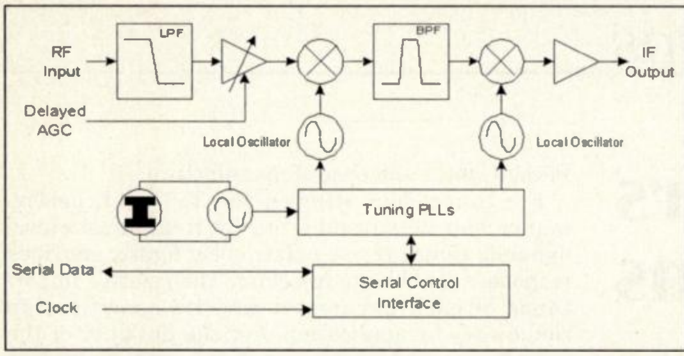


Figure 2. A double-conversion tuner with two LOs to control on-chip mixers.

decreased until the input signal level reaches 0 dB/mV, where it is set to the minimum gain. Consequently, the dynamic range adjustments for the tuner, excluding the IF amplifier, would be 20 dB for an analog cable application, 15 dB for a digital cable application, 20 dB for a combined analog/digital application, and 34 dB for a digital terrestrial application.

Quality of channel reception

In a communications system, wired or wireless, the tuner is responsible for receiving all available channels in the input bandwidth. It is also responsible for selecting a desired channel and rejecting all others, and translating the desired channel to a standard IF. Implicit in this process is retaining as much of the original signal's fidelity as possible, i.e., not adding appreciable noise and distortion. If the remaining channels are not effectively rejected, the system will experience interference. In an analog TV system this could result in unacceptable video (and audio) quality. In a digital TV system the picture could completely disappear. In a data system, interference can result in signal dropouts and reduced data transmission rates.

The quality of channel reception is greatly affected by the phase noise, noise figure, distortion, and image rejection experienced by the system. And, the relative importance of these qualities varies by application. For instance, analog TV tuners need to perform adequately for low distortion and low noise figure, while their digital counterparts must offer superior image rejection and low phase noise (see Figure 1). Analog cable requires good image rejection and low distortion, but it can tolerate phase noise and a substantial noise figure. Digital cable, on the other hand, requires a reasonable noise figure, but significantly better phase-noise performance than an analog system.

Phase noise

Because the channels are broadcast at high frequencies, it is necessary to downconvert these channels to an IF that is usable by decoding circuitry downstream from the tuner.

All tuners use mixers that are controlled by local oscillators (LO). Figure 2, for example, shows a double-conversion tuner that uses two LOs.

Although it is intended that the LOs oscillate at one pure frequency, real-world oscillators have a spectrum that, while dominated by the desired frequency, also contains undesired frequencies. Phase noise is the relative power of the undesired frequencies with respect to the one desired frequency, and is thus measured in decibels with respect to the carrier (dBc). Carrier in this case represents the desired frequency.

Phase noise has become important recently because of its impact on the reception of digitally modulated signals. One popular digital modulation technique is quadrature amplitude modulation (QAM), where the signal is divided into symbols. A symbol may be thought of as a sampling point of the signal for its amplitude and phase. For purposes of spectral efficiency, the resolution of the amplitude and phase limits the number of bits per symbol. Phase noise is the primary limiter of phase resolution in the signal and, consequently, achieves a large value of bits per Hertz, or spectral density.

The effects of phase noise on signal reception vary by application. Analog television, for instance, experiences little effect from phase noise unless it is excessive. Digital applications, such as cable modems, will not function unless phase noise is below a certain threshold. Tuners have improved recently as designers pay more attention to oscillator design and specifically target low phase noise as a requirement.

Noise figure

All tuners add some noise to the received signal. The noise figure is the measure of the noise added to a signal by passing it through a tuner. More specifically, the input signal has a signal-to-noise ratio (SNR), as does the signal output by the tuner. The difference between input SNR and output SNR is the noise figure. Noise figure is important in terrestrial applications because it defines the minimum detectable signal, or stated differently, the weakest signal receivable. It is important in receiving digitally modulated signals so that the SNR presented to downstream devices is maximized. For terrestrial applications, a 7 dB noise figure is required, while for cable applications, 10 dB usually suffices.

The nature of noise figure is that the front-end RF amplifier of the tuner dominates it. This amplifier is generally referred to as a low-noise amplifier (LNA). If it is low-noise and high-gain, it overcomes noisy downstream components such as mixers and filters. As process technologies advance, tuner noise figures should improve over time.

Spurious products

A spurious product is an undesired signal present in the output of the tuner. The source of this undesired signal may be distortion in the tuner's active circuits or an unwanted coupling from one circuit to another. If the tuner is designed well (the coupling issues are suitably suppressed), then the primary source of the spurious products is distortion.

In tuners, the distortion products that are most frequently discussed are intermodulation distortion (IMD) and cross modulation. Distortion products are generally introduced when handling large signals. For example, in terrestrial broadcast television, a typical problem is the receiving of a weak signal in the presence of a strong one. In cable systems, the problem relates to receiving one out of many relatively strong signals, where the other signals add up to create a large RF envelope.

Distortion is a bigger problem with analog in terms of its source and the visible effect. Analog TV signals are a problem because the bulk of the signal energy is concentrated near the carrier, resulting in strong peaks. By contrast, the energy in a digitally modulated signal is spread smoothly across the channel. Furthermore, the human visual perception system is acutely tuned to detecting patterns in video which is the manifestation of distortion in analog video. In digital systems, the distortion is not perceptible until the signal becomes so affected that the data stream is compromised and video is lost. For analog systems it is desirable to have all spurious products at least 57 dB below the carrier, or −57 dBc. For digital systems there is less agreement on a specific level, but −50 dBc is frequently referred to as a specification.

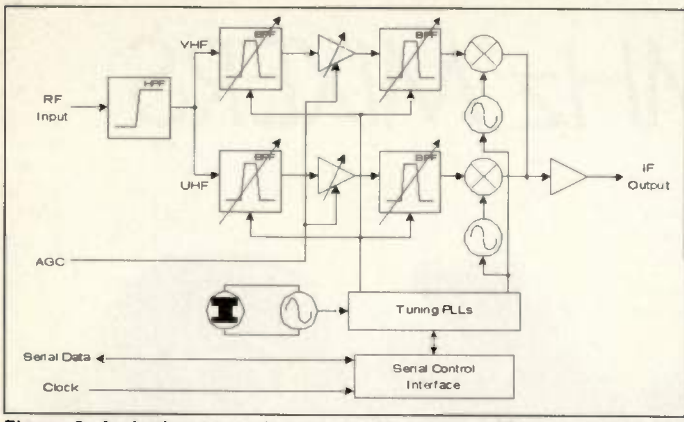


Figure 3. A single-conversion tuner with tracking filters to isolate the frequency of interest.

Image rejection

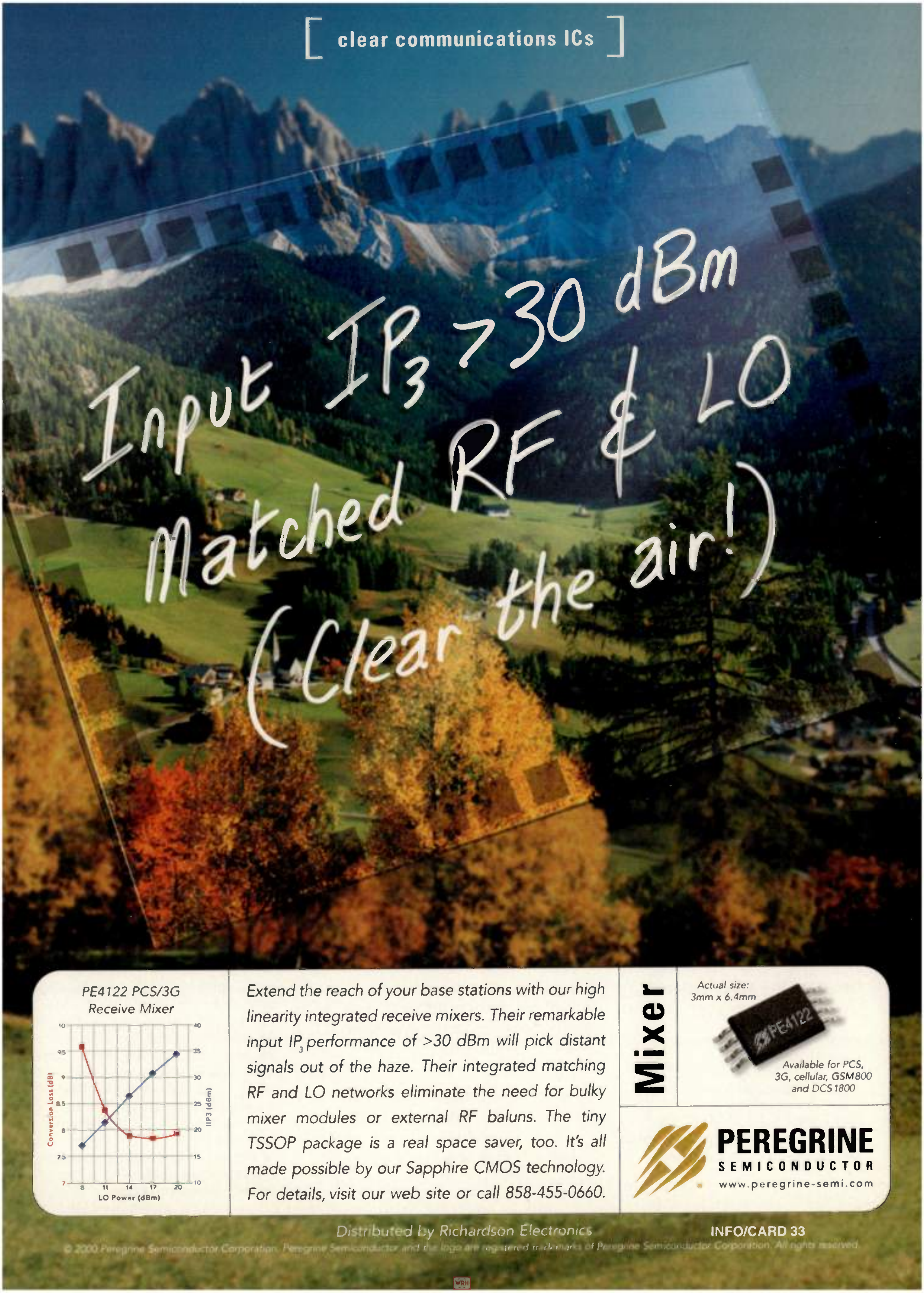
Tuning involves translating signals in frequency. Suppose that a desired channel is translated to an IF. Then a channel two times that of the IF below the desired channel will be translated to –IF. Negative frequencies differ from positive ones only in the phase of their components, and therefore, interfere with the desired channel at the positive IF. This interfering channel is known as the image channel and must be rejected to a large degree for proper reception. (In broadband systems, the desired channel is translated to –IF, such that the image channel is twice that of the IF frequency above the desired channel.)

Image rejection can be addressed with filters and/or with image-reject mixers. In a single-conversion tuner, a notch filter is used to reject the image channel prior to frequency translation. The performance of such a filter is limited to 50 to 60 dB in the UHF component of the TV band (470 MHz and up). Better performance is possible with dual-conversion tuners, where the first IF filter can suppress the image channel by an arbitrary amount, depending on the cost (and thereby performance) of the filter. For cost-effective, dual-conversion tuning systems, the preferred approach is to use a reasonably priced surface-acoustic wave (SAW) filter at first IF to achieve around 40 to 50 dB by itself, then to complement that with a specialized mixer called an image-reject mixer. Such a mixer can achieve an additional 35 to 40 dB of suppression. The combination allows for consistent image rejection in the range of better than 70 dB.

Image rejection requirements differ by application. In traditional broadcast television in the United States, the Federal Communications Commission (FCC) regulates the spectrum so that a broadcaster in a given locale is guaranteed that no other broadcaster will be transmitting at the image channel. New spectrum allocations for adding digital channels are no longer eliminating this hazard, however, so it will be possible that a digital broadcaster will have a strong analog broadcaster at his image channel. In that case, a tuner must have around 80 dB of image rejection for proper reception. In cable, the image channel nearly always exists and is at a power level similar to that of the desired channel. Consequently, 60 dB of image rejection is good for analog signals, and 50 dB is good for digitally modulated signals.

Tuner configurations

Designers of tuners suited for particular applications must balance any trade-


[clear communications ICs]
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PE4122 PCS/3G
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Mixer
Actual size:
3mm x 6.4mm
PE4122
Available for PCS, 3G, cellular, GSM800 and DCS1800
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offs among the key specifications of dynamic range, phase noise, noise figure, spurious products, and image rejection. Designers can choose from a single-conversion or double-conversion tuner design. Currently, tuners are available in subsystems, as modules or as integrated single-chip solutions.

Single-conversion tuners (see Figure 3) select the signal of interest by using a set of variable-frequency filters, called tracking filters because the center frequency of their bandpass characteristic tracks the center of the desired channel. Multiple filters are typically required because of the broad bandwidth of most systems. Because of this filtering approach, it is difficult to design a single-conversion tuner with high selectivity (good rejection of other channels.) Furthermore, these tracking filters require hand tuning during their manufacture to achieve the required filter characteristics.

Until recently, dual-conversion tuners were always used for cable applications because of their superior selectivity, necessary for handling packed (vs. sparse) cable spectrums. Furthermore, dual conversion tuners presented a better broadband impedance match to the cable, so that standing waves are reduced. Also, the LOs in a dual-conversion tuner operate outside of the input spectrum, thus minimizing the potential for interference via leakage onto the cable network. Recently, some cable designs are using enhanced single-conversion tuners because of their low cost. This is likely to be a short-lived phenomenon because nearly all single-chip tuners use dual-conversion techniques.

Tomorrow

Regardless of the application or configuration, as communications designs become smaller, faster and more complex, they are requiring next-generation tuners to offer small form factors, high reliability and compliance with emerging standards, while supporting streamlined manufacturing. As tuners have become more specialized, they are extending far beyond their original functions as tuners for analog radio or television to ones that support a wealth of applications, including multimedia PC/TVs, cable-set top boxes, cable modems and digital televisions.

RF

About the author

John Norsworthy is founder and CTO of Microtune Location. He previously served as vice president and general manager of Cirrus Logic's consumer video products division. In 1991, he co-founded Pixel Semiconductor, which developed products and technology that enabled the integration of video into the PC graphics subsystem, and served as vice president of engineering. From 1986 to 1991, Norsworthy served as vice president of advanced technologies for Visual Information Technologies (VITec). He holds 12 U.S. patents and he has a B.S.E.E. from the University of Illinois at Urbana-Champaign. He can be contacted at: **www.microtune.com**

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110.59	1.0	855659
150.0	8.0	855678
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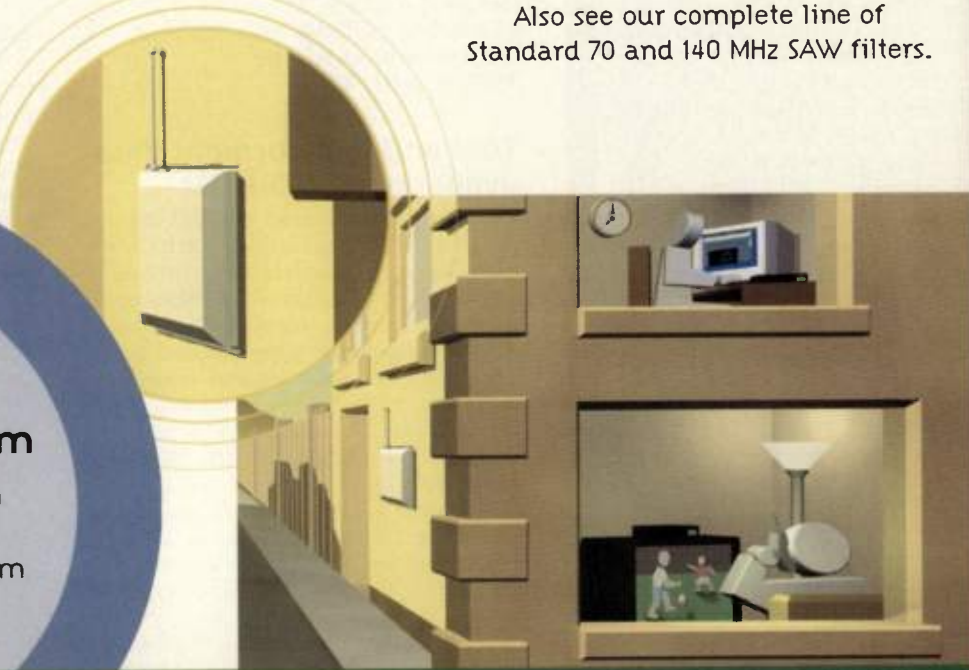

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On The Web

Updated site offers easier navigation

Space Electronics' overhauled Web site now offers easier navigation with a new search function that cross-references all three Electronic Components Group (ECG) Web sites. Also included on the updated site is a new ECG portal site and an updated, more user-friendly look. The site still offers comprehensive information on Space Electronics products, including a full product list, datasheets, information on testing services, related news and press releases, product- and technology-specific presentations, installation notes and product flows.
Space Electronics
www.spaceelectronics.com

PCB design site introduced

GE Electromaterials introduces a site for printed circuit board (PCB) design in a day. The online design center features tools for complex PCB development and integrates the life-cycle design process and a platform for supply-line management. It contains signal integrity critical-application tools and material models used for validation of interconnect performance for each step in the design life cycle process. Features of the site include a layer stack-up wizard and customized RF component libraries.
GE Electromaterials
www.GETEKconnect.com

Coaxial TEM resonator data sheet released

A two-page data sheet from Integrated Microwave describes coaxial resonators made with ceramic dielectric materials. The resonators are for use in compact frequency standards, filter elements and distributed inductive or capacitive circuit elements. Coaxial TEM resonators feature thermally stable ceramics in seven sizes (3, 4, 5, 6, 8, 10 and 12 mm) and four dielectric constants (8, 20, 36 and 98). Applications include dielectric resonating oscillators (DROs), voltage-controlled oscillators (VCOs), pagers, global positioning systems (GPS), cellular and wireless communications, bandpass/bandstop filters, narrowband/delay filters and EMI filtering.
Integrated Microwave
INFO/CARD 115

Data book covers IC line for communications

TDK Semiconductor announces its new 2001 data book/short-form catalog covering its recently introduced line of ICs for the communications and broadband markets. The volume includes technical data sheets on products released during the past eighteen months, including ICs for optical fiber network equipment, portable communications, embedded modem and digital set-top boxes. Products more than eighteen months old are highlighted in product briefs. A complete listing of worldwide sales and distribution locations is included. A CD-ROM version of the catalog will be available later this year.
TDK Semiconductor
INFO/CARD 116

Test and measurment data available on CD-ROM

LeCroy has released its 2001 catalog on both CD and hardcopy. The new catalog features complete descriptions of LeCroy's latest digital oscilloscopes, microwave communications analyzers, disk drive analyzers, probes and accessories, signal sources and waveform digitizers, as well as informative application briefs. Among the new products in the 2001 edition are the WavePro series of digital oscilloscopes, which offer faster sampling rates and longer memories in the 500 MHz to 2.0 GHz bandwidth range, and the Waverunner –2 DSOs, the second generation of LeCroy's Waverunner series. The DDA 260 disk-driver analyzer and MCA 1060 microwave communications analyzer are also featured.
LeCroy
INFO/CARD 117

Brochure covers switching components

EAO's new 180-page *Master Catalog 100* includes a 32-page color new-products section describing the latest developments in EAO switching components and integrated assemblies. New products highlighted in the updated catalog include multi-function pushbutton switches with as many as three indicator colors in one device, a new family of touch-sensitive switches, robust door pushbuttons and indicators for transportation applications, and hundreds of different LED indicators. Expanded general catalog listings now include miniature PCB key switches, vandal-resistant illuminated and non-illuminated keypads, slide and rotary changeover switches, and sealed toggles, pushbuttons and pushbuttons with LEDS that were part of EAO's Sechme Division. The catalog also includes heavy-duty Series 44 oil and watertight switching components and AS-i80 interface bus technology panels and modules.
EAO
INFO/CARD 118

Catalog features products for wireless applications

Giga-tronics has issued a short-form catalog featuring its microwave synthesizers, power meters, RF signal generators and VXI instruments. The catalog provides a quick guide to Giga-tronics' key product features, statistics and applications. A sensor selection chart is also included to help match products to exact measurement requirements.
Giga-tronics
INFO/CARD 119

Algorithm evaluation software released

NCT Group introduces ClearSpeech-PC, a Windows-based software program designed to facilitate the evaluation and testing of ClearSpeech algorithms. ClearSpeech-PC runs on Windows 95, 98 and NT and requires a PC with speakers and a default soundcard capable of playing 16-bit, 8 kHz, stereo, and Windows PCM .wav files. The ClearSpeech suite of algorithms enhances voice communications by removing the interfering noise or signal. The suite includes ClearSpeech Adaptive Speech Filter (ASF), which removes background stationary noise from speech; ClearSpeech Acoustic Echo Cancellation (AEC), which removes unwanted echo from full-duplex hands-free communications; and ClearSpeech Referenced Noise Filter (RNF), which removes noise from speech where a reference signal from the noise-producing source is available.

NCT Group
INFO/CARD 120

Parametric test software minimizes error

Keithley Instruments announces new software modules for its Automated Parametric Test (APT) systems. As add-ons for wafer characterization software, the Keithley Test Environment (KTE), Keithley Recipe Manager (KRM) and Probe Card Manager (PRM) are designed to elevate production program management and productivity. Together, these modules increase tester use by minimizing the potential for operator errors and providing historical correlation between test program changes, process control anomalies and product field failures. Keithley's KRM and PCM modules are for use on the company's APT systems running KTE v4.2 software and higher, which are used by semiconductor fabs to measure critical device characteristics. The new software modules help assure production flow efficiency through each step in a semiconductor process.

Keithley Instruments
INFO/CARD 121

Software package measures transmitter parameters

Anritsu offers a software package and hardware option for measuring 1xRTT that provides analysis of cdma2000 wireless handsets and devices. The cdma2000 1xRTT can be used with Anritsu's MT8801C or MT8802A radio communications analyzers to measure key transmitter parameters, and also supports fast forward power control and high-speed packet data testing. The Anritsu 1xRTT is capable of conducting 1xRTT test with call processing at data rates as high as 307.2 kbps with Service Option 32 (SO32) or Test Data Service Option (TDSO). In addition to supporting SO32 (TDSO), it also supports SO1, SO2, SO3, SO9, SO33, and SO55 service options. The 1xRTT hardware and software allow the MT8801C or MT8802A to simulate a 1xRTT base station and establish a call in RC1 through RC5 configurations.

Anritsu
INFO/CARD 122

Want to know the secret of success in the land of Linux and Nokia? It's APLAC. An industrial-strength simulation technology that combines the functionality of Spice with the utility of an advanced RF simulator. APLAC, and only APLAC, provides the accurate IC- and board-level models and precise methods to analyze non-linear circuit behavior demanded by top RF and analog designers. The only approach up to the complex design challenges ahead: 3G, Bluetooth, and beyond.

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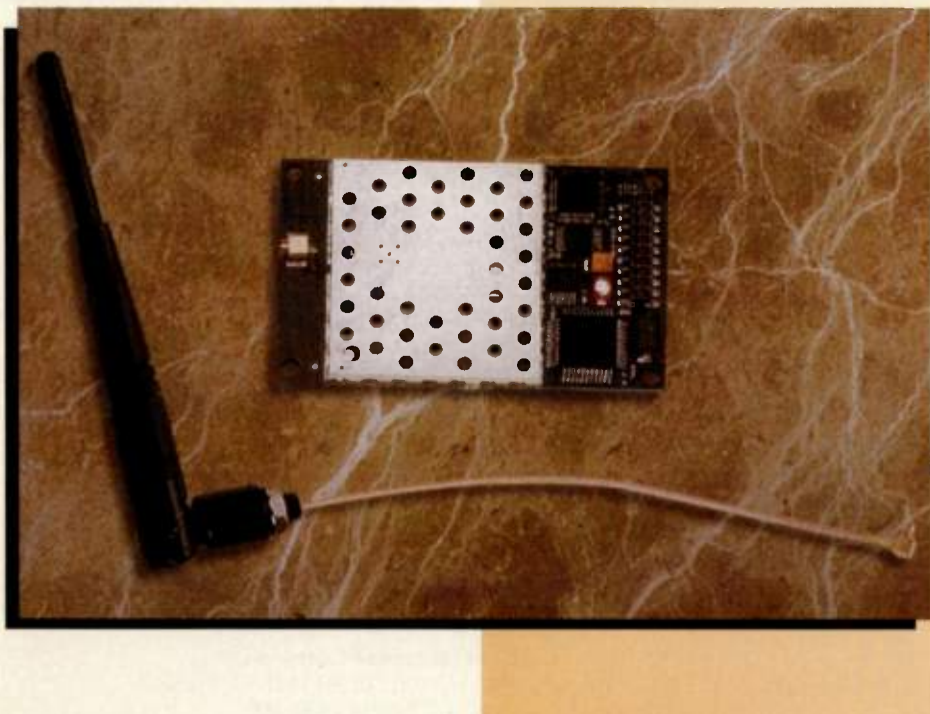
INFO/CARD 12

AEROCOMM LX 2400 SERIES

AeroComm introduces a line of 2.4 GHz frequency-hopping spread spectrum (FHSS) transceivers for OEM integration. The LX family offers manufacturers a variety of wireless capabilities, as well as comprehensive RF development tools and support for licensed and unlicensed Bluetooth and related technologies. Five versions of transceivers and two repeater models are available with varying range and network configurations. Manufacturers can choose short-range, low-power consumption versions for their battery-powered or piconet applications, and higher-power radios coupled with repeaters for miles of range. All LX transceivers have identical dimensions, connectors and software requirements, so modules are interchangeable for changing design needs. AeroComm LX transceivers are based on 2.4 GHz FHSS and are resistant to interference. With RF data rates up to 244 Kb/s, LX modules provide ample speed for most OEM applications. The product line's power output starts at 3mW for local uses (within 50 feet), and reaches 150mW for miles of range in outdoor applications or for large industrial facilities. For even longer distances, or when obstacles block the communication path, an LX repeater (or unlimited chains of repeaters) can be used to extend range. All LX transceivers are available with integral strip dipole antennas for applications not permitting external antennas. Radios with antenna connectors are also available for use with a variety of agency external antennas. AeroComm transceivers are designed for applicability to present technologies with a design track for future standards as well.

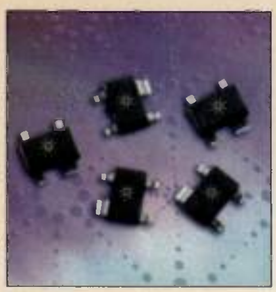
AeroComm
INFO/CARD 123



2.4 GHz spread spectrum transceivers for OEM applications

Single positive voltage supply PHEMT transistor

Agilent Technologies introduces a small-signal E-pHEMT device, the first in a family of high-gain, linear, low-noise transistors. The ATF-54143 can help simplify the design and reduce the cost of receivers in cellular/PCS base stations, multichannel multipoint distribution systems, and other applications in the 450 MHz to 6 GHz frequency range. The device offers performance, size and design advantages for applications such as tower-mounted amplifiers and front-end LNAs for base stations operating in GSM, E-GSM and W-CDMA systems at 900 MHz and 1.9 GHz. Operation is from a single +3 V supply at 60 mA current.



- Specifications at 2 GHz:
- 0.55 dB noise figure
- +36 dBm output intercept point
- 17.4 dB associated gain

Agilent Technologies
INFO/CARD 124

Low-power, broadband DACs

Analog Devices introduces the the AD975x family of CMOS-based, low-power DACs (digital-to-analog converters) for high-speed broadband communications applications. The AD9751, AD9753, and AD9755 TxDAC+ family of 10-, 12- and 14-bit converters are manufactured on a 0.35 micron CMOS that reduces chip power consumption by 40%, synthesizing input signal bandwidths up to 100 MHz and producing 300 MS/s operation. These DACs offer a noise floor of −150 dBm/Hz and over 65 dB SFDR. Using two LVCMOS-compatible data ports, the devices can interface to other standard CMOS-compatible logic devices to reduce cost and complexity. The series uses an on-chip 2x PLL/clock multiplier to eliminate the need for an external clock multiplier. The series is targeted at broadband systems where data rates rival those of optical networks, including LMDS, MMDS, satellite links and QAM systems.

Analog Devices
INFO/CARD 125

N-channel, enhancement-mode MOSFETS

Motorola announces a line of N-channel, enhancement-mode lateral MOSFET RF FETs. The devices are designed for broadband commercial and industrial communications applications at frequencies between 470 and 860 MHz. The MRF372 is a high-gain, broadband device designed for large-signal, common-source amplifier applications in 32-VDC transmitter equipment. The MRF 373A and 374A are functional replacements for their respective earlier versions (373 and 374) and provide improved performance over their predecessors.

Motorola
INFO/CARD 126

RAD hard 3 GHz PLL integrated circuits

Peregrine Semiconductor announces the PE9701 and PE9702 space-qualified 3 GHz phase-locked-loop (PLL) integrated circuits (ICs). The products are targeted for space and defense customers. Both the PE9701 and PE9702 (with and without charge pump respectively) are integer-N PLLs, featuring 10/11 dual modulus prescalers, phase comparators and counter values. The devices can be programmed through a serial or parallel interface, or can be directly hard-wired. They meet Class S-level screening and RAD hard requirements and can attain 100 Krad(Si) total dose tolerance. The devices show single-event upset (SEU) tolerances.

Peregrine Semiconductor
INFO/CARD 127

New LDMOS FET transistor

The L2721 is a 15 W, 11 dB gain 500 MHz, LDMOS FET. It is assembled in a two-lead surface-mount package. It is not internally matched and will operate from DC to above the frequency of test. It is intended to operate from 12 to 16 VDC power supplies. It is suitable for narrow-or wide-band circuits.

Polyfet RF Devices
INFO/CARD 128

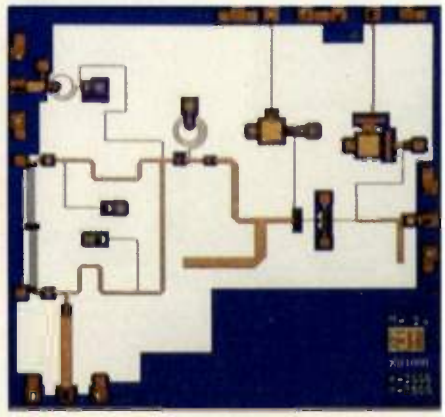
High-current 40 VDC, 2A Schottky diode

Zetex introduces a new SMT Schottky barrier diode that offers a high current capability of 2 A with a low forward voltage of (V_{Fmax}) of 500 mV. The voltage drop of the ZHCS2000 is typically 340 mV at 1A and 485 mV at 3 A. It is packaged in a SOT23-6 package that can dissipate 1.: W at an ambient temperature of 25°C in the space normally occupied by a standard S0T23 package. The device exhibits a recovery time (t_{rr}) of typically 5.5 ns and a reverse breakdown voltage of 40 VDC. Leakage current is typically 160 µA at a reverse voltage of 30 VDC. The measured capacitance for the part is quoted as 50pF at V_R of 25 VDC.

Zetex
INFO/CARD 129

Single-chip GaAs MMIC up-converter

Mimix Broadband announces an integrated GaAs MMIC up-converter on a single chip. The XU1000 is a single fundamental mixer followed by a single-stage amplifier. Using 0.15 micron gate length GaAs pseudomorphic high electron mobility transistor



(pHEMT) device model technology, the up-converter covers the 17 to 27 GHz frequency band. The device has a typical small signal gain of 0 dB with a typical third-order intercept point of +12 dBm across the band. The device features low DC power consumption and operates at +3.0 VDC. The device is used in conjunction with Mimix's XP1000 power amplifier, and the XB1000 gain block amplifier forms a complete millimeter-wave transmitter. The XU1000 enables the development of smaller size and lower cost designs and is suited for wireless communications applications such as millimeter-wave point-to-point radio, local multipoint distribution services (LMDS), SATCOM and VSAT applications.
Mimix Broadband
INFO/CARD 130

Power transistor for UMTS 2.1 to 2.2 GHz base stations

UltraRF announces the UPF2109 power transistor. The device can be used as a high-power wideband code-division multiple-access (CDMA) RF power amplifier operating in Class A or AB. It operates at 28 VDC with a broadband RF output power rating of 90 W. The discrete transistor features industry-standard packaging and an all-gold metal system for high reliability. The unit can directly replace the

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Integrated Microwave	115	74
J.W. Miller	161	97
Keithley Instruments	121	76
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KOA Speer Electronics	158	96
LeCroy	117	74
Lexra	143	92
Linx Technologies	152	95
M/A-COM	135	90
Micrel	153	96
Mimix Broadband	130	82
Morrow Technologies	146	93
Motorola	126	80
MTI-Milliren	140	92
NCT Group	120	76
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Peregrine Semiconductor	127	80
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Polyfet RF Devices	128	80
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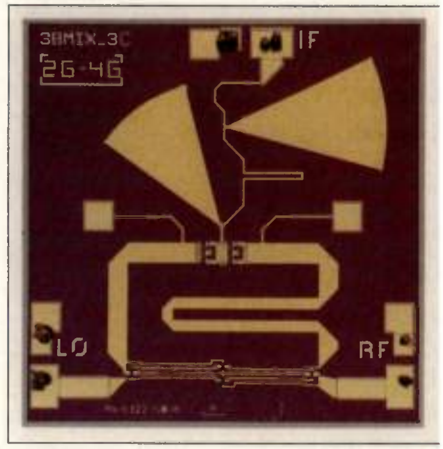
Seacor	155	96
SiGe	136	90
Stanford Microdevices	133	89
Taiyo Yuden	160	97
TDK Semiconductor	116	74
Toshiba America	151	95
Trompeter	149	94
Tru-Connector	150	94
Ultra RF	131	88
Vari-L	141	92
W.L. Gore	162	97
Zetex	129	80

Motorola MRF21090 and has compatible source and load impedance. It can be dropped into existing power amplifier designs with little or no circuit-matching changes. The device offers high linearity and is targeted for UMTS base station applications. The power transistor is internally matched to simplify amplifier design and facilitate repeatable manufacturing.
UltraRF
INFO/CARD 131



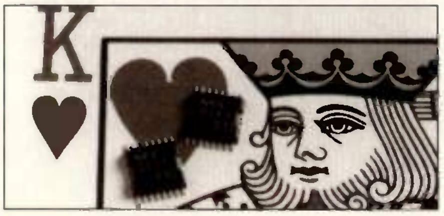
Microwave and mmwave chipsets

Raytheon's RF components division introduces a line of fully integrated radio chip-sets for point-to-point, point-to-multipoint, and LMDS applications at 23, 26 and 38 GHz. These chipsets are designed to be used in wireless radios serving as alternatives to optical fiber installations for high-speed data transmission links. They are also suitable for cellular and personal communications networks. Each chipset contains a power amplifier, low-noise amplifiers, low-noise /IF amplifiers, mixers, drivers, multipliers and buffer amplifiers. Component selection is governed by the specification of the power amplifier. The device uses a 4 V bias voltage, simplifying transmitter power supply requirements. Each device is optimized in terms of small signal performance and noise. It can be run into compression, but can be backed off for use as a lin-

ear amplifier. Frequency conversion components, consisting of doublers and triplers, are used in each set to multiply the LO signal to the required operating levels.
Raytheon
INFO/CARD 132

SiGe HBT direct-quadrature modulator

Stanford Microdevices introduces the STQ-2016, new SiGe direct quadrature modulator for the 2G/3G wireless infrastructure and fixed wireless applications. The device is a SiGe modulator operable from 800 MHz to 2500 MHz, covering 2G/3G wireless frequency bands. The part provides



carrier feedthrough typically better than −40 dBm, with low amplitude and phase error of the modulation inputs (typically < 0.2 dB and < 2 °, respectively). Output noise floor is low — typically better than −150 dBm/Hz. The STQ-2016 exhibits high linearity and is suited to 3G applications such as W-CDMA. It is specified for use with a single 5 V power supply.
Stanford Microdevices
INFO/CARD 133

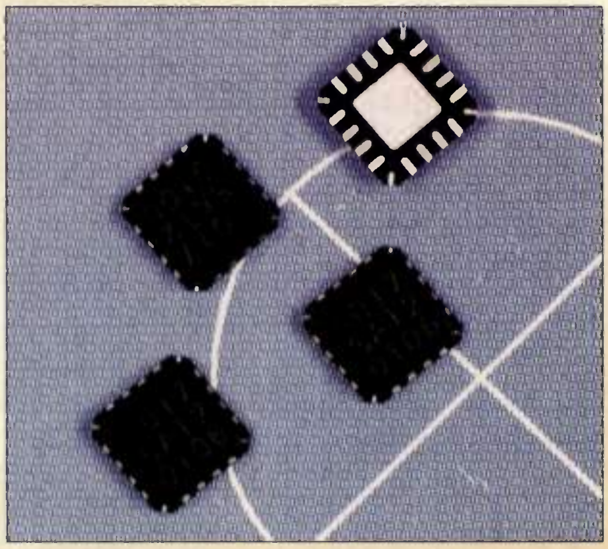
High-power, 240 W GaAs FET

Fujitsu Compound Semiconductor announces a new high-power GaAs FET with an output of 240 W for use in the 2.1 to 2.2 GHz frequency range. The FLL2400IU-2C employs a push-pull design that offers ease of matching, greater consistency and a broader bandwidth for high-power amplifiers. This product is used for solid state, base-station power amplifiers for W-CDMA and IMT 2000 communication systems Using gold metallization, along with the company's quasi-enhancement mode processing, the gain and P1dB power output characteristics of these FETs are improved.
Fujitsu Compound Semiconductor
INFO/CARD 134

Highly integrated up/downconverters for next-generation handsets

M/A-COM introduces a new series of highly integrated up- and down-converter IC's providing a complete chipset solution to meet the stringent linearity and low current requirements of next-generation CDMA and TDMA handsets. These converters integrate an IF amplifier, a mixer, a LO buffer and various other components into a small plastic package. M/A-COM's complete converter chipsets provide high linearity performance combined with low current drain in compact, rugged plastic packages. These chipsets

- Chipset solution includes:
- MD59-0043 800-900 MHz LNA/Downconverter
- MD59-0049 1800-2000 MHz LNA/Downconverter
- MD59-0054 800-900 MHz Upconverter/Driver
- MD59-0062 1800-2000 MHz Upconverter/Driver



perform better than silicon germanium (SiGe) counterparts and provide board-saving levels of integration unavailable in gallium arsenide (GaAs). The converters provide better linearity performance at the same or lower current and DC power dissipation than typical products and offer state-of-the-art performance in conversion gain efficiency and linearity efficiency, giving the system design engineer more flexibility in meeting linearity targets without compromising battery life.

M/A-COM
INFO/CARD 135

AMPLIFIERS

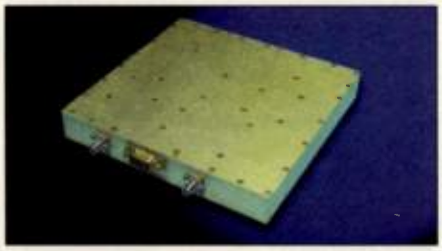
SiGe Class 1 Bluetooth power amplifiers

The PA2423L and PA2423G are the newest power amplifiers for Class 1 Bluetooth applications. The PA2423G is a compact amplifier, while the PA2423L is a compact, fully encapsulated package. They deliver +22.5 dBm output power with 45% power-added efficiency (PAE) when operated in class AB mode. This performance feature enhances operation of Class 1 Bluetooth applications because the devices can overcome insertion losses of up to 2.5dB between the amplifier output and antenna input. Both devices are low-current, consuming only 95 milliamps for +20dBm output power. The silicon germanium (SiGe) structure provides high thermal conductivity to enable a low junction temperature, as well as temperature stability better than 1dB. The devices operate from a single 3.3 V supply, and incorporate power-control and power-down modes for optimal power management.

SiGe
INFO/CARD 136

Broadband, L Band, solid-state power amplifier

Aethercomm's SSPA 1.35-1.85-30 is a high-power, broadband power amplifier. The PA operates over 500 MHz of



bandwidth from 1350 to 1850 MHz. It includes a built-in pre-distortion linearizer and offers a typical OIP3 at 1600 MHz of 58 dBm with two tones spaced at 1 MHz with a single carrier level (SCL) of 39 dBm. Minimum small-signal gain across the band is 35 dB with typical gain flatness of ±1.0 dB. Input and output VSWR is typically 1.5:1. Minimum output power is 25 W across the band. Output Harmonics are a minimum of 60 dBc. This unit operates from a typical voltage of 28 Vdc with a quiescent current of typically 6 amps. Standard features include forward and reflected power telemetry, PA temperature telemetry, Input/Output short and open circuit protection and a pre-distortion linearizer.

Athercomm
INFO/CARD 137

SIGNAL SOURCES

Electronically tunable radio frequency VTO

Paratek has developed a line of oscillator products using unique tuning capabilities. Tuning and frequency range is comparable to YIG oscillators,

while spectral purity is approaching that of dielectric resonator oscillators. Switching speed is similar to varactor-

NEW PRODUCTS

NO. 82

RF/IF MICROWAVE COMPONENTS

DC TO 20GHz TERMINATION HAS SMA MALE CONNECTOR

Mini-Circuits ANNE-50 is a broad band DC to 20GHz precision termination exhibiting return loss of 40dB typical up to 4GHz and 20dB typical from 10 to 20GHz. This low cost, off-the-shelf 50 ohm solution is capable of a broad range of applications that might otherwise require a more expensive custom design, including cellular and satellite communications. Power rating is 0.50W to 70°C ambient. Actual test data available on the Mini-Circuits web site at www.minicircuits.com.

2WAY "DO-IT-YOURSELF" SPLITTER DELIVERS COST SAVINGS

The TCP-2-25 from Mini-Circuits needs only a commercially available 475 ohm external chip resistor, and a complete 200 to 2500MHz 2way-0° power splitter is realized. Designed to lower costs through automated manufacturing, this rugged 50 ohm splitter typically exhibits 18dB isolation, 0.6dB insertion loss (above 3.0dB), and 0.8dB amplitude, 6 degrees phase unbalance. The 50/75 ohm "do-it-yourself" TCP family contains 3 units for operation within the 5 to 2500MHz band.

FEATURED PRODUCT

DC-8GHz MMIC AMPLIFIER KIT WITH FREE TEST FIXTURE

Mini-Circuits GAL family of nine different broadband MMIC amplifiers operating within the DC to 8GHz band are now available in designer's kit form. Kit number K1-GAL contains 10 of ea. model for a total of 90 units, a ***free assembled test fixture***, & complete specification and performance data. Amplifier features include InGaP HBT technology, miniature SOT-89 package, low thermal resistance for high reliability, and up to 18.2dBm (typ) output power.

IN STOCK.

DC TO 6000MHz FIXED ATTENUATOR SERIES IS COST EFFECTIVE

Mini-Circuits VAT family is a very low cost, wide band DC to 6000MHz fixed attenuator series delivering nominal attenuation from 1 to 10dB in 1dB steps, plus 12,15 ,20, and 30dB. Equipped with SMA Type Male/Female connectors, the rugged unibody construction measures only 1.42" long (.312" across hex flats) and can handle 0.5 watt power (at 70°C ambient). Ideal for impedance matching and signal level adjustment applications.

10 TO 2000MHz LEVEL 7 MIXER IS PRICE/PERFORMANCE VALUE

Mini-Circuits has introduced a very low cost high performance frequency mixer for the broad 10 to 2000MHz band. Typically at midband, the ADE-11X displays low 7.1dB conversion loss, 9dBm IP3, and excellent L-R/L-I isolation of 37dB typical. This patented mixer is housed in a low profile 0.112" SM package with solder plated leads for excellent solderability and has all-welded connections for improved reliability. The low $1.99 price includes a 2 year reliability guarantee.

1450 TO 1900MHz VCO OPERATES FROM 5V SUPPLY

Mini-Circuits new ROS-1900V is a 1450 to 1900MHz voltage controlled oscillator housed in a miniature 1/2"x1/2" aqueous washable surface mount package. The VCO offers linear tuning (tuning voltage is 0.5-20V) with low -104dBc/Hz SSB phase noise typical at 10kHz offset, 8dBm typical power output, and operates from a 5V (nominal) supply. Ideal for integration with monolithic PLL chips and commercial synthesizers. Available off-the-shelf.

Mini-Circuits®

US 47 INT'L 48
CIRCLE READER SERVICE CARD

P.O. Box 350166, Brooklyn, New York 11235-0003 (718) 934-4500 Fax (718) 332-4661 For quick access to product information see MINI-CIRCUITS CATALOG & WEB SITE

YONI™ *The Design Engineers Search Engine Provides ACTUAL Data Instantly From MINI-CIRCUITS At:* **www.minicircuits.com**

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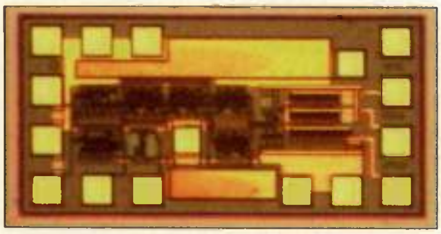
F 352 Rev.Org.

tuned oscillators. The best features of competing VCO technologies are equaled or surpassed with ETRF VTOs. Specifications include: fundamental frequencies in the 1 to 30 GHz range, multiplied outputs up to 40 GHz, tuning range beyond 20%, tuning linearity of typically 3%, no spurious and negligible microphonics – no phase hits and phase noise –105 to –120 dBc/Hz (X-band, 100 kHz offset). The devices offer custom enclosures that include surface-mount, drop-in, and connectorized

Paratek
INFO/CARD 138

SMT GaAs HBT MMIC divide-by-4 prescaler

Hittite Microwave introduces a low-noise, divide-by-4 static prescaler with



InGaP GaAs HBT technology that operates to 12 GHz. The HMC362 features a wide operating input window, a selectable output power bias option, and a low additive SSB phase noise of -149 dBc/Hz at 100 kHz offset. This device operates from DC (with a square wave input) to 12.0 GHz input frequency and operates from a single +5.0V DC supply. The HMC362 is suitable for microwave radio, fiber optic, and VSAT low-phase-noise PLL circuitry applications.

Hittite
INFO/CARD 139

High-performance 3.3v OCXO

MTI-Milliren announces a new high-performance 3.3 VDC OCXO. The 220 series measures 0.975" x 0.800" x 0.500" (LWH). With a frequency range from 4.8 MHz to 100 MHz, these devices achieve the necessary performance characteristics normally associated with larger designs by using a full-size TO-8 (HC-37) quartz housing. Features include a thermal stability of $2.0E^{-08}$ over a 100°C temperature range, warm up time is less than 5 minutes, and power consumption of 1.0 W at +25°C. Phase noise at 1 Hz offset is –85 dBc/Hz with a noise floor of –150 dBc/Hz (at 10 MHz). Devices are delivered in hermetic 16-pin DIP or surface-mount packages, or can be modified with footprint adapter boards. Applications include GPS receivers, instrumentation, rack-mounted applications, and cellular paging base stations.

MTI-Milliren
INFO/CARD 140

Commercial VCO generates from 902 to 928 MHz

Vari-L Model VCO191-915W generates frequencies from 902 to 928 MHz with control voltages from 0.4 to 2.5 Vdc. The unit typically requires 7.0 mA of current from a 3.0 V supply voltage. Typical phase noise at 100 kHz offset is -127dBc/Hz. Typical output power is -3.0 dBm. Second harmonic suppression is typically -15 dBc and third harmonic suppression is typically -20 dBc. The unit is housed in an 8mm x 6mm x 2mm surface-mount, pick-and-place/reflow-compatible package.

Vari-L
INFO/CARD 141

DIGITAL HARDWARE

Re-configurable digital converters

Analog Devices announces the VersaCOMM product family of re-configurable digital converters. The devices perform digital filtering and frequency conversion for high-speed signal processing applications. The devices designed for macro-, micro- and pico-cellular base station designs, and are field-reconfigurable for multi-standard signals including 20 and 30 cellular standards. Other suitable applications include cellular E911 location services, wireless local loop, phased array antennas, digital video, communications test equipment, and ultrasound applications. The family presently includes eight products: three digital receive signal processors (RSPs), two digital transmit signal processors (TSPs) and three quadrature digital upconversion/modulators (QDUCs).

Analog Devices
INFO/CARD 142

Synthesizable 32-bit RISC core for SoC

Lexra announces the LX4380, a 420 MHz 32-bit RISC core optimized for 0.13-micron processes. Lexra's new product targets living-room consumer electronics and communications networking applications, where designers are integrating their packaged RISC processors into SOC designs. The LX4380 enables designers to incorporate up to 64 Kb of on-board cache, and speeds data transfer more than five times over typical 32-bit CPUs. The LX4380's performance comes from three architectural innovations: a seven-stage pipeline, a new design partitioning that confines critical paths to smaller blocks of logic and a new capability called background move that eliminates the time a CPU loses waiting for data transfers to complete. The LX4380 employs a seven-stage pipeline that allocates one extra stage for instruction memory and one extra stage for data memory, permitting design teams to use off-the-shelf memory.

Lexra
INFO/CARD 143

TRANSMISSION COMPONENTS

3.5 TO 6.5 GHz I-Q modulator

Planar Monolithics debuts the model PS-360-AC, an analog-controlled phase shifter/high-speed I & Q modulator, which has been optimized for 3.5 to 6.5

GHz frequency range (other frequency ranges are also available). It is designed to have low spurious harmonics, low insertion loss and a VSWR of

only 2.0:1. It operates with ±15 VDC and its size is 3.25" x 3.25" x 0.85".
Planar Monolithics
INFO/CARD 144

TEST AND MEASUREMENT

Enhanced test capability

Anritsu has enhanced its ME7840A power amplifier test system (PATS) with additional measurement capabili-

ty, making it a comprehensive solution for real-time simultaneous tuning of power amplifier parameters. The system can now conduct ACPR, IMD, PAE, compression, harmonics, and S-parameter measurements with a single connection, thereby creating a turnkey solution that allows for unprecedented convenience, speed, and accuracy when analyzing power amplifiers.
Anritsu
INFO/CARD 145

Satcom spectrum analyzer

A new satcom spectrum analyzer with remote capabilities has been announced by Morrow Technologies. Simply locate the P9116 analyzer at your base station and operate it remotely from any location via Internet, LAN or modem. The P9116 includes a virtual

spectrum analyzer front-panel that lets you view and control your spectrum display in real time with no third-party software or hardware required. The system is a full-featured spectrum analyzer that covers the 100 KHz to 1.6 GHz frequency range. The instrument's rugged, industrial chassis includes a complete Pentium PC and Windows NT operating system.
Morrow Technologies
INFO/CARD 146

Configurable OTS microwave switching system

Keithley Instruments announces the Model S46 microwave switch system. The system can be ordered in a wide variety of configurations for off-the-

shelf delivery. Systems can be built with as many as eight SPDT coaxial microwave relays and as many as four multi-pole relays. Any of the multi-pole relays can be specified as either SP3T, SP4T, SP5T, or SP6T. These relays can be connected to create switching systems that are configured as multiplexers, matrices, independent relays or any combination that meets users' needs. For added versatility, coaxial relay connections are provided on the front panel and the physical position of each relay can be specified by the customer. Moreover, as test requirements change, relays can be easily changed or added to the system to create a new switch configuration. The resulting 19-inch rack package comes in a mounting height as low as 2U (3.5 inches).
Keithley Instruments
INFO/CARD 147

Bluetooth protocol analyzer

Frontline Test Equipment announces its SerialBlue analyzer for the HCI UART, L2CAP, and BCSP Bluetooth protocols. Unlike existing Bluetooth protocol analyzers that monitor RF data flows between Bluetooth products, the system captures and analyzes serial Bluetooth data as it travels between a host and host controller. The system can shrink product development cycles and software debug intervals, and also reduce Bluetooth R&D expenses and improve time to market. Features include: real-time data capture, time stamping, filtering and error analysis. The product has the ability to fully decode data frames on the link between a Bluetooth host and host controller at the bit level. This full decoding capability enables Bluetooth product developers to rapidly detect and isolate even the most minute and intermittent problems associated with their product designs. SerialBlue runs on any Windows 95/98/Me/NT/2000-compatible PC. In its standard configuration, using two standard PC serial ports, the product supports data rates up to 115.2Kbps. With an optional high-speed serial card, also available from Frontline, SerialBlue supports speeds up to 921.6Kbps.
Frontline
INFO/CARD 148

INTERCONNECT/ INTERFACES

Surface-mount F-connector

Trompeter introduces a new edge-mount circuit board F-connector, the CBJE130. The unique design of this new connector features a center pin in-

line with the plane of the board for signal integrity. This side-launch design approach also features a lower profile than standard right-angle jacks for reduced board space requirements. The connector was designed for applications such as broadcast and cable-box products and is part of the new high-frequency PCB coax connector series.
Trompeter
INFO/CARD 149

Coaxial connectors feature high temperature dielectrics

A series of RF coaxial connectors designed for high-power military applications and high-temperature environments such as aircraft engines is introduced by Tru-Connector. The HN-, N-, SC-, SMA-, and TNC-Series connectors can now be manufactured with Fluoroloy H dielectrics, which allow them to operate at up to 5,000 W and temperatures to 200°C, depending on connector type. Configurations include straight, right-angle, and bulkhead connectors in both male and female designs for use with semi-rigid and flexible cables. Fluoroloy H dielectric connectors are manufactured in accordance with MIL-STD-348 and MIL-C-39012 specifications.
Tru-Connector
INFO/CARD 150

Ultra-compact RFICs for Bluetooth/mobile commo

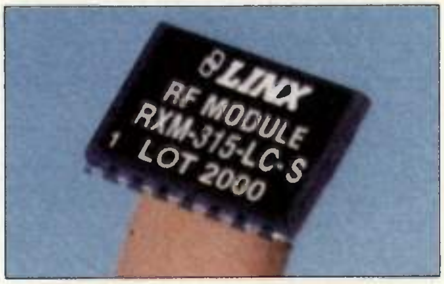
Toshiba America Electronic Components announces the expansion

of its family of RFIC packages called Cell Packs. They offer designers a quick and easy way to implement a variety of important functions required for mobile communications products and Bluetooth-enabled devices. Each pack eliminates the need for multiple discrete devices required to perform specific functions such as wide band amplifier, mixer, attenuator, and single-pole double-throw (SPDT) switch. The family now includes three-wideband amplifier Cell Packs for low-voltage operation, crystal oscillator functions, and SPDT switch functions.
Toshiba America
INFO/CARD 151

SUBSYSTEMS

Addition to LC Series RF module line

Now available in 315 MHz, the second generation of Linx's LC receiver modules features enhanced performance, reduced size and lower cost. When paired with an LC-Series transmitter, these tiny SMD modules provide a cost-effective solution for the wireless transfer of serial, control, and command information over distances in excess of 300 feet. The modules



interface directly to virtually any data source, including microcontrollers and decoder chips, making them ideally suited for use in applications such as remote control, keyless entry, and periodic data transfer. No external components are required (except an antenna), allowing for easy design integration, even by engineers without previous RF experience.
Linx Technologies
INFO/CARD 152

Single-chip transmitter simplifies design

An integrated, single-chip ASK RF transmitter designed specifically to meet the needs of low-cost loop antenna transmitters is now available from Micrel Semiconductor. The MICRF102 QwikRadio transmitter is a highly integrated SOIC-8 package that

requires only five external components. It incorporates transmit power control and unique automatic antenna tuning. Included within the power control function is the ability to set the transmit power by controlling the voltage on a power control pin The device operates off a 5 VDC supply, consumes 7.5 mA (mark) and 4 mA (space) of supply current and has a shutdown pin to further conserve power. It features a 300 MHz-470 MHz transmit frequency, 20 kb/s (maximum) data rate, closed-loop power control and ASK modulation.
Micrel
INFO/CARD 153

RF components for wireless infrastructure market

Conexant Systems announces its first devices in a new product portfolio of broad-range, high-performance RF and microwave components for wireless infrastructure equipment. The products are designed to meet the performance, functionality and cost requirements of current- and next-generation cellular base stations. They include the CX42053 and CX42054 diversity downconverter/receiver front-ends designed for cellular PCS base station receivers and new ultra-linear PA drivers, CX65001, CX65002 and CX65003 for cellularPCS, and other wireless transceivers operating up to 2.500 GHz.
Conexant
INFO/CARD 154

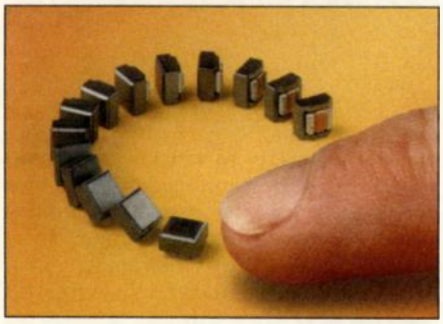
PASSIVES

55 kVDC film capacitors

An assortment of high-voltage film capacitors operating to 55 kVDC is now available from Seacor. The Series MVDF/MVDO offers capacities from 220 to 1 0,000 pf and voltages from 10 to 55 kVDC. Special ratings are also available on request. Small and lightweight, these new devices function from –25°C to +100°C without derating. They offer high humidity resistance and incorporate proprietary compound construction. Round or flat shapes with axial leads are supplied. Key applications include: high-voltage power supplies, laser-focusing circuits, test & measurement equipment, high-resolution TV and video monitors, copying machines and high-voltage circuits.
Seacor
INFO/CARD 155

Tiny power inductors are alternative to costly designs

With a footprint of just 3.7 x 3.7 mm, Coilcraft's new 1008PS Series of surface-mount power inductors is an attractive alternative to larger and



more costly shielded parts. These inductors are 2.6 mm high, making them suitable for applications requiring magnetic shielding, the smallest possible size and low cost. Typical uses include notebook computers, PC cards, wireless communication and handheld devices. The specially designed ferrite cover provides magnetic shielding and the best possible surface for pick and place handling. The series includes 26 models with inductance values from 1 to 1000 μH and saturation current ratings up to 3 Amps.
Coilcraft
INFO/CARD 156

High-current SMT flat coil

Pulse unveils a series of power inductors for high-current applications. Used in DC/DC converters for high-speed desktop computers, notebook computers, servers, and power supply modules, the power inductor provides a low-profile (6 mm or 0.236" max.) compact and robust design for easy SMT mounting. The power inductor has a newly designed base with lock-in metal clips and a mini-coat on the spring coil to enhance the robustness of the inductor. The flat-top, self-leaded design with a fixed clip is suitable for easy pick-and-place applications and higher current ratings. They are magnetically shielded for reduced electromagnetic interference (EMI). The series is available in six inductance values from 0.90 to 4.95 μH with DC resistance values from 1.75 to 12.40 mΩ (typ). The power inductor also provides a high current application as high as 20 ADC (Isat). Operating temperatures are 0°C to 100°C.
Pulse
INFO/CARD 157

Wire wound chip inductor for high frequency apps

KOA Speer Electronics' has added the KQ 0402 surface-mount inductor to its KQ line of open-core, wire-wound, chip inductors. The KQ series is designed with a high Q factor and high self-resonant frequency. The inductor offers a flat op design and is suitable for high-speed pick-and-place components and automatic machine insertion. The KQ series offers tolerances of ± 2%, ±5%, ±10 and ±20% with a nominal inductance range of 1.8 nH to 10 μH. The quality factor range is 16 to 65 with a self-resonant frequency range of 90 to 6000 MHz. The KQ series is available in sizes 0402, 0603, 0805 and 1008. The inductors are designed for use in telecommunications equipment, mobile telephone components and other wireless equipment applications.
KOA Speer Electronics
INFO/CARD 158

Crystals and oscillators added to distributor line

Fox Electronics has added nine new crystals and oscillators. The new additions include 1.0 MHz oscillators in both full- and half-sized metal DIP packages, and several new standard frequencies in the F4100 and F3345 range of 5 x 7 ceramic oscillators. Other products include the VCS25AXT VCXO series, the FOX801BE TCXO and the JITO-2 programmable oscillators. Fox can supply its JITO-2 programmable oscillators with standard or custom frequencies from 340 KHz to 250 MHz.
Fox Electronics
INFO/CARD 159

Chip capacitor arrays offer up to 1μ F capacitance

Taiyo Yuden's Multilayer Chip Capacitor (MLCC) Array series offers the company's materials and miniaurization technologies to integrate he functionality of up to four individ- ıal chip capacitors into a single com- ›onent. The new MLCC Arrays are .vailable in 0630, 0805, or 1206 case izes with X5R or Y5V temperature atings. Designed for EMI/RFI noise nitigation, smoothing and bypass decoupling) applications, the new ИLCC Array components reduce ›CB real estate needs by as much as ;0% and onboard component counts ›y up to 75%.

'aiyo Yuden
NFO/CARD 160

;ore RF chokes feature ıigh current capacity

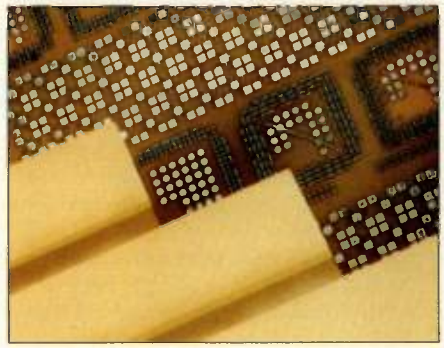
J.W. Miller has expanded its line of ›obbin core RF chokes with the intro- luction of the 1110, 1120, 1130 and .140 RF Choke Series. The new addi- ions offer high current capacity and a vide inductance range. The inductance ıas a range from 1 to 15,000 μH, and ın operating temperature range of -55 o +105°C.

J.W. Miller
INFO/CARD 161

MATERIALS

Prepreg for critical signal layers

W. L. Gore offers a new mixed dielectric using its high-performance SPEEDBOARD@ C prepreg. This approach meets the needs faced today by electronic systems designers for increased signal speeds and densities. SPEEDBOARD C prepreg affords the electrical advantages of faster signals, thinner and lighter PWBs, increased signal integrity, and wider traces. It is comprised of expanded PTFE that has been impregnated with a modified BT resin. The air space inside the ePTFE is replaced with resin and the ePTFE membrane becomes the carrier or delivery system for the resin. In addition to its use in mixed dielectric constructions, it performs well in high-speed digital applications. This prepreg material features a dielectric constant of 2.6 and a low-loss tangent of 0.0036 that is stable over both frequency (1 MHz to 10 GHz) and temperature (–55°C to +200°C). It uses standard thennoset processing as opposed to fusion bonding



and exhibits adhesion to all laminates (FR-4, GETEK@, hydrocarbon, and PTFE). It is UL94-VO certified and NASA-approved for space applications.

W.L. Gore
INFO/CARD 162

GLOSSARY OF TERMS USED IN THIS ISSUE OF RF DESIGN

2G – second generation of wireless communications systems
3G – third generation
A/D – analog-to-digital
AC – alternating current
ACPR – adjacent–channel power ratio
ADC – analog–to–digital converter
AGC – automatic gain control
AMPS – advanced mobile phone system
ASIC – application–specific integrated circuit
ASK – amplifier shift keying
ASP – application service provider
ATM – asynchronous transfer mode
AWGN – additive white gaussian noise
BER – bit error rate
BPSK – binary phase shift keying
CCRR – co–channel rejection ratio
CDMA – code–division multiple access
CDPD – cellular digital packet data
CMOS – complementary metal-oxide semiconductor
CMRR – common-mode rejection ratio
CPU – central processing unit system
CW – continuous wave
DC – direct current
dBc – dB relative to the carrier power
DCS – distributed communications system or digital cellular system
DDS – direct digital synthesis
DECT – digital european cordless telephone
DSP – digital signal processor
DUT – device under test
EEPROM – electrically erasable programmable read-only memory
EMC – electromagnetic compatibility
EMI – electromagnetic interference
ESD – electrostatic discharge
ETSI – european telecommunications standards institute
FCC – federal communications commission
FDD – frequency division duplex
FEC – forward error corrections
FER – frame error rate
FET – field-effect transistor
FHSS – frequency-hopping, spread spectrum
FIFO – first-in, first-out
FIR – finite impulse response
FM – frequency modulation
FSK – frequency shift keying
GaAs – gallium arsenide
GaN – gallium nitride
GFSK – gaussian filtered frequency shift keying
GHz – gigahertz
GMSK – gaussian minimum shift keying
GPIB – general-purpose interface bus
GPRS – general packet radio service
GPS – global positioning system
GSM – global system for mobile communications
HBT – heterojunction bipolar transistor
HEMT – high electron mobility transistor
Hz – Hertz
HSCSD – high–speed circuit-switched data
HTTP – hypertext transfer protocol
I and Q – in–phase and quadrature
I/O – input/output
IC – integrated circuit
IF – intermediate frequency
IM – intermodulation
IMD – intermodulation distortion
InGaP – indium gallium phosphide
InP – indium phosphide
IP – internet protocol
IR – infrared
ISM – industrial, scientific, and medical
kB – kilobyte
LDMOS – laterally diffused metal oxide silicon
LMDS – local multipoint distribution service
LNA – low–noise amplifier
LO – local oscillator
LOS – line of sight
LPF – low–pass filter
LSI – large scale integration
LTCC – low–temperature co–fired ceramic
MDS – multipoint distribution systems
MHZ – megahertz
MMDS – multichannel multipoint distribution service
MMIC – monolithic microwave integrated circuit
MOSFET – metal–oxide semiconductor field–effect transistor
MS/s – million of samples per second
NTC – negative temperature coefficient
OEM – original equipment manufacturer
OXCO – oven controlled crystal oscillator
PA – power amplifier
PAE – power added efficiency
PAR – peak-to-average ratio
PCB – printed circuit board
PCM pulse-code modulation
PCN – personal communications network
PCS – personal communications system
PDA – personal digital assistant
PDC – pacific digital cellular
PECL – positive emitter-coupled logic
PHEMT – pseudomorphic high-electron-mobility transistor
PLL – phase–locked loop
PSK – phase shift keying
QAM - quadrature amplitude modulation
QASK – quadrature amplitude shift keying
QPSK – quadrature phase shift keying
RFI – radio frequency interference
RFIC – radio frequency integrated circuit
RISC – reduced instruction set computing
ROM – read–only memory
SDH – synchronous digital hierarchy
SCM – signal code modulation
SiGe – silicon-germanium
SMR –specialized mobile radio
SMS – short messaging service
SMT – surface–mount technology or surface–mount toroidal
SNR – signal–to–noise ratio
SOIC – small-outline integrated circuit
SONET – synchronous optical network
SPDT – single–pole double–throw
SSB – single side band
SSPA – solid state power amplifiers
TCP – transmission control protocol
TDD – time division duplex
TDMA – time–division multiple access
TETRA – trans european trunked radio
TTL – transistor –transistor logic
TXCO – temperature–compensated crystal oscillator
UART – universal asynchronous receiver transmitter
UHF – ultra high frequency
UMTS – universal mobile telecommunications service
PVCO – voltage-controlled oscillator
VCXO – voltage–controlled crystal oscillator
VOFDM – vector orthogonal frequency division multiplexing
VSAT – very small aperture terminal (satellite service)
VSB – vestigial side band
VSWR – voltage standing wave ratio
WAP – wireless application protocol
W–CDMA – wideband code–division multiple access
WLAN – wireless local area network

RF literature and product showcase

INFO/CARD 250

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INFO/CARD 251

RF Microwave Components:

Typcial products: Combiners/Splitters Price from:$20
Models: from 2 way to 48 way
Frequency range: 800MHz-2200MHz(option: 0.1-6GHz)
Isolation: 20(dB) Insertion Loss: 0.6(dB) VSWR: 1.3:1
Connecter: N(F)/SMA/TNC

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HSA-TV2 (hermetic)
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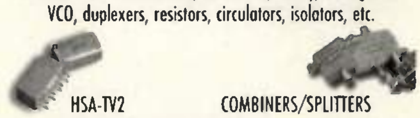

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±0.80 dB @ 20dB	±1.60 dB @ 60 dB
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10 - 30 dB @ ±0.50 dB Max	50 - 60 dB @ ±1.5 dB Max
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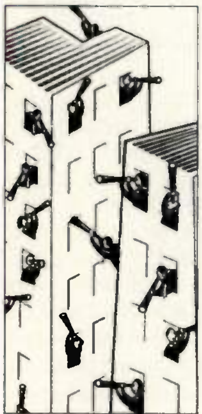

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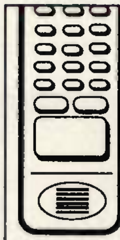

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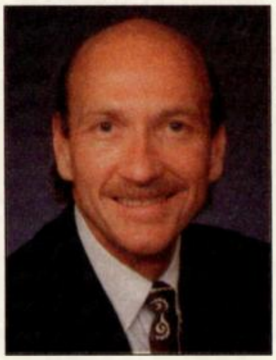
A concept thats time has come



by Ernest Worthman
technology editor
eworthman@intertec.com

From time to time an idea comes along that is so phenomenal that it just sends goose bumps down my spine. Such is the case for disposable cell phones.

Not too long ago, I came across a bit of information about this concept. Being the type of individual who is blacklisted by the insurance companies when it comes to replacing my cell phones, this totally piqued my interest.

No more worries — Yahoo! Finally, I thought. Now all I have to do is buy a cheap disposable cell phone for a few bucks and if I lose it, buy another one. Great concept, but not quite ready for prime time, I'm afraid.

Too bad — As I saw visions of styrofoam or some other biodegradable and disposable typical footprint format cell phones dancing in front of my eyes, I had my rude awakening. As I was ready to rush down to the closest Worst Buys to pick up a dozen or more of these devices, I realized I (like I often do) had left reality.

Well, disappointed to the max, I returned to reality. But, by the time you read this, there may indeed be a disposable cell phone on the market.

It's made of — The technology intrigues me. Currently, there are a couple of manufacturers getting close to product. But what excites me is not the fact that this is tailor-made for me, but the technology used to implement this. The disposable phones have a couple of unique issues that have to be dealt with. First of all, to keep the price down, component count has to be kept to a minimum. And PC boards, LCDs and cases have to be earth-friendly and safely disposable or recyclable. Second is the power source.

There are currently a few ways of doing this. One potential player has decided that the intelligence will reside on the network – exactly the opposite of the current trend. The substrate will be a paper-based with components imbedded directly into the paper. Metallic inks will be used as interconnects, instead of wires or circuit traces. Various other features, such as the keypad graphics and non-imbeddable components will be on separate surfaces that will be connected with ribbon or other inexpensive interconnect cables and connectors. All layers will be sandwiched together in the proper order. The power source can be alkaline, LiON, Ni-MH, or other various battery technologies in a separate pack with quick connect/disconnect connections. The packs would be inexpensive, standard and easily replaceable if lost. If not, they can last as long as the user keeps the phone, or if the phone is lost, the pack simply continues service on the next phone. It can also be assumed that progress will continue on power conservation. And, once voice recognition gets real, the need for a keypad and the associated circuitry disappears, lowering the cost even more. Also, the user will keep the earpiece/mic combination and just recycle it to each new phone.

The implications — This has some far-reaching implications. For example, it could pound another nail into the pay phone coffin. Cell phones have already been sealing it. With comfortably priced disposable cell phones, rather than feeding dollars into a pay phone, one could simply drop a $5 or $10 bill into a vending machine and get a disposable phone with an hour, or whatever, of talk time.

With a little more advancement in voice technology, stolen phones could be a thing of the past as well. It would be pretty cool to simply activate a cell phone with your voiceprint each time you want to use it. If you lose it, you wouldn't have to worry about unauthorized calls and wouldn't even have to notify the carrier of the loss. Simply go drop another one out of the vending machine. And, no more worries about credit checks, long-term contracts and lost-phone police reports.

You also wouldn't have to worry about phone configuration. As I had mentioned earlier, with disposable phones, the intelligence is in the system, so each phone would be a clone of your personal profile kept on the network.

The rest of the story — Disposable cell phones are only the tip of the disposable society iceberg. Some may argue that such disposability isn't the ideal direction for society, but I think it has its merits. First of all, not everything will be headed for disposability. And some things should have been made disposable a long time ago (automobiles and computers).

There is also talk about disposable credit cards, for example. Once they expire, poof...they become useless. Also, if we can tie fingerprint ID to the cards, who cares if the are lost?

Personally, I think we need disposable eyeglasses, and if you ask me and our associate editor, Megan, you'll get a positive on disposable keys as well.

Editor's note: Ern and Megan both recently lost their keys. This proves disposable keys already exist.
-Roger

Ern

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ADE-1	4	0.5-500	+7	5.0	55••	15	1.99▲
ADE-1ASK	3	2-600	+7	5.3	50••	16	3.95
ADE-2ASK	3	1-1000	+7	5.4	45••	12	4.25
ADE-6	5	0.05-250	+7	4.6	40	10	4.95
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ADE-4	3	200-1000	+7	6.8	53••	15	4.25
ADE-14	2	800-1000	+7	7.4	32	17	3.25
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ADE-30	3	200-3000	+7	4.5	35	14	6.95
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ADE-42MH	3	5-4200	+13	7.5	29••	17	14.95
ADE-1H	4	0.5-500	+17	5.3	52••	23	4.95
ADE-10H	3	400-1000	+17	7.0	39	30	7.95
ADE-12H	3	500-1200	+17	6.7	34	28	8.95
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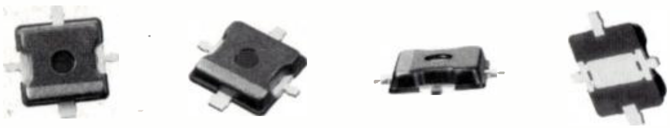

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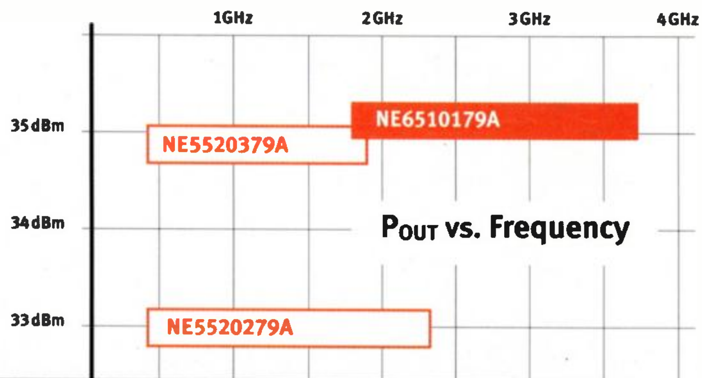
- High output power, high linear gain and high efficiency.
- Low thermal resistance lets you drive your designs harder for higher linearity.

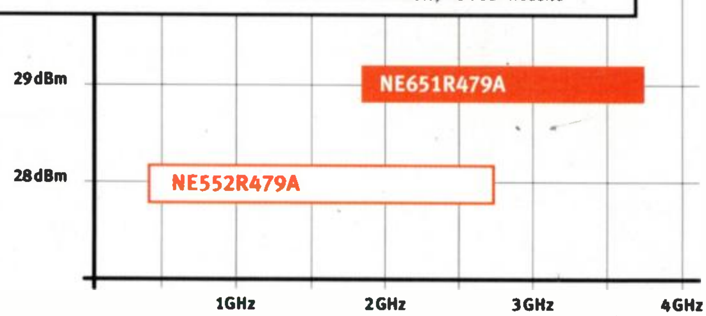


Typical Performance @ 2.3 GHz, V_{DD} = 5V

Part Number	Description	P_{1dB} (dBm)	G_L (dB)	R_{TH} (°C/W)	Freq (GHz)
NE6510179A	GaAs	35	11	5	1.8 – 3.7
NE5520279A	LDMOS	33	10	7	0.4 – 2.35
NE651R479A	GaAs	29	12	30	1.8 – 3.7
NE552R479A	LDMOS	28	11	10	0.4 – 2.7

*Other devices available, visit our website

- Low voltage operation and miniature size make these devices ideal for wireless modems, wireless LANs, mobile radios, cordless phones, cellular phones pagers and other handheld designs.



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