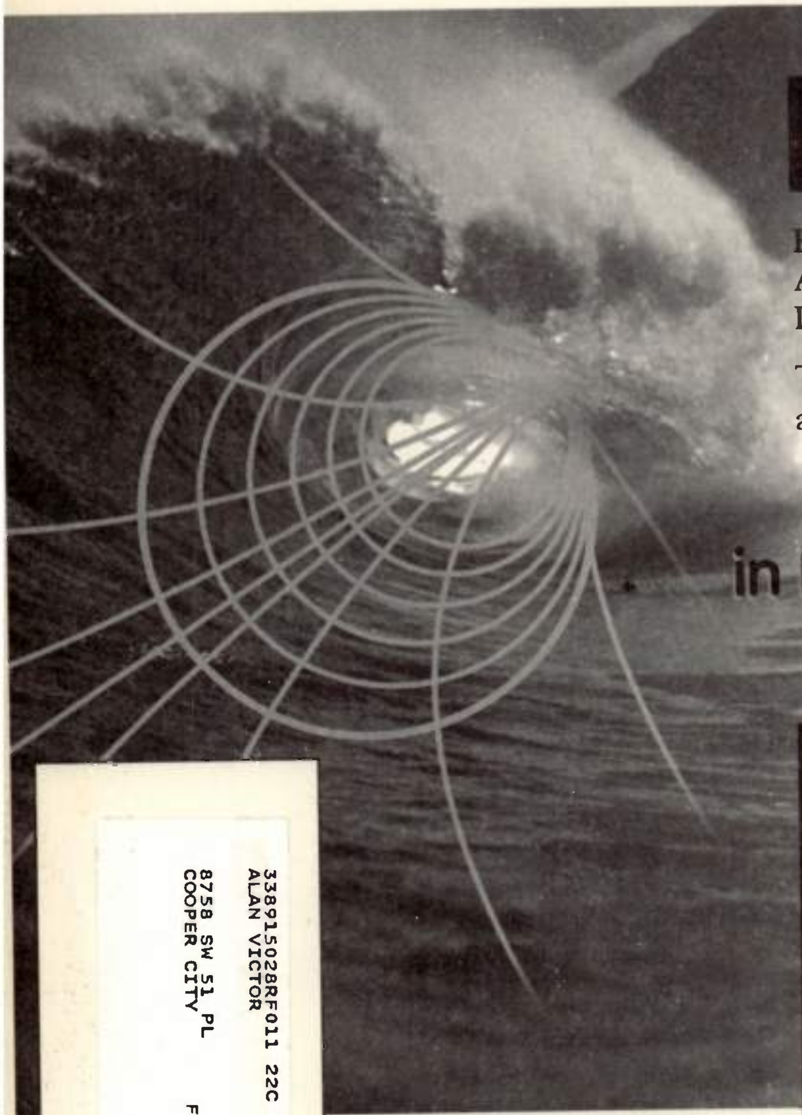
An INTERTEC®/K-III Publication

Periodicals

RF design®

engineering principles & practices

March 1997



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SCHEDULE at-a-GLANCE*

T1 track one: Engineering Fundamentals (Instructional Level: Introductory)

T2 track two: Engineering Applications (Instructional Level: Intermediate)

T3 track three: Advanced Applications (Instructional Level: Intermediate to Advanced)

Monday, April 21, 1997

7:00am - 6:00pm Registration Open

8:00am - 5:00pm

T11 RF and Wireless Engineering Part I: Fundamental Concepts

T21 Practical High-Frequency Filter Design

T31 Digital Modulation and Spread Spectrum for Wireless Communications

Tuesday, April 22, 1997

7:00am - 5:00pm Registration Open

8:00am - 5:00pm

T12 RF and Wireless Engineering Part II: Fundamentals of Amplifier Design

T22 Oscillator Design Principles

T32 RF Power Transistors and Amplifiers: Principles and Practical Applications

Wednesday, April 23, 1997

7:00am - 5:00pm Registration Open

8:00am - 5:00pm

T13 RF and Wireless Engineering Part III: Amplifier Design

8:00am - 4:00pm

Special Course for Management, Marketing and Purchasing Personnel

MT1 Management: Wireless Communications For Non-Engineers Track *(Instructional Level: Non-Technical)*

* *Subject to change.*

See the registration form on the inside back cover. For complete course descriptions and IWCE program information, call our FAX-ON-DEMAND line at 800-601-3858. Or Intertec Presentations at 303-220-0600. You can also FAX your request to 303-770-0253.

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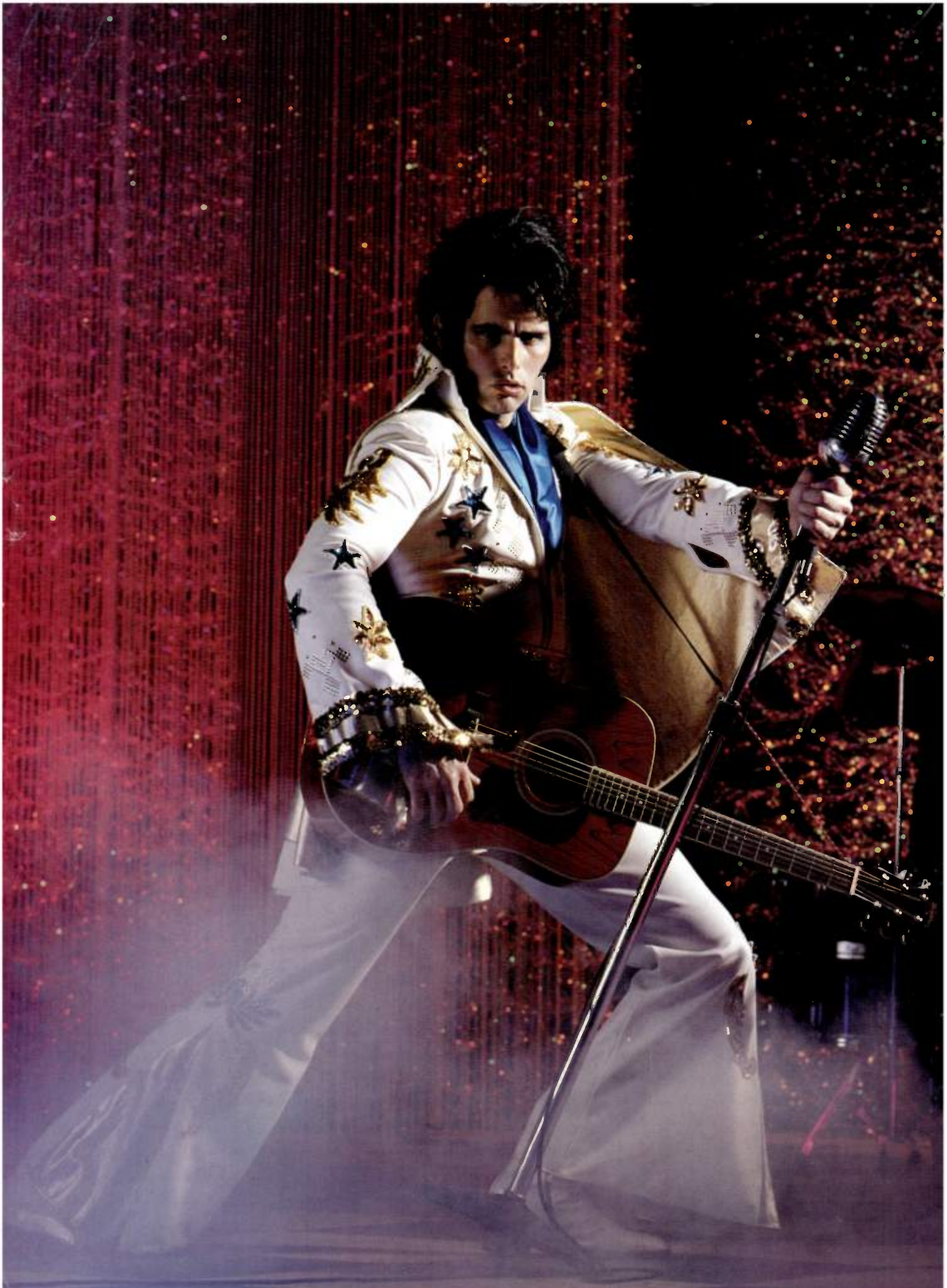
March 1997

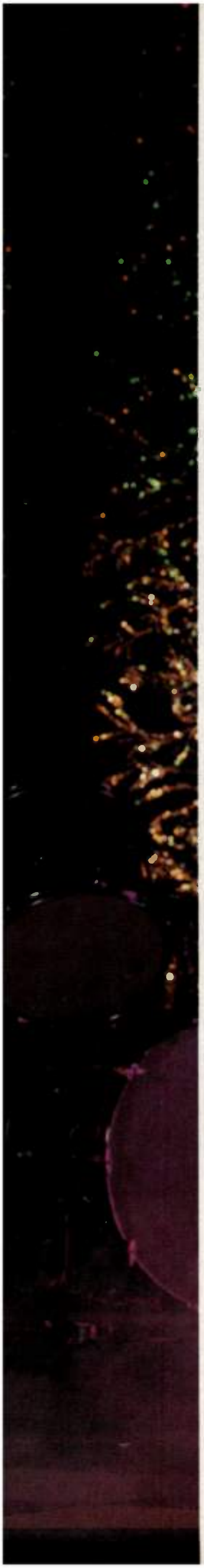
RF design 97 Conference & Expo

September 10-12, 1997
Santa Clara, CA

New GPS hardware and software improve applicability and services



Pictured above is Dale Sydnor, President of Noise Com, finally getting his "fifteen minutes of fame."



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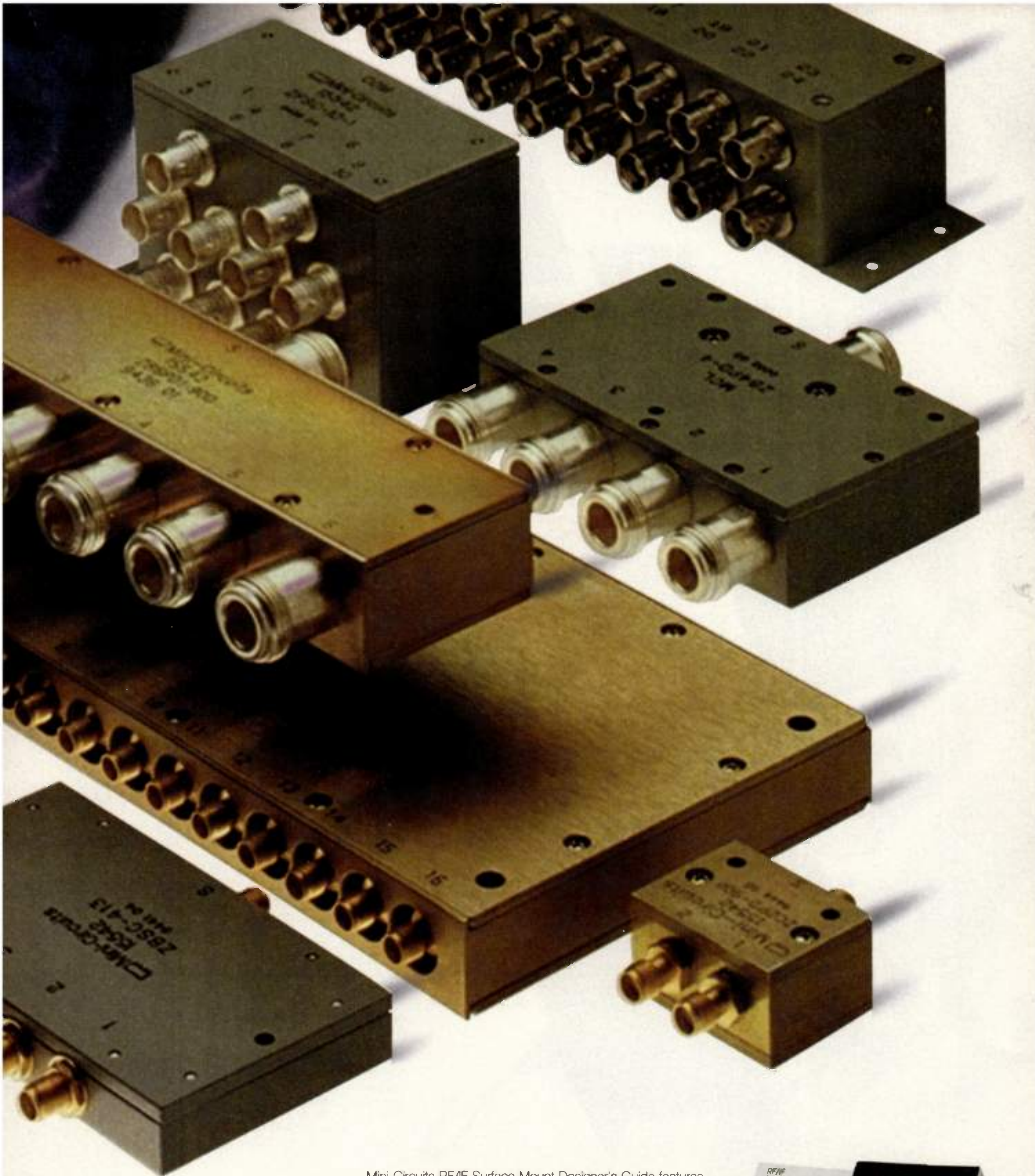

RF design
engineering principles and practices

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March 1997

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cover story

tutorial

departments

This month's cover art shows the GPS satellite constellation (provided by Trimble Navigation) sending signals to a GPS receiver (provided by Eagle).

Coming in April

- DSP
- Tutorial: Crystal oscillators
- Cover story: Integration
- Product forum: Modulators

Wireless networks that croak

By Don Bishop
Editorial Director

The cover of Uyless D. Black's latest book, *Mobile and Wireless Networks*, has pictures of frogs. You see, the science of telecommunications imitates nature— the communication skills of the tiny *coqui*. Quoting (and condensing) Black:

"In the tropical rain forests of Puerto Rico, where the coqui makes its home, there are many frogs in each part of the forest. Because of population density, coqui share audible frequency spectrum with other frogs. Each frog uses a variety of frequencies when it croaks. This little animal is capable of *spectrum-sharing* and *frequency-division multiplexing*. The listener filters extraneous signals and processes relevant signals.

"These frogs exhibit *time-division multiplexing* capabilities, too. Different species 'place calls' (emit croaks) at particular times to make better use of the limited frequency spectrum and to reduce 'co-channel interference.'

"The coqui uses these time slots not only during a particular time of day but during each moment of the day. Each frog knows when (and when not) to croak to reduce or eliminate possible interference with a neighbor. In mobile and wireless jargon, this capability is called a *talk spurt*. Perhaps we can dub the time-division multiplexing capabilities of these frogs as *croak spurts*.

"...[S]ome frogs add redundancy to their transmissions just as we do in some wireless systems. Certain species of frogs can produce a 'periodic stereotyped' call that exhibits redundancy in the 'signals.' If certain frogs happen to interfere with each other's croaks, extra information is provided in the signal to help the listener figure out the information contained in the croak.

"Finally, certain frogs are tuned to the tone and characteristic periods of a specific croak. In case some frog is croaking at the same time as its neighbor, the listener can discern the desired signal from the undesired signal. In other words, these little animals have selective coders and decoders so that they can glean the relevant information in a composite accumulation of croaks.

"...Our society has invested extensive research and committed an immense amount of money in developing the sophisticated technology called mobile wireless communications. Yet, in the reaches of a primitive rain forest, the frog has been developing and refining these communications capabilities for millions of years."

Many people benefit from what Black says about frogs. Scientists who study frogs and who sometimes receive criticism from people who think such studies are silly can cite Black's book and say, "Ah-hah!"

People who think engineers charge too much to design communications networks can cite Black's book and say, "A frog can do it cheaper!"

At Prentice Hall, they can say, "Look at the ink it got us!"

Uninspired magazine editors who nevertheless must deliver a commentary about telecommunications can quote Black's comments about frogs extensively. The book can be ordered by your book dealer by ISBN 0-13-440546-3, or call the publisher at 201-236-7000.

Call for papers

Come to the RF Design '97 Conference & Exposition Sept. 10–12 at the Santa Clara Convention Center, Santa Clara, CA. Visitors are welcome, but you have an opportunity to take an active part, too. Look for our call for papers on page 42. Your colleagues will be there for the presentation of technical papers, and your paper can be among them. RF

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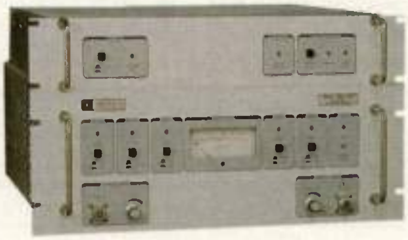

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.50	20	35.0	-	-	1	1.0	50	3000//0
4	TE5090	3	7.50	30	17.5	-	-	2	2.0	75	3300//0
6	TE5100	6	7.50	60	22.5	-	-	3	2.0	90	3300//0
8	TE5110	6	7.50	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	-	-	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	-	-	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	-	-	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

21.4 MHz

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2	TE5180	3	3.75	15	12.5	-	-	2	1.0	50	850//+6
4	TE5190	3	3.75	30	12.5	-	-	3	2.0	70	850//+5
6	TE5200	6	3.75	60	12.5	-	-	4	2.0	90	850//+5
8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850//+5
2	TE5220	3	6.50	15	20.0	-	-	2	1.0	50	1300//+2
4	TE5230	3	6.50	30	22.5	-	-	3	2.0	70	1400//0
6	TE5240	6	6.50	60	22.5	-	-	4	2.0	90	1400//0
8	TE5250	6	6.50	60	17.5	80	22.5	4	2.0	100	1400//0
2	TE5260	3	7.50	15	25.0	-	-	2	1.0	50	1500//0
4	TE5270	3	7.50	30	25.0	-	-	3	2.0	70	1600//0
6	TE5280	6	7.50	60	25.0	-	-	4	2.0	90	1600//0
8	TE5290	6	7.50	60	20.0	80	25.0	4	2.0	100	1600//0
2	TE5300	3	15.0	15	50.0	-	-	2	1.0	45	3000//0
4	TE5310	3	15.0	30	45.0	-	-	3	2.0	60	3000//-1
6	TE5320	6	15.0	60	45.0	-	-	3	2.0	90	3000//-1
8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000//-1

45.0 MHz

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2	TE9420	3-OT	3	3.75	18	16.0	3	1	40	2000//-1.0
4	TE9310	3-OT	3	3.75	30	12.5	3	1	70	2000//-1.0
2	TE7420	3-OT	3	7.50	18	28.0	2	1	40	3000//-1.0
4	TE7430	3-OT	3	7.50	40	30.0	3	1	70	3000//-1.0
2	TE7440	3-OT	3	15.0	15	47.0	2	1	40	8000//-1.5
4	TE7450	3-OT	3	15.0	30	50.0	3	1	70	8000//-1.5
2	TE7730	FUND	3	15.0	15	50.0	2	1	40	1100//+1.5
4	TE7740	FUND	3	15.0	40	60.0	3	1	70	800//+1.0

70.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10400	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10410	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10420	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10430	3-OT	3	10	35	40	80	-910	3	1	2500//-1

90.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10440	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10450	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10460	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10470	3-OT	3	10	35	40	80	-910	3	1	2500//-1
4	TE10480	3-OT	3	15	30	50	80	-910	3	1	4000//-1

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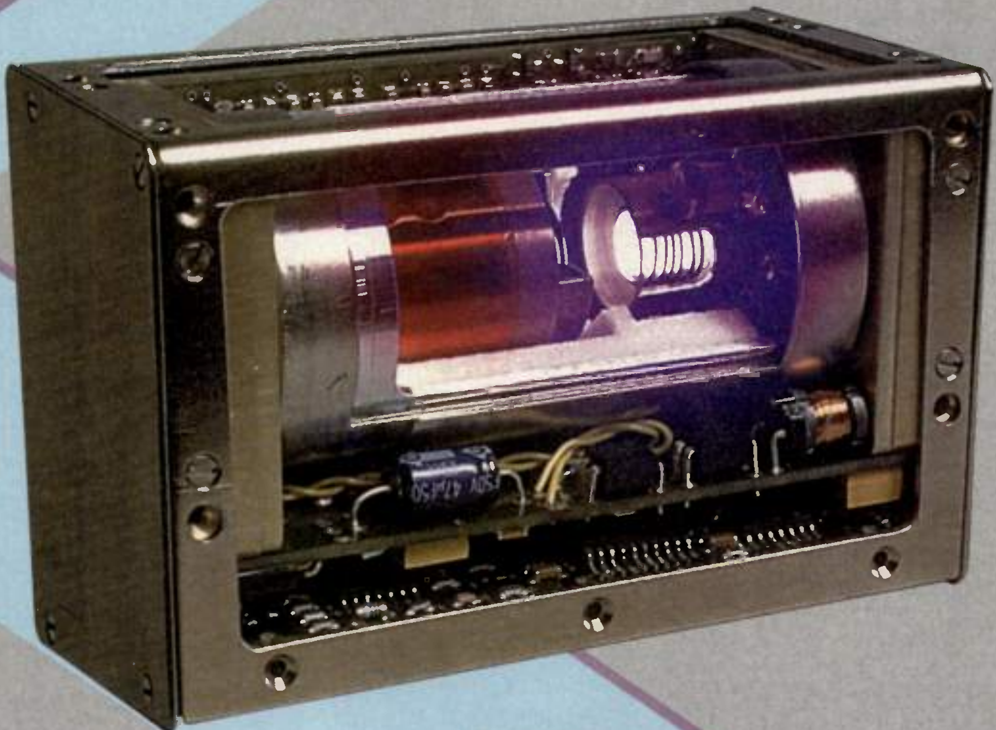

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RF letters

Coaxial standard and other notions

Thank you for your November 1996 editorial "Coax, Haarps, Solutions ...and More Harping." I enjoyed the letters about of the origins of the 50-ohm lines. Please continue these discussions. As a senior engineer, I get asked these questions from less-experienced engineers, and it is good to have some kind of answers. A few years ago, the IEEE *Microwave and Guided Wave Letters*, and *Microwaves & RF* magazine had a spirited discussion on whether a signal could propagate faster than the speed of light. This type of discussion is both entertaining and may inspire creative ideas for new products.

Mark McDonald
Milpitas, CA

Basic physics

I continue to be amazed at the way beta (current gain) hangs around for devices for which beta is meaningless. It is a secondary parameter for bipolar devices and totally meaningless for FET-type devices. Basic physics shows that a modification of the diode equation applies for all these devices. All it requires is application of the diode equation to a pair of simultaneous linear equations to get the correct result. I even have had letters in *RF Design* and many other publications showing the validity of the fact that g_m is the proper parameter, and it is proportional to the output current for two-port semiconductor-type devices.

The equation takes the form: $g_m = K \times (q/kT) \times I_o$, where K is an efficiency factor that is particularly important with electron tubes and FET devices; q is electron charge, k is Boltzmann constant, and T the absolute temperature. Beta *where* it applies is a geometrical factor that depends very precisely on matching of device doping to match g_m and identity of specific design to achieve the kind of operation required for such microprocessor as the 80X86 and pentium chips.

The false application of beta continues unabated, nonetheless. See, for example, Mr. Staudinger's article on p. 20 of the December 1996 issue. I had to do a bit of research to be sure that the beta he mentioned in the article for a MESFET was in fact a current gain. Like a rube operated with negative grid bias, the input current is almost entirely a charging current. (See equation on the bottom right of p. 24.) But the most amazing thing about this equation is the fact that a partial de-

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☐	EMS-1X	Base Station	10-1000	6.0dB	30dB
☐	ESMD-C1	Base Station	1-1000	6.5dB	40dB

TRANSFORMERS

Send me more info	Part Number	Application	Frequency MHz	Ratio	Insertion Loss
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☐	ETK4-2T	Wireless	2-1000	4:1	3 dB
☐	ETC1.6-4-2-3	Wireless	500-2500	4:1	3 dB
☐	ETC4-1-2	Wireless	2-800	4:1	3 dB
☐	ETC9-1	Wireless	70-220	9:1	2.5 dB

COUPLERS

Send me more info	Part Number	Application	Frequency MHz	Ratio	Insertion Loss
☐	ESDC-7-2-75	CATV	5-800	7±1	2.8 dB
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rivative of I_{ds} is taken with respect to I_{ds}. It is important that we make certain that our basic physics is sound (one of the editor's jobs), as these kinds of flubs do not help the reputation of the magazine.

Keats A. Pullen Jr.
Kingsville, MD

Staudinger replies

I'd like to thank Mr. Pullen for pointing out a small typographical error that appeared in the article, "Bias-circuit Methods to Reduce Process and Temperature Variations in GaAs MESFET ICs Are Compared," in the December 1996 issue. Clearly, Equation 4, which defines output conductance of an FET, is given by the derivative of I_{ds} with respect to V_{ds} (i.e., $g_{ds} = \partial I_{ds} / \partial V_{ds}$) and not as the derivative of I_{ds} with respect to I_{ds} as was erroneously stated. Equation 4 is correct in all other aspects. I hope this typographical error was obvious to the magazine's readership. Mr. Pullen's other comments regarding β, are at best difficult to interpret, especially in the context of this article. Many of the *empirical* MESFET large signal models that have been proposed as well as ones that have found wide acceptance in RF and microwave circuit design, use a parameter β to describe, in part, the nominal square law relationship between drain current and applied terminal voltages, e.g.:

$$I_{ds} = \beta(V_{gs} - V_{TO})^2(1+\lambda V_{ds})\tanh(\alpha V_{ds}).$$

It should be emphasized that thi model is empirical in nature and tha model parameters such as β, α, and should not be assigned undue physica significance. Nonetheless, variations o β, λ, V_{TO}...can be used to evaluate varia tions in underlying physical parameter of the device, as well as estimate hov these variations influence circuit perfor mance. Lastly, regarding current gain a simple derivation of current gain for MESFET (defined in terms of hybri parameter h_{21}) shows a contributio caused by large signal model paramete β.

Joe Staudinger
Phoenix, AZ

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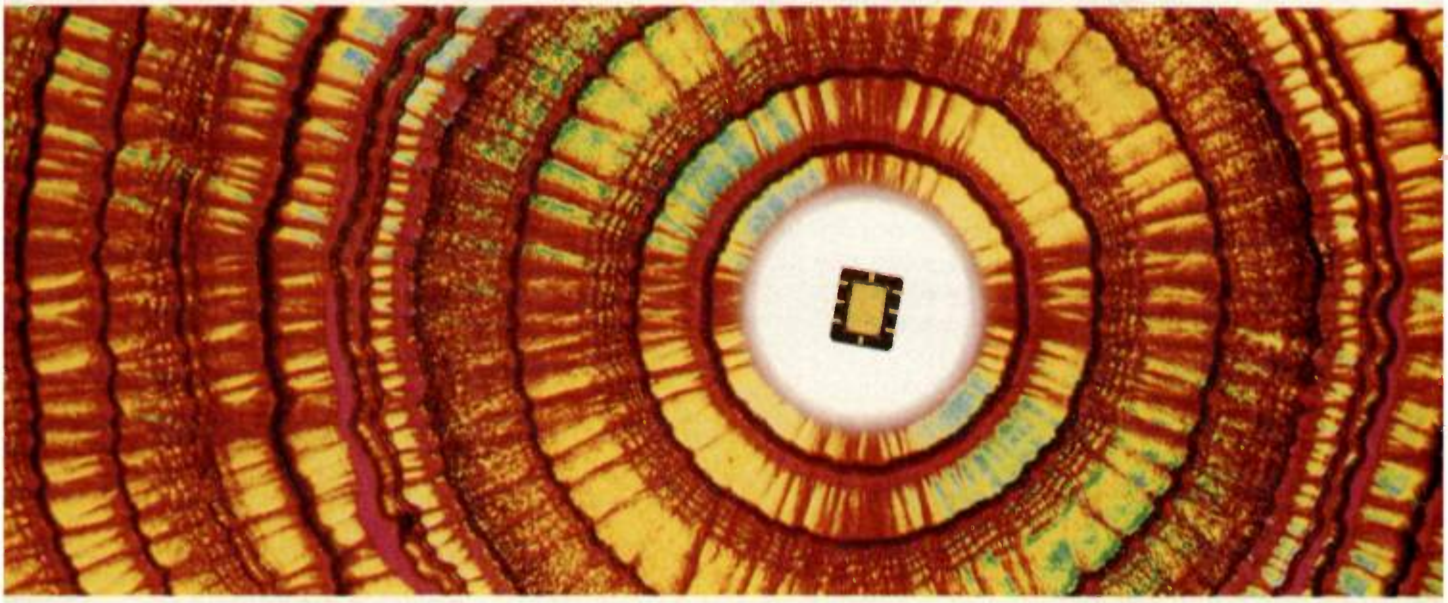

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RF calendar

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25–27 **Communication Design Engineering Conference**—*Washington,* DC. Information: Conference, 600 Harrison St. 4th Floor, San Francisco, CA. Tel. 617-821-9219; Fax 617-828-8198; E-mail cdec@exporeg.com.

April 2 **Surface Mount Technology Association Exhibition and *Sessions***—*Atlanta.* Exhibiting information: Lynda Kelly, Tel. 770-518-9039; Fax 770-642-1288; Sessions information : SMTA National. Tel. 612-920-7682; Fax 612-926-1819; E-mail smta@ smta.com; Web site http://www.smta.org.

14–17 **International Conference on Antennas and Propagation**—*Edinburgh.* Information: ICAP Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, United Kingdom. Tel. +44 (0) 71-344-5467/5473; Fax +44 (0) 71-240-8830; E-mail lhudson@ iee.org.uk or mswift@iee.orguk.

21–23 **RF Design Seminar Series**—*Las Vegas.* Information: Intertec Presentations, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel. 800-288-8606 or 303-220-0600; Fax 303-770-0253.

22–24 **International Wireless Communications Expo**—*Las Vegas.* Information: Intertec Presentations, 6300 S. Syracuse Way, Denver, CO 80111. Tel. 800-288-8606 or 303-220-0600; Fax 303-770-0253. RF Pavilion—Manufacturers' exhibits within IWCE. Components, test equipment, software and services for RF equipment manufacturing.

22–24 **Convergence Tech and IC Expo for microelectronics, communications and computer professionals**—*Dallas.* Information: Electronic Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045. Tel. 800-877-2668, ext. 243; Fax 310-641-5117.

23–26 **Broadcast Technology**—*Jakarta, Indonesia.* Information: Eileen Lavine, Information Services, 4733 Bethesda Ave., Suite 700, Bethesda MD 20814. Tel. 301-656-2942; Fax 301-656-3179.

May 5–7 **Vehicular Technology Conference for cellular & mobile wireless communications**—*Phoenix.* Information: Wendy Rochelle, Registrar, IEEE Conference Service, 455 Hoes Lane, P.O. Box 1331 Piscataway, NJ 08855-1331. E-mail w.rochelle@ieee.org.

6–8 **Electronics Industries Forum of New England**—*Boston.* Information: Linda Hanson. Tel. 914-779-0696.

13–15 **ID Expo, bar code and automatic data capture trade show**—*Philadelphia.* Information: Advanstar Expositions. Tel. 800-331-5706 or 218-723-9130.

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JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
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Besser Associates & Microwave Online Services—*RF and Wireless Made Simple*—May 5–6, Los Altos, CA; May 12–13, Dallas; *RF Measurement Techniques*—May 5–8, Dallas; *Wireless RF System Design*—May 5–9, Dallas; *RF Component Modeling*—May 12–16, Dallas; *Applied RF Techniques I*—May 12–16, Los Altos, CA; *Applied RF Techniques II*—May 12–16, Dallas; *RF Productivity Improvement*—May 14–16, Dallas; May 19–22, Los Altos, CA; *EM Field Simulator Made Practical*—May 15–16, Los Altos, CA; *Nonlinear RF Circuit Design and Modeling with Spice*—May 19–20, Los Altos, CA; *Digital Signal Processing*—May 21–23, Los Altos, CA. Information: Annie Wong, Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel. 415-949-3300; Fax 415-949-4400. RF and Microwave courses are available on the World Wide Web exclusively through RF Globalnet. The courses will be given over a two-to-three week period. *Applied RF Techniques I: Linear Circuits.* Information: Larry Black, Microwave Online. Tel. 303-415-9233; E-mail larry@mosco.com; Web site http://www.rfmicrowave.com.

CKC Laboratories—*EMC for Medical Electronics*—April 22–23; *Immunity to ESD*—May 12, Orange County, CA; *CE Mark Design and Compliance Routes*—May 13–14, Orange County, CA; *Core EMC Design*—June 17–18, Hillsboro, OR; . Information: Linda Grunow or Todd Robinson, CKC Laboratories, 5473-A Clouds Rest, Mariposa, CA 95338. Tel. 800-500-4362 or 209-966-5240; Fax 209-742-6133; E-mail lgrunow@ckc.com.

Georgia Tech Continuing Education— *Advanced Electronic Warfare Principles*—March 25–28, Atlanta; *Radar Cross Section Reduction*—March 25–28, Atlanta. Information: Department of Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel. 404-894-2547; E-mail conted@gatech. edu; Web site http://www.conted.gatech.edu.

Mead Microelectronics—*Architectural and Circuit Design for Portable Electronics Systems* (3-days digital, plus 3-days analog)—March 31–April 5; *RF IC Design for Wireless Communication Systems*–May 12–16; *Data Communication ICs*–May 14–16. Information: Mead Microelectronics, 7100 Grandview Dr., Corvallis, OR 97330. Tel. 541-758-0828; Fax 541-752-1405. In Europe, contact Mead Microelectronics, Venoge 7, 1025 St. Sulpice, Switzerland. Tel. +41-21-691-0244; Fax +41-21-691-0245; E-mail mead@netgate.net or valence@mead.ch; Web site http://www.netgate.net/~mead.

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National Institute of Standards and Technology—*Boulder*, CO. *Time and Frequency Seminars: Introduction-Level I*, June 23–24; *Fundamentals-Level II*, June 25–27. Information: Wendy Ortega Henderson, National Institute of Standards and Technology. Tel. 303-497-3693; Fax 303-497-6461; E-mail ortegaw@boulder.nist.gov. Web site: http://www.boulder.nist.gov/timefreq.

Northeast Consortium for Engineering Education—*Principles of Electronic Counter-countermeasures*—April 8–10, Atlanta; *Infrared Technology and Applications*—April 15–18 Atlanta; *Phased-array Radar System Design*—April 29–May 2, Atlanta; *Antennas: Principles, Design and Measurements*—May 19–22, St. Cloud, FL. Information: Kelly Brown, Northeast Consortium for Engineering Education, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel. 407-892-6146; Fax 407-892-0406.

***RF Design* Seminar Series**—April 21–23, Las Vegas; Sept. 9–11, Santa Clara, CA. Information: Intertec Presentations, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel. 303-220-0600 or 800-288-8606; Fax 303-770-0253. RF Pavilion—Manufacturers' exhibits within IWCE. Components, test equipment, software and services for RF equipment manufacturing—April 22–24, Las Vegas.

University of Missouri-Rolla—*Grounding and Shielding Electronic Systems: How to Diagnose and Solve Electrical Noise Problems*—May 19–21, Boston; June 25-27, Research Triangle Park, NC, August 13–15, Dallas; *Circuit Board Layout to Reduce Noise Emission and Susceptibility*—May 21, Boston; June 27, Research Triangle Park, NC; August 15, Dallas; *Electromagnetic Compatibility Certificate Program*—Combination home study 40-hour video taped lecture and one-week on-campus laboratory course. Information: Continuing Education Coordinator, University of Missouri-Rolla, 103 ME Annex, Rolla, MO 65409-1560. Tel. 573-341-4132. Fax 573-341-4992; E-mail buddyp@ shuttle.cc.umr.edu; Web site http://www.umr.edu/~conted.

Z Domain Technologies—*DSP Without Tears*—April 9–11, San Jose, CA; April 30–May 2, Atlanta. Information: Z Domain Technologies, 555 Sun Valley Drive, Suite A4, Roswell, GA 30076. Tel. 800-967-5034 or 770-587-4812. Fax 770-518-8368; E-mail dsp@zdt.com; Web site http://www.zdt.com/~dsp.

NTIA offers strategic spectrum planning program

A report from the Commerce Department's National Telecommunications and Information Administration (NTIA) proposes spectrum management measures to allow more user access to the 3–30 MHz, high-frequency portion of the electromagnetic spectrum. *High Frequency (3–30 MHz) Spectrum Planning Options* discusses future U.S. high-frequency spectrum requirements and long-range planning options to satisfy additional high-frequency spectrum requirements.

Current high-frequency spectrum is congested in many countries and various parts of the world, and the number of systems is expected to increase in the next ten years. There are increased demands for safety for maritime and aeronautical operations, for greater international broadcasting capabilities and for more recreational uses of the high-frequency spectrum.

The report discusses four long-range spectrum planning options: 1) making more efficient use of the current allocations; 2) reaccommodating incumbent spectrum users; 3) using non-spectrum technologies such as fiber optic cable instead of radio spectrum; 4) using higher frequencies.

Copies of all NTIA reports are available on the Internet at http://www.ntia.doc.gov. A limited number of printed copies are available from NTIA's Office of Public Affairs at 202-482-3999.

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A high-accuracy I-Q demodulator and modulator

By David I. Hoult

A circuit performs with considerable accuracy both in-phase and quadrature (I-Q) demodulation and modulation at intermediate frequencies of the order of 5 MHz. Errors of phase and amplitude are reduced to less than 0.1% over a 10 kHz bandwidth so that small "ghost" signals associated with this type of detection become comparable to other spurious signals produced by harmonic distortion. A phase-shift method essentially removes ghosts and harmonic signals and also can be used with direct intermediate frequency digitization. Use the circuit to implement a novel narrow bandpass filter.

Despite the prevalent trend in radio architecture toward digitization of the intermediate frequency signal, a good argument can be made for retaining the classic superheterodyne design when the ultimate in dynamic range over a narrow bandwidth is needed. This is the case in the author's field—medical nuclear magnetic resonance. In this equipment, a huge dynamic range of 120 dB occasionally can be encountered in the received signal. Formally, at least, magnetic resonance equipment is similar to that employed in radar. The received radio frequency (RF) signal is a transient response to a short-amplitude or phase-modulated RF pulse. The latter may be a 10 kW peak burst of 50 μs duration at 130 MHz that elicits responses from various resonant atomic nuclei in a sample. (See Figure 1.) These nuclear magnetic resonances (sometimes as many as 100 or more) may have frequencies spread out over a bandwidth of 10 kHz and may have incredibly high Q-factors—occasionally as great as 10^9. As a result, the ringdown following the impact of the pulse may last anywhere from a few milliseconds to seconds, depending on the sample under study (sometimes a human being). Coupling to the nuclear resonances is achieved by mutual induction because the phenomenon is magnetic. It follows that, unlike radar, the signal is not received as a radio wave, but by near-field induction in an optimally matched pickup coil. The induced electromotive force (EMF) from the hydrogen atoms in water in a human head initially may be as large as 10 mV at 130.0005 MHz. The signal from the hydrogen atoms in some low-concentration metabolite may be as small as 10 nV at 130.0002 MHz, and comparable to the ever-present background thermal noise. From then on, though, the received signal is treated just as any other. It is passed to a low-noise preamplifier and then down converted in one or two stages until final quadrature (I-Q) demodulation and filtering results in two audio-frequency transient decays that are digitized and analyzed with a computer.

Our task is to pick out accurately and unambiguously the small in the presence of the large, and not surprisingly, to reduce the dynamic range. Many fancy tricks of physics that "torture" the atomic nuclei have evolved over the years. At bottom though, the experimental results are only as good as the transmitter and receiver electronics and the digital signal processing that follows. The heart of the computing process that classifies the various nuclear signals is a fast Fourier transform (FFT) that delivers the spectrum from the two audio-frequency de-

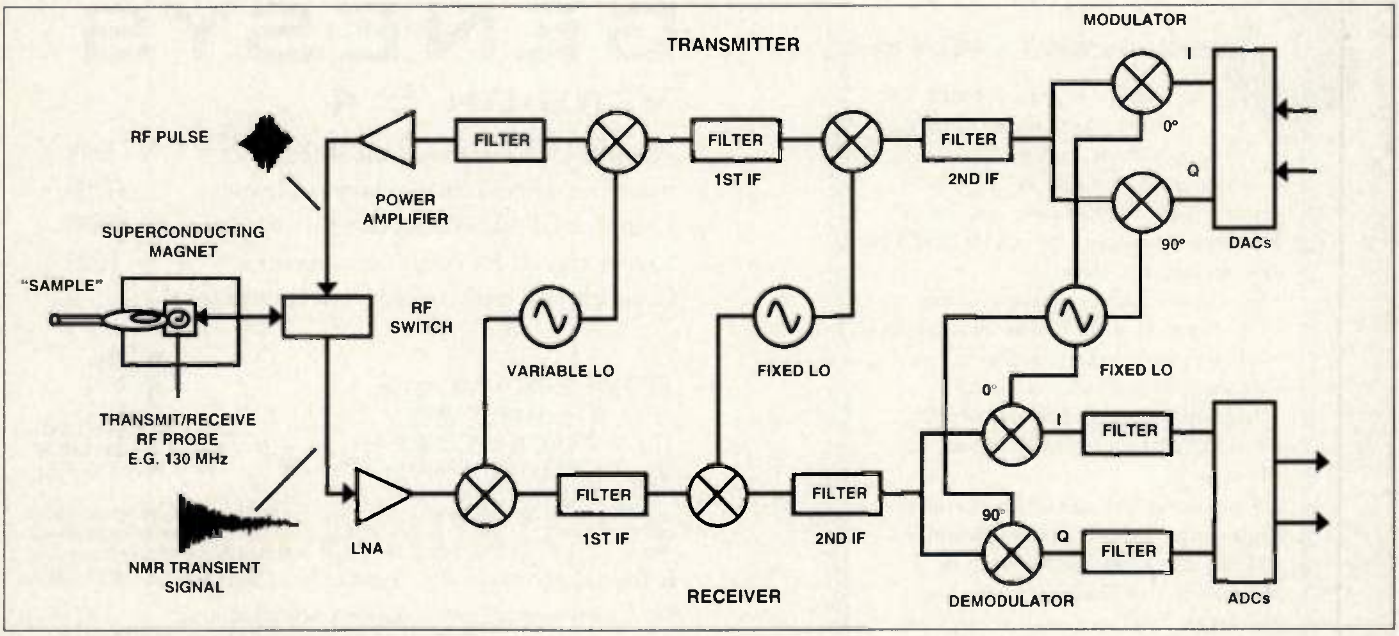

Figure 1. Nuclear magnetic resonance spectrometer architecture.

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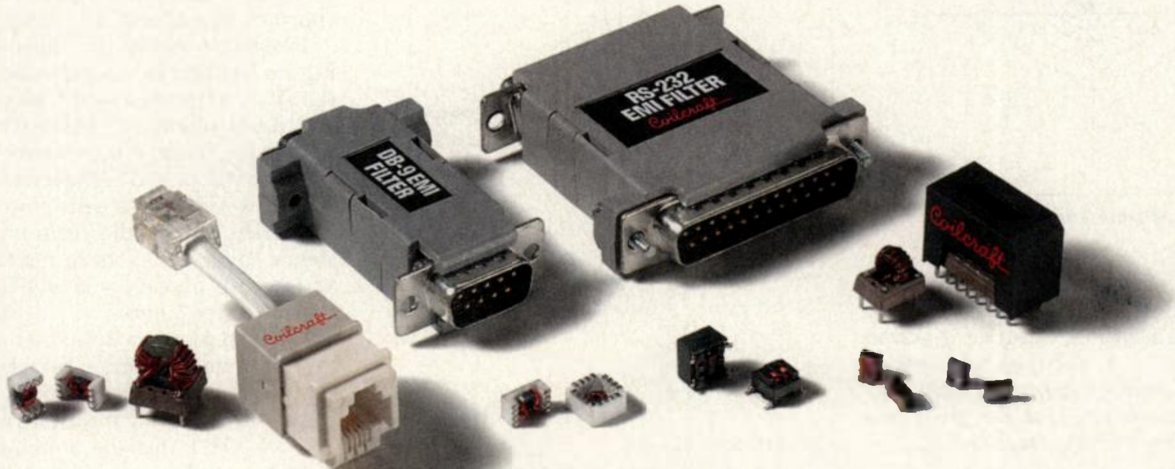

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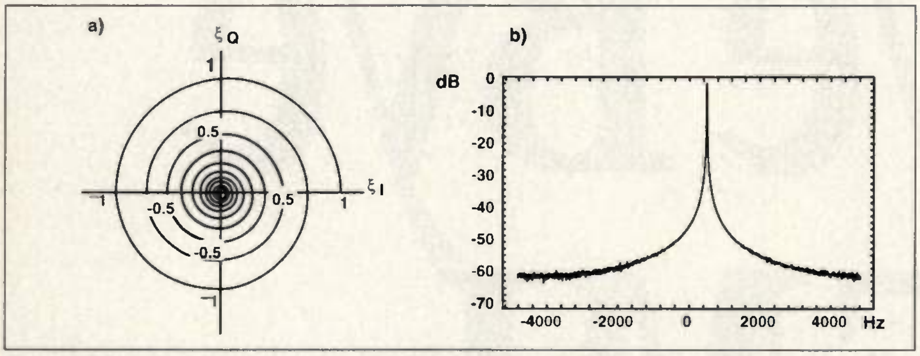


Figure 2. a) Lissajous figure of the outputs of a perfect I-Q demodulator, b) the corresponding spectrum.

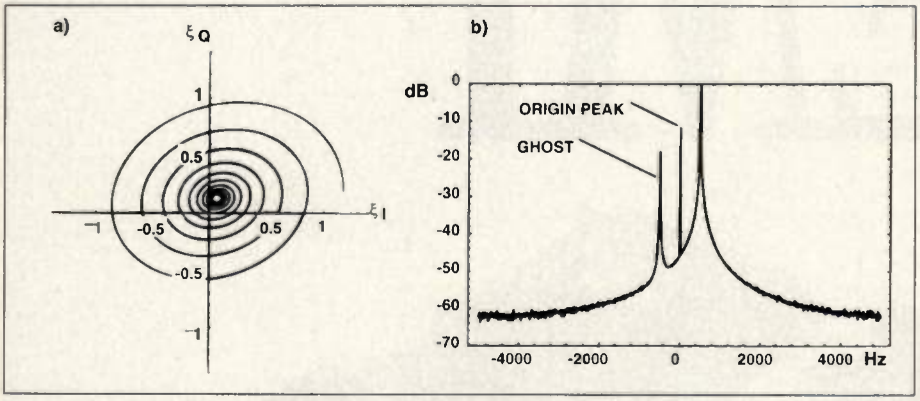


Figure 3. a) Imperfect I-Q demodulator output, b) the corresponding spectrum.

cays. It is worth spending a few moments looking at the mathematics of the signal reception, because it paves the way for a graphical method of looking at some of the problems that arise. Let us assume that the pulse has excited a strong single resonance ξ:

$$\xi = A\exp\left\{\frac{-t}{T_2}\right\}\cos(\omega_o t + \phi) \qquad (1)$$

where A is the initial amplitude of the signal; T_2 is the time constant of the decay, typically 100 ms; ω_o is the angular radio frequency (say 2π times 130.0005 MHz); phase ϕ is a function of the electronics (cables lengths and the like) and generally is known and adjustable. For example, and most importantly for what follows, ϕ may be changed by altering the phase of the transmitter pulse.

The essence of the I-Q demodulation process is the multiplication of this signal by in-phase and quadrature sine waves at the transmitter frequency ω_t (say 130 MHz), followed by a filtering out of high frequencies, giving us the two signals:

$$\xi_I = A\exp\left\{\frac{-t}{T_2}\right\}\cos(\delta\omega t);$$
$$\xi_Q = A\exp\left\{\frac{-t}{T_2}\right\}\sin(\delta\omega t) \qquad (2)$$

where $\delta\omega = \omega_o - \omega_t$.

It is convenient to regard these two analog signals both mathematically and graphically as components of a single complex signal $\xi_C = \xi_I + j\xi_Q$, where $j = \sqrt{-1}$. The resulting complex signal is:

$$\xi_c = A\exp\left\{j\delta\omega - \frac{1}{T_2}\right\}t \qquad (3)$$

If we draw this on an Argand diagram—essentially a Lissajous figure of the two signals—it is simply a decaying spiral. (See Figure 2a.) When the time constant of the decay is long, the Argand plot of one cycle essentially is a circle. When we digitize this complex signal and Fourier transform it, we obtain a Lorentzian line:

$$F(\omega) = A\int_0^\infty \exp\left\{j\delta\omega - \frac{1}{T_2}\right\}t\cdot\exp\{-j\omega t\}dt$$
$$= \frac{AT_2}{\{1 - j(\delta\omega - \omega)T_2\}} \qquad (4)$$

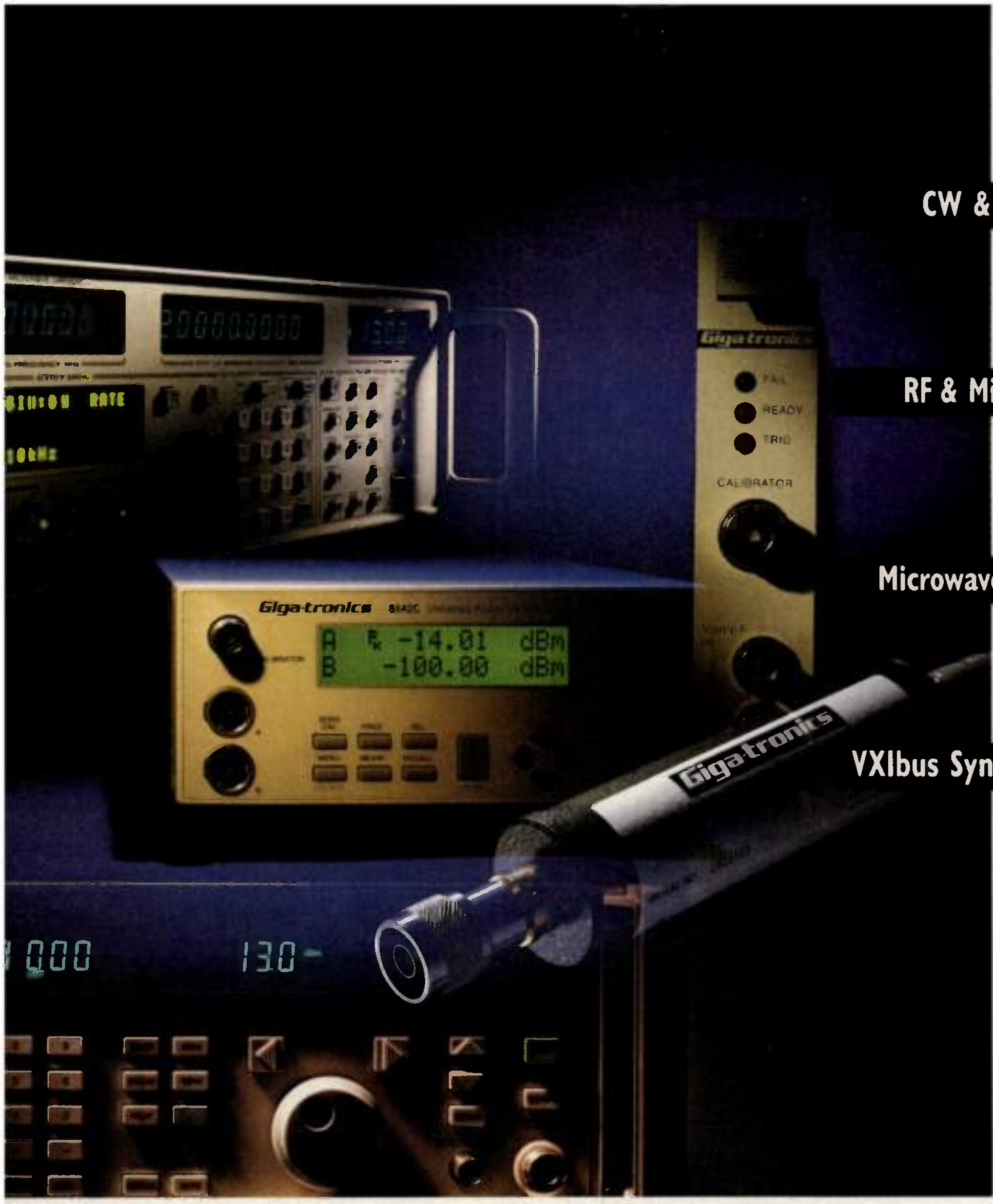
(See Figure 2b.) As mentioned earlier, a typical magnetic resonance spectrum might contain many such spectral lines, some large, some small.

A better I-Q demodulator

It has been known for many years that quadrature demodulation is problematic, especially when a large dynamic range is needed. It is partly to get around these problems that the communications industry is turning to intermediate frequency sampling. Problem number one is that zero frequency—DC—is part of the spectrum. The spiral of Figure 2a changes direction when $\delta\omega < 0$. If $\delta\omega = 0$, the plot does not spiral at all. It simply decays in a straight line. This does not present difficulties if the decay is to the origin, but if there is a DC offset, as shown in Figure 3a, the computer interprets the offset as a genuine signal and gives a spurious line at zero frequency.

Problem number two, also shown in Figure 3a, but in exaggerated form, is that it is notoriously difficult with conventional phase-sensitive demodulators (e.g., diode-ring mixers) to preserve a 90° phase difference between the two sinewave references, and hence the two audio frequency transients. Over time, there may be drift, and even a phase error of 1° can have serious consequences.

Problem number three is that even if the I-Q demodulator is perfect, the filters that must follow (typically 4-pole Butterworth with cut-off frequencies of about 5 kHz) must be matched closely if the two audio signals are to be orthonormal, i.e., exactly equal in amplitude and 90° apart in phase over the required bandwidth. Problems two and three both generate a spurious signal or "ghost" at the negative frequency, as shown in Figure 3b, and we will consider these errors in more detail later. The problems disappear when the final demodulation is avoided and the intermediate frequency is sampled. Excellent articles on this subject may be found in the September 1994 and May 1995 issues [1,2]. On the other hand, the price paid is a loss of dynamic range because the resolution of analog-to-digital converters (ADCs) drops considerably with the necessary higher sampling frequencies needed. For example, two 16-bit ADCs, such as Analog Devices' AD1382, sampling quadrature data may have a maximum sampling rate of 500 kilosamples per



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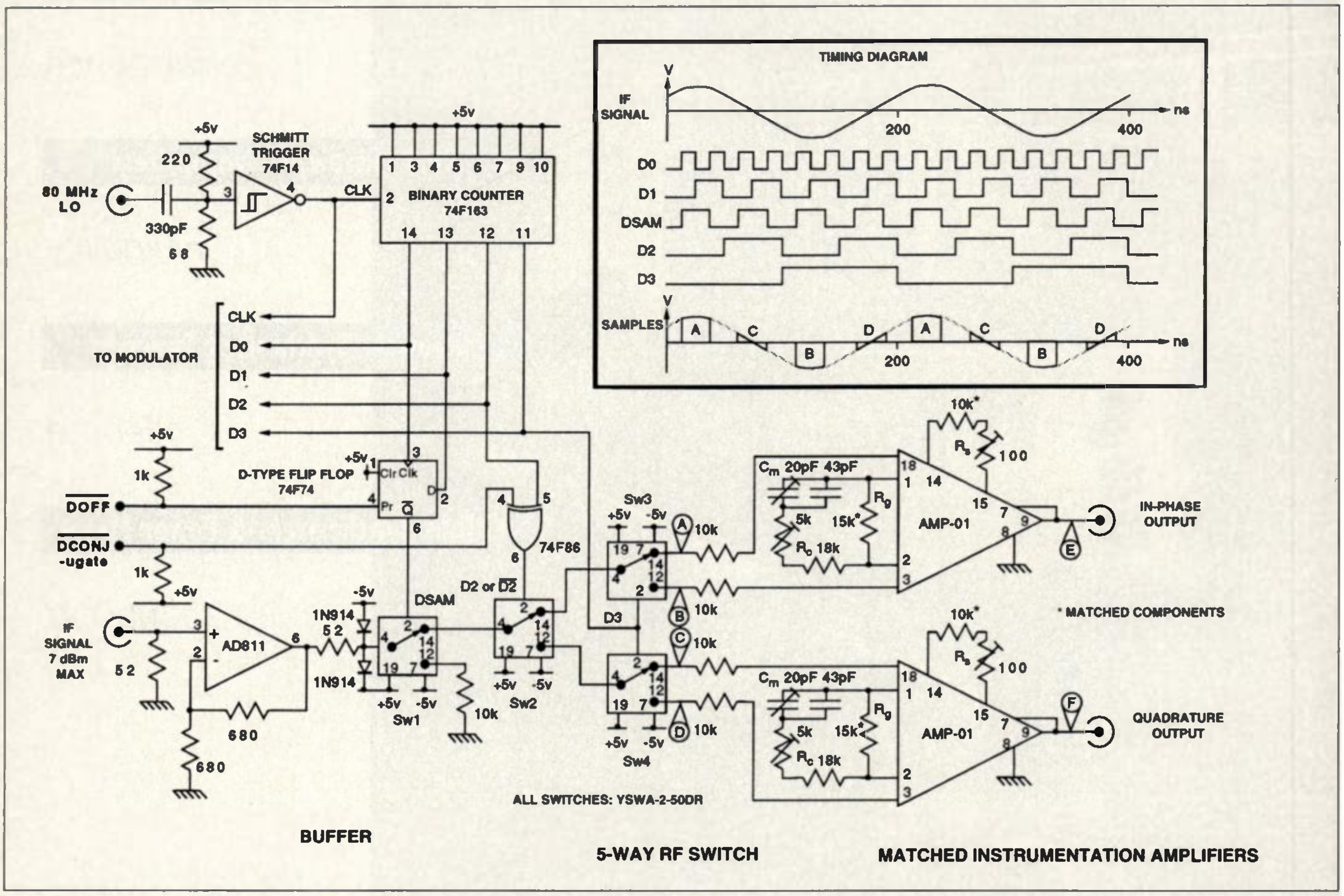

Figure 4. The I-Q demodulator circuit.

second (kSPS) and a bandwidth of ± 50 kHz, i.e., 500 kHz. When sampling at 40 megasamples per second (MSPS) with a single ADC, a bandwidth 40 times greater can be accommodated, and thus the noise floor is √40 times, or nearly 3 bits, larger. To maintain the same dynamic range, a 13-bit ADC is needed. Typically, only a 10-bit ADC can function this quickly, so there is a loss of approximately 20 dB in dynamic range.

The challenge was to build a better I-Q demodulator, not forgetting that to generate the transmitter pulse, a good I-Q modulator was also needed. The design finally adopted was to sample *analogically* the IF signal at four times its frequency, then to route the analog samples appropriately to two instrumentation amplifiers with the aid of GaAs RF switches. The principle is shown in Figure 4. Using fast transistor-transistor logic (TTL) and an IF of 5 MHz, the 74F163 binary counter divides an 80 MHz reference signal from a high-stability source (in our case, a frequency synthesizer) to obtain three controls, DSAMple, D2 and D3, for the Mini-Circuits YSWA-2-50DR absorptive GaAs single-pole, double-throw (SPDT) switches. (The use of 80 MHz with division by 16, rather than 40 MHz with division by 8, will be discussed later.) These switches have rise times of typically 1.5 ns, so the routing performed by switches Sw 2, 3 and 4 is well in place before switch Sw1 opens for 25 ns. The figure shows how the four outputs from the switches, points A–D, each receive a sample of the IF signal once per cycle. Because the instrumentation amplifiers (Analog Devices AMP-01) that follow cannot pass frequencies higher than about 200 kHz, they put out the *average* of the differences A minus B, and C minus D—in other words, direct voltages. Clearly, with the phase relationship shown between the IF signal and the switch controls (essentially the reference for the detector), there is a large output at E and no output at F. On the other hand, at a later time, if the IF is not exactly 5 MHz, the phase relationships will have altered, and with them the voltages at points E and F. A full analysis shows that a normal I-Q demodulator has been made where the outputs at E and F depend on the cosine and sine of the phase differences between the reference and the IF. The key point is that because of the exactly repetitive sampling, the two outputs are exactly 90° apart in phase. Further, because the instrumentation amplifiers are high-quality, low-noise devices with little offset and drift, the outputs are highly stable.

The circuit is not without its subtleties. Like many demodulators, it responds to odd harmonics of the IF, so a good filter (not shown) is needed ahead of the demodulator to remove any higher-frequency spurious signals or noise. Second, the RF switches must pass only small currents to avoid nonlinear, on-resistance effects; thus, 10 kΩ resistors are included between the switches and the instrumentation amplifers to minimize high-frequency cur-

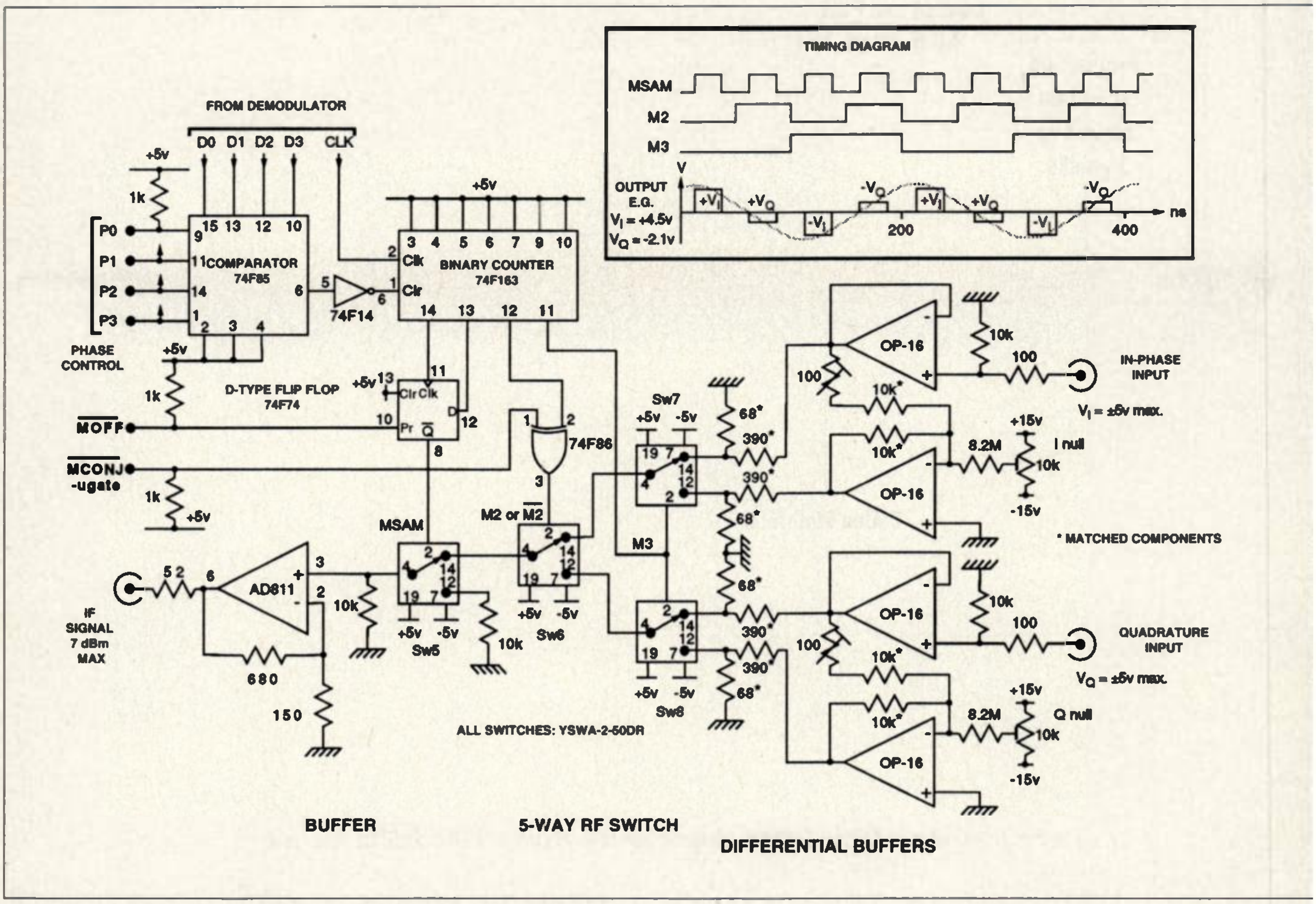

Figure 5. The I-Q modulator circuit.

rents that could be passed by the amplifiers' input capacitances. On no account should capacitors-to-ground be placed at points A through D in an attempt to filter the signals. The capacitors charge and discharge through the switches, and if the various time constants in the two channels differ, the average voltage output amplitudes also will differ. At first sight, this behavior seems bizarre, because any lowpass filter surely must give the average voltage as its output. But, because switching can change the source impedance, a simple linear analysis cannot be applied. The circuit is non-linear, and the high-frequency behavior "mixes" with the low-frequency behavior.

At higher frequencies, the phase difference can depart from 90° if the instrumentation amplifiers are unmatched. Lack of matching causes errors to be greatest at high frequencies close to the amplifer cut-offs of about 100 kHz. On the other hand, the greatest dynamic range is obtained by running the ADCs at or close to their maximum sampling rate, and then using digital signal processing methods to reduce the bandwidth to that desired. For this reason, a compensation network is included with the gain resistor R_g of each amplifier. Adjustment of the capacitor C_m changes slightly the mid-frequency gain and phase (at about 50 kHz) while the value of the resistor R_c determines the bandwidth.

With the values shown, the conversion gain is about +17 dB with a maximum output of ±5 V (see below) and a cut-off frequency (–3 dB) of 250 kHz. The addition of resistor R_c not only tailors the frequency response to the ADC Nyquist frequency, but also prevents 5 MHz components in the switched signals from overloading the front ends of the amplifiers. The spectrum of magnetic resonance signals (a range of 10 kHz was mentioned earlier) typically covers less than the available ADC bandwidth of 500 kHz, and within the spectral range, the phase error was less than our current measurement accuracy of ±0.2°. (A 16-bit data acquisition system is being built.)

A 1% gain trim is included with the instrumentation amplifiers, and this allows the gains of the two signal channels to be set to considerably better than 0.1%. The DC offsets on the outputs were –2.7 mV and –0.3 mV and changed by less than 0.3 mV when the demodulator was turned off with the aid of control line DOFF/. If desired these offsets easily can be annulled by the addition of a potentiometer to the AMP-01s' nulling connections, but an alternative method of dealing with them is described below. Finally, the mostly low-frequency noise on the outputs was less than 200 µV.

An I-Q modulator and precision RF filter

Exactly the same sampling principles can be used in reverse to build a precision, narrow-band I-Q modulator. The schematic is shown in Figure 5

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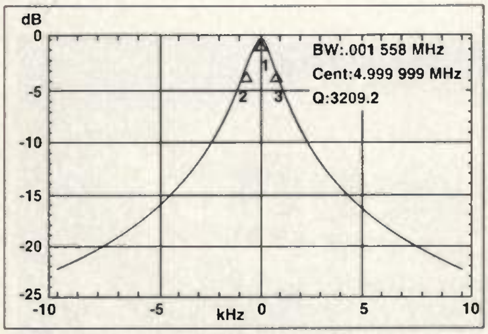

Figure 6. The transfer function of the down-up conversion passband filter.

Once again, three control lines MSAMple, M2 and M3 govern the switches, selecting in turn, from the OP-16 buffers, voltages $+V_I$, $+V_Q$, $-V_I$ and $-V_Q$ to be presented to an AD811 buffer amplifier. This is followed by a six-pole elliptic bandpass filter (not shown) at the intermediate frequency. Note that the OP-16 outputs are attenuated as a simple means of protecting the switches from excess voltage. The in-phase and quadrature components of the 5 MHz signal are accurately 90° apart, and the trims on the high-quality op-amps allow adjustments of the two components' amplitudes to better than 0.1%.

The output of the modulator was adjusted to 7 dBm into 50 Ω, obtained when either input voltage was ±5 V, and the non-linearity at this level, as assessed by a test described below, was less than 0.5%. It is possible to change the phase of the modulator output in discrete increments, relative to the phase of the I-Q demodulator reference, by clearing the counter in Figure 5 at different points in the counting cycle of the counter in Figure 4. The 74F85 comparator compares the state of the demodulator count D0–3 with a 4-bit control byte P0–3 and when the two are the same, the 74F163 counter is reset. This allows the phase of the modulator output (and ultimately that of the transmitter) to be set in 22.5° increments and is the reason for the 80 MHz reference frequency. The maximum frequency tolerated by the logic circuitry was about 100 MHz, so if the incremental phase-shifting is not desired, the IF can be as high as 12.5 MHz.

An unusual application of the modulator and demodulator together is to make an RF bandpass filter with a specific symmetrical transfer function that can be as narrow as desired. The minimum bandwidth is determined only by the stability of the 80 MHz local oscillator. The trick is to pass the I and Q outputs of the demodulator through identical lowpass filters having half the desired transfer function and thence to the two inputs of the modulator. Figure 6 shows the transfer function obtained with a network analyzer when two simple resistor-capacitor (RC) filters of nominal time constant 200 µs were used. The transfer function is Lorentzian with the expected width at the –3 dB points of 1.56 kHz and an effective Q-factor of 3,200. Of course, one must be careful not to reverse the I and Q connections. This will cause the frequency spectrum to be reversed, and an input at 5.001 MHz will be passed as 4.999 MHz with interesting consequences. The error can be corrected by asserting low either the MCONJ/ or DCONJ/ signals, which reverse the operation of the middle switches and generate the complex conjugate signal, albeit with a phase shift.

A sensitive test of the combined modulator and demodulator can be made with the above configuration at maximum bandwidth, i.e., no RC filters, and with the IF input set at various frequencies $\delta\omega$ + 5 MHz. Any gain or phase imbalance in the two audio frequency chains of the combined system will result in amplitude modulation of the output at a frequency of $2\delta\omega$. It easily is seen that a difference of α in the amplitudes of the two channels produces a peak-to-peak modulation of $\alpha \cos(2\delta\omega t)$, while a phase error of β radians gives a peak-to-peak modulation of $\beta \sin(2\delta\omega t)$. Triggering an analog oscilloscope off one of the demodulator outputs and viewing the bandpass-filtered 5 MHz modulator output at high oscilloscope gain, no modulation at frequency $2\delta\omega$ could be seen, to an accuracy of 50 dB between DC and $\delta\omega$ = 10 kHz. At 100 kHz, an error of 5 mV in 1 V that could not be removed by gain adjustment was seen, implying a phase error of 0.5°. Reversal of the modulator inputs gave the same results. Although it is possible that phase errors in the demodulator and modulator could cancel, this is unlikely, and the results confirm that the sampling principle gives excellent phase performance. This test also revealed at full output power 0.5% second-order distortion giving modulation at frequency $4\delta\omega$. This disappeared with 3 dB attenuation. A note of caution with regard to the measurement: because the modu- lator output is designed to be fed into a 5 MHz filter, no serious attempt was made to zero the common mode output of the differential buffers. Any such output is fed directly to the output buffer, and because the latter is DC-coupled, if the filter is lowpass rather than bandpass, measurement accuracy can be compromised.

Distortion and its removal

Over the desired bandwidth of ±10 kHz, the sampling demodulator described above effectively reduces the ghost and DC artifacts to less than –60 dB relative to the signal, the residual error mainly being caused by drift in the gains of the two channels. This is a good performance, but in the face of a possible (and probably unattainable) dynamic range of 120 dB, it is not good enough. Indeed, when confronted with such a dynamic range, we also must seriously consider the effects of distortion. Thus, the buffer amplifier prior to the switches is a high-speed, low-noise video op-amp (AD 811) with a high third-order intermodulation distortion intercept (IMDI) at 5 MHz of about 47 dBm. The buffer consistently sees a high impedance from the switches, for the switches are terminated with at least 10 kΩ at 5 MHz. In this manner only a small current flows through them to eliminate the distorting effects of current-dependent switch on-resistance.

Using a carefully constructed two-tone test [3] with signals at ω_1 = 5.015 MHz and ω_2 = 5.021 MHz, each having a power of 1 dBm into 50 Ω, all distortion products were at least 60 dB down—the limit of our current measuring ability. Even so, with a power increase of only 2 dB, many harmonic spurs were seen, those dominant being fifth-order at only 44 dB less than the tones. Thus, no accurate measure of third-order intercept could be made nor, in fact, would such a measure be meaningful. Not surprisingly, the source of the distortion was the RF switches that were beginning to bottom on the negative RF signal peaks. Accordingly, a limit of 7 dBm was set as the maximum input power. With the RF switch insertion loss of 12 dB and the amplifier gain set to 29 dB (× 30, the value shown in the schematic), a maximum permissible output of 24 dBm, or about 10 V p-p (available without clipping over a bandwidth of ±

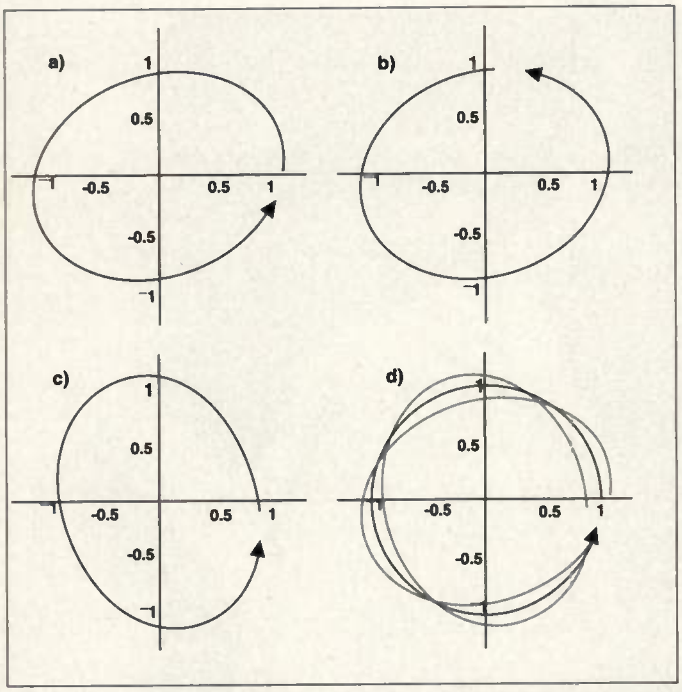

Figure 7. Correction of first-order distortion with a 90° phase shift in the signal.

100 kHz), was obtained. This is suitable for many ADCs.

When our 16-bit quadrature digitizer is complete, we should be in a better position to assess distortion further. No matter what the results, there is a technique that can be used successfully to remove distortion harmonics when small signals must be detected in the presence of only one large tone, rather than the two normally used for test purposes. Let us assume that the single, large magnetic resonance signal from water has an initial amplitude of 7 dBm at the input to the demodulator. The harmonics of the output are then about the same size as the ghost (< 60 dB), and there is a real sense in which the ghost can be considered a consequence of distortion—a minus first harmonic is created. This interpretation fits well with the graphical representation of the transient signal shown in Figure 2a. Figure 7a shows the Argand plot obtained when large I-Q gain and phase errors of about 10% are present. Normally, this should not occur, but it highlights the fact that a gain or phase error is really a form of distortion. The circle of Figure 2a has become an ellipse, and along with this distortion goes the corresponding spurious harmonic ghost signal in the spectrum, as was already shown in Figure 3.

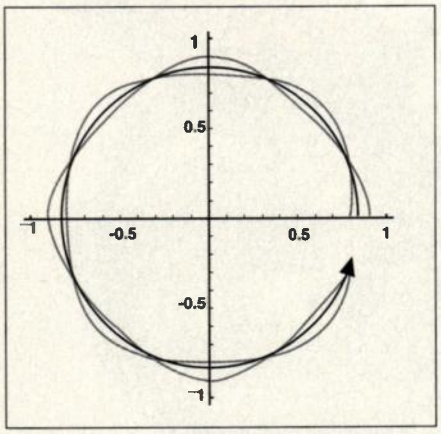

Figure 8. Correction of third-order distortion with a 45° phase shift in the signal.

Provided the ghost is small (e.g., –60 dB), the way to remove it was discovered many years ago [4,5]. The method relies on the fact that the signal, as in radar, is a response to a stimulation and contains phase information. Suppose we store the transient decay in the computer, then repeat the experiment with the phase of the transmitter changed by 90°—an easy thing to do with the I-Q modulator described above. The Argand diagram of the signal again would be an ellipse. Following the transmitter, it too would suffer a 90° phase change and would start near the y axis, as shown in Figure 7b. We again store this signal in the computer. But now, we digitally rotate it –90° (Figure 7c) so that it is back in phase with the original transient. We then add it to the latter. As may be seen from the average in Figure 7d, the result is a perfect circle, and the ghost vanishes. (A little thought will show that the three lines not intersecting perfectly is correct.) Mathematical analysis is simple, though tedious. In the first experiment, we may write the two outputs of the demodulator as:

$$\xi_{I1} = A(1+\alpha)\exp\left\{\frac{-t}{T_2}\right\}\cos(\delta\omega t+\beta);$$
$$\xi_{Q1} = A(1-\alpha)\exp\left\{\frac{-t}{T_2}\right\}\sin(\delta\omega t-\beta) \qquad (5)$$

Here, α is the mean fractional gain error and β the mean phase error in the two signals. In complex form, this first transient is then:

$$\xi_1 = A\exp\left\{\frac{-t}{T_2}\right\}\{\exp(j\delta\omega t)+\exp(-j\delta\omega t)(\alpha-j\beta)\} \qquad (6)$$

where we have assumed that α and β are << 1. For the second experiment,

$$\xi_{I2} = A(1+\alpha)\exp\left\{\frac{-t}{T_2}\right\}\cos\left(\delta\omega t+\beta+\frac{\pi}{2}+\gamma\right);$$
$$\xi_{Q1} = A(1-\alpha)\exp\left\{\frac{-t}{T_2}\right\}\sin\left(\delta\omega t-\beta+\frac{\pi}{2}+\gamma\right) \qquad (7)$$

Phase γ is a small error possibly introduced by the modulator—we assume it does not change phase by exactly 90°. In complex notation, and with the (exact) back rotation of 90° introduced by the computer, the second transient is:

$$\xi_2 = A\exp\left\{\frac{-t}{T_2}\right\}\{\exp(j\delta\omega t)(1+j\gamma)-\exp(-j\delta\omega t)(1-j\gamma)(\alpha-j\beta)\} \qquad (8)$$

When equations (6) and (8) are added, we obtain:

$$\xi = A\exp\left\{\frac{-t}{T_2}\right\}\{\exp(j\delta\omega t)(2+j\gamma)-\exp(-j\delta\omega t)(j\gamma)(\alpha-j\beta)\} \qquad (9)$$

and the amplitude of the ghost—the term in $\exp(-j\delta\omega t)$—is, to first order, a factor $0.5\gamma(\alpha^2+\beta^2)^{1/2}$ less than that of the main signal. In other words, if α, β and γ are of the order of 10^{-3}, then the ghost is of the order of 120 dB down relative to the signal, and we have accomplished our objective. In particular, high accuracy in the transmitter phase shift ($\gamma \cong 0$) can override gain and phase errors in the demodulator.

Considering the ghost to be caused by distortion gives insight into the reduction of other distortion products [6]. For

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example, Figure 8 shows, greatly exaggerated, the effects of third-order distortion on the Lissajous display of the strong water signal. For a single signal at audio frequency $\delta\omega$, the distortion produces a third harmonic at frequency $-3\delta\omega$. Here it is clear from the figure that a rotation of 45° in the transmitter, followed by a back rotation of 45° in the computer, will tend to cancel the distortion. This principle of Fourier synthesis relies on the fact that if a signal changes phase by ϕ, its n^{th} harmonic changes phase by $n\phi$. A subsequent phase correction of $-\phi$ restores the original phase of the signal, but leaves the harmonic with a phase of $(n-1)\phi$. If this is 180°, then the harmonic cancels. In general, to cancel n^{th} order harmonics, phase shifts of:

$$\phi = \frac{(2p+1)\pi}{(n-1)}; \; n \neq 1 \tag{10}$$

are needed, where p and n are positive or negative integers, including zero. "Zeroth order distortion" is not excluded. A zeroth order signal is a DC offset. Repeating the transmitter pulse with inverted phase followed by subtraction of the new transient from the old removes the offset. (This explains why no effort was made to annul the small offsets on the demodulator outputs.)

Figure 9a shows the frequency spectrum obtained when two tones, one large at ω_1 = 600 Hz and one small at ω_2 = –650 Hz, pass through a non-linear demodulator. The true small signal is impossible to discern. Even so, when eight data sets, having phases in 45° increments, are phase-corrected in the computer and then averaged, the resulting spectrum is as shown in Figure 9b, and the true signals are revealed. Of course, if the large signal can be placed "on resonance" at 0 Hz, harmonics do not arise, but this strategy cannot always be adopted. It might have been expected from Equation 10 that a phase shift of 60° would have been used to cancel harmonics of order n = –2. However, advantage has been taken of the integer p in the equation. By setting it to three, the more convenient value of 180° has been used. As many higher-order harmonics as is necessary can be annulled with this technique of cyclically ordering the phases. (CYCLOPS is an acronym commonly used in magnetic resonance.) The 45° increments used above cancel as high as sixth order. Using 16 22.5° increments (the value chosen for digitally phase-shifting the I-Q modulator described earlier), as high as 14th-order distortion may be annulled, while 11.25° (π/16) increments annul as high as 30th order. Unfortunately, the method does not cancel spurious signals whose phases change by the same amount ϕ as that of the transmitter. For example, the satellites produced in a third-order intermodulation distortion test are at frequencies $2\omega_1$-ω_2 and $2\omega_2$-ω_1, and the net phase change here is 2ϕ-ϕ = ϕ. On the other hand, if one tone (ω_1) is larger than the other (ω_2), as is common in magnetic resonance, only one satellite easily is visible at frequency ($2\omega_1$–ω_2). If the latter is x dB down, the other satellite is 2x dB down and normally in the noise. The visible satellite, which can be seen at +1.85 kHz in Figure 9b, easily is distinguishable as such, because it is negative. (A small positive offset was deliberately left in the plot.) This is a consequence of the minus sign that normally accompanies the odd order terms in a Taylor expansion of the transfer function of an overloaded electronic device.

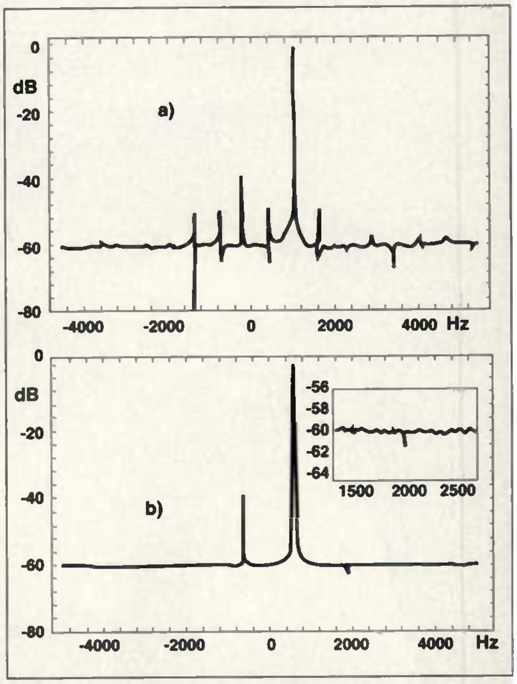

Figure 9. a) Distorted two-tone spectrum, b) after use of an eightfold phase sequence.

Conclusion

With careful design and the use of modern fast RF switches, the disadvantages of I-Q demodulation—namely lack of orthogonality of the two outputs, unequal gain and DC offsets—can be minimized. Slower, high-resolution ADCs with superior dynamic range can be employed. When the phase of an incoming signal can be controlled, it also has been shown that harmonics caused by distortion (including the ghost signal caused by lack of demodulator orthonormality) can be removed by appropriate digital manipulation. Although the details of the latter method have been known to magnetic resonance practitioners for some years, they appear not to be known widely outside that discipline. I hope that this article will help to rectify the situation. Finally, the use of the demodulator and modulator in combination with audio frequency filters has been shown to create at the intermediate frequency a narrowband RF filter with characteristics as easy to control as those of any audio frequency filter. *RF*

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David Hoult is a senior research officer with the National Research Council in Winnipeg, Manitoba, Canada. He received a B.A. and D.Phil. from Oxford University. The author of numerous papers on the physical engineering of magnetic resonance, he is one of the original investigators of MR imaging.

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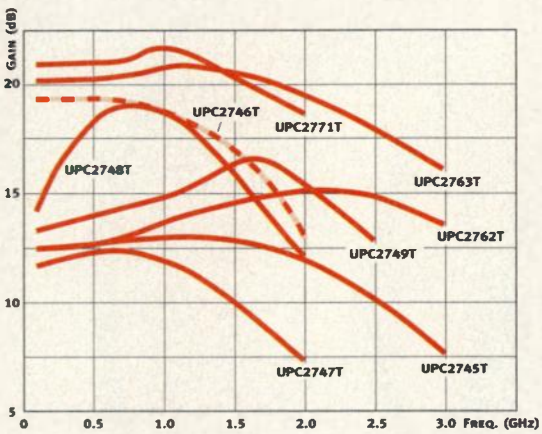

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UPC2748T	200 MHz–1.5 GHz	19	2.8	-8	6	900 MHz
UPC2749T	100 MHz–2.9 GHz	16	4	-12.5	6	1.9 GHz
UPC2762T	100 MHz–2.9 GHz	14.5	7	7	27	1.9 GHz
UPC2763T	100 MHz–2.4 GHz	19.5	5.5	6.5	27	1.9 GHz
UPC2771T	100 MHz–2.1 GHz	21	6	11.5	36	900 MHz

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0.01–200	AU-1442	35	0.5	2.0:1	1.2	1.2	1.2	5	$315
0.01–200	AU-1447	56	0.5	2.0:1	1.2	1.2	1.2	12	$325
0.01–500	AU-1310	30	0.5	2.0:1	1.3	1.4	1.5	8	$325
0.01–500	AU-1332	45	0.5	2.0:1	1.3	1.4	1.5	10	$350
0.01–1000	AM-1300	27	0.75	2.0:1	1.4	1.6	1.8	8	$350
0.01–1000	AM-1431	35	0.75	2.0:1	1.4	1.6	1.8	10	$375
0.02–1000	AM-1551	38	1.0	2.0:1	1.4	1.6	1.8	16	$375
1–100	AU-3A-0110	53	0.5	2.0:1	1.2	1.2	1.2	12	$300
1–200	AU-1464	35	0.5	2.0:1	1.2	1.2	1.2	6	$275
1–200	AU-1494	56	0.5	2.0:1	1.2	1.2	1.2	12	$300
1–500	AU-2A-0150	30	0.5	2.0:1	1.3	1.4	1.5	8	$275
1–500	AU-3A-0150	45	0.5	2.0:1	1.3	1.4	1.5	10	$300
1–500	AU-4A-0150	60	0.5	2.0:1	1.3	1.4	1.5	10	$325
1–1000	AM-2A-000110	27	0.75	2.0:1	1.4	1.6	1.8	8	$300
1–1000	AM-3A-000110	35	0.75	2.0:1	1.4	1.6	1.8	9	$325
1–1000	AM-4A-000110	52	1.0	2.0:1	1.4	1.6	1.8	9	$350
5–300	AU-1021	24	0.5	2.0:1	2.4	2.5	2.7	20	$275
5–300	AU-1525	61	0.5	2.0:1	1.2	1.2	1.3	20	$350
100–1000	AM-1412	35	0.75	2.0:1	1.4	1.6	1.8	15	$350
100–2000	AM-1526	9	1.0	2.0:1	4.5	4.0	4.2	20	$350
200–2000	AMMIC-1427	20	1.5	2.0:1	4.2	4.3	4.6	14	$450
500–1000	AM-2A-0510	24	0.5	2.0:1	1.4	1.5	1.6	0	$300
500–1000	AM-3A-0510	36	0.5	2.0:1	1.4	1.5	1.6	9	$325
500–2000	AM-3A-0520	30	0.75	2.0:1	1.4	1.9	2.4	3	$395
1000–2000	AM-3A-1020	30	0.5	2.0:1	1.8	2.1	2.4	10	$350
1000–2000	AM-1477	39	1.0	2.0:1	1.8	2.1	2.4	15	$350
1000–2000	AM-4A-1020	40	0.75	2.0:1	1.8	2.1	2.4	10	$375

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INFO/CARD 60

Global positioning systems—technology into the 21st century

By Ernest Worthman
Contributing Editor

The Global Positioning System (GPS), long the domain of government and law enforcement, is beginning to come into the mainstream. Seven years ago, some of the first hand-held GPS receivers from Trimble Navigation drew a lot of attention at the International Mobile Communications Expo (now the International Wireless Communications Expo). Once reserved for only high-end geographic applications, this technology is finding its way into K-Mart, Wal Mart, sporting goods stores and mail-order catalogs.

Cutting-edge technology in component development and large-scale integration, as well as new markets, have brought about this change.

Satellite-based GPS technology relies upon differential satellite positions to pinpoint locations. The navigation satellite timing and ranging (NAVSTAR) system was built and is maintained by the U.S. Department of Defense. The system employs 24 satellites, plus two spares, operating in six orbital planes 20,200 kilometers (10,900 nautical miles) above the earth at inclination angles of 55 degrees with a 12-hour period. The satellites are positioned precisely so that a minimum of four will always be in view of the user with a position dilution of precision (PDOP) of six or less.

Each satellite transmits on two L-band frequencies, L1 (1,575.42 MHz) and L2 (1,227.6 MHz). The satellites transmit on exactly the same frequency. Even so, each satellite's signal is Doppler-shifted by the time it reaches the user. L1 carries a precise (P) code and a coarse and acquisition (C/A) code. L2 carries only the P code. A navigation data message is superimposed on these codes. The same navigation data message is carried on both frequencies. The P code normally is encrypted, so that only the C/A code is available to civilian users. Some information can be derived from the P code. When encrypted, the P code is known as Y code.

To locate positions, readings are taken from different satellites. Three satellites can be used for a two-dimensional position fix, although four are required for height above terrain (HAT). The receiver takes the data from each of the four known transmission sources (the satellites) and determines the time it takes for the signal to arrive at the receiver. This range is called the pseudorange. With this data known, the formula in Figure 1 is used. Solving for the four unknowns, Ux, Uy, Uz, and Cx (clock bias) will provide a differential time element based upon the time it took for the signals to reach the receiver, relative to each satellite's position. The data received determine the device's position in latitude, longitude, altitude and time. These data are then converted into latitudinal and longitudinal coordinates from which position, speed and time can be determined. Because the elements are time, speed and distance, GPS works in a three-dimensional plane, allowing for land, sea and air applications.

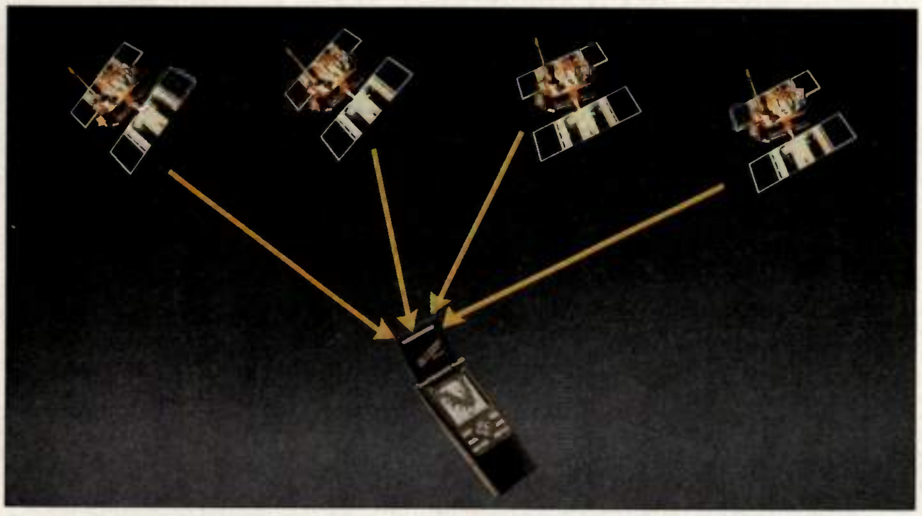

Figure 1. The GPS satellites transmit position data signals and a GPS receiver captures them and computes the time it took for the signals to reach it. From these calculations, four satellites will provide enough data to pinpoint a single position location.

Accuracy levels

GPS comes in two flavors: standard positioning service (SPS) and encoded precise positioning service (PPS). The PPS is a non-reduced service, primarily used by agencies that require extremely accurate readings, such as the military, emergency services, public works, fleet vehicle services and United States Geological Survey (USGS) mapping. PPS is capable of providing accuracy of 16 meters or less in positioning, 0.1 meters in velocity and 90–100 nanoseconds in time.

The SPS commonly is used in the commercial environment and provides position only to an accuracy of 100 meters. SPS signal accuracy intentionally is reduced by the government to protect U.S. national security interests. This reduced accuracy, called selective availability (SA), controls the availability of the system's full capabilities.

As is often common to encryption or encoding technology, methods soon develop to defeat these schemes. Commercial manufacturers have developed a way to eliminate most of the effect of SA by what is called differential GPS (DGPS). DGPS adds stationary receivers at known points on the ground. The receivers transmit low-

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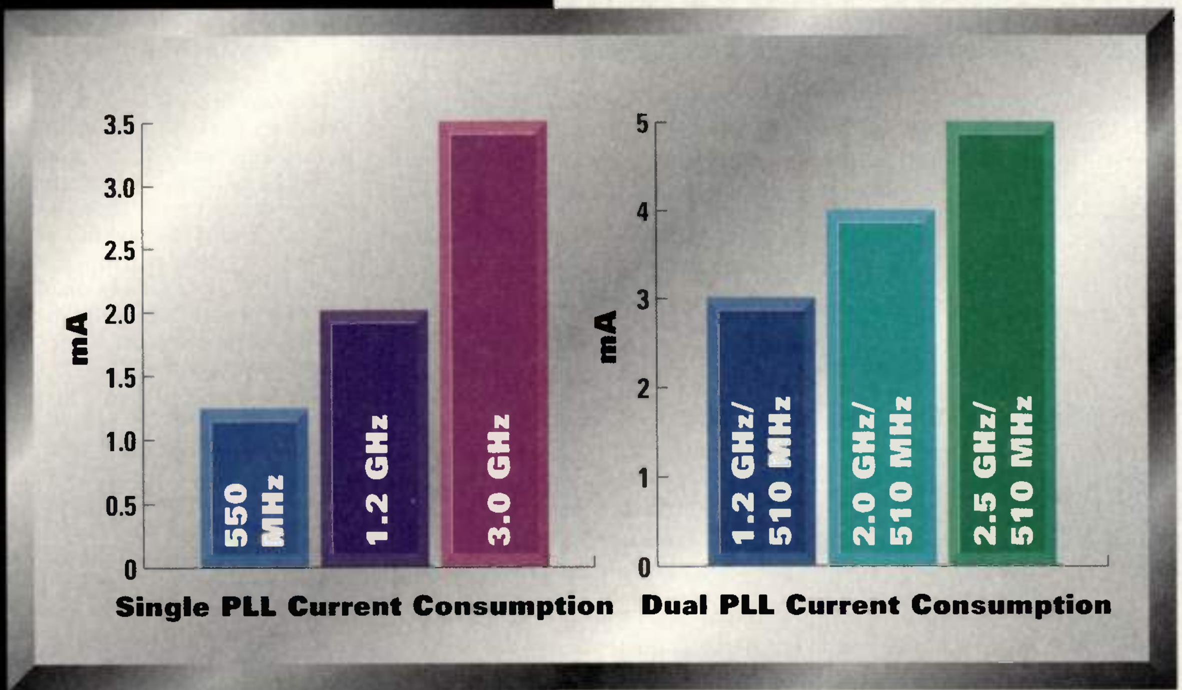

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rate (100 bits per second [bps]) data at a frequency of about 300 kHz. These transmitters relay error-correcting data to the receiver and improve accuracy to 10 meters or better. Accuracy also can be affected by ionospheric conditions and ground clutter such as woods, buildings and multipath signal reflections. Under ideal conditions, GPS receivers are limited to an accuracy of about 2 meters.

Technology

For a long time, commercial GPS receivers were stand-alone components. The end user would be required to make the receiver integrate with his application. Today's GPS systems are becoming more complete. When a GPS system is deployed, it tends to include support data, such as navigation maps and post-processing software. For example, a commercial GPS unit sold to the automotive market may contain not only the receiver, but also localized data maps of city streets for the locality in which the unit is being sold. It also may contain a car power adapter kit, external antenna and even software to allow the user to modify the database. In this application, the GPS becomes a component of a complete product. This trend also is being seen in commercial markets such as airline navigation and law enforcement.

Greg Turetzky, a product manager at SiRF Technology, a GPS front-end component manufacturer, says the approach that SiRF is taking is to apply very large scale integration (VLSI) techniques to shrink the module and lessen component count. From a design standpoint, the GPS modules of the receiver comprise two chips, an RF converter module, largely made up of bipolar components and the complementary metal oxide semiconductor (CMOS) digital signal processor (DSP). The front-end section is the RF portion that takes the signal, which can be as low as –160 dBm at approximately 1.5 GHz, processes it and converts it to digital output of a few hundred millivolts. The digital output is fed to the DSP section. The function of the hardwired DSP is to remove the spread-spectrum encoded portion of the signal and to convert it into basic in-phase and quadrature (I and Q) for running track loops.

This "component only" approach is typical of many manufacturers in many industries in which a standard component approach is being developed. Turetzky points out that SiRF is attempting to make the inclusion of a GPS module easy for the product developer. SRF's approach "is to make it easy for people who aren't skilled in RF design to be able to add this to their product." By adding a handful of external components, such as filters, reference crystals and bias components, for example, a complete GPS front-end is made available. It is this technology approach that opens doors for commercial GPS.

As this technology develops, it is likely that we will see an integrated RF and DSP unit that will offer advantages such as improved signal-to-noise

ratio, simplified interconnect and better immunity to environmental interference.

Applications and markets

In the early days of GPS, the principle markets were surveying and timing. Before GPS systems were used for marine navigation, the compass and the Loran coastal navigation system were the only ways mariners could navigate. The compass was not particularly accurate, and Loran only worked in coastal waters. GPS provided a one-stop application for both open seas and coastal navigation and even provided a precise location device for inland wate bodies. A secondary benefit was tha the user did not have to learn to navi gate by stellar. It was necessary only t take a GPS position fix, look at th charts and point the boat to a compas heading. GPS was simpler than previ ous methods.

GPS is beginning to find its way int people and package tracking. This typ of application has a unique set of issue that other tracking objects do not First, these objects are not always out doors. Second, the object is not alway moving. Once a person or a package i surrounded by massive brick and stee structures, signal reliability drop markedly. And if the object is not mov ing, there is no Δt. The challenges tha GPS faces in this arena are to try an improve sensitivity to these minima signal levels, which may be –160 dBn or greater, and to intelligently sens whether the object has stopped moving A rather ambitious project would be t attach a GPS receiver to package delivered by courier services.

Another technology area that is likel to see the implementation of GPS i personal communications. Cellula phones are a prime candidate for inte grated GPS. In an emergency, dialing 9 1-1 from a cellular phone does nothin to determine the actual location of th phone (unlike landline phones, which can be traced). According to one indus try source, the FCC is putting out a mandate that will call for a certain per centage of cellular 9-1-1 calls to b locatable to within a defined area with in a set time period. GPS in-building un reliability, blocked signal areas an movement requirements are likely t cause debate over this type of technology

The obstacles

Although there have been a grea number of advancements, GPS technol ogy needs further refinements. To us an industry vernacular, GPS signa strength at the receiver is in the mud Imagine you are an integrated, ultra sensitive receiver in a cellular tele phone looking for a signal at –160 dBm All of a sudden a 3 W, 900 MHz cellula transmitter, only a few inches from you, fires off. Even though the trans mitted frequency is 600 MHz away, th power radiated at the proximity of th receiver is close enough to affect an sometimes damage the sensitive front end of the GPS receiver. Even with today's high isolation and rejectio

devices, the super-sensitive GPS receiver is always affected by the proximal power.

GPS manufacturers face power consumption also. Today's GPS receivers have a steady-state power drain of around 750 mW. Hand-held GPS receivers usually use 6 V battery configurations (1.5 V × 4 cells). Some quick math reveals that the current drain is 125 mA. Although this does not seem like much of a drain, add in the display (popularly the liquid crystal type) and some other support electronic, such as DSP, and it becomes an issue. To improve upon time between batteries (TBB), various power-saving scheme are supported, such as powering dow the components when the unit is not i actual receiving mode, or after a specif ic period of non-use is sensed. Even sc to maintain updated positioning infor mation, the receiver still has to mak periodic calculations, even when th unit is in standby mode. Ways to re duce power further are still high on th list of developmental projects.

The sensitive nature of current RI technology still requires careful shield ing, grounding and component place ment. Although digital component have relative immunity to spuriou emissions, high-frequency RF circuitr does not have that luxury. High-spee microprocessors running at 100 MH and higher can wreak havoc with ultra sensitive RF sections.

The future

As was mentioned earlier, most o the movement in the GPS industry is i trying to improve receiver sensitivit and to deal with stationary receivers This translates into the issue of reliabl signal levels. At a certain level, th receiver loses phase lock but not neces sarily the signal itself. Although main taining phase lock is required for guar anteed accuracy, some of the industr designers are looking at methods tha can sense signal, even though it is to low for the phase-lock-loop (PLL) cir cuitry's threshold. Even at signal level of –180 dBm and below, if the signa can be identified, some information stil can be gleaned. And if this informatio can be identified by using digita enhancement technology, it is possibl to make it useful in position location As Turetzky indicates, part of the new technology is a paradigm shift in per ception.

If the position fix is not accurate to 1 meter, but is accurate to 100 meters does that make the data unreliable? I you were hiking and got lost in the woods, would you throw away you GPS receiver just because it could fi only to 100 meters rather than to 10 meters, or would you rather have the rescuers looking all over the mountain? Even though the information is not 100% reliable, it is not useless. Consider today's cellular technology. If it had to be 100% reliable, none of us would be using it.

GPS appears to be a means to an end, rather than the end itself. As a stand-alone product, latitude and longi-

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tude are not hot items outside of surveying and navigation. But as an integral part of 21st century technology, GPS has a bright future. For example, consider using a GPS receiver to keep your portable computer's time set or as a small light emitting diode (LED) on your automobile's instrument cluster that will locate you on any street or highway and then reference the data to the local geography and populous.

As with any high-technology industry, GPS is benefiting from the developments in surface-mount technology (SMT), VLSI, low-power devices and other advancements. Another area of development is in embedded processors and microprocessors in general. It is likely that many of the functions, once exclusively the domain of discrete components, can be emulated and will be provided in a digital format as processor speeds approach 500 MHz and eventually break the GHz barrier. Faster processors will allow the elimination of much of the frequency manipulation and conversion techniques we now use to bring the frequencies into workable digital range. The future chipsets likely will look like a universal asynchronous receiver and transmitter (UART), just another component that gets called into use when needed. Microsoft is designing a position navigation application program interface (PNAPI), so computers with the operating system can position-tag data if needed, and GPS manufacturers can easily design software driver interfaces for their products.

The final limiting technology in GPS is the satellite system itself. Although it is relatively easy to update earth-bound hardware, to do the same to satellites involves launching new ones (or going aloft and updating the existing ones). Either way, it is an expensive proposition, but the power of the commercial market is one that is hard to ignore. RF

What's new in GPS?

Low-cost GPS technology may be the next opportunity for consumer GPS systems. At any time, under ideal conditions, a GPS receiver could gather data from 12 satellites. On the other hand, the system is designed to permit precise positioning calculations with data from only four. Somewhat less precise positioning can be obtained using three satellites. The application of digital technology and VLSI, coupled with innovative design, makes it possible to continue calculating position "fixes" with only a single satellite's signal.

Single-satellite positioning is critical for GPS use in urban areas. Automobiles equipped with GPS-based accessories often travel on narrow streets surrounded on two sides with tall buildings. Such "urban canyons"

can obliterate reception of more than one satellite's signal. With single-satellite positioning, positions can continue to be fixed and are corrected as soon as an automobile enters an intersection, permitting multiple-satellite signal reception.

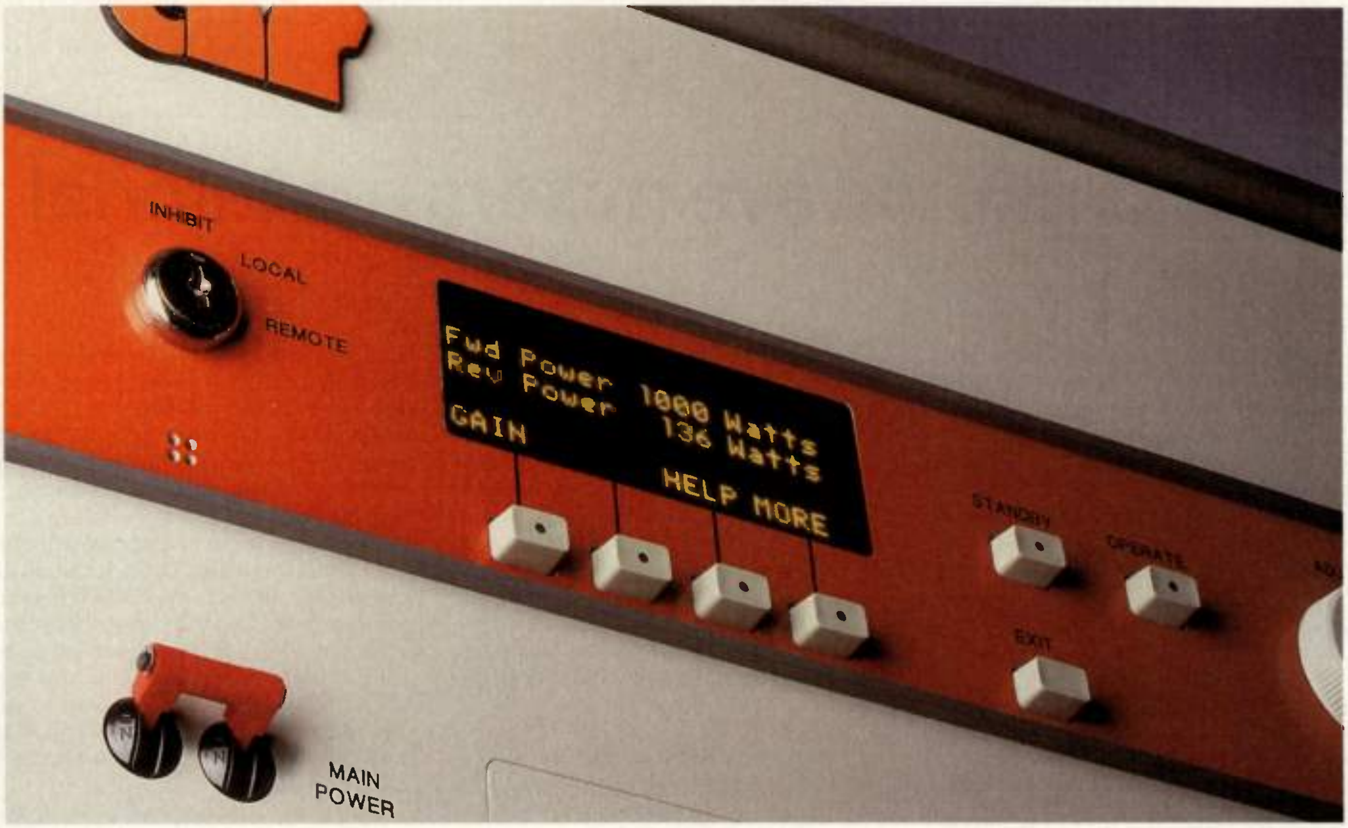
A key to this capability is the receiver's ability to access quickly satellite signals while the automobile proceeds through the intersection. As a result of the fast movement of vehicles combined with the canyon effect, acquisition technology must be designed to achieve re-acquisition in as little as 100 msec, much faster than the more typical two to three seconds required by other technologies. To achieve such quick re-acquisition time, new technological developments, such as parallel spectrum search involving 20 code samples at a time, are needed, and parallel spectrum search is coming on the market.

Another problem associated with urban areas is multipath errors. Signals bounce off buildings, introducing time delays in reception and positioning errors. Dual-multipath rejection technology eliminates low-level reflected signals and filters higher-level reflected signals. Without multipath rejection, such errors can produce large-scale positioning deviation.

Finally, GPS originally was designed for use with expensive, spread-spectrum receivers. The standard signal threshold is –160 dBw. This permits receiving a signal that is much reduced in power. Consumer products with GPS capabilities often will be used in wooded or tree-lined areas, where signal levels tend to fall below this threshold. Therefore, increased sensitivity receivers allowing sensitivity 20 dB lower than the threshold are needed for these conditions. Thus, signals that are indistinguishable to other GPS receivers easily are detected.

These issues and their resolutions are only part of the developing market for GPS. The combination of GPS innovations and consumer demand is making it possible to provide fast, accurate, personal positioning information in almost any outdoor location, using handheld consumer products that sell for less than $200.

—Greg Turetsky
SiRF Technology



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RF and microwave mixers—a tutorial

By Roland K. Soohoo

Mixers are an essential and fundamental part of high-frequency communication systems. This tutorial describes a mixer's function, its theory of operation, semiconductors used, types of mixers and key specifications. This article is meant as an introduction for individuals who want a general understanding of mixer principles and operation.

A mixer is a component that acts as a frequency converter by "mixing" two input signals together to produce a desired output signal. Figure 1 shows the receiver portion of a typical RF communications system.

An RF signal is received at the antenna and typically is amplified by an RF amplifier because the amplitude of the input signal is usually small. The amplified signal is applied to one of the input ports of a mixer, and a second local oscillator signal is applied to the other input port. What emerges from the output port is a sum or difference of the input frequencies and other frequency components.

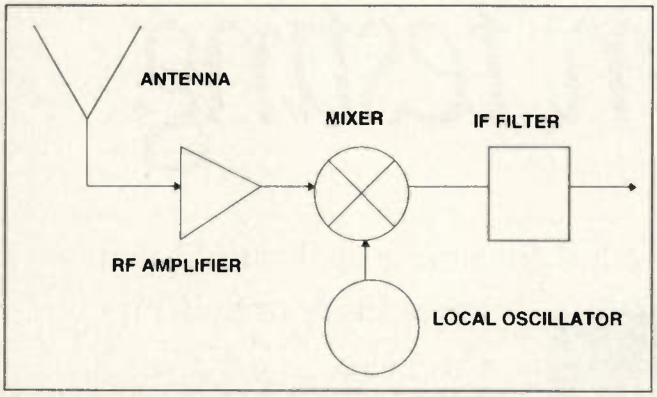


Figure 1. Typical receiver (downconverter) of an RF communications system.

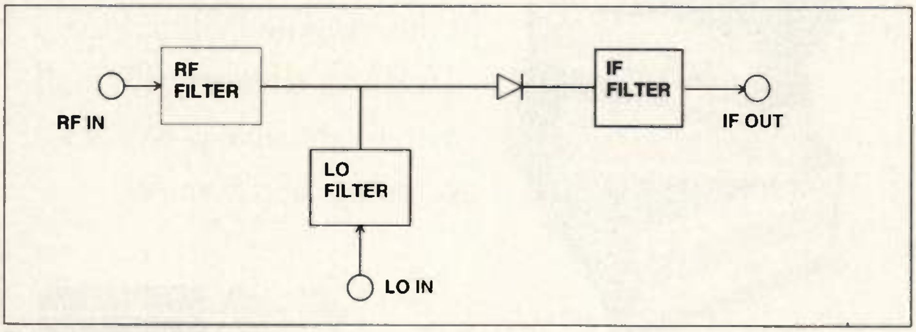


Figure 2. Single-ended mixer.

Typically, in a receiver it is the difference frequency that is the desired signal. A filter is used at the output of the mixer to remove the undesired signals and to pass the desired signal to the rest of the communication system. As an example, if an 850 MHz signal were received by the antenna and an LO (local oscillator) signal of 800 MHz were applied to the mixer, a signal at 50 MHz would emerge from the output port, which is much lower in frequency than the original signal received at the antenna.

This type of RF receiver also is known as a downconverter because it converts a higher-frequency signal to a lower one. The signal is downconverted to enable the signal to be processed at frequencies that are easy to work with. The sum frequency at the output typically is used in a transmitter. Because the RF input frequency is converted into a higher frequency, this is known as upconversion. Matching and filtering networks are used to enhance either the sum or difference frequencies, depending on the application. The mixer is fundamentally a frequency conversion device that uses the non-linear and switching properties of semiconductor devices to achieve the mixing action.

Operation theory

A mixer can use a wide variety of semiconductor devices, such as bipolar junction transistors (BJTs) and field-effect transistors (FETs), nevertheless, diodes are one of the most popular. The simplest mixer uses a single diode, but some mixers use as many as 12. A typical single-diode mixer (also known as single-ended) is shown in Figure 2.

The three ports of the mixer include the radio frequency (RF), local oscillator (LO) and intermediate frequency (IF) ports and usually are abbreviated as R, L and I. For a single-ended mixer, all the ports are connected to the same diode. Filters typically are used to electrically separate input and output signals to achieve port-to-port isolation as shown in Figure 2.

Assuming that the inputs to the mixer are sinusoids and are applied to the R and L ports and that they can be represented in the form of:

$$A \cos(\omega t) \tag{1}$$

where: A = amplitude of the input signal voltage and current and $\omega = 2\pi f$, where f is the input frequency.

The current-voltage relationship for a diode is:

$$I(V) = I_s \exp(-V_a/V_t) \tag{2}$$

where:

V_a = applied AC and DC voltages
V_t = KT/q
K = Boltzmann's constant
T = Ambient temperature
q = electron charge

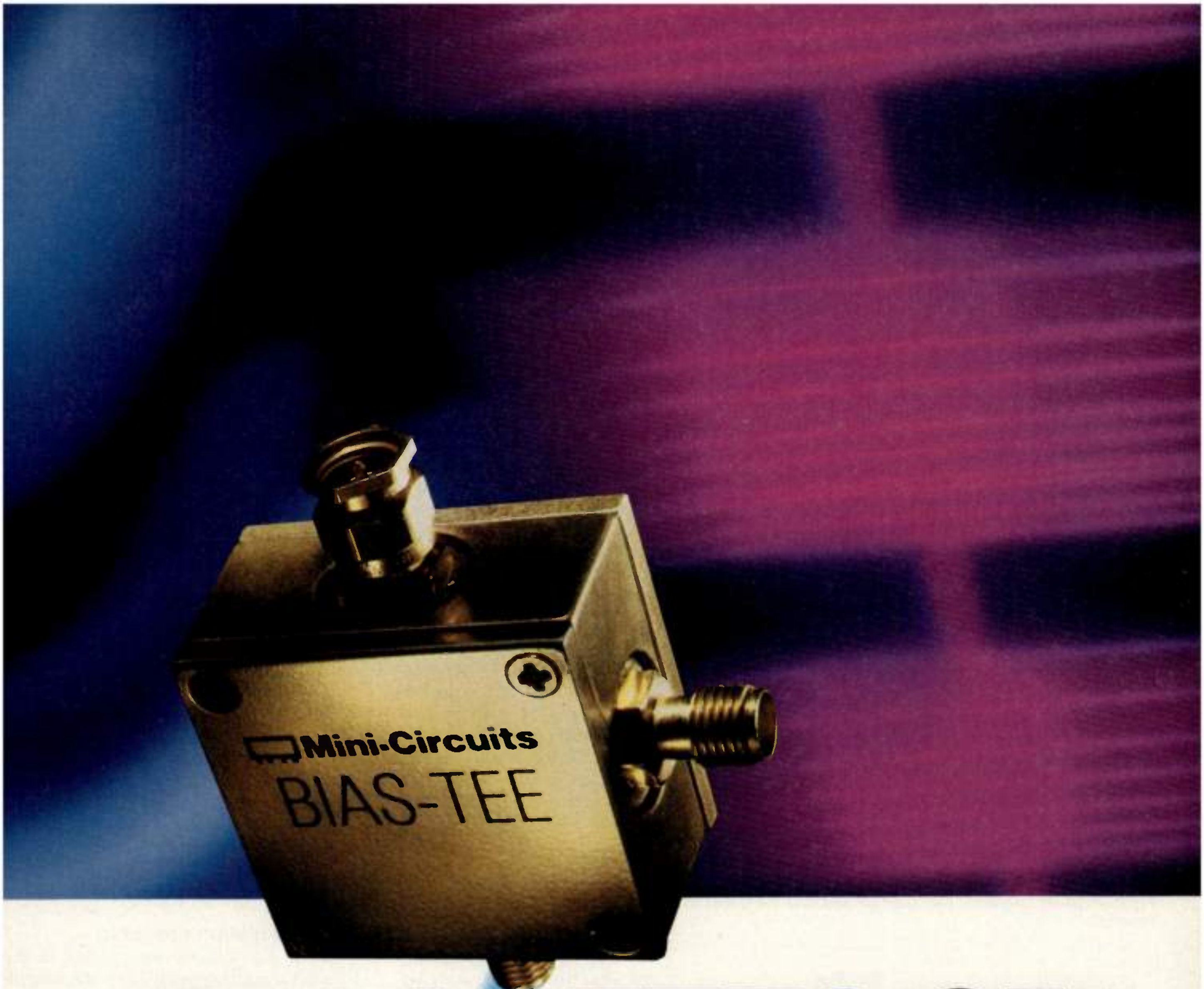
Because two sinusoidal input signals (R and L) are being applied simultaneously, the output current of the diode will be in the form of:

$$I(V) = I_s \exp(-(A_1 \cos \omega_r t + A_2 \cos \omega_l t)/V_t) \tag{3}$$

Because of the nonlinear nature of the diode, as evidenced by the exponential form of the current, intermodulation products are created. If the expression in equation (3) is expanded by using a Power series then:

$$I = I_s \sum_{n=1}^{\infty} (A_1 \cos\omega_r t + A_2 \cos\omega_l t)^n / V_t^n n! \tag{4}$$

Separating DC and AC terms:



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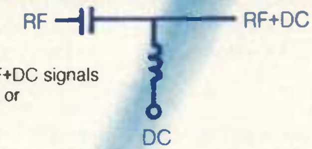

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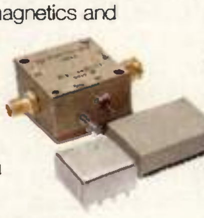

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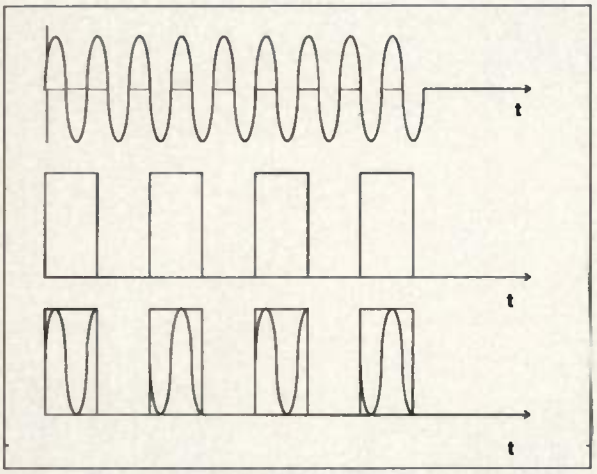

Figure 3. RF and LO ideal waveforms and resultant IF multiplication waveform from LO switching action.

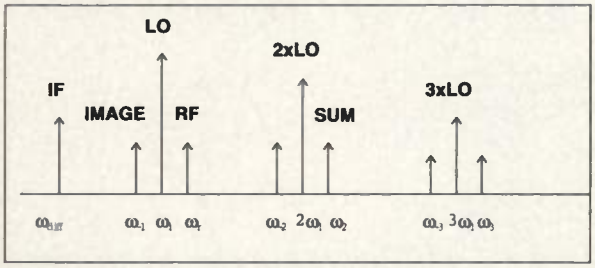

Figure 4. Mixer frequency spectrum (IF and first three spurious groups).

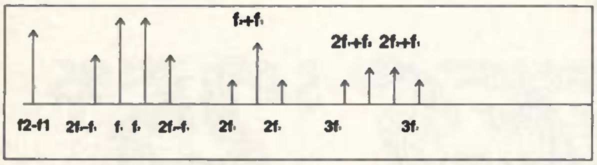

Figure 5. Mixer spectrum caused by two RF input signals and or saturation.

$$i(v) = I_o + I_s \sum_{n=1}^{\infty}(1/V_t^n\, n!)\,((A_1\cos(\omega_r t) + A_2\cos(\omega_l t)) \quad (5)$$

The magnitude of the higher-order terms of this expression will be controlled by the amplitudes A_1 and A_2 and naturally will decrease due to the n! term in the denominator, assuming small signal operation. If the Fourier transform is taken of Equation 5, the expression of the frequency components will be in the form:

$$\omega_x = m\omega_r + n\omega_l \quad (6)$$

where x, m and n are positive or negative integers, m and n can take on an infinite amount of integer values, and they represent all the different frequency components that will be contained in the mixer output signal. Obviously, some frequency components will be higher in magnitude than others, because the corresponding voltage and current vary as described in Equation 4.

In addition to the mixing action caused by the nonlinear characteristics of the diode, the diode can be switched on and off to enhance the mixing action. If the LO power >> RF power and the LO power are sufficient to saturate the mixer diode, than the mixer diode switches on and off. The amount of RF power needed to saturate the diode is dependent upon the potential barrier and ideality factor of the device. Because the LO signal amplitude is much greater than the RF signal amplitude, as is the usual case for a mixer, the LO port can be viewed as a switch.

What actually causes the switching action in the diode is a time-varying conductance that is caused by the large LO signal. The instantaneous conductance of the diode g(t) can be found by taking the differential of Equation 3 with respect to voltage (g(t) = di(v)/dv). After taking the differential and performing a series expansion on the quantity, the expression for the diode current can be expressed in the form:

$$\cos((m\omega_l + \omega_r)\,t) + \cos(m\,\omega_{diff.} - \omega_r)\,t) \quad (7)$$

where: $\omega_{diff.}$ is the frequency difference between RF and LO frequencies (usually the IF freq.) and m = 1,3,5,7 because the square wave is an odd function.

Figure 3 shows the RF input waveform as well as the LO port waveform. Because the diode is switched on and off by the LO signal, the RF signal is sampled at the rate of the LO frequency. The resultant IF waveform is a multiplication of the RF waveform and the LO waveform.

Therefore, the form of the current obtained in Equation 7 can be obtained by noting that the product (mutiplication) of two cosine functions is of this form by trigonometric identity. Because the LO signal is much greater in magnitude than the RF signal, the difference frequency between the LO and RF signals appears as sidebands on either side of the LO signal and its harmonics. This results in an output spectrum as shown in Figure 4.

The LO signal and its harmonics have sidebands spaced at multiples of the difference of the RF and LO frequency, which is the IF frequency. This frequency spectrum is valid only if the RF amplitude is small compared with the LO amplitude. Additional intermodulation products are created when both signals are large.

Referring to Figure 4, Saleh developed a useful nomenclature for numbering the various IF output signals with positive and negative subscripts indicating a difference between frequencies residing above or below the LO signal and its harmonics.

These signals are of most concern:

- ω_l is the local oscillator frequency.
- ω_{diff} is the IF or difference frequency.
- ω_r is the RF signal frequency.
- $\omega_1 = -\omega_l + \omega_{diff.}$, which is also known as the image frequency.
- $\omega_2 = 2\omega_l + \omega_{diff.}$, which is the sum frequency of RF and LO.

Intermodulation products

If two RF signals are applied to the RF port simultaneously, or if the magnitude of the RF signal is not negligible as compared to the LO signal, then additional intermodulation (IM) products are generated by these extra signals. The general form of the IM products follow Equation 6 if the RF signal is not negligible compared with the LO signal. If two tones (two closely-spaced signals) are injected into the mixer, the situation gets worse and results in a frequency spectrum as shown in Figure 5.

This frequency component can be expressed in the form:

$$f_{IM} = +/- \; pf_1 +/- qf_2 +/- nf_{LO}$$

where: p,q and n are integers; f_1 is the first frequency; f_2 is the second frequency.

This is an undesirable spectrum to have in a receiver because it is difficult to filter out all these products. For this reason, most receivers operate in the linear (non-saturated) state, and substantial filtering is used to remove unwanted signals before and after the mixer. The next section describes different types of mixers and the techniques for intermodulation suppression using symmetric diode structures.

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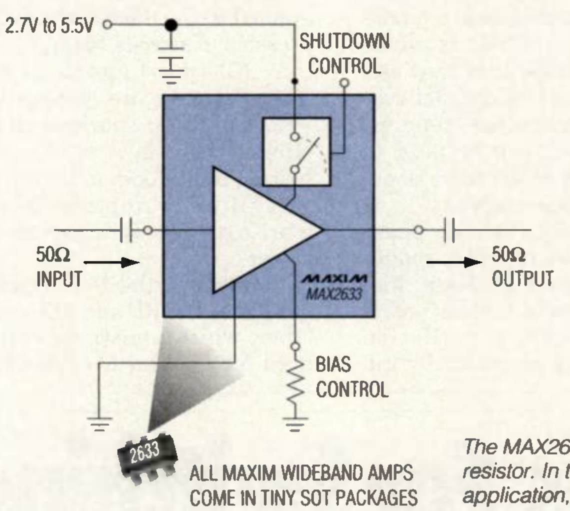

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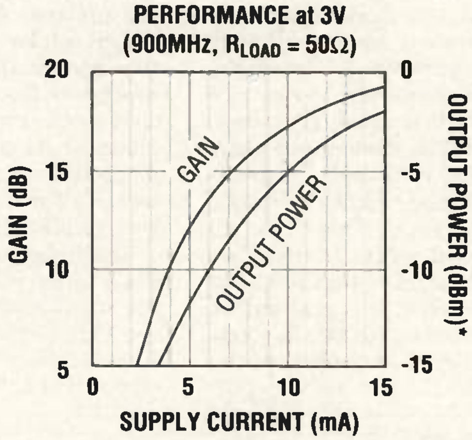

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MAX2631**	2.7 to 5.5	6.6	14 @ 900	Shutdown	SOT23-5	Maxim Proprietary
MAX2632	2.7 to 5.5	6.6	14 @ 900	Bias Control	SOT23-5	Maxim Proprietary
MAX2633**	2.7 to 5.5	6.6	14 @ 900	Shutdown + Bias Control	SOT23-6	Maxim Proprietary
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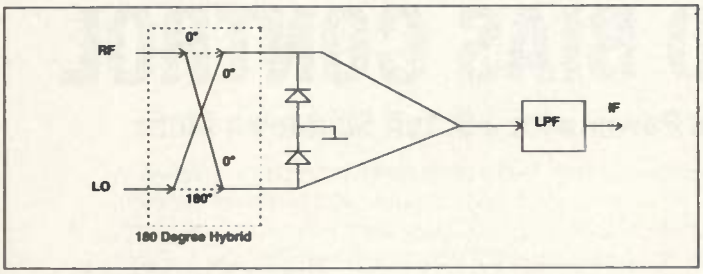


Figure 6. Single-balanced mixer.

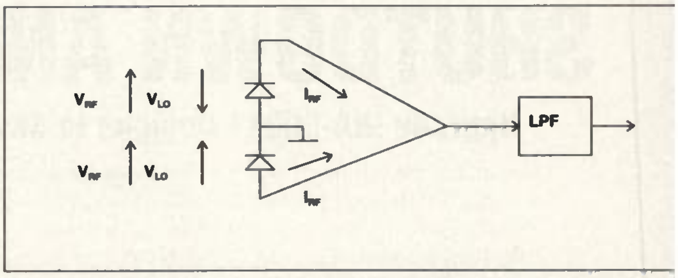


Figure 7. Phase of LO and RF signals at mixer diodes results in IF addition at output.

The single-balanced mixer

To improve port-to-port isolation and to reduce the magnitude of some of the spurious signals, two diodes commonly are used to create a single-balanced mixer as shown in Figure 6. The single-balanced mixer essentially is two single-ended mixers that are interconnected in a configuration that causes some spurious products to be reduced greatly. This type of mixer typically uses a four-port, 180° hybrid. This takes an input signal and splits it into two equal-amplitude signals that are 180° apart in phase when a signal enters one input port and creates two in-phase signals, if the signal is applied to the other input port.

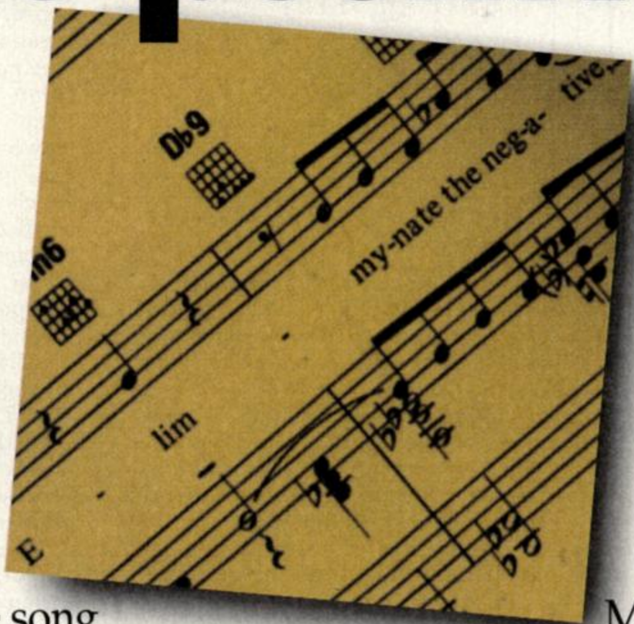
Figure 6 shows the four ports of a typical 180° hybrid and the corresponding phases. At RF frequencies, the hybrid can be constructed with a ferrite core and small wires. Bifilar is multi-conductor transmission line that can split and carry signals in parallel conductors. At microwave frequencies, printed coupled lines can be used to create a hybrid. One of the most popular hybrids is the Lange coupler.

The hybrid is used to achieve phase cancellation in the two diodes to reduce the number of spurious products. Figure 7 shows the phase at various points in the system and how cancellation occurs. The LO signal enters the hybrid at port 1 and creates an in-phase signal at port 2 and a 180° out-of-phase signal at port 3. Because of the phase shift applied to the diodes, the following suppression of signals occur,

1. All (m,n) spurious responses, where m and n are even are eliminated.
2. The (m,n) spurious responses are eliminated if m is even and n is odd, but not if m is odd and n is even.
3. kth order responses where k = m + n arise only for voltage terms of the kth power.

In addition, the 180° hybrid inherently isolates the RF and LO port from each other, which sometimes eliminates the need for RF and LO filtering. Because

Accentuate the positive.



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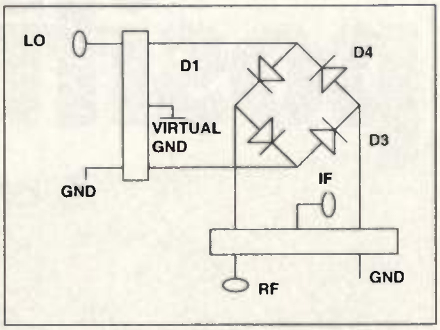

Figure 8. Double-balanced mixer implemented with Baluns.

the signal is split into equal amplitudes in the hybrid, 3 dB of additional power is required compared to a single-ended mixer for the diodes to receive the equivalent LO power. Despite the additional RF power requirement and the added complexity, this type of mixer is popular because it suppresses spurious emissions and it reduces or eliminates the need for RF and LO filters.

The double-balanced mixer

A double-balanced mixer configuration is shown in Figure 8. This configuration consists of four diodes and two balanced mixers connected in a "ring" configuration. Double-balanced mixers use two hybrids instead of one, which improves both the L-R port and R-I port isolation.

A special type of hybrid called a balun commonly is used in this type of mixer. It still provides a phase shift between ports, but also provides a virtual ground for the LO signal from which the IF frequency can be extracted. Because of its highly balanced and symmetrical configuration, this type of mixer theoretically supresses 75% of the possible IM products. The only products that have appreciable content are those with either odd LO or odd RF harmonics (harmonics with n or m = 1,3,5,7....).

In the double-balanced mixer, two diodes are on and the other two are off because of the way the signals are applied to the diodes. Referring to Figure 8, D1 and D2 are on, and D3 and D4 are off. This causes the switching action that was described before that

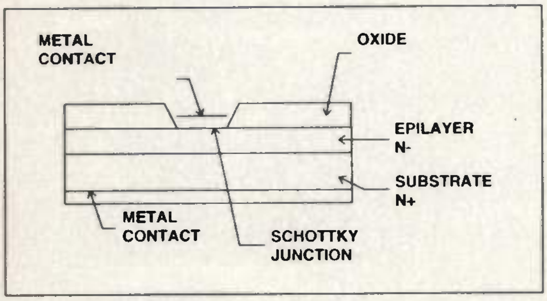

Figure 9. The Schottky diode structure.

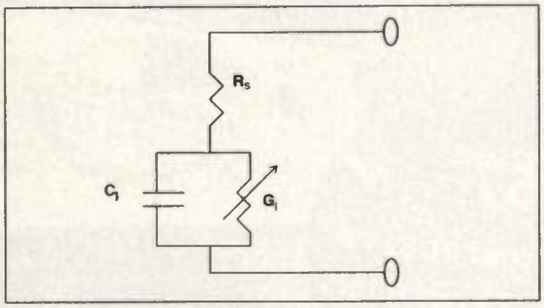

Figure 10. Equivalent circuit of a Schottky diode.

causes mixing. Of course, because two hybrids are involved, an additional 3 dB of both LO and RF power is required to drive this type of mixer as compared to a single balanced mixer.

In addition, the diode characteristics should match each other closely or spurious suppression will decrease. To achieve good results, the diodes usually are fabricated together on one piece of semiconductor and interconnected at the chip level.

Semiconductor devices

As described before, the nonlinearity and switching in the mixer device are the two most important factors for mixing. Not all types of semiconductor devices are suitable for this type of component. The most commonly used devices for mixers include Schottky diodes, FETs and small-signal bipolar devices.

• *The Schottky diode* — The Schottky diode is a specially fabricated device that is formed from a metal-semiconductor interface. Because the diode is not made from a p - n junction, it is a majority carrier device. Figure 9 shows a typical configuration of the device. The semiconductor consists of a heavily-doped substrate with a thin, lightly-doped layer deposited on top of it and a layer of metal deposited on top of the thin layer.

This configuration allows the diode to have low capacitance caused by the metal semiconductor interface as well as low resistance, caused by the highly-doped n+ layer. Because the mixer switches at the LO frequency and relies on the fast-switching action for mixing, it is imperative that the diode has low junction resistance and capacitance because its switching time is related directly to the RC time constant of the junction. The expression for the diode capacitance is:

$$C(V) = C_{jo}/(1 - V/\phi_{bi})$$

where: C_{jo} is the zero bias capacitance; ϕ(bi) the built-in potential of the metal semiconductor junction

The small-signal diode model is shown in Figure 10. It consists of a junction capacitance in parallel with a junction conductance and a series resistance. The typical capacitance for a Shottky diode for mixer diodes is typically 1 pf or less, and the typical series resistance is 10 Ω or less.

Because the capacitance and resistance are quite low, the diode switches at a fast rate, which facilitates the mixing action. A figure of merit that often is used is the cutoff frequency, which is:

$$f_c = 1/2\pi R_s C_{jo}$$

The rule of thumb for mixer operation is for the LO frequency to be 1/40th of the cutoff frequency to ensure a quick switching time. As a result, the RC time constant of the diode would be low compared to the LO frequency time constant of one RF cycle.

Key mixer parameters

Conversion loss— Figure 11 shows a single diode mixer with matching circuits at the RF and IF ports. An ideal match theoretically can be achieved at one frequency. Nevertheless, in practice there is always a finite mismatch that results in RF loss. This RF loss is known as the conversion loss, which is defined as:

Conversion loss = available RF signal input power/available IF output power

There are two components to conversion loss, one caused by circuit mismatch and another caused by intrinsic loss in the diode because finite junction capacitance and resistance (along with the series resistance) exsist.

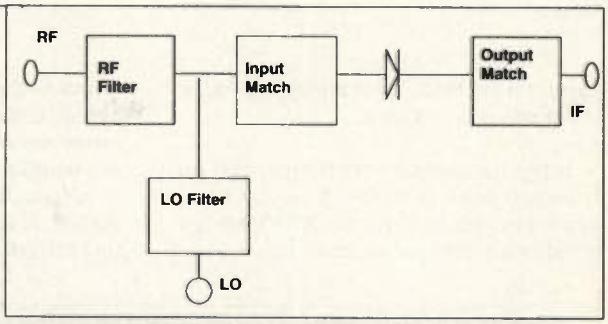

Figure 11. Single-diode mixer with input and output matching.

The approximate conversion loss caused by the diode itself is:

$$L_{Diode} = 1 + R_s/R_j + \omega^2 C_j^2 R_s R_j$$

where R_s and R_j and C_j are the series and junction resistance and capacitance, respectively.

Assuming a matched condition at the RF and IF ports, the total mixer conversion loss in dB is approximated by:

$$L_{total} = 10 \log (Z_o(g_o + g_2 + 1/Z_o))\,((g_o(g_o + g_2 + 1/Z_o) / g_1^2) - 2)$$

where: Z_o is the RF, LO and IF port impedance; g_0, g_1 and g_2 are the first three terms from a series expansion of the diode conductance.

If the image and sum frequencies are reflected back into the diode by using open and short circuits at those frequencies, then conversion loss can be improved because this reflected power can be converted into IF power. Typical conversion losses of a single diode mixer are usually in the range of 3–4 dB.

•*Noise figure* — The other key parameter of a mixer is noise figure. The noise figure of a cascade of circuits is defined as:

$$F = F_1 + F_2 - 1/G_1 + F_3 - 1/G_1G_2$$

This equation indicates that the first-stage noise figure controls the noise figure of a system. A typical downconverter cascade that included a mixer was shown in Figure 1. Oftentimes, in receivers, the first stage is a mixer. In practice, the noise figure always is measured in reference to a characterized noise source. Noise in mixers is caused by the resistive components as well as noise sources related to electron flow in the diode. Typical noise figures for a diode mixer are in the 3–4 dB range.

•*Port-to-port isolation* — Port-to-port isolation is defined as the amount of RF power that leaks from one port of a mixer to another. These include RF-to-IF, RF-to-LO, LO-to-IF, LO-to-RF, IF-to-LO and IF-to-RF isolation. This isolation nomenclature defines the first port as the source of the RF signal and the second port as the port where the leakage is measured. As described earlier, single-diode mixers have the worst isolation, and double-balanced mixers have much better isolation. This is caused by the natural isolation of a hybrid or balun in balanced configurations. This parameter also can be improved by using filters at the various ports to improve the frequency selectivity of the mixer. Typical isolation values range from 10 to 25dB depending on the type of mixer.

Conclusion

An introduction to mixers has included the role of mixers, theory of operation, types of mixers, semiconductor devices and key figures of merit.

About the author

Roland Soohoo received a B.S.E.E. from the University of California at Davis and an M.S.E.E. from the University of Santa Clara. He worked for Avantec in R & D and program management functions from 1985 - 1991. He was a project leader for Hewlett-Packard from 1991-1996 leading a cross functional team for development of RF and MMW subsystems and components. He joined Fujitsu Microelectronics in late 1996, where he is product marketing manager in the wireless products group in San Jose, CA.

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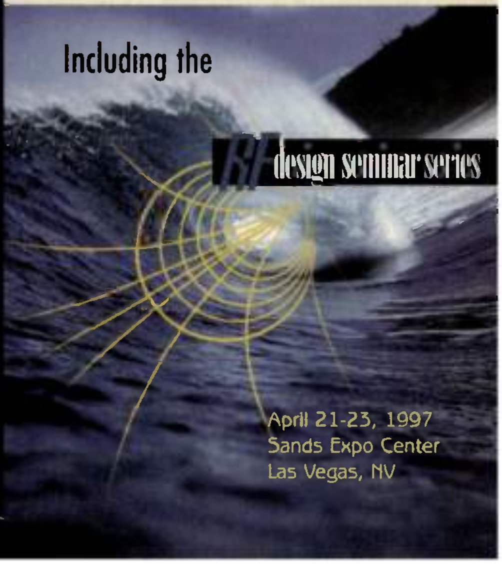

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- 220 MHz Systems
- Overcoming Market Barriers to Small Business
- Using the Internet for Business Development: *How to Use the Internet to Attract Customers and to Find Products and Services*
- Electromagnetic Energy Awareness: *How to Implement the FCC Guidelines*
- Specialized Antennas for Public Systems
- Tower & Site Acquisition and Management

TECHNICAL:
- Using Digital Computer-based Test Equipment
- Technician Recruiting and Training
- Resolving RF Interference Problems

PAGING:
- Reselling Paging
- Two-Way Paging: *What it Means to the Industry*
- Upcoming Paging Technologies

Schedule-At-A-Glance*

Monday, April 21, 1997

Registration Open	7:00 am - 6:00 pm
RF Seminars	8:00 am - 5:00 pm

Tuesday, April 22, 1997

Registration Open	7:00 am - 5:00 pm
Opening General Session	8:30 am - 9:45 am
RF Seminars	8:00 am - 5:00 pm
Exhibits Open	10:00 am - 5:00 pm
IWCE Sessions	11:00 am - 12:15 pm
IWCE Sessions	1:00 pm - 3:45 pm
International Reception *(Invitation only)*	5:00 pm - 7:00 pm

Wednesday, April 23, 1997

Registration Open	7:00 am - 5:00 pm
RF Seminars	8:00 am - 5:00 pm
IWCE Sessions	9:00 am - 11:45 am
Exhibits Open	10:00 am - 5:00 pm
IWCE Sessions	1:00 pm - 3:45 pm

Thursday, April 24, 1997

Registration Open	7:00 am - 1:00 pm
Exhibits Open	9:00 am - 1:00 pm
IWCE Sessions	9:00 am - 10:15 am
Closing General Session	10:30 am - 11:45 am

**schedule subject to change.*

As today's wireless marketplace continues to be redefined, so do the strategies that are needed to bring these products to market. This RF Design Seminar, co-located with the International Wireless Communications Expo (IWCE), provides design engineers, engineering managers and other RF professionals the tools and information they'll need to stay ahead of the changing marketplace.

Co-located@IWCE

Track One: Engineering Fundamentals (Instructional Level: Introductory)

Track One Instructors: Dr. Robert Feeney and Dr. David Hertling, Georgia Institute of Technology. These professors bring both academic substance and practical design experience to the classroom. CEUs are available from tthe Continuing Education department at Georgia Tech.

(M-1) RF and Wireless Engineering Part I: Fundamental Concepts

Monday, April 21 — 8:00 am - 5:00 pm

This first day reviews fundamental ideas and terms that are important in all RF and wireless systems. Concepts such as gain, bandwidth, noise figure, dynamic range, resonance and Q are presented. Practical components are discussed and models are developed. Fundamentals of electromagnetic theory are reviewed; transmission lines are discussed. The Smith chart is developed from transmission line theory and its use in RF design is introduced.

(M-2) RF and Wireless Engineering Part II: Fundamentals of Amplifier Design

Tuesday, April 22 — 8:00 am - 5:00 pm

This second day begins by designing lumped element and distribution impedance transformation networks using the Smith chart. Next, the course presents a unique approach to impedance matching network design that facilitates the design of T-, Pi- and L-networks for specified phase shift or Q. The same procedure is extended to design resistive attenuators and balanced networks. Active device models are then introduced and important concepts such as power gain, noise figure and stability are reviewed. The theory, physical meaning and measurement of S-parameters are presented, then graphical and analytical techniques for amplifier design using S-parameters are developed. The fundamentals of computer-aided analysis and optimization are then summarized.

(M-3) RF and Wireless Engineering Part III: Amplifier Design

Wednesday, April 23 — 8:00 am - 5:00 pm

The third day uses all the theory and techniques developed during the first two days to design several representative RF amplifiers. In addition to the basic RF design, practical topics such as bias network design, out-of-band stability, decoupling network design, the effect of microstrip discontinuities and the use of RF design software are presented.

Track Two: Engineering Applications (Instructional Level: Intermediate)

(T2-1) Practical High-Frequency Filter Design

Monday, April 21 — 8:00 am - 5:00 pm

HF Filter Design is a detailed review of low pass, hi pass and band pass filter designs employing passive L-C and transmission line components. Choice of transfer functions, filter order, amplitude and phase responses will be examined. Emphasis will be placed on simple computer-aided analysis and how to use the literature to produce practical designs.

Some of the topics to be explored are filter realizability, insertion loss, design with "real" components, tuning, filter match, various types of coupled resonators, and simple crystal filters.

Instructor: Dr. Dennis Sweeney is a professor at Virginia Polytechnic Institute and State University, Department of Electrical Engineering. He has taught undergraduate and graduate courses in Radio Engineering and is co-author of three reference papers on digital signal processing and satellite slant path propagation. His credentials include work in land/mobile communications systems for General Electric, research at the National Radio Astronomy Observatory at Green Bank, WV, and GPS projects at the JPL. He designed and built a major part of the 12.5/20/30 GHz beacon receiver for the Olympus Satellite experiment program at Virginia Tech and has built and deployed microwave propagation measurement systems at Vermont.

(T2-2) Oscillator Design Principles

Tuesday, April 22 — 8:00 am - 5:00 pm

A unified approach to oscillator design is presented which describes how to create high-performance oscillators using any type of resonator and any type of active device. Oscillators are demystified and fully understood so that design is no longer based on copying or modifying existing units. A complete understanding provides for known oscillation margins and design optimization for state-of-the-art performance. Both negative-resistance and open-loop Bode response design techniques are described. Gain margin, matching, starting, limiting, output level and harmonics are discussed. The theory is then applied to several practical oscillator circuits using L-C, SAW, transmission line and quartz crystal resonators with bipolar, FET and MMIC devices. Broadband tuning VCOs, general purpose and low-noise, high-stability oscillators are covered.

Instructors: Dr. Robert Feeney and Dr. David Hertling, Georgia Institute of Technology. These professors bring both academic substance and practical design experience to the classroom.

Track Three: Advanced Applications

(Instructional Level: Intermediate to Advanced)

T3-1 Digital Modulation and Spread Spectrum for Wireless Communications

Monday, April 21 — 8:00 am - 5:00 pm

This advanced class covers the essential topic of digital communications for wireless applications, exploring the principles and system architectures for complex I-Q, QPSK, Π/4-DQPSK, GMSK and FQPSK and other modulation schemes. Standards for specific cellular and PCS systems are introduced. A theoretical analysis of spectral and power efficiency follows, including adjacent channel interference, linear/nonlinear amplification and BER performance. Advanced modulation techniques such as 16-QAM, trellis coding, enhanced performance GFSK and multi-level FM are presented. TDMA radio design is also highlighted. The near-far problem and operation in mobile environments are discussed.

This course assumes substantial experience in RF design. All course participants will receive a complimentary copy of K. Feher's book, "Wireless Digital Communications: Modulation and Spread Spectrum Application," published by Prentice Hall in 1995.

Instructor: Dr. Kamilo Feher is a professor at the University of California, Davis and is also Vice President of Digicom's Consulting/Licensing and Technology Transfer Group.

(T3-2) RF Power Transistors and Amplifiers: Principles and Practical Applications

Tuesday, April 22 — 8:00 am - 5:00 pm

This practical one-day workshop covers RF power transistors from their design and unique characteristics to circuit selection, circuit construction and impedance matching concepts usable in their application. The course generally follows the book, "RF Power Transistors—Principles and Practical Applications," written by Norman Dye and Helge Granberg.

Specific topics range from what's different about power transistors to fundamental characteristics, choosing between FETs and bipolar transistors, tips on construction and biasing, amplifier design, wideband impedance matching, and the concepts of frequency compensation and negative feedback.

Attendees will walk through: Selecting a transistor for a particular application • Selecting the type of circuit best suited to the application • Building a microstrip circuit • Biasing the transistor • Matching the transistor.

Instructor: Norman Dye is an engineering consultant with extensive experience in Motorola's power transistor division. Along with Helge Granberg, he recently authored the book, "Radio Frequency Transistors."

Management Track

(Instructional Level: Non-Technical)

SPECIAL COURSE FOR MANAGEMENT, MARKETING AND PURCHASING PERSONNEL

(MT-1) Wireless Communications For Non-Engineers

Wednesday, April 23 — 8:00 am - 4:00 pm

The morning session introduces the key concepts of signals, circuits, systems and the radio spectrum. These technical concepts are presented in an intuitive, visual manner, relating them whenever possible to familiar activities of everyday life.

Demonstrations and class participation will help non-technical personnel grasp these concepts, which are essential for "speaking the language" of wireless communications. The session continues with descriptions of common communications systems: AM and FM radio broadcasting, television broadcasting, cable TV and traditional two-way radio.

The afternoon session begins with a quick review of the principles, with additional notes on how radio waves travel from one point to another. Then, modern wireless communications systems are described, including remote control systems, cordless telephones, the cellular phone system and satellite systems. Finally, the newest areas of communications: Personal Communications Services (PCS), wireless local-area networks (WLAN), automatic toll collection and RF identification (RFID) tags are described.

Instructor: Gary Breed is president of Noble Publishing Corporation, publishers of technical books, instructional video courses, and *Applied Microwave and Wireless* magazine. He has 25 years experience in the wireless communications industry, including 11 years as Editor of RF Design magazine. He has experience in the design, construction and operation of a wide range of communications systems.

CEUs available for Track 1 and Track 2 sessions.
See registration form.

Travel & Housing Information

Make your Las Vegas reservations today! DESTINATIONS, INC. is the official housing and travel coordinator for IWCE '97. Hotel and airline reservations at the lowest possible rate can be made by calling 1-800-325-2078 or 404-584-0599 ext. 260. You may fax your request to: 404-584-0685. Please include arrival/ departure dates, housing needs, and credit card number for confirmation. Make sure to mention you're with IWCE '97.

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*Partial list as of 1/97

REGISTRATION FORM

21st Annual IWCE '97

April 22 - 24, 1997
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Mobile Radio Technology
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(1) GENERAL INFORMATION

First Name ____________________

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Address ____________________

City ____________________

State/Province ____________________

Zip ____________ Country ____________________

Phone* ____________________

Fax* ____________________

*International guests, please include country and city codes.

☐ Please contact me regarding barrier-free accommodations.

(2) PROFILE

(ALL QUESTIONS MUST BE ANSWERED)

1) YOUR JOB FUNCTION? (check one)

A ☐ President/Owner
B ☐ Director/VP
C ☐ Sales & Marketing
D ☐ Communications Manager
E ☐ Technician/Service Supervisor
F ☐ Engineer
G ☐ Purchasing Manager
O ☐ Other ____________

2) WHAT BEST DESCRIBES YOUR BUSINESS? (check one)

A ☐ End User of Radio Communications Equipment
B ☐ Business Radio Dealer
C ☐ Business Radio Dealer/Service Shop
D ☐ RCC/PCP
E ☐ Cellular Operator/Carrier
F ☐ Cellular Agent/Dealer
G ☐ Public Safety
H ☐ Exporter/Importer
J ☐ Manufacturer's Rep
K ☐ Distributor
L ☐ Consultant
M ☐ Manufacturer
N ☐ SMR
P ☐ Association/Academia
Q ☐ Press
O ☐ Other ____________

3) WHAT ARE YOUR PRIMARY PRODUCT INTERESTS? (check ALL that apply)

A ☐ International Markets
B ☐ RF Engineering
C ☐ Mobile Data
D ☐ PCS
E ☐ Pagers
F ☐ Two-Way Radio
G ☐ Antennas
H ☐ Systems Integration
J ☐ Satellites
O ☐ Other ____________

4) WHAT BEST DESCRIBES YOUR PURCHASING ROLE? (check one)

A ☐ Make final purchasing decisions
B ☐ Recommend/Specify
C ☐ No purchasing role

Source Code AD

(3) PROGRAM REGISTRATION

1) IWCE Registration

	On or Before 4/1/97 (circle)	After 4/1/97 (circle)
S ☐ IWCE Sessions & Exhibits (all three days)	$425	$495
T ☐ Exhibits & Tues. Sessions	$175	$225
W ☐ Exhibits & Wed. Sessions	$175	$225
R ☐ Exhibits & Thurs. Sessions	$175	$225
E ☐ Exhibits Only (all three days)	$25	$45

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2) RF Design Seminars - Advanced Registration Before 4/1/97 Only

Register for individual sessions, or save BIG when you register for an entire seminar track. Includes IWCE exhibit pass and course materials. To make your selections, cross-reference the codes below with the course descriptions found on the pages 3 and 4.

Monday, April 21 (check one): ☐ T1-1 ☐ T2-1 ☐ T3-1

Tuesday, April 22 (check one): ☐ T1-2 ☐ T2-2 ☐ T3-2

Wednesday, April 23 (check one): ☐ T1-3 ☐ MT-1

☐ One course/one day (as checked above) @ $295
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CEUs are available for tracks 1 and 2 through the Georgia Institute of Technology School of Electrical Engineering for those attendees registering on or before to 4/1/97. ONLY. Certificates of attendance will be granted. If you will be applying for CEUs, please enter your Social Security Number here: SS# ____________

(4) METHOD OF PAYMENT

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Test equipment manufacturers make equipment smaller, faster

Each month, the product forum gives companies manufacturing the products used by RF engineers the opportunity to offer their opinions regarding today's marketplace and trends in the industry without editorial interpretation. This month, the product forum highlights test and measurement equipment. The information in this section was compiled by Gregg Miller, Technical Editor.

Marconi Instruments

The introduction of cellular radio networks and the opening up of national telecommunications markets is creating a boom in the use of microwave radio links. These microwave radio links are divided into two categories: trunk links and mini-links. Marconi believes the worldwide market for radio links, especially mini-links, will continue to grow rapidly between now and the year 2,000. The majority of these links will be used to establish new or expanded cellular radio networks. The pressure to expand networks faster than ever before has put intense pressure on network managers to install links in the shortest time possible. When any new radio is installed, its power and frequency must be measured to verify that the radio is performing correctly and is complying with the regulatory authorities' licensing terms. Marconi Instruments will continue its worldwide commitment to being a leading supplier of test instruments for RF and microwave radio links.

Keithley Instruments

Devoting special attention to the test needs of communication equipment producers has resulted in the introduction of several new test instruments designed for these applications. High-quality electrical measurements are required as manufacturing processes become more sophisticated, the components become smaller and use less power and as quality and reliability benchmarks advance. Communications products pose special manufacturing challenges because they are complex devices produced in high volumes. Manufacturers need to make reliable, accurate measurements at various stages of assembly to ensure that the products are within specification at every step of the process. Keithley has addressed these issues by introducing several new instruments with smaller footprints, higher throughput and better reliability than ever before.

Programmed Test Sources

Currently the test and measurement market is being driven largely by the requirements of the wireless communications sectors. As a manufacturer of signal sources, PTS products are being pushed on two fronts. First, wireless communications are taking place at higher and higher frequencies and therefore the design and test phases require signal sources capable of producing higher and higher output frequencies. At the same time, higher levels of performance in the baseband signal-processing ICs are driving a requirement for higher-quality low and medium frequency sources with SFDR of –70 dBc and better at very close-to-the-carrier offsets. PTS serves both segments of this market through a broad line of high-performance frequency synthesizers, which now cover as high as 3.2 GHz.

Anritsu Wiltron

We envision the future of test and measurement to be the creation of smaller, portable, more easily configurable instruments that meet the customers needs and may be upgraded in the future. To achieve this, Anritsu Wiltron has developed equipment based on a "platform" concept. The base unit may be configured for multiple frequency ranges, power ranges or applications. Customers purchase only the capabilities that they require but are assured of an easy upgrade path, should their needs change.

Noise/Com

Test equipment alone is insufficient. The RF engineer requires a partnership in which the vendor will add valuable technical and applications knowledge to speed the RF engineer through the test process. RF design engineers must dedicate time to engineering the core product within their businesses. The cellular, PCS and satellite communications industries are growing so much that even test engineers need help doing what they do best: making test solutions quickly. We listen to the RF and test engineers to conceive the perfect test system guaranteeing overall accuracy, consistency, user-friendly procedures that will uphold our reputation.

Tektronix

Digitally-modulated standards have created requirements for higher performance and flexibility in test equipment for wireless design. Signal generators and spectrum analyzers must support application-specific modulation and allow for evolving standards. Flexibility and range of support is important to designers who must support multiple technologies. The complexity of the digitally modulated waveform, increased spectrum use and the needs of manufacturing all call for greater precision in signal generation, digital modulation and spectrum analysis. Making applications-specific measurements with great precision is key to the latest standards, speeds the design task and allows for greater tolerances in manufacturing.

Telecom Analysis Systems

Wireless communications is one of the most demanding applications for the telecommunications equipment designer. Typical signals at the receiver have delay spreads in the tens of microseconds and experience multipath fading resulting in large signal power fluctuations. The only way to ensure that a system is able to function properly in an environment with these characteristics is through extensive testing. TAS has developed an RF channel emulator with a flexible model that can be used to emulate an RF channel.

Amplifier Research

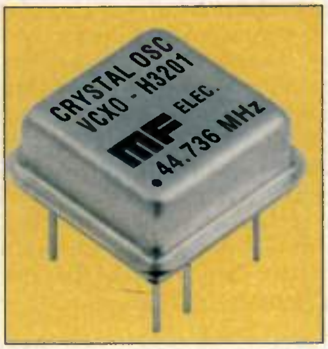
We have been preaching the value of pre-compliance testing for more than 25 years. Testing from the design level up results in reduced time-to-market and redesign costs; the requirement that electronic products sold in the EU carry the CE mark that formally endorses our position, at a time when manufacturers realize precompliance testing can no longer be left to chance. EMC test systems let you conduct precompliance immunity and emissions testing in your lab, under your control. Test cells are small enough to use on-site but large enough to accept test objects as high as 1.5 meters per side. RF

40 W UHF RF amplifier system operates from 225–600 MHz

A 40 W UHF RF amplifier system that operates from 225–600 MHz features a solid-state protection system that provides rugged, reliable operation under conditions of abuse. Designed for EMI and EMC applications, it features automatic current limiting, a condition that may arise when operating into a high-voltage standing wave ratio (VSWR). The system is designed not to shut down during current limiting conditions. Performance specifications include a 40 W CW RF ouptput power, 40 dB gain, a nominal input and output impedance of 50 Ω and –60 dBc spurious. The system also offers protection against bad loads and a margin of safety from voltage breakdown. It includes a power supply, amplifier module, cooling, front panel indicators and monitors. It can be used as a stand-alone on a bench top or be rack mounted. The system is priced at $5,950.
LCF Enterprises
INFO/CARD 159

Wide pull VCXO for ATM applications

The H3201 series of voltage controlled crystal oscillators (VCXOs) provide data and clock recovery for ATM applications. The units guarantee frequency capture (pull) to ±150 ppm and cover the frequency range from



3–175 MHz. Guaranteed performance eliminates the user's need to make allowance for temperature, aging and supply voltage variations—protection measures that increase cost, not accuracy. The VCXOs measure 0.5″ × 0.5″ × 0.23″ and are priced from $9.50 for ±50 ppm capture range units to $12 for versions with ±150 ppm capture range.
MF Electronics
INFO/CARD 160

Patch antenna for GPS applications

The DAK series is a miniature dielectric ceramic antenna element for use in global positioning systems

(GPS). This flat patch antenna incorporates a rectangular micro-strip design to be used for GPS C/A right-hand circular, polarization wave reception. It features a 25 mm² surface area and a height of 4 mm. The DAK antenna is designed to be used at 1,575 MHz for L1 reception. It uses a high-performance proprietary ceramic material, which provides excellent temperature stability and sensitivity, has an impedence of 50 Ω, a 9 MHz minimum bandwidth and a temperature coefficient of 20 ppm/°C maximum.
Toko
INFO/CARD 161

Switch system for telecomm testing

The Model 7116-MWS switch system introduced by Keithley has a design optimized for computer automated testing of telecommunication devices such as portable cellular phones. It provides 18 GHz bandwidth switching for IEEE 488-based test systems, with a switch configuration that can be arranged as either a single 1 × 16 multiplexer or five independent 1 × 4 multiplexers. The 7116-MWS switch configuration ensures that all 16 signal paths are of equal length so

that every channel has similar transmission line characteristics. There is less than 0.5 dB loss at 3 GHz in each of the signal paths. The 7116-MWS features a 50 Ω impedance and SMA connectors.
Keithley Instruments
INFO/CARD 162

High-frequency trimmer capacitor

A miniature solid dielectric trimmer capacitor features a range of 0.45–3.5 pF

and positive stops at minimum and maximum. It features more than seven turns of tuning and measures 0.155″ in diameter and 0.4″ in length. The solid dielectric design offers linear tuning and no chance of internal shorting or microphonics. It is rated at 250 V working and at 500 V withstanding. The Q is more than 2,000 at 100 MHz. High-voltage and 40 psi sealed versions are also available. The trimmer capacitor is priced at $3.15 in 100-piece quantities and $1.70 in 50,000-piece quantities.
Voltronics
INFO/CARD 163

SIGNAL SOURCES

OCXO features excellent phase noise

The FE-102A-100 oven-controlled crystal oscillator (OCXO) features operation at 100 MHz, low phase noise of –172 dBc, voltage-controlled oscillator (VCO) range of ±5 ppm for 0–10 V and a low profile of 1". Temperature stability is 2 ppm per year and 5 ppm per 5 years. The unit was designed for applications such as low-noise frequency synthesis, doppler radar, low-noise phase-locked loop (PLL) systems and phase noise reference.
Frequency Electronics
INFO/CARD 164

Universal surface-mount crystal packages

Three surface-mount crystal packages are designed to match several industry standard crystal footprints. The 49SUB, 49SAB and 49SNC com-

bine proven metal packaged AT- or BT-cut strip resonator crystals with a precision-molded base design to meet strictest coplanarity specifications. The crystals are available in frequencies ranging from 3.5–66.6667 MHz with a standard frequency tolerance of ± 50 ppm from –20 to 70°C. They are priced from 80 cents to $1.50 in quantities of 1,000.
Saronix
INFO/CARD 165

PDRO targets emerging wireless technologies

A line of phase-locked sources tageting the needs of Satcom converters, digital radio and the emrging wireless technologies features single-loop fractional multiplication techniques, 0.5–18 GHz operation phase-locked to a 10 MHz crystal reference, low phase noise of –107 dBc at 1 kHz offset from a 1,112.5 MHz carrier and low DC power consumption. These PDRO sources require only a single loop for reliable phase-lock performance with a 5 or 10 MHz crystal reference oscillator, resulting in the PDRO 601 source delivering 6.1 GHz phase-locked output from a 10 MHz reference oscillator.
Elcom
INFO/CARD 166

Plastic surface-mount oscillator

The EC1400SJ series of plastic surface-mount J-lead oscillators uses an industry standard footprint suitable for network products, modems, telecommunication applications or any application requiring a rugged surface mount design. The EC1400SJ series has a frequency range of 1.025–66.667 MHz. Options for this series includes ±50 ppm stability, tight duty-cycle and tri-stae function.
Ecliptek
INFO/CARD 167

Low-profile 25 W terminator

Part numbers 12-3022SM and 12-3022SF are low-profile, 25 W, coaxial flange-mount terminators with a lightweight aluminum housing and a stainless steel SMA connector. The device features a voltage standing wave ratio (VSWR) of less than 1.25:1. They are available from DC to 18.0 GHz and in SMA male and female versions.
Florida RF Labs
INFO/CARD 168

SIGNAL PROCESSING COMPONENTS

Low-loss surface-mount hybrid couplers

A low-loss line of surface mount hybrid couplers has been added to the Xinger brand surface-mount component line. These models offer a specified insertion loss of less than 0.2 dB. They feature a footprint size of 0.48″ × 0.65″ and are priced at $3.98 in large quantities and $6.89 in quantities less than 1,000.
Anaren Microwave
INFO/CARD 169

Flange mount microwave detector

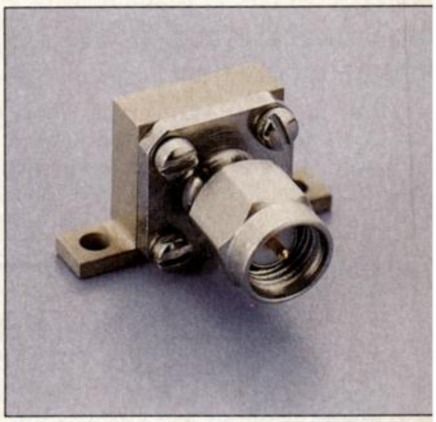
The DFN series of flange-mount microwave detectors are designed to meet connectorless assembly requirements. The DFN0140 covers a frequency range of 0.01–4.0 GHz with a minimum voltage sensitivity of 500 mV/mW, maximum flatness of 0.30 dB and a typical voltage standing wave ratio (VSWR) of 1.3:1. The flange size is 0.375″ square with a total length of 0.530″. The unit is hermetically sealed in a micro-module and integrated into a custom housing to meet broadband frequency requirements. The micro-modules can be manufactured with planar tunnel diodes, zero bias Schottky diodes and biased Schottky diodes.
Device Technology
INFO/CARD 170

Voltage variable attenuators for mobile digital wireless

The SM-515003 and SM-516003 voltage variable attenuators (VVA) are designed to be compact, easily implementable HF, VHF and UHF attenuator pads. They are cascadable with additional 50 Ω VVAs, without the need for any external components, to provide even higher levels of signal padding. The SM-515003 is designed to cover the frequency range from 3–300 MHz and typically can provide 70 dB of isloation at 100 MHz. The SM-516003 is designed to cover the frequency range from 300–3,000 MHz at the same time typically providing 40 dB of isolation at 2.4 GHz. Both attenuators operate on 5 VDC and require two analog voltages per section with a range from –3.0 V to 0.0 V at less than 15 µA each. The models are priced at $2.10 and $2.40 in quantities of 10,000 respectively.
Samsung Microwave
INFO/CARD 171

YIG-based frequency synthesizer

The YTS series frequency synthesizer uses a YIG voltage-controlled oscillator (VCO) to achieve superior phase noise. This synthesizer has



broadband operation from 4–8 GHz and an output power of 13 dBm minimum. It also can be designed with optional L-band frequencies and a fixed L-band output often used as a second local oscillator (LO). Custom design options are available.
Miteq
INFO/CARD 172

Filter prevents cellular interference

Model 3322-806/860(40) lowpass filter is designed to protect UHF off-air antenna reception from cellular telephone and mobile communications interference. The unit has a passband from 50–806 MHz and an insertion loss that typically is less than 1 dB. A minimum of 40 dB rejection is acheived for signals in the 860–1,000 MHz band. The filter measures 1″ × 1.5″ × 4″ and is priced at $125.
Microwave Filter
INFO/CARD 173

RF literature

High-performance IC data books available

The *Linear Products* and the *Mixed Signal Products* data books feature more than 60 high-performance and mixed-signal products, plus complete product descriptions, applications tips, performance graphs, detailed specifications and ordering information for the company's line of high-performance linear and mixed-signal products.

The *Linear Products* data book contains more than 1,300 pages and offers designers more than 35 high-performance linear products including operational amplifiers; instrumentation amplifiers; isolation products; power operational amplifiers; optical sensors; references and regulators; and demonstration boards.

The *Mixed Signal Products* data book has more than 1,100 pages of high-performance data conversion products for design projects. Products profiled include: A-to-D and D-to-A converters for industrial, instrumentation and communications applications; data acquisition components; digital audio converters; sample and hold amplifiers; V/F converters; multiplexers; and demonstration boards.

Both data books are free.

Burr-Brown
INFO/CARD 155

Catalog is source for automation products

Intelligent Instrumentation's 32-page factory view catalog features more than 100 of the company's hardware, software and interface tools for a variety of applications for data acquisition, test measurement, machine and process monitoring and control, data collection, and operator interface and control.

The factory view catalog showcases a line of PC-based data acquisition hardware and software tools including plug-in boards that provide analog and digital I/O combinations and sampling rates as high as 100 MHz. Intelligent Instrumentation's complete line of I/O hardware offers products for any data acquisition, control system, test and measurement application. External parallel port data acquisition systems are featured to accommodate portable PC applications. A selection of analog and digital signal conditioning and termination accessories are available to accommodate interfacing to real-world signals and sensors.

The catalog has been translated into French, German, Hungarian, Italian and Japanese.

Intelligent Instrumentation
INFO/CARD 156

Catalogs feature amplifiers

A 66-page catalog contains information on MITEQ's AFS series low-noise and medium-power amplifiers from moderate bandwidth to ultra-wideband. Also included are a variety of special amplifier designs such as surface-mount, microstrip, variable gain, cryogenic, temperature compensated and dual output, as well as information about high reliability and space qualified units.

A 42-page catalog contains information on the company's AMF series low-noise medium-power and power amplifiers. Included are Satcom, temperature compensated, cryogenic and sloping amplifiers.

The bipolar amplifier catalog details performance as high as 2 GHz for low-noise, medium-power amplifiers. The brochure includes a detailed listing of specification definitions, available options and typical test data.

Miteq
INFO/CARD 157

On-line

Product, sales information on web site—Signal Processing Technologies' product and sales information site on the World Wide Web features product information including on-line data sheets for SPT's high-speed analog-to-digital converters (ADCs), digital-to-analog converters (DACs), comparators and track-and-hold amplifiers. The web site also maintains a complete database of the company's sales and distribution network and corporate information. On-line product charts contain information such as resolution, sample and conversion rate, power dissipation, linearity and I/O specifications. On-line data sheets are available in Adobe Acrobat format. The site is located at http://www. spt.com.

Signal Processing Technologies
INFO/CARD 158

RF software

Tools, libraries enhance system-to-ASIC design flow

Alta Groups' tools and libraries enhance the system-to-application specific integrated circuit (ASIC) top-down design flow for communications product design. Offerings include the fixed-point communications library, the filter module generator and a new version of Alta's Hardware Design System (HDS). This combination supports a communications design flow from floating-point communications design to fixed-point design and finally to ASIC implementation through synthesis and library mapping. These tools and libraries further enhance the offerings within the new EnWave wireless design release and Alta's library-based system design, which enables design reuse by capturing fixed and variable implementation libraries.

The fixed-point communications library is $8,000 and requires users to own HDS. The filter module generator is available at no charge to filter design system users who are under a maintenance contract. The filter design system is $4,000. The Hardware Design System ver. 3.5.1 is available at no cost for users under maintenance contract. The price for new users is $10,000.
Alta Group
INFO/CARD 153

Graphical user interface builds test equipment

BenchTop Plus is a graphical user interface. It provides engineers with a set of virtual instruments and the tools necessary to build an automatic test equipment (ATE) system. Add-ons to the BenchTop Lite software package are included with all instruments sold by PC Instruments.

Included in BenchTop Plus are a pulse analysis function for analyzing pulse responses and digital communication signals; a fast Fourier transform (FFT) for frequency domain analysis; a phase-angle meter for use in analyzing power distribution, filters, gyroscopes and other mechanical sensors; a boundary test function for comparing acquired waveforms to high and low limits; and the automated test language for controlling a sequenced set of tests. The operator's manual contains sample programs. BenchTop Plus sells for $495.
PC Instruments
INFO/CARD 154

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PRODUCT SUPPORT ENGINEER - ASIA

Responsible for support major Asian customer accounts in a software support organization. Responsible for supporting users of our software products with hardware and software related issues; installation of software; training customers on software products; defining feature enhancements to software and serving as liaison to our clients. Requires a working knowledge of UNIX with an emphasis on Sun workstations. The successful candidate will possess 3 - 5 years engineering experience in the telecommunications industry, and excellent written and verbal communication skills. Experience in cellular or microwave engineering highly desirable. Extended travel throughout Asia is required. Bilingual a plus (Korean).

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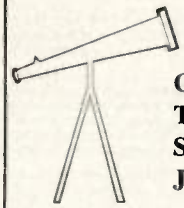

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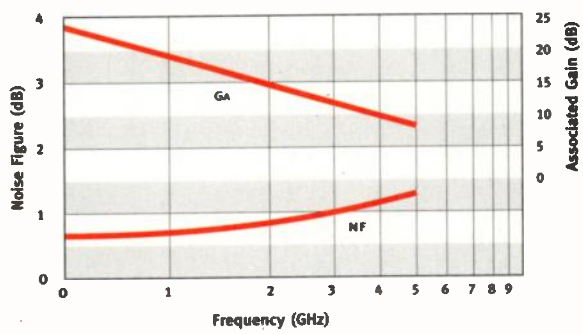

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