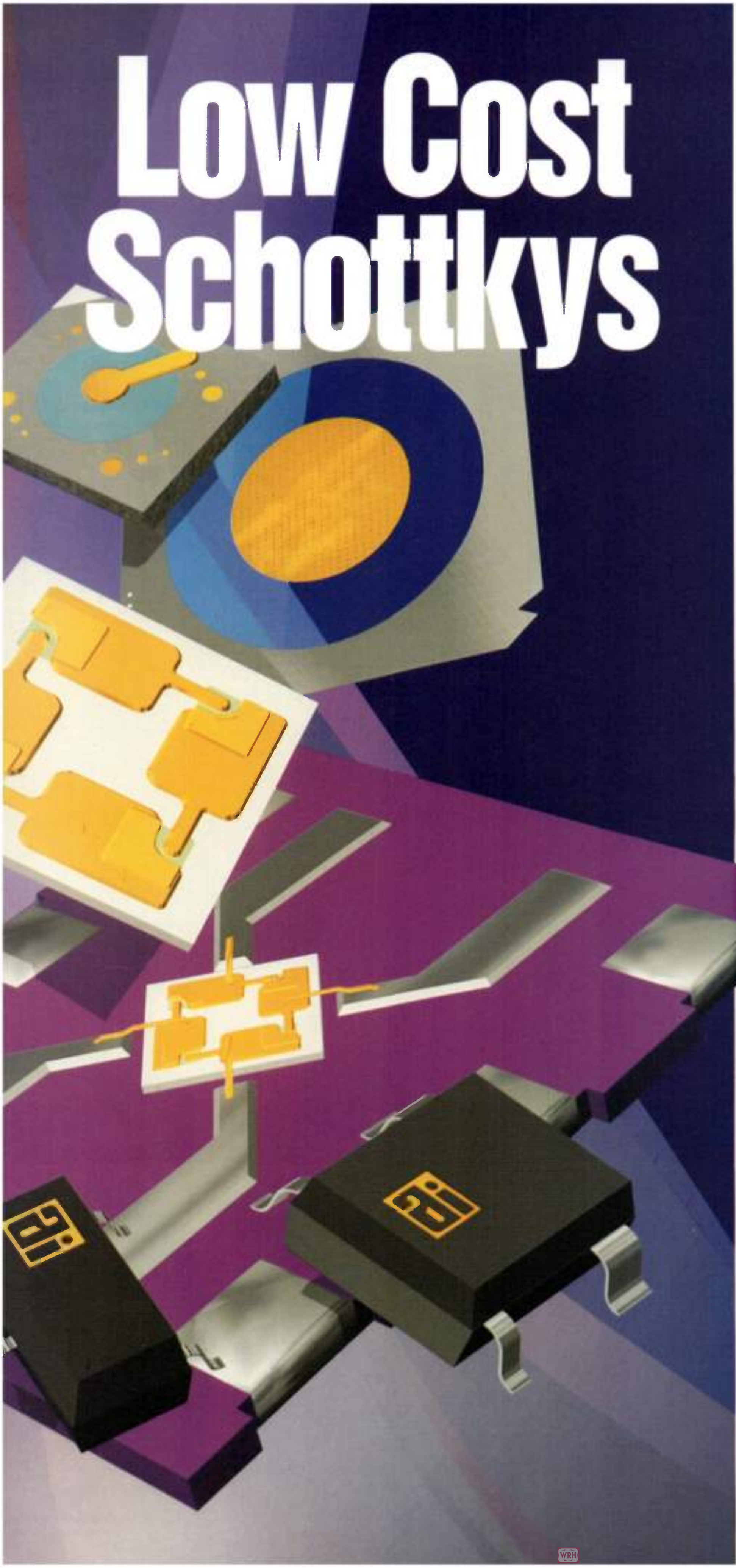
RF design

engineering principles and practices

Cover Story
New Chipset Aids Designs for DECT

Featured Technology
Spread Spectrum



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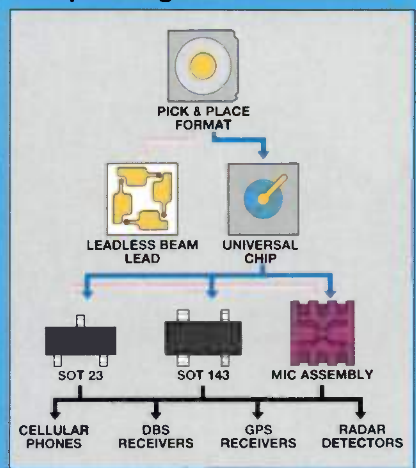

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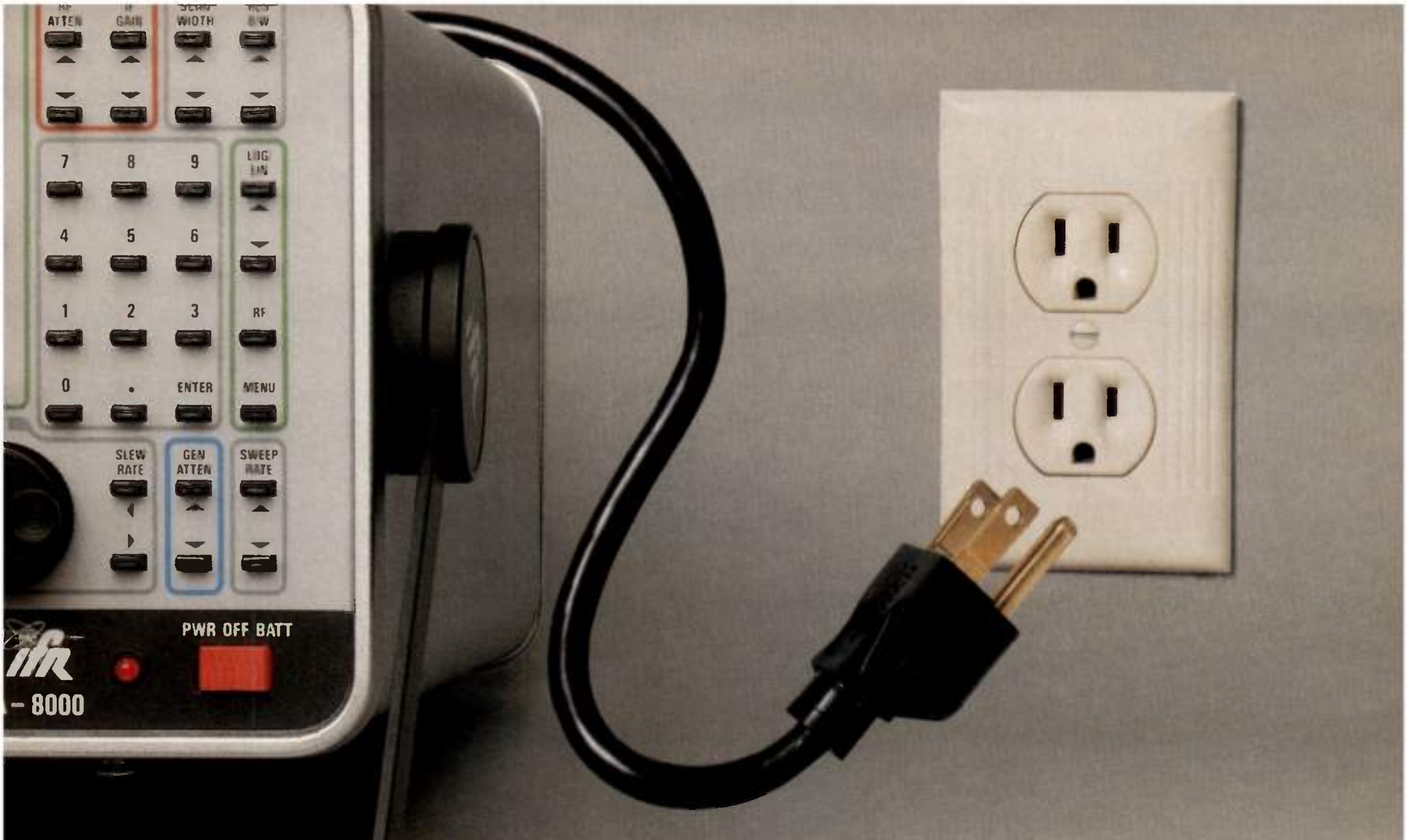

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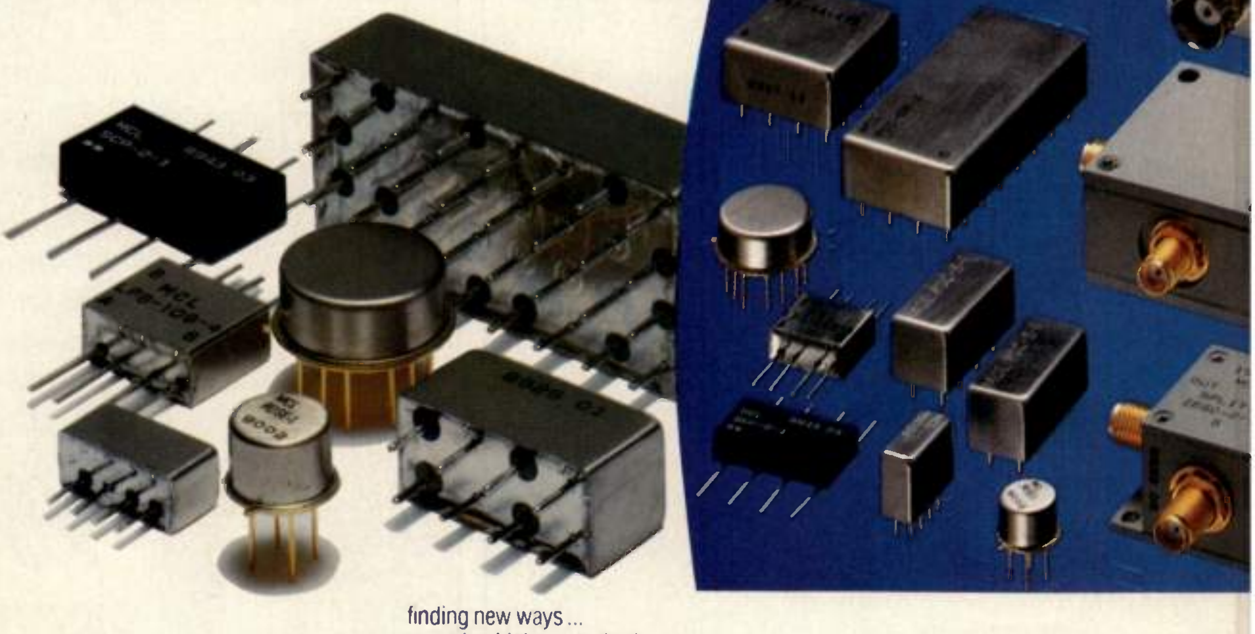

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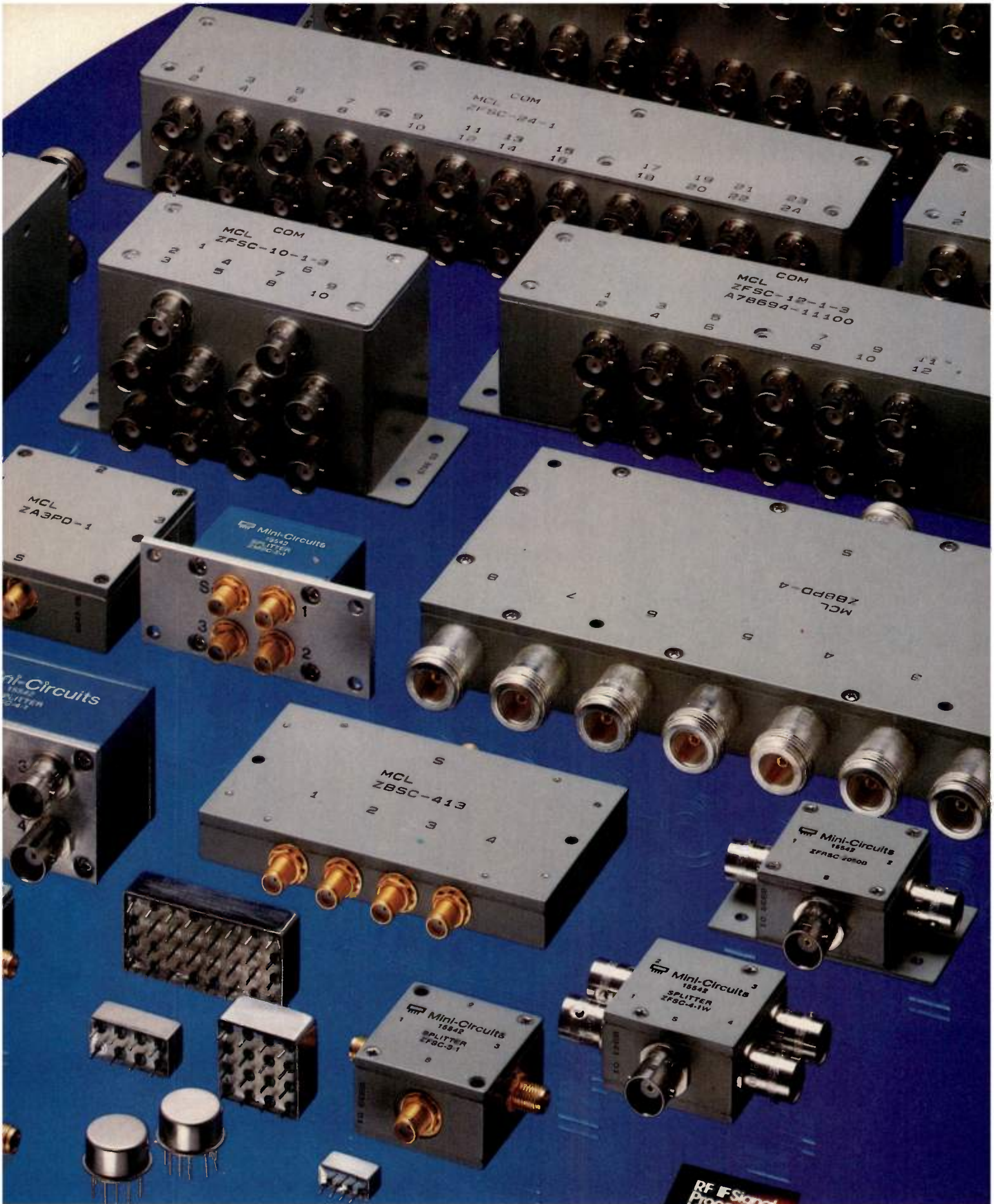

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• THOMAS REGISTER Vol. 23 • MICROWAVES PRODUCT DIRECTORY
• EEM • MINI-CIRCUITS' 740-pg HANDBOOK.

INFO/CARD 3

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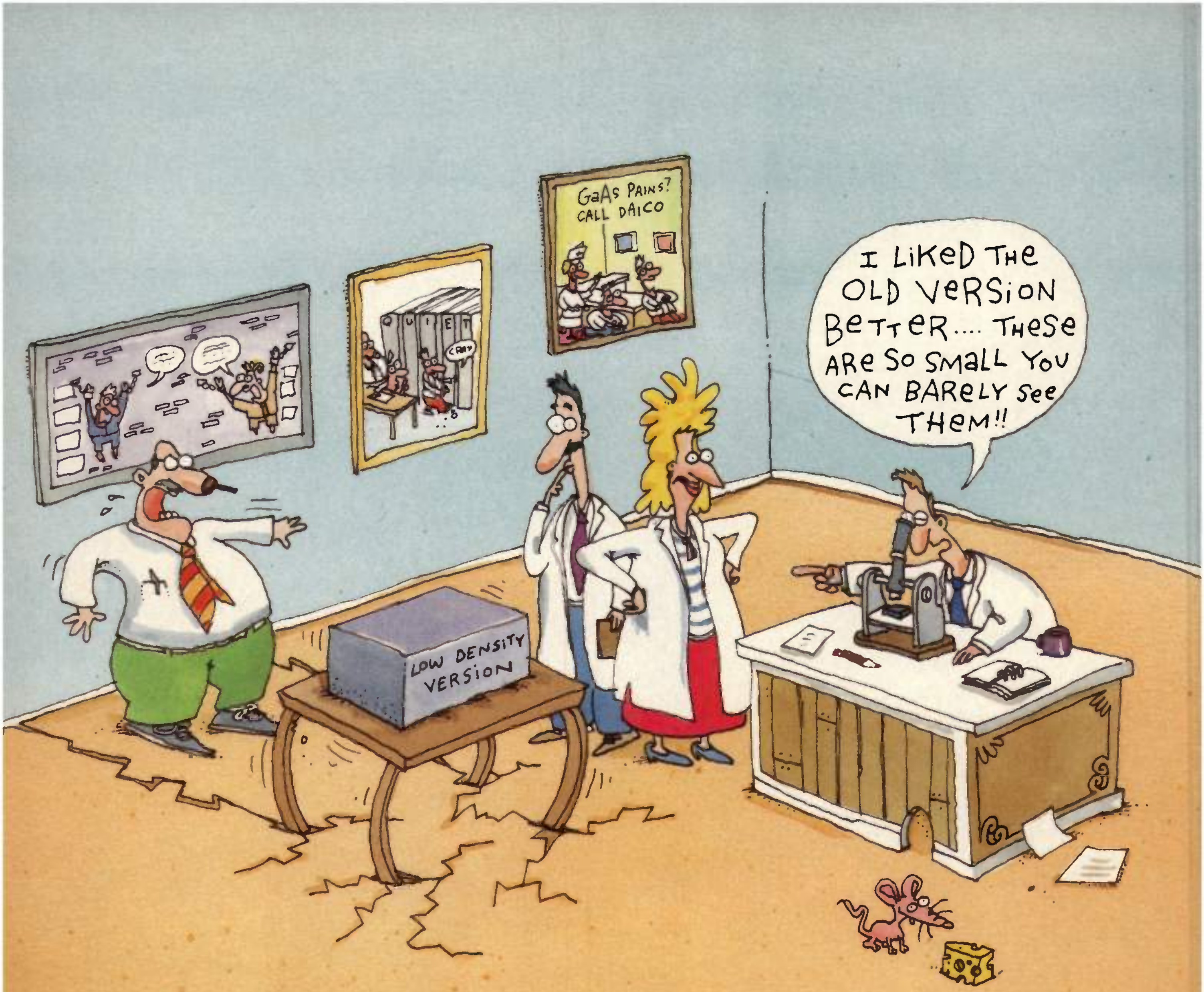
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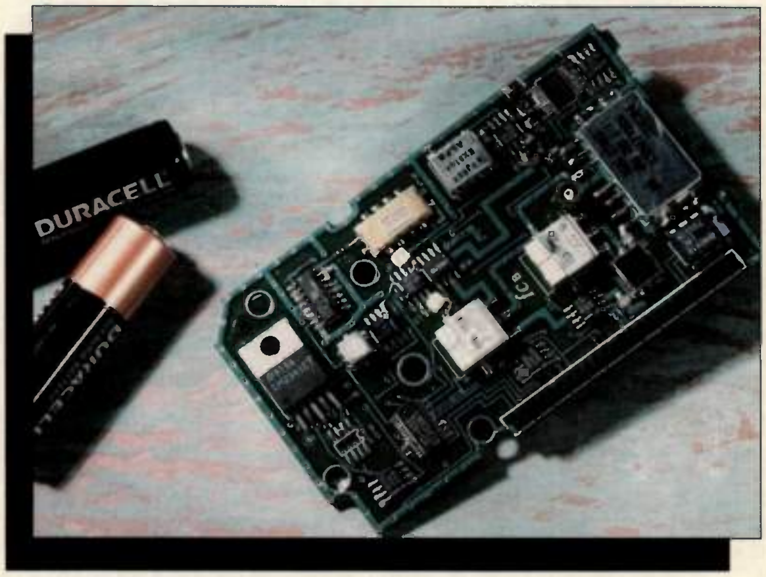
RF design

contents

April 1993



featured technology

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tutorial

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departments

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It All Comes Down to Price

By Gary A. Breed
Editor

This issue has an emphasis on spread spectrum (SS) systems and other wireless applications. Engineers with a military background will consider SS to be a high performance, complex transmission method that is jamming-resistant and secure. Researchers investigating commercial applications will be concerned with error-correction algorithms and interfaces with existing analog and digital systems.

However, design engineers are working on a very different problem — making such communications systems fit cost goals! Everyone in the industry has spent plenty of time examining the huge new markets for RF-based devices. The result is a classic case of "Catch 22" — many markets won't materialize unless the prices are low enough, and the prices can't be lowered without the economy of scale offered by a large market.

I believe that many companies looking for a piece of these new RF markets have not yet realized how overwhelming the issue of cost really is!

For example, let's look specifically at remote utility meter reading. Here is a promising application that could result in the manufacturing of tens, or hundreds, of millions of transponders. Fewer employees are needed for meter reading, and bills can be sent (and collected) on a shorter turn-around, improving cash flow. But for the utilities to justify the investment, a certain cost-per-unit threshold must be crossed. At present, however, it is not possible to make these transponders at the right price.

So, how does a design team get the cost cut down to the right size? A little bit can come from elegant design, creative manufacturing methods, and efficient management — but that's not enough. The real cost savings will come from reducing the component count; integrating the RF, baseband and digital functions into as few devices as possible. The first steps of this process are underway, as manufacturers reduce the size of devices, and offer new devices addressing front-end, signal processing and local oscillator design with higher levels of integration. But a big step is yet to be taken before we reach the $15 remote meter reader or the $99 pocket-sized personal communicator.

Designing circuits to be directly built on silicon (or other substrates) is the focus of major efforts in ASIC design, analog/RF/digital modeling software and passive component design (e.g., filters). An optimist will say that the road is clear, its just a matter of the speed that we can muster. A more cautious observer could easily point out that the necessary commitment to specific markets and technology has not yet been made.

I'll take the optimist's view. Enough visionary entrepreneurs in companies large and small are ready to make a commitment — the rewards are too big to let them go by. Let's encourage them to go ahead at full speed; assuring that the RF boom contirues.

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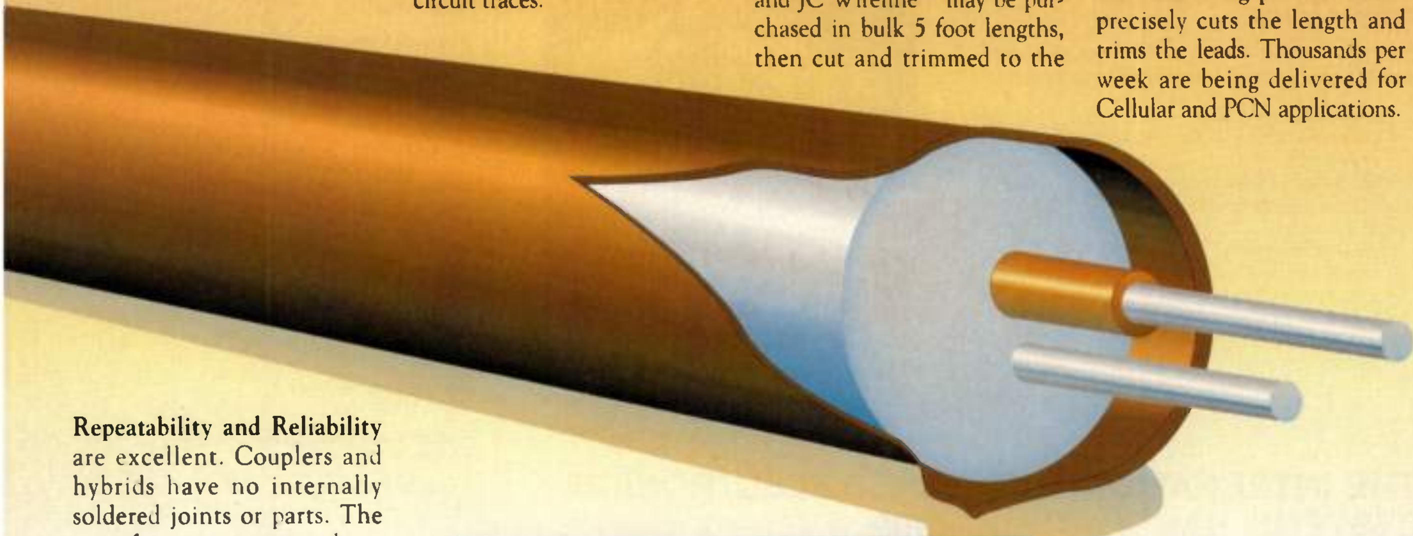

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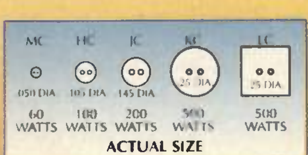

DROP-IN QUADRATURE COUPLERS AND HYBRIDS

PART NUMBER	FREQ MHz	MID BAND COUPLING	INSERTION LOSS	ISOL.	VSWR
HCB2-5.93	225-400	2.7 dB	0.3 dB	20 dB	1.2:1
JCB2-2.80	440-880	2.7 dB	0.3 dB	20 dB	1.2:1
LCB2-2.63	500-1000	2.7 dB	0.2 dB	30 dB	1.1:1
MCB3-2.06	810-990	2.8 dB	0.3 dB	20 dB	1.2:1
GCB6-2.06	810-990	10.25 dB	0.1 dB	30 dB	1.2:1
KCB2-1.31	1000-2000	2.7 dB	0.2 dB	30 dB	1.1:1
HCB1-1.00	1710-1990	3.0 dB	0.2 dB	20 dB	1.2:1
GCB6-1.00	1710-1990	10.25 dB	0.1 dB	30 dB	1.2:1
MCB3-0.14	9000-18000	2.8 dB	0.3 dB	20 dB	1.2:1

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RF letters

Letters should be addressed to: Editor, RF Design, *6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.*

We received many letters responding to the questions raised by Robert Orban concerning EMI rejection by consumer equipment and the power level of amateur radio transmitters ("Letters," February 1993 issue). Excerpts of several letters are included here.

— Editor

Who's to Blame?

Editor:

. . .It's easy to blame the messenger for the bad news, and the "news" is that the consumer is often victimized by poorly engineered, shoddily manufactured electronic equipment that works great on the showroom floor but fails in short order when exposed to the real world environment. Blaming ham radio operators for this problem is akin to American automakers blaming their loss in consumer confidence on the Japanese auto industry, and not the least bit on themselves.

W.J. Kleronomos
Lyons, CO

Editor:

. . .The situation is not as simple as Mr. Orban suggests. A recent study done by the FCC Field Operations Bureau concluded that "Operating power of 100 watts or less can cause interference," and "Simply reducing the transmitter power does not alleviate interference in most cases." The problem is not one of excess power overwhelming an otherwise well designed device, but rather it is the situation where a poorly designed device collapses in the face of an RF field.

To limit amateur operations to a power level of 10 watts would be contrary to the FCC's stated intentions, impractical and punitive to the amateurs, and the 20 dB reduction in near-field levels would not guarantee elimination of interference problems.
Stephen J. Root
South St. Paul, MN

The Cost of EMI Performance

Editor:

. . .Low EMI susceptibility, like quality, is free when it is an initial design consideration. It is quite expensive when it is an afterthought implemented with multiple layers of cut and try bandaids.

James Long
Consulting Engineer
Sunnyvale, CA

Editor:

. . .I am appalled by the dominant concern about increasing costs for manufacturers, when they make little effort to design their products with anything other than minimum parts count and minimum cost as their goals. I have, on numerous occasions, solved both amateur and commercial broadcasting interference problems to consumer equipment with the addition of parts that would cost a manufacturer no more than two dollars to incorporate in their designs . . .

The real issue is not whether we should "tax" consumers to "permit a few amateurs to indulge in their hobby," but rather, why should all of us be subjected to inferior designs which not only increase susceptibility to interference from all sources, but give consumers products which perform marginally! These same products routinely cause interference to licensed services, both amateur and commercial.

Jack Parker
KOZZ AM and FM
Reno, NV

Editor:

. . .I recall one experience designing automotive electronics, where company policy was more "leave it out" than "minimize cost sensibly." This led to numerous design cycles when the real world intruded into the wide open receiver front end. It also led to a half-million dollar loss; all because materials cost was "God." Spend the lousy few bucks up front, and you will save big later on by having a conservative design that will be producible.

Harold Chase
VVH Electronics Co.
Pepperell, MA

How Do You Get 9 V/m?

Editor:

. . .I calculated that 9 V/m could only be obtained by approaching the driven element of a 20-meter antenna to within three feet. I thought my answer must be off so I took my field strength meter to the local 1 kW broadcast station on 920 kHz and had to approach the antenna within three feet to obtain the same 9 V/m. The 9 V/m came from some stupid statement made by the FCC in *RF Design* for August 1992. [Actually, it is a worst-case measurement, as noted in Robert Weller's August article — *Editor*]

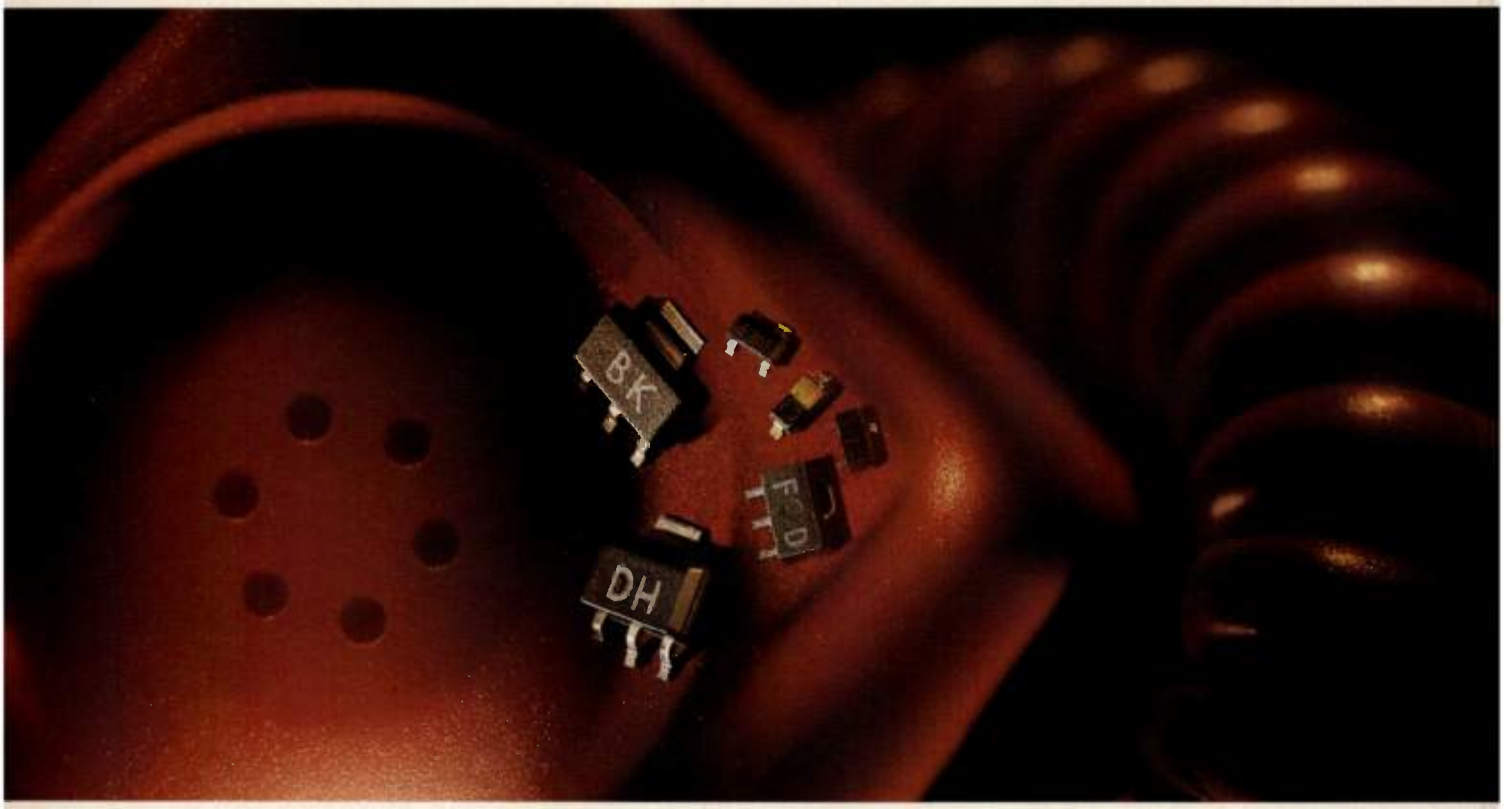
As for the cost of such a product, Mr. Orban is way off base here. I agree with his eight-to-one cost of component to retail price ratio, but in one of my designs the prototype RFI components cost seven dollars, for which the production cost was fourteen cents. If a manufactured product cannot be engineered to FCC Class B standards for less than that, get a new design engineer who understands RF protection!

Cliff Buttschart
Estero Systems
Morro Bay, CA

Photo Credit

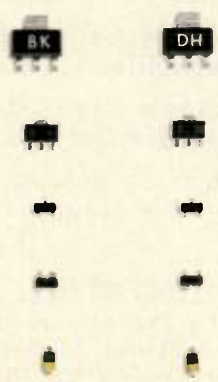
The photo accompanying March's "Product Report" should have been credited to AVX Corporaton. The photo depicts the ACCU-F chip capacitor from AVX.

SIEMENS



RF Performers

Announcing the most advanced RF transistors for mobile communications.



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BFP 280	8	10	1.4	18	1 to 5/0.2 to 8 low-noise pager
BFP 181	12	20	1.3	19	1 to 8/0.5 to 10 low-noise pre-amplifier
BFP 182	12	35	1.2	19	1 to 8/1 to 20 low-noise amplifier
BFP 183	12	65	1.2	19	1 to 8/2 to 28 low-noise amplifier
BFP 193	12	80	1.2	19	3 to 8/5 to 40 low-distortion output stage
BFP 196	12	120	1.4	18	3 to 8/30 to 100 low-distortion output stage

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SCI-1009

INFO/CARD 8

RF calendar

April

18-21 **The 4th IEE Conference on Telecommunications**
Manchester, UK
Information: ICT 93 Secretariat, Conference Services, IEE, Savoy Place, London, WC2R 0BL, United Kingdom.
Tel: 071-240 1871. Fax: 071-497 3633.

18-22 **Symposium on Ceramics for Wireless Communication**
Cincinnati, OH
Information: Henry O'Bryan, AT&T Bell Laboratories.
Tel: (908) 582-6980. Fax: (908) 582-2521.

18-22 **The NAB Multimedia World Conference and Exhibition**
Las Vegas, NV
Information: NAB, 1771 N Street, NW, Washington, DC 20036-2891. Tel: (202) 429-5350. Fax: (301) 216-1847.

19-22 **SUPERCOMM '93 International Conference and Exhibition**
Atlanta, GA
Information: Henry Weinland, United States Telephone Association (USTA), 900 19th Street NW, Suite 800, Washington, DC 20006-2190. Tel: (202) 835-3100.
Fax: (202) 835-3248.

26-28 **41st International Relay Conference**
Newport Beach, CA
Information: Pete Walsh, (202) 457-4932 or National Association of Relay Manufacturers (NARM) Headquarters, (414) 351-4548.

27-29 **Electro**
Edison, NJ
Information: Electronic Conventions Management, PO Box 92275, Los Angeles, CA 90009-2275.
Tel: (800) 877-2668. Fax: (310) 641-5117.

27-30 **IEEE 1993 International Conference on Acoustics, Speech and Signal Processing**
Minneapolis, MN
Information: Professor A.K. Katsaggelos. Tel: (708) 491-7164. Fax: (708) 491-4455.

28-30 **EMC/ESD International 1993**
Denver, CO
Information: Renae Fierros, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.
Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

May

2-6 **IEEE Circuits and Systems Symposium**
Chicago, IL
Information: Electronic Industries Association, EIA Components Group, 2001 Pennsylvania, Avenue N.W., Washington, DC 20006-1813. Tel: (202) 457-4930.

3-6 **1993 Annual Assembly Meeting of the Radio Technical Commission for Maritime Services**
San Diego, CA
Information: Katy Ackland, RTCM Assembly Publicity Chairperson, COMSTAT Mobile Communications, 950 L'Enfant Plaza SW, Washington, DC 20024. Tel: (202) 863-6097. Fax: (202) 488-3814.

RF courses

Digital Switching
April 29-30, 1993, Madison, WI
Cellular Radio
May 11-14, 1993, Madison, WI
Information: The University of Wisconsin-Madison, Engineering Information. Tel: (800) 462-0876.

Phased-Array Radar System Design
April 13-16, 1993, Atlanta, GA
Infrared/Visible Signature Suppression
May 18-21, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Synthetic Aperture Radar: Design, Processing, and Applications
April 26-30, 1993, Los Angeles, CA
Advanced Communication Systems Using Digital Signal Processing
April 26-30, 1993, Los Angeles, CA
Microwave/Millimeter-Wave Monolithic Integrated Circuits
May 11-14, 1993, Los Angeles, CA
Integrated Services Telecommunications Networking: High-Speed, Local, Metropolitan, and Wide-Area Networks
May 17-21, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825- 1047. Fax: (310) 206-2815.

Microwave System Engineering
April 26-30, 1993, Washington, DC
Simulation Modeling
April 13-16, 1993, Washington, DC
Communication and Radar Systems: Detection, Estimation, & Geolocation Techniques
April 14-16, 1993, Washington, DC
Grounding, Bonding, Shielding and Transient Protection
April 20-23, 1993, Washington, DC
Ionospheric Radio Propagation: Principles and Application
April 20-23, 1993, Washington, DC
Communication Satellite Engineering
April 26-30, 1993, Washington, DC
Satellite Communication Engineering Principles
May 5-7, 1993, Washington, DC
New HF Communication Technology: Advanced Techniques
May 10-14, 1993, San Diego, CA
Modern Radar Technology: Monopulse Tracking Techniques and Other High-Performance Developments
May 24-28, 1993, Washington, DC
Analog/RF Fiber Optic Communications
May 26-28, 1993, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Chemical Vapor Deposition for Microelectronics
April 21-23, 1993, Princeton, NJ
Modern RF & Microwave Techniques
April 20-23, 1993, Phoenix, AZ
Navstar/GPS
June 2-4, 1993, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335. Fax: (818) 995-2932.

Vacation School on Digital Techniques in Radio Systems
April 18-22, 1993, Leeds, UK
Sixth Vacation School on Microwave Measurements
May 2-7, 1993, Malvern, UK
Information: The Institution of Electrical Engineers, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071-240 1871. Fax: 071-497 3633.

Satellite Communication Systems: Techniques and Technology for Communications and Broadcasting
April 19-23, 1993, Cambridge, United Kingdom
Combined Coding and Modulation Techniques
April 20-21, 1993, Cambridge, United Kingdom
Mobile Telecom Systems: Switching, Fixed Network Interconnections
April 20-23, 1993, Cambridge, United Kingdom
Aspects of Modern Military and Commercial Radar
May 10-14, 1993, Cambridge, United Kingdom
Low Sidelobe Antennas
May 10-14, 1993, Cambridge, United Kingdom
Modern Microwave Techniques: Measurements, Signal and Network Analysis, Microwave Products and Systems Characterization
May 10-14, 1993, Cambridge, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

Radar Simplified EW Receivers
May 4-6, 1993, Washington, DC
ELINT Analysis, Adaptive ECCM Signal Processing for Radars, Electromagnetic Propagation
May 11-13, 1993, Washington, DC
ELINT Interception, Elint/EW Applications of Digital Signal Processing, Radar Vulnerability to Jamming
May 18-20, 1993, Syracuse, NY
ELINT/EW Data Bases
June 1-3, 1993, Washington, DC
Information: Research Associates of Syracuse, Incorporated, Hancock Army Complex, 510 Stewart Drive, N. Syracuse, NY 13212. Tel: (315) 455-7157.

1993 High-Speed Digital Symposium
April 22, 1993, Los Angeles, CA
April 27, 1993, San Diego, CA
April 29, 1993, Phoenix, AZ
May 25, 1993, Dallas, TX
May 27, 1993, Austin, TX
June 2, 1993, Burlington, MA
June 4, 1993, Washington, DC
June 8, 1993, Boca Raton, FL
Information: Hewlett-Packard Company, Microwave Instruments Division (MID). Tel: (800) 765-9200.

DSP Without Tears
June 7-9, 1993, Norcross, GA
June 16-18, 1993, San Jose, CA
June 23-25, 1993, Chicago, IL
Advanced DSP With a Few Tears
June 10-11, 1993, Norcross, GA
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967-5034, (404) 664-6738. Fax: (404) 442-1210.

PENSTOCK RF/MICROWAVE "WIRELESS"

DESIGN CONTEST

Contest will be judged on: Originality, Technical Content, Marketability and Ease of Manufacturing.
JUDGED BY: PENSTOCK—Dr. Bob Craig • RF DESIGN—Gary Breed • MICROWAVE JOURNAL—Harlan Howe, Jr. • MICROWAVES & RF—Jack Browne • TOKO AMERICA—Steve Skiest

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PENSTOCK is looking for the most unique RF and MICROWAVE wireless application using TOKO (TOKO AMERICA, INC.) parts. Your design should be laid out in Block Diagram form.

The design should show where PENSTOCK TOKO AMERICA components fit in. A brief description (under 500 words) of the product and application (along with a color photo of a working prototype) should be included.

Your design will be judged on its application to any of the following technologies: COMMUNICATIONS/TELECOM, INSTRUMENTATION, MEDICAL, COMPUTER & NETWORKING, CONSUMER APPLICATIONS, DATA SYSTEMS, AUDIO, VIDEO & IMAGING SYSTEMS.

Your entry should include the coupon below (photocopy OK). PENSTOCK reserves the right to exclude any company and/or person from this contest due to conflict of interest. Once the Grand Prize and 2nd Place Winners are chosen, PENSTOCK may publish the Design/Applications in journal form to be used as training and promotional materials.

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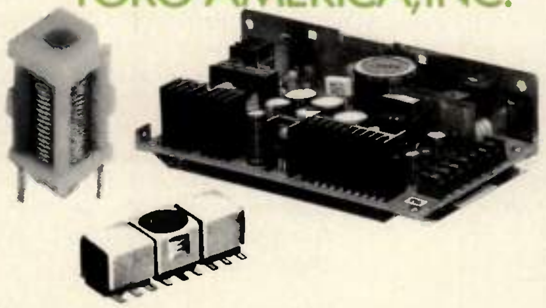

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CONTEST DEADLINE—JUNE 1, 1993. Winners will be announced by SEPTEMBER 1, 1993.

Name ______________________________

Company ______________________________

Address ______________________________

City ______________ State __________ Zip __________

Phone ______________ FAX ______________

Design/Product/Application ______________________________

____

I agree to all the terms of the Design Contest and have not worked for any Penstock distributed companies past or present. I am including a description & photo of a working prototype. I agree to allow Penstock to publish my work.

Signature ______________________ Date __________

Send To: PENSTOCK DESIGN CONTEST 520 Mercury Drive, Sunnyvale, CA 94086

INFO/CARD 11

RFID Used to Track Ostriches and Emus

Texas Instruments RFID technology is now being used by breeders of ostriches and emus to track and identify the valuable birds, without putting either handler or bird at risk. The Texas Instruments Registration and Identification System (TIRIS) is being used to identify the birds, which can sell for as much as $40,000 for a breeding pair of ostriches, for medical, genetic and insurance purposes. Traditional methods of restraint and identification can not only be harmful to the animal, but also places breeders themselves at risk of the bird's powerful kick.

A small transponder, 32.5 x 3.85 mm, is implanted under the skin beneath the bird's wing when it is as young as three months old. The transponder transmits an unalterable identification number back to a hand-held reader when interrogated. A specially designed long-reaching antenna on the reader, combined with a read range in excess of two feet, allows the breeder to accurately identify the bird from several feet away. Dr. Robbins of the Folsom Animal Hospital, who is training breeders to use the RFID technology, notes that while the bird must initially be restrained when the transponder is inserted, it feels no pain because the transponder is injected into an unsensitive part of the skin.

Dr. Brent Robbins of Folsom Animal Hospital identifies an emu using TIRIS™ radio frequency technology from Texas Instruments.

Ostrich breeding is becoming a rapid growth industry in the U.S., and the TIRIS technology allows serious breeders to protect their investment.

Call for Papers — The Sixth International Conference on Electronic Engineering in Oceanography has issued a call for papers for their conference to be held July 19-21, 1994 in Cambridge, United Kingdom. Some of the suggested topics for papers include: monitoring networks; remote sensing, satellite, radio and acoustic; satellite navigation and communication; sonar systems and applications; and standards, measurements and test facilities. A synopsis of approximately 500 words, indicating the appropriate topic area and key novel aspects of the paper, should be sent to Louise Bousfield, EEO '94 Secretariat, Conference Services, Savoy Place, London WC2R OBL, United Kingdom by August 15, 1993.

Search and Rescue Effort Aided by Satellite System — The lives of eleven fishermen were saved when the 406 MHz Emergency Radio Beacon on board their vessel, *Cape Aspy*, was triggered, and began transmitting an emer-

Wireless Communication at Your Finger Tips

Low Cost Convolutional Encoder/Veterbi Decoder
With a 9 Mbps data rate, the STEL-2030A operates at code rate 1/2 and also with punctured codes at code rates 2/3 and 3/4. It also includes a V.35 Scrambler and BER Monitor.

Highest Flexibility PN Coder
At 30 Mcps and up to 32 taps per coder the STEL-1032 Triple PN Coder has the flexibility to generate a wide variety of codes for many spread spectrum applications.

Digital Downconverter and Carrier Tracker
The STEL-2130 can be used to convert signals digitized at up to 40 Msamples/sec at I.F. directly to baseband.

Highest Performance Digital Matched Filter (cascadable)
The STEL-3310 64-Tap Cascadable Complex Matched Filter provides fast acquisition of PN sequences up to 256 chips/symbol at up to 22 Msamples/sec.

Low Cost and Low Power Digital Matched Filter (noncascadable)
The STEL-3340 can be used in many of the same application as the STEL-3310 with a limit of 64 chips/symbol, with lower cost and power consumption.

Phase Modulated NCO
The STEL-1175 Numerically Controlled Oscillator offers 32-bit resolution and 80 MHz clock frequency. It can be phase modulated at up to 20 MHz.

Highest Speed Accumulator/Correlator
At a 70 Msamples/sec accumulation rate, and up to 32K samples without overflow, the STEL-2410 is the industry's highest speed signal accumulator/correlator for spread spectrum applications.

Bit Synchronizer and BPSK/QPSK Demodulator
The STEL-2110A can be used for coherent and differential demodulation of signals up to 6 Mbps BPSK and 12 Mbps QPSK. The STEL-2110A includes a symbol timing NCO and phase or frequency discriminator for carrier tracking.

Differential BPSK/QPSK Demodulator
The STEL-2120 can be used for differential demodulation of signals up to 8 Mbps BPSK and 16 Mbps QPSK in both spread spectrum and clear channel applications.

Fastest Burst Data Acquisition
The STEL-2210 Block Phase Estimator allows fast acquisition of burst data. It can be used for coherent demodulation of signals at up to 8 Mbps BPSK and 16 Mbps QPSK without a carrier tracking PLL.

For further information on these or any other product offered by Stanford Telecom, please contact us today.

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INFO/CARD 13

gency signal on January 31, 1993 when the ship sank. The signal was detected by a geostationary satellite and transmitted to the ground station STARLUT, developed and manufactured by CAL Corporation and owned by the Department of National Defence. The information was decoded and the identity of the ship confirmed. Another Local User Terminal manufactured by CAL was able to further assist the rescue efforts by informing rescue personnel of the location of the vessel within two kilometers. The search and rescue team, using the beacon information, were able to locate and pull the survivors to safety from a life raft.

New Commercial GPS System On-Line — Pinpoint, a nationwide high-accuracy satellite positioning and location network was recently announced. The new service will permit a user to determine his geographic position within a few feet using signals from U.S. navigational satellites and transmissions from local FM radio stations. Pinpoint is the result of a strategic alliance between Magnavox Electronic Systems Company and CUE Network Corporation. Magnavox and CUE have developed techniques to enhance the accuracy of GPS for high-precision applications, by measuring errors in the satellite signals and transmitting correction data through FM radio subcarriers. Subcarriers are unused frequencies within a station's licensed bandwidth. This technique is called "differential GPS" and it improves positioning accuracy from about 300 feet to 15 feet or better. Following the initial introduction in Los Angeles, the service will be expanded to as many as 50 additional cities across North America.

ICC Approved by ITI — International Compliance Corp recently received their Certificate of Assessment from Interference Technology International, Ltd. to provide EMC testing services for specifications implemented by the EC Directive on EMC.

Scientific Atlanta Division To Receive ISO 9001 Certification — Scientific Atlanta's Electronic Systems Division has passed the final assessment audit for certification to ISO 9001. ISO 9001 is a stringent international quality standard initially designated as a requirement for suppliers to meet in order to do business in the European Economic Community.

TRM Awarded Million Dollar Contract — Technical Research and Manufacturing, Inc. has been awarded a million dollar contract from Allied Signal Aerospace Company, Bendix Communications Division, for the design, development and manufacture of RF and microwave components in the RF Distribution Unit for the E-Scan Secondary Surveillance Radar. The first field use of the ESSR system is for precision runway monitoring at U.S. airports.

Park Air To Supply FAA with Emergency Transceivers — The Federal Aviation Administration recently

So, moving up to this level of speed and performance is as easy as unplugging that old box and plugging in your new 8540 Series Power Meter. Think about what all this will do for your ATE productivity as well as for your company's bottom line.

FAST, EASY PEAK POWER MEASUREMENT

Now, an easy-to-use CW power meter can also measure pulsed RF signals with the simple addition of a peak power sensor.

There are no time-consuming, unreliable duty cycle corrections, and you'll get the same accuracy and speed you'd get with a much-more-expensive dedicated peak power meter.

View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.

You'll be confident of your peak power readings without sacrificing any of the benefits of having an incredibly fast CW power meter.

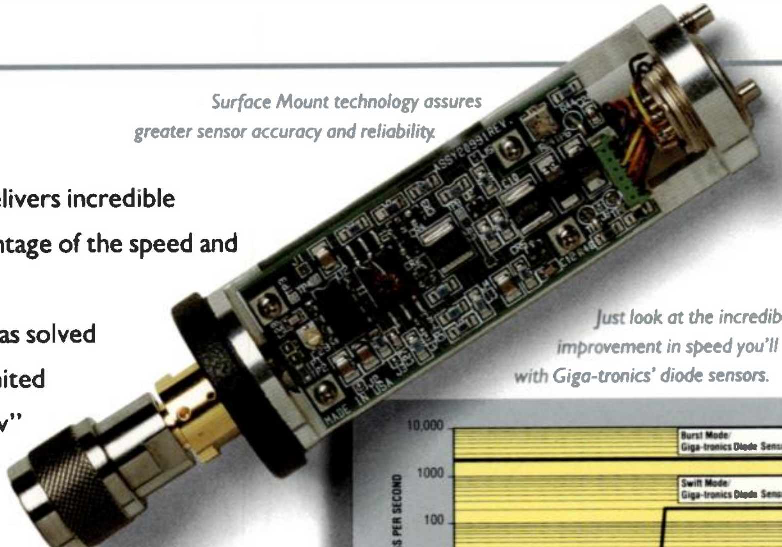
ONE OR TRUE TWO CHANNEL OPERATION

If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, prepare yourself for a pleasant surprise: With the Model 8542, you can see readings from both channels *simultaneously*.

Use one line for CW and the other for peak measurements, and see both readings at the same time. Or display readings in dBm on one line and mW on the other for gain and output power.

The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

What's more, Giga-tronics has solved the challenge that previously limited diode sensors to the "square law" region—below −20 dBm—by utilizing a built-in power sweep calibration system. So you get speed and dynamic range without sacrificing accuracy.

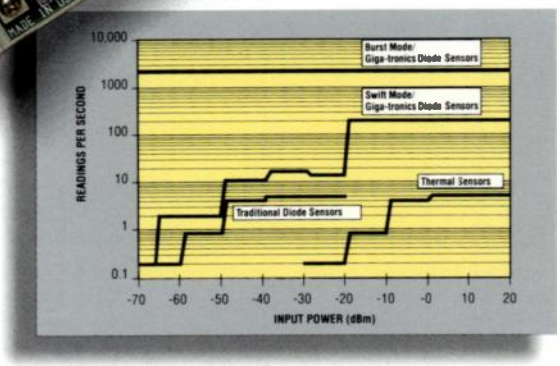
For incredible versatility, we offer a full range of CW and Peak sensors tailored to your specific needs. Our sensors cover frequency ranges from 10 MHz to 40 GHz with up to 90 dB dynamic range. There are 1 to 50 Watt high power sensors, too.

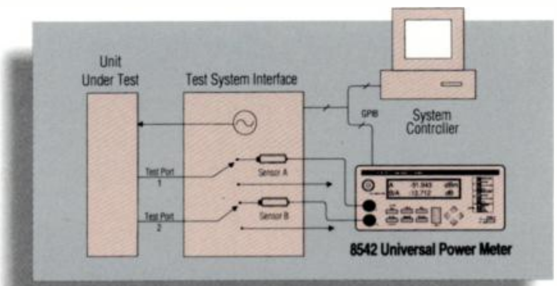
Use our diode-based True RMS sensors to accurately measure the 1 dB gain compression power of amplifiers, while our low VSWR sensors give you unequalled CW measurement accuracy. Or connect a Precision Return Loss Bridge and measure return loss by using just a single channel.

So why wait? Get the truly incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.

Call us toll free at **1 800 726 GIGA.** Outside the U.S. call your local Giga-tronics representative, or call 001 408 734 5780. We'll send you more information or arrange for an incredible hands-on demonstration.

Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.

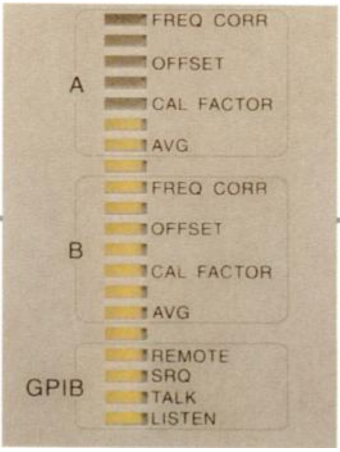




Cut your measurement time and reduce switching complexity by using one Giga-tronics 90 dB sensor instead of the two 50 dB sensors you were once forced to use.

Giga-tronics

Giga-tronics Incorporated
488 Tasman Drive
Sunnyvale, California 94089
Telefax: 408 747 1265



SIMPLE, INTELLIGENT OPERATION

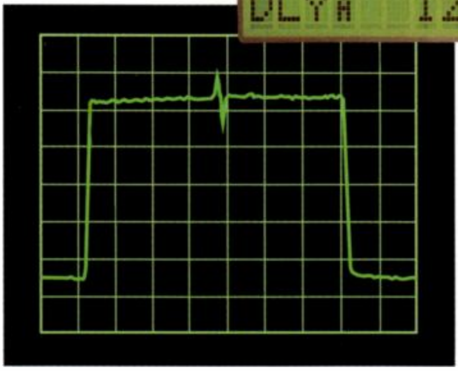
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

Twenty LEDs confirm selections, show status of GPIB operation and function as an electronic peaking meter.

For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

A two-line LCD display shows prompts for instrument settings, as well as measurements. While a 20-segment LED bargraph shows instrument status and also acts as a peaking display.

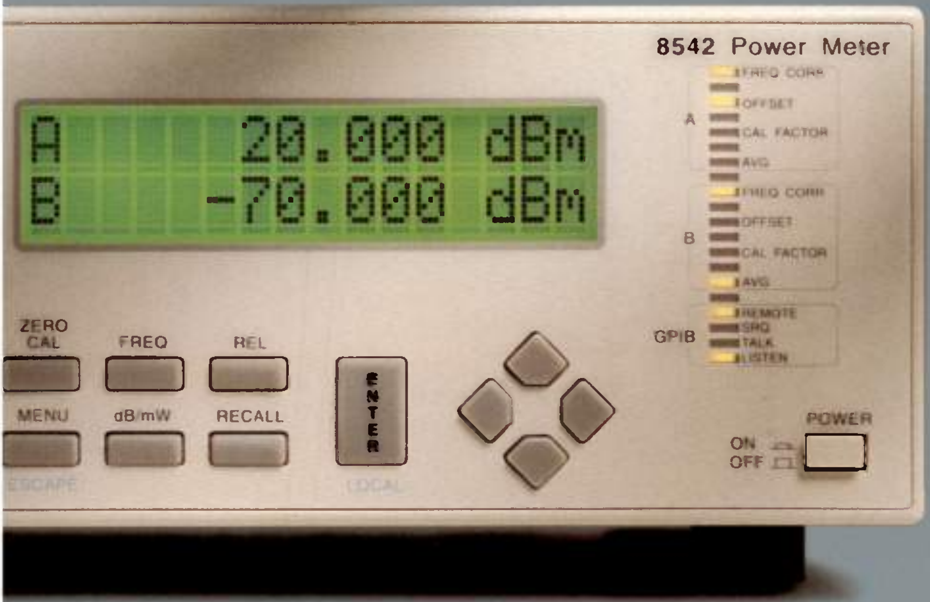
A two-line back lit LCD display provides you more data in less time.



The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulse signals.

Giga-tronics

Incredibly Fast CW Measurements From −70 to +20 dB Without Changing Sensors.



An Incredible Power Meter At A Truly Incredible Price.

UNIVERSAL POWER MEASUREMENT

Incredible is credible when describing the 8540 Series of Universal Power Meters.

From Giga-tronics, the new power in power meters.

For the very first time, you can make CW *and* peak power measurements quickly and accurately with a single meter—a Universal Power Meter.

And all for about the same price you'd pay for the competitor's CW *only* power meter.

POWER MEASUREMENTS INSTANTLY

Imagine seeing display updates instantly: measurement speeds over the GPIB exceeding 200 readings per second and an exclusive Burst Mode capturing more than 2,000 readings in the same tick of a clock.

The two-line display also lets you set the desired resolution and select either Lin or Log readout for each line.

And because the 8540 Series uses diode sensors, you can measure all the way from −70 to +20 dBm with the same sensor, and without range changing delays.

If you're worried about having to write new code for your computer controlled testing, don't be: The 8540 Series uses the same GPIB command set as HP's 436A, 437B and 438A.

GIGA-TRONICS

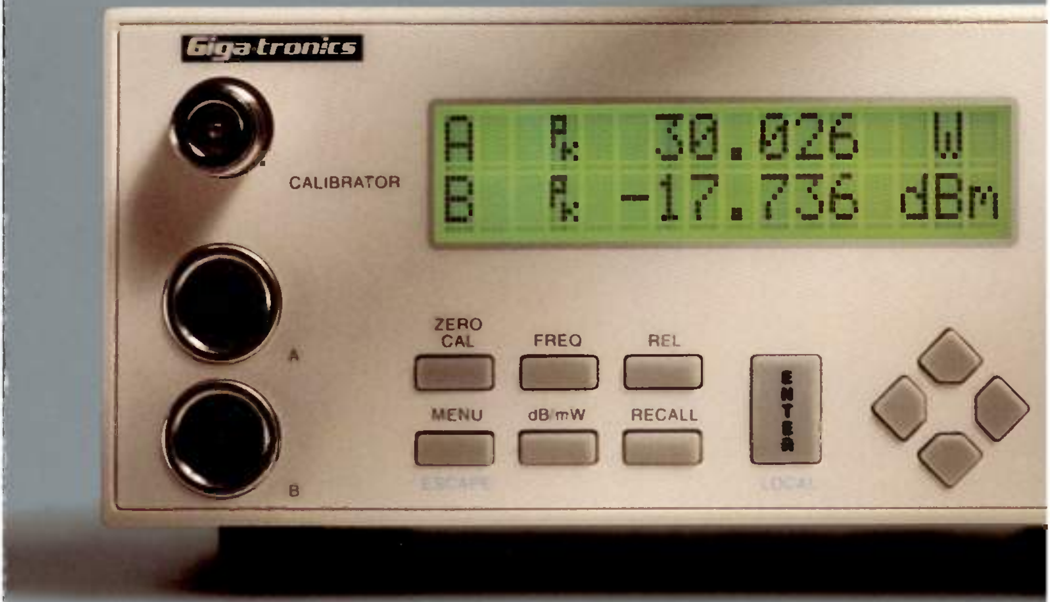
8540

UNIVERSAL

POWER

METER

Incredibly Accurate Peak Measurements Just By Changing Sensors.



chose Park Air to supply the new Denver International Airport with UHF and VHF transceivers and specially designed remote control units. The units will provide the air/ground emergency communications for the airport in the event of a primary communications failure.

Scientific-Atlantic Receives Contracts — Scientific-Atlanta recently received two contracts. The first contract was for $3.2 million from NASA to modify one of three transportable ground stations currently being produced for NASA. The modified equipment will be deployed by NASA as a participant in the Canadian Government's Radarsat microwave radar earth imaging program and will be used to receive microwave radar image data from polar orbiting satellites.

The second contract, with CEMEX calls for the initial installation of a Very Small Aperture Terminal network to link CEMEX's headquarters with 50 of its cement and concrete distribution centers and manufacturing facilities throughout Mexico. CEMEX will operate a private satellite master station at its Monterrey headquarters. Installation of the private hub and the 50 VSATs will be completed by June.

Micro-Coax Awarded Follow On Contract — Micro-Coax has been awarded a follow-on contract for In-A-Cable filters by the Defense Systems and Electronics Group of Texas Instruments. The contract award is an extension of a long-standing contract with TI for the High-speed Anti-Radiation Missile (HARM) program. The filters are designed into the missile's guidance system. The amount of the contract award was not released.

Radian Receives DOE Contract — Radian Corporation recently completed negotiations with a U.S. Department of Energy laboratory to supply a 50 MHz radar wind profiler system. This is the first radar profiler that DOE has purchased for use at Atmospheric Radiation Measurement Program (ARM) sites. ARM is a segment of the DOE Global Change Research Program. Multiple 50 MHz and 915 MHz profilers will be required at each ARM Cloud and Radiation Testbed (CART) site. The CART sites, located worldwide, will be developed at the nominal rate of approximately one every 18 months.

Flam & Russell Delivers System to Sandia — Flam & Russell, Inc. recently delivered a turnkey Antenna and Radar Cross Section (RCS) Measurement System to Sandia National Laboratories. The system, an advanced version of the FR959, makes use of the new HP 85330 Multiple Channel Controller to perform fully polarimetric RCS measurements and multiple port antenna measurements.

KW Microwave Relocates — KW Microwave has announced the relocation of their facility to Carlsbad, California. The new address is 1985 Palomar Oaks Way, Carlsbad, CA 92009. Tel: (619) 929-9800. Fax: (619) 929-9899.

Lucas Awarded Contracts — Lucas Aerospace was recently awarded two contracts. The first was from Hughes

Aircraft Co., Systems Sector to provide low-power and medium-power amplifiers. Under the terms of the contract, which is valued in excess of $3 million, Lucas Zeta will manufacture, integrate and test the amplifiers. The second contract was awarded to Lucas Aul, a subsidiary of Lucas Aerospace Communications Business. The follow-on contract calls for Lucas to provide AN/SRQ-4 Radio Terminal Sets to the U.S. Navy's Light Airborne Multi-Purpose System (LAMPS) program. This brings the total value of the contract to date to $25.7 million.

Amp Achieves ISO Certification — The AMP Aerospace and Government Systems Sector's Federal Systems business group has received International Standards Organization (ISO) 9001 certification. The sector joins 10 other business units in AMP as adherents of the ISO 9000 international quality management and assurance standard.

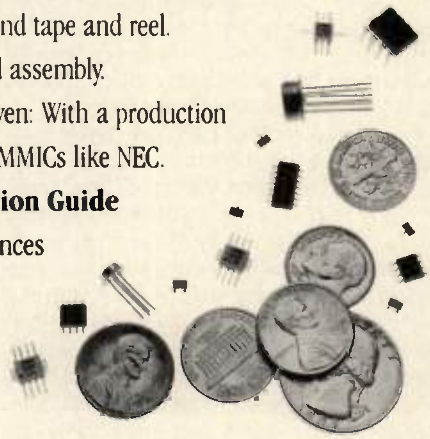
AMD and Rochester Sign Afterlife Pact — Advanced Micro Devices, Inc. and Rochester Electronics, Inc. have entered into an agreement that will allow AMD's customers to continue to buy IC products that AMD has discontinued. The agreement provides Rochester with all residual products (i.e., wafers, dice, finished goods) and access to selected tooling and process recipes for discontinued AMD devices. The agreement covers military and commercial sole-source and multiple-source products.

SBIR Awarded to HiTc — HiTc recently announced that they have received a Small Business Innovation Research Program (SRIB) grant from NASA for a process to produce useful quantities of bulk high temperature superconductors. The process uses HiTc's 3-P technique which starts with melt textured growth generated powder which is subsequently magnetic-pinning-site refined to provide completely phase-pure, clean grain boundary bulk materials.

Ray Proof Ltd Moves — Ray Proof Ltd has moved its RF interference technology operation from Enfield, North London to new premises near Stevenage, Herts. The 47,000 square foot facility houses all aspects of the company's activities, including design, manufacturing, marketing and support services. Their new address is Boulton Road, Pin Green Industrial Area, Stevenage, Herts SG1 4TH, England. Tel: (44) 0438 747477. Fax: (44) 0438 747170.

MPD Moves Operations — Microwave Power Devices recently announced its move to an 85,000 square foot facility. The new address is 49 Wireless Boulevard, Hauppauge, New York 11788. The telephone number remains the same and the fax number is (516) 231-8081.

LEMO Rf Established — LEMO recently announced the establishment of LEMO Rf as a company to meet the requirements of the specialty RF/microwave high frequency connector/component market. They may be reached at 345 Tesconi Circle, Santa Rosa, CA 95401. Tel: (707) 578-8811. Fax: (707) 578-0869.



The Technology of Personal Communications

By Gary A. Breed
Editor

The promise of huge new markets for RF-based personal communications is the main force driving the RF industry right now. The development of these new markets depends on many factors — government regulations, national and international frequency allocations, technical operating standards, and of course, eventual acceptance by the customer. Rather than reexamine the political and regulatory side of personal communications, this short report will note the changes in engineering techniques and RF components that are required if engineers are to design cost-effective personal communications products.

Digital Cellular

The Pan-European GSM system has brought digital cellular from the lab to production. Digital modulation techniques have created new requirements beyond the current analog FM cellular systems. Linearity is the main change — low level and power amplifiers must utilize class A or AB to accurately transmit the QPSK digital modulation that is part of GSM, as well as coming U.S. and Japanese systems. Transistor manufacturers and power amplifier module makers have been busy creating products with the required linearity, while addressing the additional matter of efficiency for reduce power consumption.

Filtering is another part of a digital cellular system that needs a change from analog transmission. Delay characteristics are now of paramount importance. Uniform group delay (linear phase) places new constraints on filter designs, especially because this type of performance results in less selectivity for the same order of filter. The flexibility, small size, and high frequency operation of SAW filters appears to give that technology an edge in digital cellular IF filtering.

Modulators are one more area of development. Matched, monolithic modulators and demodulators are generally the best solution for high performance I and Q channel processing. Preamplifiers, buffers and baseband filtering on the same chip adds a degree of control and consistency.

Mobile Voice and Data

Transmission of data on channels normally used for voice can be viewed in two ways — short term and long term. In the long term, digital cellular will be a truly digital system, and traditional multiplexing techniques will allow data and voice to share the same system.

In the short term, however, transmitting data on an analog cellular system raises some problems that are not significant with voice-only communications. Data rate is one area of concern. The limitations of the modulation bandwidth and IF performance are part of the problem, placing a firm upper limit on transmission channel capability. Multipath propagation (signals bouncing around and arriving at different times) is another problem. Digital systems will have error-correction schemes to alleviate this problem, but any digital-on-analog system must include it external to the radio equipment.

These problems, the protocols of sharing voice and data on the same channels, plus provisions for error-checking are included in several systems already in use (e.g., Mobitex and others). These interim solutions provide at least modest capabilities until a fully digital cellular infrastructure is in place.

DECT/CT-2/PCS/PHP

This issue's cover story on page 48 announces one company's commitment to the Digital European Cordless Telephone (DECT) system. The British CT-2, Japanese Personal Handy Phone (PHP) and the roughly outlined U.S. Personal Communications Service (PCS) are all in a period of rapid development. Many companies and independent analysts believe that this type of service will be the best market possibility in the near future — it fits between cellular and cordless phones in cost and capability.

The problems are similar to digital cellular, since these are all digital systems with RF portions that are not radically different from one another, or from digital cellular systems. The unique areas of engineering problem-solving are in power consumption and cost. Where a cellular telephone can be considered a "professional" product, this family of products is certainly intended to reach the consumer market. Price of the finished product is a major factor, while the convenience of long operating time is another. One additional factor is that most of these systems will operate in the 1.7-1.9 GHz range, twice the frequency of the cellular band.

High level integration, 3-volt device technology, miniature packaging and highly automated manufacturing have seen dramatic increases in activity in companies pursuing advanced cordless telephone products. All this must be accomplished in systems that operate at around 2 GHz. Cellular technology required 900 MHz devices, but at twice the frequency, refinements or new processes must be brought to bear.

Multi-Function Systems

Recently, we have seen some highly-publicized announcements of units called "personal digital assistants" or something in a similar vein. The most publicized are from Apple Computer and EO Systems, but others are in the works, as well. These systems integrate notebook computers, cellular telephones and software intended to put all the information needs of a busy business person in one package. From an RF engineer's perspective, most of the same concerns noted above are in play, along with increased pressure for miniaturization and low power consumption. Another new addition is interface requirements among the various functional components — radio link, computer, phone, data network, and operating system software.

All of these new applications present new problems for RF engineers and their digital, mechanical, software and manufacturing counterparts. For most companies, the challenges are welcome because, if met properly, they can mean profitability and success. **RF**

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By Greg Magin
Intellon Corporation

In 1985 the Federal Communications Commission revised the Part 15 rules and regulations governing non-licensed RF communications devices to permit coherent spread spectrum operation. The FCC allows power output levels of up to 1 Watt spread over at least 500 kHz of bandwidth in three different frequency bands: 902-928 MHz, 2400-2483.5 MHz, and 5725-5850 MHz (1). The high power spread spectrum operation can provide reliable non-licensed communications while reducing the potential for harmful interference.

The FCC Part 15 rules allow the use of direct sequence, frequency hopping, and hybrid spread spectrum methods. A new digital signaling technique has been developed for operation under the current spread spectrum provisions which is called Spread Spectrum Carrier Technology™. The approach is optimized for low cost Carrier Sense Multiple Access (CSMA) RF packet networks where many users or nodes share the same channel frequency. Based on direct sequence spreading, the system features very fast synchronization for rapid channel access. The spreading and correlation techniques have been designed to relax the carrier frequency accuracy requirements of the radio link.

A simple limiter based receiver can be used to recover the transmissions. The transmission and reception processing can be performed by a single digital Application Specific Integrated Circuit (ASIC), Figure 1.

There are many potential applications for Spread Spectrum Carrier (SSC) Technology ranging from commercial to residential environments. The system is optimized for the rapid transmission of moderate length control type messages in a packet format. Early applications include: remote utility meter reading, load management, security systems, Heating Ventilation and Air Conditioning (HVAC), and vending machine networking. The growing demand for data communication and control systems will provide many future opportunities for this new approach.

The Electronic Industry Association (EIA) has adopted SSC as the Consumer Electronics Bus (CEBus) home automation communication standard for RF links (2). CEBus is intended to provide a low cost packet based control network environment over which consumer appliances and devices can communicate. The CEBus committees have defined standards for various media including: radio frequency, AC power line, coaxial cable, infrared, and twisted pair. The protocols employed over the various media have a common interface to reduce software requirements and to facilitate bridging between media.

Signal Generation

The CEBus network transfers information from point to point in the form of short packets, as shown in Figure 2. The packets typically begin with a brief On Off Keyed (OOK) preamble during which channel access and contentions are resolved. Once the channel is secured the packet body containing address, control, or data is sent using Phase Shift Keying (PSK). The CEBus RF standard utilizes pulse width encoding to represent the symbols "1", "0", End Of Field (EOF), and End Of Packet (EOP). The minimum symbol period is defined as the Unit Symbol Time (UST). The duration of the UST is 100 mS dur-

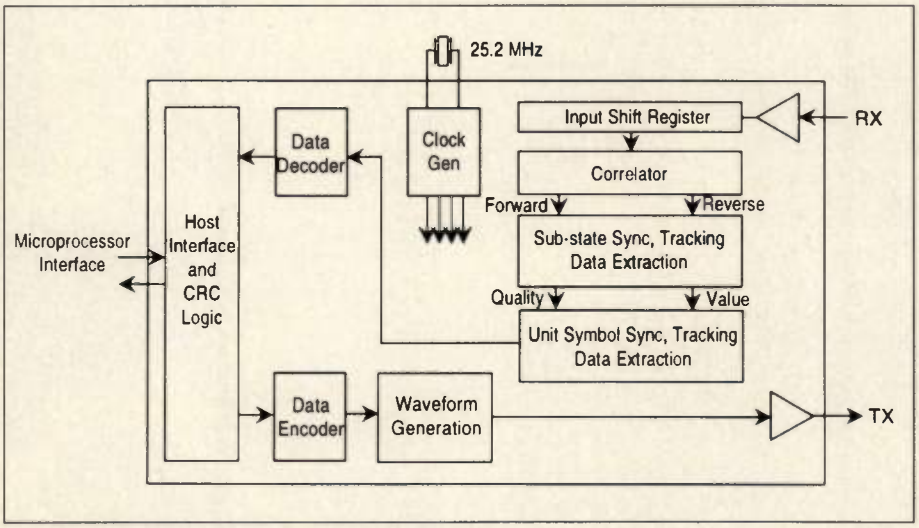

Figure 1. Block diagram of modem ASIC.

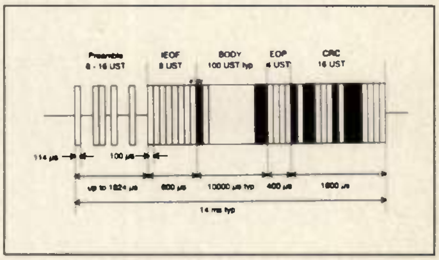

Figure 2. Typical packet timing.

FILTERS

dc to 3GHz from 11^{45}

lowpass, highpass, bandpass

• less than 1dB insertion loss • greater than 40dB stopband rejection • surface-mount • BNC, Type N, SMA available
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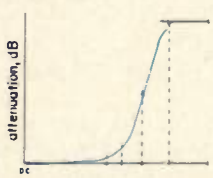


low pass, Plug-in, dc to 1200MHz

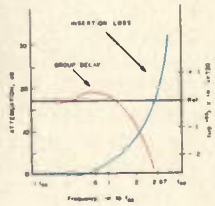
Model No	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB	Model No	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
★LP-5	DC-5	8-10	10-200	★LP-250	DC-225	320-400	400-1200
★LP-10.7	DC-11	19-24	24-200	★LP-300	DC-270	410-550	550-1200
★LP-21.4	DC-22	32-41	41-200	★LP-450	DC-400	580-750	750-1800
★LP-30	DC-32	47-61	61-200	★LP-550	DC-520	750-920	920-2000
★LP-50	DC-48	70-90	90-200	★LP-600	DC-680	840-1120	1120-2000
★LP-70	DC-60	90-117	117-300	★LP-750	DC-700	1000-1300	1300-2000
★P-90	DC-81	121-137	167-400	★LP-800	DC-720	1080-1400	1400-2000
★LP-100	DC-98	146-189	189-400	★LP-850	DC-760	1100-1400	1400-2000
★LP-150	DC-140	210-300	300-600	★LP-1000	DC-900	1340-1750	1750-2000
★LP-200	DC-190	290-390	390-800	★LP-1200	DC-1000	1620-2100	2100-2500

Price, (1-9 qty), all models plug-in $14.95, BNC $32.95, SMA $34.95 Type N $35.95

Surface-mount, dc to 570MHz

SCLF-21.4	DC-22	32-41	41-200	SCLF-190	DC-190	290-390	390-800
SCLF-30	DC-30	47-61	61-200	SCLF-380	DC-380	580-750	750-1800
SCLF-45	DC-45	70-90	90-200	SCLF-420	DC-420	750-920	920-2000
SCLF-135	DC-135	210-300	300-600				

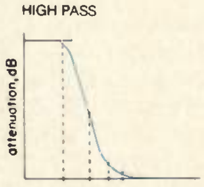
Price, (1-9 qty), all models $11.45



Flat Time Delay, dc to 1870MHz

Model No	Passband MHz loss < 1.2dB	Stopband MHz loss > 10dB	Stopband MHz loss > 20dB	VSWR Freq. Range, DC thru 0.2fco X̄	VSWR Freq. Range, DC thru 0.6fco X̄	Group Delay Variations, ns Freq. Range, DC thru fco X̄	Group Delay Variations, ns Freq. Range, DC thru 2fco X̄	Group Delay Variations, ns Freq. Range, DC thru 2.67fco X̄
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.7	4.0	5.0
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.9
★BLP-156	DC-94	312-416	416	0.3:1	1.1:1	0.3	1.1	1.5
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.4	1.3	1.6
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.2	0.6	0.8
★BLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
▲BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
▲BLP-1870	DC-850	3740-6000	5000	1.45:1	2.9:1	0.05	0.1	0.15

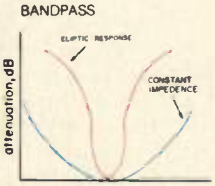
Price, (1-9 qty), all models plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE ▲ -933 and -1870 only with connectors, at additional $2 above other connector models



high pass, Plug-in, 27.5 to 2200MHz

Model No	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband, MHz loss < 1dB	VSWR Passband Typ	Model No	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband, MHz loss < 1dB	VSWR Passband Typ
★HP-25	DC-13	13-19	27.5-200	1.8:1	★HP-400	DC-210	210-290	395-1600	1.7:1
★HP-50	DC-20	20-26	41-200	1.5:1	★HP-500	DC-280	280-365	500-1600	1.8:1
★HP-100	DC-40	40-55	90-400	1.8:1	★HP-600	DC-350	350-440	600-1600	2.0:1
★HP-150	DC-70	70-95	133-600	1.8:1	★HP-700	DC-400	400-520	700-1800	1.6:1
★HP-175	DC-70	70-105	160-800	1.5:1	★HP-800	DC-445	445-570	780-2000	2.1:1
★HP-200	DC-90	90-116	185-800	1.6:1	★HP-900	DC-520	520-660	910-2100	1.8:1
★HP-250	DC-100	100-150	225-1200	1.3:1	★HP-1000	DC-550	550-720	1000-2200	1.9:1
★HP-300	DC-145	145-170	290-1200	1.7:1					

Price, (1-9 qty), all models plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95



bandpass, Elliptic Response, 10.7 to 70MHz

Model No	Center Freq. (MHz)	Passband I.L. 1.5 dB Max (MHz)	3 dB Bandwidth Typ (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
★BP-10.7	10.7	9.6-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
★BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
★BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
★BP-60	60.0	55.0-67.0	49.5-70.5	44 & 79	4.6 & 190-1000
★BP-70	70.0	63.0-77.0	68.0-82.0	51 & 94	6.0 & 193-1000

Price, (1-9 qty), all models plug in $18.95,
BNC $40.95, SMA $42.95, Type N $43.95

Constant Impedance, 21.4 to 70MHz

Model No	Center Freq MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR 1.3:1 Total Band MHz
★IF-21.4	21.4	18-25	1.3 & 150	DC-220
★IF-30	30	25-35	1.9 & 210	DC-330
★IF-40	42	35-49	2.6 & 300	DC-400
★IF-50	50	41-58	3.1 & 350	DC-440
★IF-60	60	50-70	3.8 & 400	DC-500
★IF-70	70	58-82	4.4 & 490	DC-550

Price, (1-9 qty), all models plug-in $14.95,
BNC $36.95 SMA $38.95, Type N $39.95

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F132 REV. C

Symbol	Duration	Length in UST's
one	100 us	1
zero	200 us	2
EOF	300 us	3
EOP	400 us	4

Table 1. Periods of four CEBus encoded symbols.

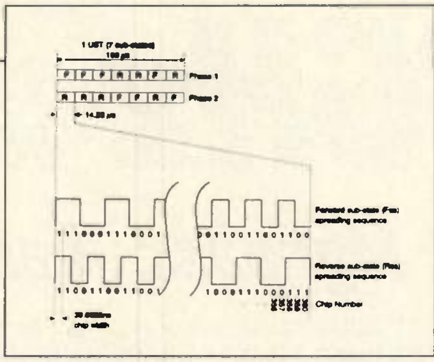

Figure 3. UST and sub-state timing.

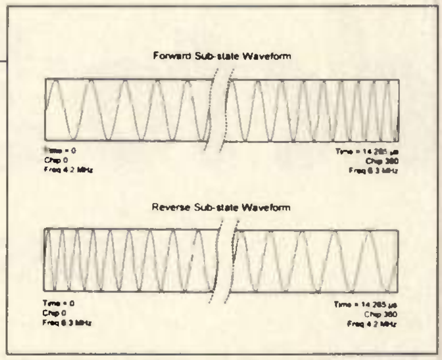

Figure 4. Forward and reverse sub-states.

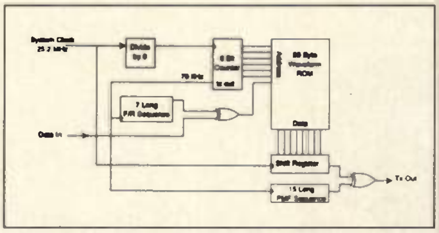

Figure 5. Transmit processor.

ing the packet body. The four CEBus encoded symbols have the periods shown in Table 1.

The channel rate is specified at 10,000 USTs per second. The variable length symbol encoding means that the effective bit rate is dependent on the message content.

For the RF media, symbols are encoded using two UST states, Phase 1 or Phase 2. Each state is divided into seven sub-states, Figure 3. The sub-states each consist of a specific carrier spreading sequence of 360 chips. Therefore, each UST contains 2520 chips used to spread the carrier. The basic chip rate is 25.200 MHz which yields a chip time of 39.6825 nS. Each resulting sub-state is 14.2857 mS duration.

There are two possible sub-states. The 360 chip sequence can be transmitted in either a forward or reverse direction. The complete UST is encoded using one of two complimentary maximal length coding sequences of forward (F) or reverse (R) sub-states. Phase 1 is F F F R R F R and Phase 2 is R R R F F R F. The 360 chip sequence was designed to distribute the transmit energy very evenly. The resulting waveform approximates a frequency sweep from 4.2 to 6.3 MHz, as shown in Figure 4. The sequence has good correlation characteristics and is tolerant of moderate frequency offsets.

An additional Phase Modulation Function (PMF) is used to further modify the sub-state sequence of each UST. The 15 bit PMF sequence is the maximal length code 1 0 1 0 1 1 0 0 1 0 0 0 1 1 1. Each bit of the PMF is XORed with the chips of a sub-state sequence causing all the bits of that sub-state to be inverted or normal. This results in 180 degree phase shifts which further improve the spectral energy distribution.

The combined effect of the main 360 chip sequence, the 7 chip forward-reverse sequence, and the 15 chip PMF sequence is a smooth spreading of the transmit energy over a 2.1 MHz bandwidth. The overall spectral distribution is quite rectangular. The energy peaks are approximately 3 dB above the average transmit energy. A transceiver employing these spreading techniques has been reviewed and certified by the FCC.

Figure 5 is a detailed block diagram of the signal generator. The 25.2 MHz system clock is divided and provides all of the timing for the generation process. A 6 bit counter is used to address a ROM which contains both of the 360 chip (bit) waveform sequences. The data input is used to pass or invert the output of the 70K forward/reverse sub-state sequence generator. This line is used to select which sequence will be clocked out of ROM by controlling the MSB. A shift register is used to convert the parallel output of the waveform ROM into a 25.2 MHz serial bit stream. This bit stream is passed or inverted by the 15 long PMF sequence to form the TX output. This simple state machine can be fabricated using any of the modern logic techniques.

The ASIC has a single logic level TX waveform output bit. The output waveform is over-sampled at a rate of 25.2 MHz. The majority of the output energy lies between 4.2 and 6.3 MHz. A simple L-C filter is adequate to clean up the ASIC output. This filtered output is referred to as the transmit baseband and is capable of driving 50 ohm loads directly.

Signal Correlation and Detection

The ASIC provides all of the correlation, synchronization, tracking, data decoding, and CRC functions required to recover the spread signal, see Figure 1. The baseband signal from the receiver front end is first bandpass filtered by a simple L-C filter and then amplitude limited. The limited signal is applied to a zero crossing detector which converts the signal into a serial digital stream. These bits are shifted into a shift register which forms the input of the correlator. The correlator provides detection of both the forward and reverse direction sub-state waveforms.

The correlation information is used to establish timing synchronization and maintain tracking of the sub-state symbols. Estimations of the sub-state symbol quality and value are determined and are passed to the unit symbol processing circuits. Here the unit symbol timing synchronization and tracking occur. The UST value is passed to the data decoder which determines if a ONE, ZERO, EOF, or EOP was transmitted. The CRC is verified at the end of the packet and the information is delivered to the micro-processor interface.

The digital signal processing which occurs within the ASIC permits high performance reception with simple receiver front ends. The spreading sequence and the correlator work together to permit relaxation of the transmitter and receiver local oscillator frequency accuracy requirements. The RX input to the ASIC can be off frequency by as much as plus or minus 350 kHz before performance begins to degrade. This means that low cost Surface Acoustic Wave (SAW) resonators can be used for both the transmit carrier oscillator and the receive local oscillator. The correlation is also phase insensitive and a moderate amount of phase jitter is not catastrophic.

The FCC Part 15 rules also specify that the receiver of a spread spectrum system be able to coherently re-assemble the spread information (3). The rules call for at least 10 dB of processing gain in direct sequence based systems. There are many different ways to define process gain (4). The FCC however, is interested in the ability of the receiver to reject interference. Therefore, the

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process gain is typically determined by measuring the jamming threshold with the spreading code enabled and disabled. These new signaling techniques are highly integrated and the system can not function with the spreading disabled. This makes verification of the process gain a more complex problem.

A transceiver employing SSC technology at 915 MHz was submitted for examination by the FCC. The Authorization and Evaluation Division of the FCC cooperated to arrive at a suitable measurement method. The system was able to demonstrate more than the required 10 dB of process gain. With this ground work completed, certification of future products should be straightforward.

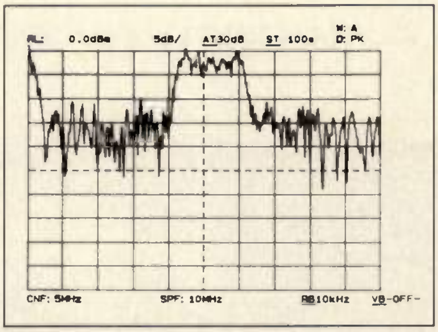

Figure 6. Baseband transmit spectrum before filtering.

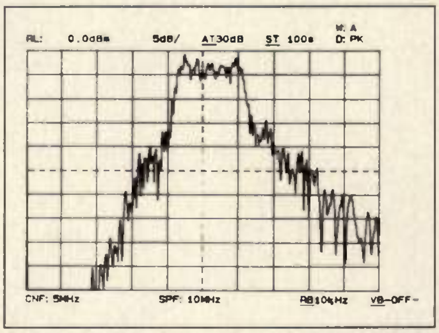

Figure 7. Baseband transmit spectrum after filtering.

Frequency Conversion for Transmission

The spread spectrum baseband signal is generated in the 4.2 to 6.3 MHz range. This low frequency baseband signal can be transmitted directly across various media such as AC power wiring or coaxial cable. The energy spreading, interference rejection, and quick network access features improve the performance of non-radio applications as well. To use SSC technology for radio links, conversion of the baseband signal to higher frequencies is essential.

The TX output bit of the ASIC is basically the output of a one bit over-sampled D to A converter. The sampling rate is equal to the system clock frequency of 25.2 million samples per second. The resulting output spectrum shown in Figure 6 contains a fair amount of energy outside of the desired 4.2 to 6.3 MHz frequency range. The ASIC output driver is capable of driving a 50 ohm load to 2.5 Volts peak-to-peak. A simple 50 ohm passive 5 pole bandpass filter, like that shown in Figure 7, can be used to clean up the unwanted energy. Low Q surface mount parts with 10 percent tolerance are sufficient. The resulting filtered baseband signal is ideal for directly driving a passive double balanced diode mixer.

The filtered baseband signal is typically applied to the IF port of the balanced mixer, Figure 8. The Local Oscillator (LO) port of the balanced mixer is driven with the desired carrier or center frequency. In this manner, the spread spectrum baseband signal PSK modulates the RF carrier. The resulting modulated signal, shown in Figure 9, contains energy sidebands located approximately 5 MHz below and above the carrier frequency. The LO signal feed through is attenuated by the balance of the mixer. Typically both sidebands are amplified by a linear power amplifier before transmission via the antenna.

Both of the sidebands contain similar copies of the spread spectrum information. Only one sideband signal is actually required to convey the message. The classic single sideband generation techniques of filtering or image rejection may be used to eliminate one of the two sidebands. The signal will then occupy one half of the original bandwidth. Since the single sideband signal contains only one frequency component at any instant in time, class C amplification may be used. The resulting efficiencies could offset the additional complexities incurred for SSB operation.

Down Conversion for Reception

To recover the spread spectrum message, the radio frequency signal must be down converted to the 4.2 to 6.3 MHz baseband range. In this system, it is not necessary to phase lock the receiver LO to the transmitter's carrier. The time typically required to lock onto carriers of differing frequency would add considerable delays to the network. Since the receive LO is not phase locked to the carrier of the incoming DSB signal, destructive cancellation of the sideband information will occur. For this reason, selection of one sideband and rejection of the other sideband is essential. Any of the classic single sideband down conversion techniques may be employed. The unwanted sideband need only be suppressed by 20 dB to avoid destructive cancellation of the information.

The requirement for single sideband down conversion carries some additional

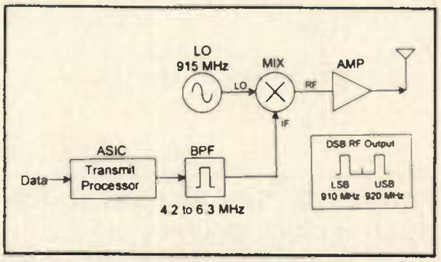

Figure 8. Transmitter block diagram.

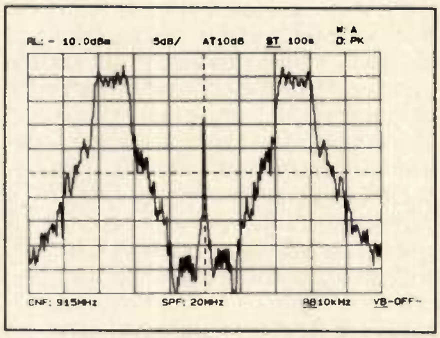

Figure 9. Typical modulated DSB spectrum.

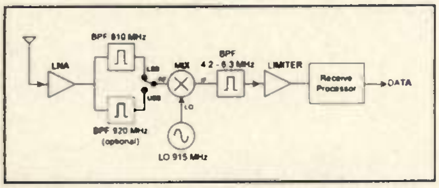

Figure 10. Single conversion receiver block diagram.

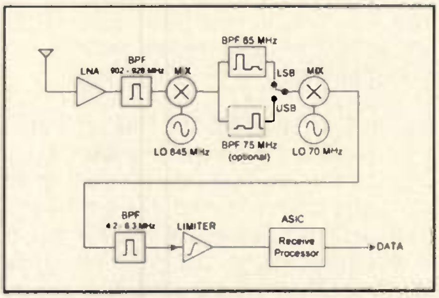

Figure 11. Double conversion receiver block diagram.

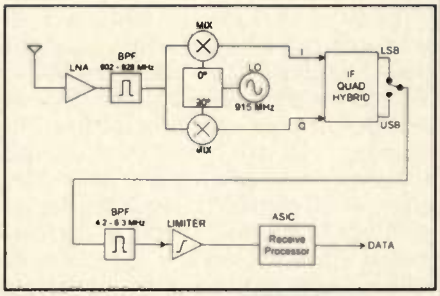

Figure 12. Image rejection receiver block diagram.

benefits. Once one of the sidebands has been selected and converted down to baseband, only a simple bandpass filter and a limiter are required to complete the receiver analog chain. Single sideband reception carries the additional system benefit of frequency diversity. The lowest cost transmitter implementations will transmit both sidebands which are separated in frequency by 10 MHz. This frequency separation provides relief from multi-path nulls and interference. A selectable sideband receiver can pick the sideband which yields the best performance at the time. The frequency offset of the sidebands from the carrier by 5.25 MHz provides a tremendous benefit in terms of RF link LO accuracy requirements. If the sidebands started right at the carrier (as they do in classical direct sequence PSK systems), a slight LO frequency offset would cause "folding" of the sideband energy and destructive cancellation of the message information. With the sideband energy spaced far from the carrier, moderate frequency offsets are not destructive.

There are several methods that can be used for single sideband down conversion of the RF signal. Probably the simplest implementation is the single conversion scheme in Figure 10. This approach relies on a low loss narrow band RF filter for sideband rejection. A low cost helical or SAW based filter can be employed to achieve sideband selection. The LO can be implemented with a simple SAW resonator oscillator circuit. The mixer may be active or passive and need not be balanced due to the wide separation in RF and baseband frequencies. A simple L-C bandpass filter passes the 4.2 to 6.3 MHz baseband signal and determines the receiver noise bandwidth.

The double conversion method places the sideband selection filter at a lower

Getting on the CEBus

The signaling techniques described in the article were developed by Intellon Corporation. The RF Modem ASIC is available from:

Intellon Corporation
5150 West Highway 40
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Tel: (904) 237-7416

CEBus Product Development Seminars are offered by:

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intermediate frequency, Figure 11. Simple L-C or helical bandpass filters fabricated for the 50 to 300 MHz range may be employed. The double conversion process can yield much better rejection of the unwanted sideband and other spurious signals. The additional complexity may be acceptable for high performance receivers.

Image rejection down conversion is another possible method of sideband selection, as shown in Figure 12. This technique employs two mixers driven by a quadrature LO at the center frequency. The resulting I and Q baseband signals are combined in a low frequency quadrature hybrid circuit. Either sideband may be selected by switching the output of the baseband hybrid. The LO hybrid could be a printed micro-strip structure or a surface mount pre-packaged phase shift network. The baseband hybrid can be made using small torroid cores or active phase shift circuits. The unwanted sideband can be rejected by the required 20 dB using low cost components. The amplitude and phase balance requirements on the I and Q path set the limit of sideband rejection at about 30 dB.

All of the receiver implementations require a high gain limiter stage to provide the majority of the system gain. The baseband frequency range of 4.2 to 6.3 MHz is easily accommodated by many of the available IF amplifier/limiter ICs. Gain of 90 dB is typical and is sufficient for a high performance receiver. By placing most of the gain at low frequencies, receiver parts' cost can be kept very low. The limiter is able to respond to rapid variations in signal levels more easily than AGC based designs. The limiter stage provides a low impedance output of 100 mV peak-to-peak to the ASIC RX input.

As discussed, the suppression of the unwanted sideband should be at least 20 dB to prevent destructive cancellation of the message information. This will yield good performance in low cost implementations. A strong interfering signal located in the area of the unused sideband can cause performance degradation. Typically, this will be a Part 15 device with moderate transmit power and the process gain of the system will overcome the interference. System performance can be improved by better rejection of the unwanted sideband.

RF Modem ASIC

All of the functions required to implement this new signaling technique have been integrated into a single low cost ASIC shown in Figure 1. The ASIC is housed in a 28 pin small outline surface mount package. A single power supply of +5.0 V and 25 mA is required. The ASIC has a total of 11 lines for interface to the host microprocessor. The ASIC provides all of the timing requirements specified by the EIA CEBus physical Layer Specification. The host controller is not responsible for any physical layer timing. During transmit the only responsibility of the host is to provide the ASIC with encoded data 8 USTs at a time, and to monitor the status indicators provided. In receive mode, the ASIC demodulates the incoming signal and presents the recovered data to the host, 8 USTs at a time. During receive, the host must decode the incoming USTs and monitor the status indicators provided.

Conclusion

A new signaling method which has been optimized for operation under the FCC Part 15 direct sequence spread spectrum rules has been developed. A complete baseband modem which fully implements the CEBus RF Physical Layer has been designed in the form of a single low cost ASIC. The relaxed requirements for transmitter and receiver designs permit low cost implementations. The high level of integration achieved enables the design of compact RF data products with robust performance. ***RF***

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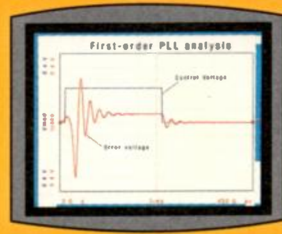

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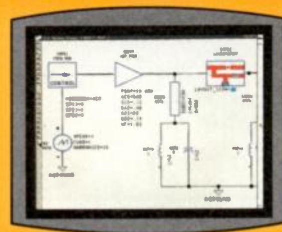

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Frequency Translation of a Baseband Signal

By Eric A. Adler, Edward A. Viveiros and John T. Clark
Army Research Laboratory

In communications and radar systems, baseband signals are often upconverted to higher frequencies for transmission where typically only the upper or lower sideband is transmitted. Baseband signals described in this paper are broadband and do not extend down to DC. Frequency translation of baseband signals results in the local oscillator (LO) leakage signal and an undesired sideband very close in frequency to the desired sideband. The frequency separation between the desired sideband and the LO leakage signal at the intermediate frequency (IF) output of the mixer is equal to the frequency separation between the lower extent of the radio frequency (RF) mixer input signal and DC. The undesired sideband will be twice that distance from the desired sideband. This creates a problem in filtering the desired sideband from the other sideband and LO leakage.

This problem is further complicated when the upconverted signal is required to have a flat amplitude and linear phase response over a wide bandwidth. This is the case for some radar systems, which use linear frequency modulation (FM) (or chirp) waveforms. Any distortion to the amplitude or phase response will degrade the system performance by spreading the main lobe (decreasing resolution) or generating undesired sidelobes (false targets) in the range response.

This paper describes a technique for obtaining a single IF with minimum amplitude and phase distortion and without the other sideband or LO leakage. The system described will translate a baseband chirp signal, 15 to 45 MHz bandwidth, 100 ms period, and 50 percent duty cycle to an IF centered at 180 MHz.

System Design

The most common problem in frequency translation is the generation of intermodulation products (spurious signals) which fall within or just outside the desired IF band. Multiple frequency conversion schemes, in which an IF is generated, are often used to achieve better spurious signal levels than can be obtained in single conversion architectures. The IF products out of a mixer include the upper and lower sidebands (LO ±RF), the intermodulation products, and leakage from the RF and LO signals.

The RF input power to a double-balanced mixer is usually set at a low level to minimize the spurious products which result from the mixing process. The two-tone, third-order intermodulation products are of particular concern since they often fall within the desired IF passband. This two-tone intermodulation problem primarily occurs in receiving systems with multiple simultaneous input signals, but can also occur in an exciter design when using waveforms that have close-in spurious signals, such as direct digital synthesizers. This problem can be controlled by setting the maximum RF input

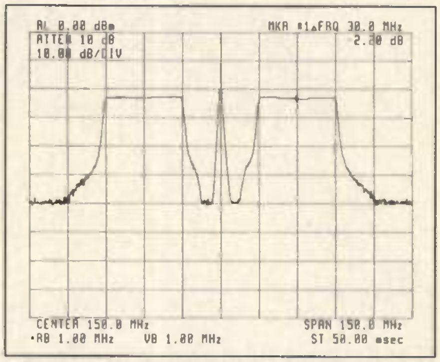

Figure 1. IF output of standard double balanced mixer.

power to the mixer at a level to achieve a required spurious-free dynamic range (SFDR). A rule of thumb is to set the RF power 20 dB below the input 1 dB compression point to achieve 60 dB SFDR. In a typical mixer, this results in the upper and lower sidebands having approximately the same power level as

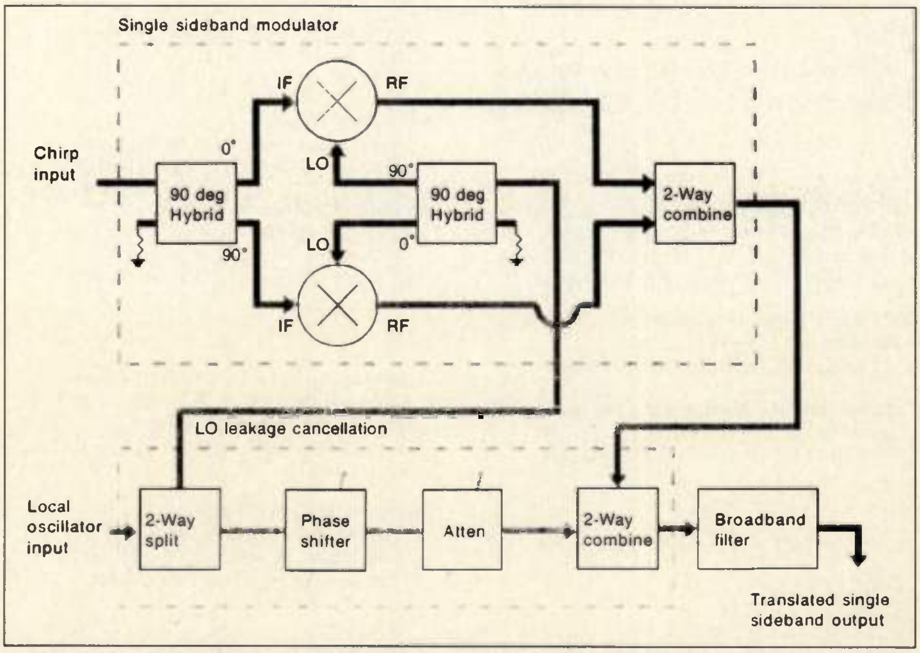

Figure 2. Block diagram of SSB upconverter including single sideband modulator, LO cancellation circuit, and broadband filtering.

the LO leakage signal. SFDR in this context relates the power level of spurious signals within the desired IF band relative to the IF signal power.

The selection of the intermediate frequency is important since filtering becomes more difficult as the baseband signal is converted to a higher IF. This occurs because the desired IF signal will have the same frequency separation from the LO leakage and undesired sideband. Designing a filter with a wide, flat passband, linear phase response and sharp rolloff to sufficiently attenuate the undesired, close-in signals is an extremely difficult task and will be addressed later.

Single Sideband Modulation

Translating the 15 to 45 MHz chirp signal to an IF centered at 180 MHz with a standard double-balanced mixer produces an IF spectrum, as shown in Figure 1. The input power of the chirp was selected at –15 dBm for SFDR considerations. The IF signal from 165 to 195 MHz is the desired sideband. Note that the power levels of the two sidebands are equal and the 150 MHz LO leakage signal is also approximately the same amplitude.

Since the undesired sideband is only 30 MHz away from the desired sideband, it is difficult to design a filter to effectively remove it (>60 dB suppression) without distorting the phase or amplitude response of the chirp. The undesired sideband can be attenuated by using a single sideband (SSB) modulator (1) instead of the double-balanced mixer. An SSB modulator with LO cancellation is illustrated in Figure 2.

In the SSB modulator, the LO signal is split between two mixers and the RF is split with a quadrature hybrid, which shifts one of the outputs by 90 degrees. The mixer outputs are combined through another quadrature hybrid, which shifts one of the IF's by an additional 90 degrees for a total phase difference of 180 degrees. The input and isolation ports of the second quadrature hybrid are used as output ports of the SSB modulator. One of these ports provides the upper sideband as the output. The port that is not used should be terminated in a matched load. The output spectrum of the SSB modulator is shown in Figure 3. This plot shows the undesired IF signal is suppressed by 20 dB. The LO leakage signal is suppressed by approximately 25 dB, while the desired IF signal is undistorted.

LO Leakage Cancellation

The LO leakage, which is still present after SSB modulation, can be further attenuated by inserting a notch filter after the mixer. Designing a notch filter with a narrow rejection bandwidth, a wide passband (30 MHz), and minimal phase and amplitude distortion within the passband is extremely difficult. The LO leakage signal can be reduced by injecting a sample of itself 180 degrees out of phase and equal amplitude into a power combiner, as shown in Figure 2. The vector addition of these two signals results in cancellation of the LO signal. The degree of cancellation is strongly a function of the amplitude and phase of the LO leakage signal and the injected sample. The LO leakage signal is sampled at the LO input of the SSB modulator and passed through a variable attenuator to match the amplitude of the LO leakage signal at the output of the SSB modulator. It is then passed through a variable phase shifter, which is used to tune the sampled signal to 180 degrees out of phase with the leakage signal. The two signals are then summed in a standard two-way combiner. The amount of LO leakage cancellation attained varies, depending on how well the amplitude is matched and how accurately the phase is tuned to 180 degree difference. For our example, the LO leakage is well below the spectrum analyzer noise floor, as illustrated in Figure 4. Further analysis showed the LO cancellation is on the order of 40 dB. The LO cancellation circuit also resulted in an additional 3 dB insertion loss in the desired IF.

Linear Phase Filtering

Additional filtering is required to remove all out-of-band intermodulation products and to further attenuate the undesired IF sideband after the single sideband modulation and LO leakage cancellation circuitry. Phase linearity, amplitude flatness, and stopband attenuation are all critical and somewhat conflicting parameters in selecting a filter. Filters with linear phase response typically exhibit poor attenuation rolloff and passband amplitude response (Bessel and Gaussian), while those filters with good rolloff and passband amplitude response have poor phase response within the filter passband (Butterworth and Chebyshev)(2). Additional poles could be added to increase the stopband attenuation but would result in increased phase distortion. The phase response of filters becomes nonlinear in the transition region from passband to stopband and can be seen as a deviation from flatness in the group delay response of the filter since the group delay is the derivative of the phase response.

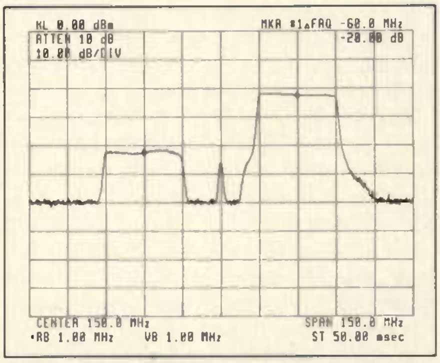

Figure 3. IF output of SSB modulator.

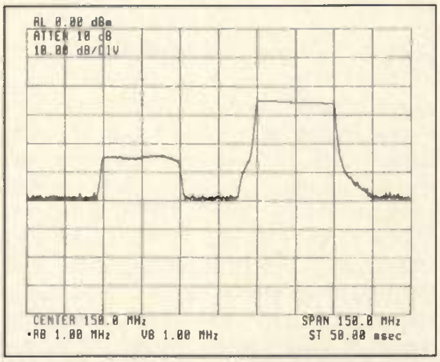

Figure 4. IF output of SSB modulator with LO cancellation.

In our system, the LO leakage will be 15 MHz away and the other sideband will be 30 MHz away, regardless of the choice of IF. The selectivity of bandpass filters is related to the ratio of its 3 dB bandwidth to its center frequency. As the ratio decreases, so does the selectivity. Since the bandwidth of the baseband signal is fixed, a higher IF (and center frequency) results in decreased filter selectivity.

Linear phase filtering is an active subject in filter design but is outside the scope of this paper. A simple approach to linear phase filtering is to use a filter

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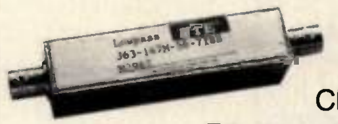

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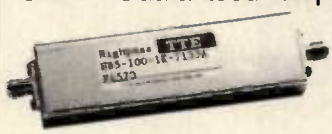

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DL-30	32MHz	35MHz	61MHz	DK-40-5P	40.0MHz	5%F_o
DL-50	48MHz	55MHz	90MHz	DK-50-5P	50.0MHz	5%F_o
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DL-90	81MHz	90MHz	157MHz	DK-70-5P	70.0MHz	5%F_o
DL-100	98MHz	108MHz	189MHz	DK-100-5P	100.0MHz	5%F_o
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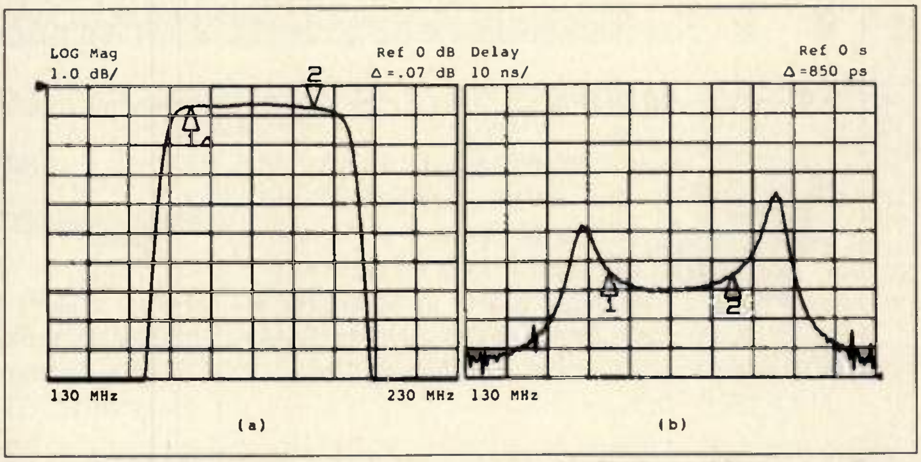

Figure 5. Bandpass filter response a) log magnitude and b) group delay.

with a wider 3 dB bandwidth than is required and minimum passband ripple. The worst amplitude and phase deviations for a filter with good rolloff occur at the 3 dB points of the particular filter. By operating only in a smaller portion of the filter bandwidth, amplitude flatness and phase linearity can be achieved. For the desired 30 MHz signal, a filter with a wide bandwidth, 0.1 dB ripple, centered at an appropriate frequency, is used as shown in Figure 5a. The group delay deviation is shown in Figure 5b. The desired IF band can be placed within the filter bandwidth, away from the nonlinear transition regions, as the markers denote. This filter attenuates the undesired IF band and other intermodulation products. The output of the SSB modulator and bandpass filter is shown in Figure 6. Note the undesired sideband has also been attenuated below the spectrum analyzer noise floor.

Conclusion

A combination of techniques have been used to translate a 30 MHz bandwidth chirp to a higher IF. SSB modulation was used to isolate the desired IF sideband from both the undesired sideband and LO leakage. The LO leakage is further attenuated by injecting a sample of the LO, from the input of the SSB modulator, which is matched in amplitude and 180 degrees out of phase with the LO leakage into a summing network. Finally, an IF filter with bandwidth wider than is required was used to preserve amplitude flatness and phase linearity of the chirp, as well as to attenuate the other sideband, LO leakage, and all intermodulation products outside the desired IF band. ***RF***

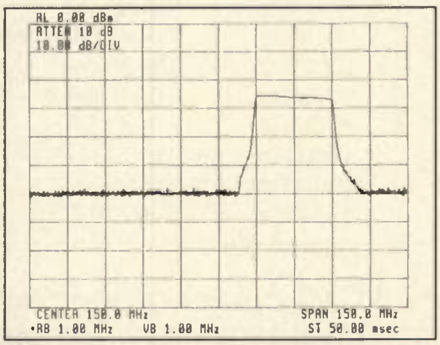

Figure 6. IF output of SSB modulator with LO cancellation and filter.

Acknowledgement

The authors would like to thank Phuong Phu of the Army Research Laboratory for her assistance in assembling and testing the hardware.

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About the Authors

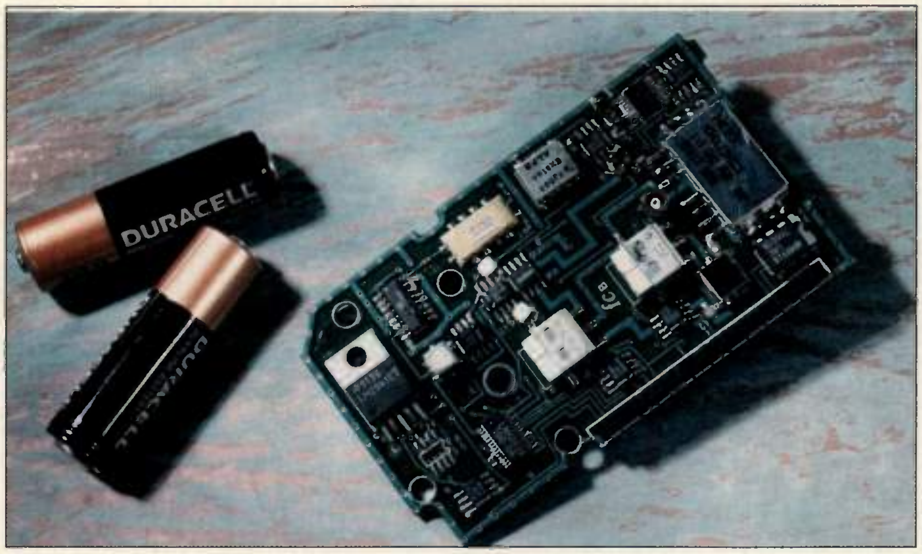
Eric Adler is a senior design engineer currently working on a multi-mode radar receiver/exciter for SAR/MTI applications. His past experience includes narrowband frequency excision techniques for wideband COMINT systems. Edward Viveiros is currently a senior design engineer. He has worked on several radar (MTI and SAR) and COMINT systems in receiver, exciter, frequency synthesizer, and system design. John Clark is a design engineer, working on MTI and SAR systems.

This work is part of the wideband receiver/exciter design for the moving target technology program. All authors work in the Signal Processing Systems Branch of the Army Research Laboratory, 2800 Powder Mill Rd., Adelphi, MD 20783-1197. Tel: (301) 394-4160.

Radios for the Future: Designing for DECT

By Benny Madsen and Daniel E. Fague
National Semiconductor

Nations worldwide are implementing standards that will transform today's conventional land line phone system to an untethered network of services. A first step in achieving this goal is the implementation of digital telephony standards for both cellular and cordless telephones. One of the first digital cordless standards to be implemented is the Digital European Cordless Telecommunications (DECT) standard, a pan-European standard designed to connect all of Europe with a common digital cordless system. This article introduces a new 2 GHz radio architecture which implements the Digital European Cordless Telecommunications (DECT) standard.



The DECT standard presents the radio designer with a number of challenges. For example, a DECT phone must provide all the capabilities of a standard cellular phone at roughly twice the operating frequency, and satisfy user demand for portability: small size, lightweight, long talk and standby times. These requirements make a strong case for integrated RF solutions. Table 1 shows a summary of the key specifications for DECT as well as other digital cordless standards.

The DECT system is based on Time Division Duplex, Time Division Multiple Access (TDD/TDMA). This technique of duplexing uses separate timeslots on a single carrier for transmitting and receiving signals. Figure 1 shows the structure of the DECT TDD/TDMA frame. The complete frame is 10 ms in duration, with a 5 ms transmit and a 5 ms receive sub-frame. Each sub-frame is divided into 12 timeslots of 480 bit times in duration (416.67 ms). When a communication link is made, a transmit and receive timeslot is assigned to the users. Normally the same timeslot number (i.e., T3 and R3) is used for both transmit and receive. Multiple users are accommodated by assigning different timeslots to different users. This method requires the transmitter to be turned off while receiving and the receiver to be turned off while transmitting. This method of duplexing places strict requirements on the Phase Locked Loop (PLL). It must switch fast and have low spurious noise.

Applicability to the U.S.

In the United States, the FCC has proposed allocating 20 MHz within the Emerging Technologies (ET) band for "non-licensed" Personal Communications Systems (PCS), with 10 MHz allocated for voice type (isochronous) services. The U.S. can choose to modify a standard already developed, like DECT with a frequency almost identical to that proposed by the FCC, or develop its own. The following explains how DECT could be modified to meet the proposed U.S. requirements.

Modulation method — DECT employs a frequency modulation scheme, called Gaussian-shaped frequency shift keying, which could be used in the PCS ET band. Frequency modulation schemes have a number of advantages, including the use of more efficient power amplifiers to lower power consumption. In addition, FM techniques allow the use of limiter/frequency discriminator structures for demodulation. These structures are simple to use and build, as opposed to fully linear receivers requiring sophisticated gain control.

Channel bandwidth vs. data bit rate — The DECT bit rate (1.152 Mb/s) and channel spacing (1.728 MHz) limits spectral efficiency to about 0.67 bits/sec/Hz. As a single parameter, the spectral efficiency is not sufficient for determining the capacity in systems like DECT or any other small cell system. System capacity is a function of cell size (transmitted power), frequency reuse distance, available frequency range, and spectral efficiency. The frequency reuse distance is determined by the receivers' ability to cope with interference from unwanted transmitters; interference on the same channel (co-channel) or adjacent channels. A receiver's ability to cope with same channel and adjacent channel interference is determined by the chosen modulation scheme, as well

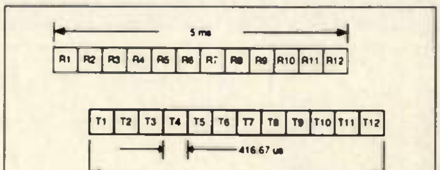


Figure 1 - TDD/TDMA frame structure for DECT.

Parameter	DECT	CT-2	USER PCS	PHP
Origin	ETSI	UK	US	Japan
Access	TDMA & FDMA/TDD	FDMA/TDD		TDMA/TDD
Modulation	GFSK	GFSK		π/4-DQPSK
Baseband Filter	$B_bT = 0.5$ (Gaussian)	$B_bT = 0.5$ (Gaussian)		$\alpha = 0.5$ (Root Nyquist)
Data Rate	1.152 Mb/s	72 kb/s	≈1.0 Mb/s	384 kb/s
FM Deviation	288 kHz	14.4-25.2 kHz		
RF Channels	0: 1897.344 MHz 9: 1881.792 MHz	1: 864.15 MHz 40: 868.05 MHz	1910-1930 MHz	1895-1911 MHz
No. of RF Channels	10	40	4	52
Channel Spacing	1.728 MHz	10 kHz	1.25 MHz	300 kHz
Synthesizer Switching Speed	30 us (BS) 450 us (HS)	1 ms (ch.-ch.) 2 ms		30 us (BS) 1.5 ms (HS)
Frequency Accuracy	50 kHz	10 kHz		3 ppm
Speech ch./RF ch. (Full/Half Rate)	12/24	1/1		4/8
Speech CODEC	32 kb/s ADPCM	32 kb/s ADPCM		32 kb/s ADPCM
Frame Length	10 ms (12 Tx + 12 Rx)	2 ms (1 Tx + 1 Rx)		5 ms (4 Tx + 4 Rx)
Peak Power	250 mW	10 mW		100 mW

Table 1 - A comparison of digital cordless telephony standards

as receiver structure and design.

The narrower 10 MHz frequency band proposed for U.S. isochronous services will require a higher spectral efficiency than DECT. This can be obtained by narrowing the (Gaussian) modulation filter or by using other modulation schemes. Narrowing the filter will limit receiver sensitivity, while other modulation schemes may sacrifice the simple transceiver structure.

Radio Design Issues

Single conversion receiver, direct modulation transmitter — Digital modulation can be done in a couple of ways. Quadrature (I and Q) modulation is highly accurate, and is the method of choice for the digital cellular standards. A quadrature down conversion provides maximum flexibility in the baseband section for any type of demodulation. However, the added circuits that accompany these types of demodulators are costly. Normally, a Digital Signal Processor (DSP) must be used in the backend, which increases current consumption, and some automatic gain control circuits must be used to ensure that the baseband A/D converter dynamic range is fully utilized. In some systems, a carrier recovery loop such as a Costas Loop must be implemented to recover the carrier's phase.

The TDD/TDMA nature of DECT allows for two simplifications that drastically reduce the radio's complexity. First, the transmitter is on for short bursts of 424 bits, or about 380 ms. This, and a moderate frequency drift tolerance, allow for an open loop, direct modulation of the voltage controlled oscillator (VCO). Second, the receiver is only on for the same short length of time. The argument can be made that in such a short burst of data, a carrier recovery loop is not required, because the phase difference of the local oscillator (LO) will not deviate much in such a short time. However, the coherent solution will suffer from the frequency drift allowed in DECT. The large frequency drift makes an unlocked LO unacceptable. But, the drift is not too large that a discriminator can't be used. In fact, the drift allowed is about 10 percent or less of the discriminator's bandwidth, making the discriminator a good choice for DECT.

Traditionally, 2 GHz radio designs were limited to at least two down conversions and one up conversion. Figure 2 shows a typical receiver. The second down conversion is necessary because limiting and demodulating at the high IF is difficult and modulating a low frequency VCO is more practical.

For DECT, National Semiconductor chose a single conversion receiver, direct modulation transmitter architecture. This is shown in Figure 3. There are several key features that make this architecture possible. The signal is received at the antenna and filtered with a low loss (typ. 1 dB) antenna filter. The filter is used on the receive side to reduce possible out-of-band interfering signals. On the transmit side the filter is used to attenuate harmonics of the transmitted signal and to reduce possible wide band noise.

The signal then passes the duplexer which can either be a fast switch or circulator. The insertion loss in the duplexer should not exceed 1 dB. From the duplexer the signal enters a low noise amplifier, image reject filter and mixer structure. National Semiconductor provides various IC solutions for receiver front ends, which combine the low noise amplifier (LNA) and mixer circuit. One of them is the LMX2216 shown in Figure 3.

Before the DECT standard was finalized, a group of component and equipment manufacturers had worked out standards for key DECT components, within the framework of a group called ECTEL (European Community Telecommunications). A standard intermediate frequency of 110.592 MHz was chosen by this group. Consequently, a number of manufacturers provide highly selective Surface Acoustic Wave filters for this frequency with low insertion loss (typ. 3 dB) and reasonable group delay characteristics. The adjacent channel selectivity of most of the filters permits FM demodulation on the output of the filter.

Two ways to do FM demodulation are: (1) the limiter/discriminator structure as shown in Figure 3; and (2) PLL tracking of the instantaneous frequency deviation. By first limiting the signal in a high gain amplifier, then mixing the signal

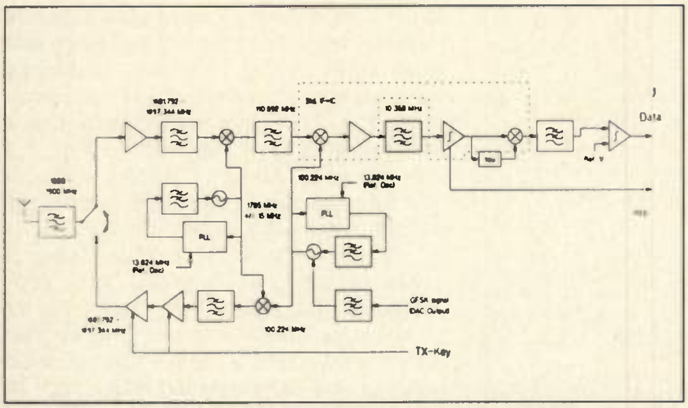


Figure 2 - A block diagram of a typical 2 GHz radio.

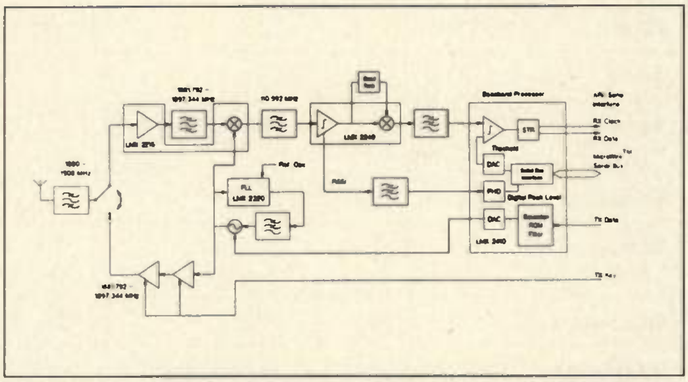


Figure 3 - Block diagram of National Semiconductor's single conversion receiver architecture.

with a 90-degree phase shifted replica of itself, the instantaneous frequency deviation is derived. The quadrature tank shifts the phase of the signal 90 degrees, but only at the center frequency. The important parameter of the phase-shifter is a large frequency dependence of the phase-shift. Normally a linear phase-shift dependency of the frequency is desirable. Thus a steep linear phase characteristic is desired. The steepness, though, is determined by the Q of the quad tank, which limits the design possibilities.

For demodulating FM signals at relatively high frequencies a phase locked loop is a common solution. The PLL tracks the instantaneous phase and thereby the instantaneous frequency deviation of the received signal. Such solutions are commonly used in satellite TV demodulators. Some of the drawbacks of PLL FM demodulator solutions are less immunity to interference (such as co-channel interference), a significantly higher level of power consumption and increased component cost.

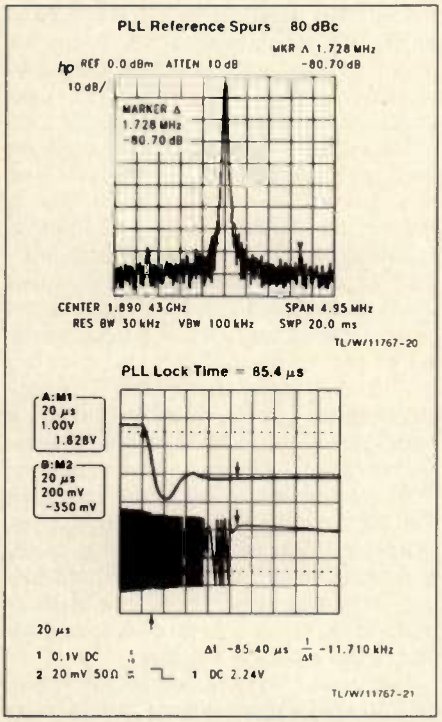


Figure 4 - Lock time and spurious noise performance of National's PLL for the DECT transceiver.

As the demodulated signal leaves the LMX2240, a lowpass filter removes remnants of IF frequency and its harmonics, and improves the signal-to-noise ratio before the signal enters the data comparator. Depending on the signal level at the input of the comparator, the comparator can become crucial to single conversion design. National Semiconductor therefore integrated a high performance data comparator into the LMX2410 baseband processor. The output of the comparator is a binary signal, corresponding to the transmitted binary signal. To derive timing information, the signal is processed by a symbol timing recovery circuit. The symbol timing can either be based on continuously phase locking to the edges of the signal, or burst wise correlating with the known burst signal. The LMX2410 uses the digital PLL method, but simulations show that either of the methods are usable and yield about the same performance.

On the transmit side, the baseband processor includes the Gaussian filter needed to pulse shape the incoming serial binary transmit signal. The baseband processor internally regenerates the transmit clock, eliminating the need for an extra transmit clock line from the digital back-end. The Gaussian filter is implemented by ROM table lookup instead of a conventional FIR or IIR filter. This ROM table and the filter can be mask programmed to meet other filter requirements, e.g. to support U.S. standards for the PCS and ISM bands. From the filter, the signal is fed to a DAC, then directly to the frequency synthesizer VCO, so the signal directly modulates the VCO. This principle is commonly used in analog cellular and cordless telephones. In order not to introduce distortion, the filter bandwidth of the loop must be significantly lower than the lowest frequency components of the modulating signal. But a low loop frequency leads to a slow switching speed, which is undesirable in a DECT transceiver. In order to modulate the VCO directly, the phase locking needs to be disabled, corresponding to opening the loop while modulating. The principle is simple, but puts requirements on the components used. There are two basic needs. The loop must be opened quickly, without causing a sudden jump in the VCO frequency, and the VCO frequency must be stable during a data burst. The National Semiconductor PLL, LMX2320, in combination with an active loop filter, accomplishes this.

The synthesizer needs to span the DECT frequency band both transmitting and receiving. A wide VCO tuning range and a fast switching speed — in the order of 400 us (max.) for a frequency jump of 130 MHz — are needed. The modulated signal from the VCO is fed to the power amplifier. The power amplifier delivers a maximum of 250 mW peak, plus whatever loss is introduced between the PA and the antenna (typically in the order of a couple of dB). The signal is finally fed to the antenna through the duplexer and antenna filter.

Supply voltage and power consumption are also key parameters in the design of handheld and pocket equipment. Today, cordless telephones use as few as three rechargeable battery cells, either Nickel Cadmium (NiCD) or Nickel Metal Hydride (NiMH). To avoid costly voltage conversion circuits, all parts of the cordless telephones need to operate at 3 volts. The DECT chipset from National Semiconductor will operate down to 2.85 V. Several manufacturers offer 3 volt wideband VCOs with a control voltage range of 0.5 to 2.5 volts, as well as RF power amplifiers in GaAs technology, operating from a 3 volt supply. GaAs, however, requires a negative bias supply. Power amplifiers can also be built using discrete bipolar transistors. These run from a typical supply voltage of 3.6 volts, with a lower power efficiency than GaAs modules. Both solutions can be used and the choice between them depends on cost, power consumption and size requirements.

Phase Locked Loop Issues

Lock time and low spurious noise — A single conversion radio architecture requires fast synthesizer switching speed in order to transmit and receive on as many as 24 timeslots per frame. In addition, in a single conversion transceiver, the synthesizer needs to make a large jump in frequency between transmitting and receiving, typically in the order of 110 MHz.

Lock time is defined as the time it takes a PLL to switch from one frequency to another within a given tolerance. Spurious noise, also called reference spurs, is defined as the level of the reference sidetones in relation to the desired tone. Together, these two parameters form a major tradeoff in radio design. Faster lock time almost certainly means higher spurs and vice versa.

Besides designing a proper loop filter, another key to obtaining simultaneous fast switching and low spurs is a well balanced charge pump circuit on the phase detector output. This charge pump circuit must have excellent balance to reduce or eliminate reference frequency spurs, and its dead band region must be at a minimum, even at high reference frequencies, to ensure good noise performance and fast switching speed. The LMX2320 was designed to meet these requirements. For a DECT transceiver, the PLL must also have a wide tuning bandwidth at a high reference frequency. This requires a 64/65 prescaler to achieve legal divide ratios. The prescaler and PLL must be kept as low current as possible to help preserve battery life. An example of a typical DECT PLL performance is shown in Figure 4. The figure shows settling time and reference spurs for a simple passive loop filter.

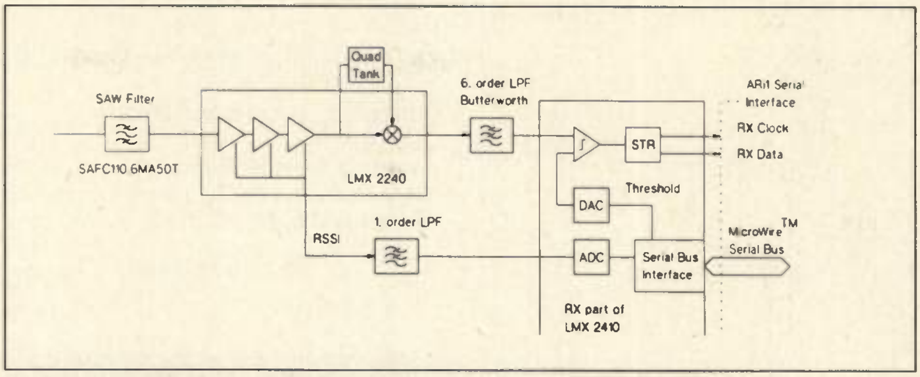


Figure 5 - Simplified block schematic for a DECT transceiver.

Open loop modulation — Another key element of this architecture is open loop modulation. In this mode, the PLL loop is opened while modulating the VCO, allowing the VCO to run free. The loop filter voltage is preserved by putting the PLL's charge pump circuit into a high impedance state. An active loop filter can be chosen to counteract possible loop discharge due to leakage current from the VCO tuning varactor diode. Alternatively, the PLL chip can be powered down, which also brings the charge pump circuit into a high impedance state. In this case, a buffer may be needed to limit an unintended jump in frequency due to possible load pulling of the VCO.

The modulating signal is added to the loop voltage at the input of the VCO, and the result is a modulated carrier. The modulating signal must have a stable midband voltage before opening the loop, in order to avoid a frequency offset during modulation. Such a frequency offset will occur if the loop is stabilized when the modulation signal is either at the negative or positive peak voltage. In these cases, as the loop is opened and the modulation commences, the resulting center frequency will be a sum of the intended center frequency set by the PLL, and an unintended frequency offset equal to plus or minus half of the peak to peak frequency deviation of the modulated signal.

The LMX2410 Baseband Processor is designed to provide an accurate midband voltage. When the loop is opened and the modulation starts, the deviation correctly swings above and below the intended center frequency. The drift in the carrier is kept to a minimum because the burst in DECT is very short. The DECT standard specifies a frequency drift of less than 13 kHz/ms. A DECT PLL based on the LMX2320 can be designed to have orders of magnitude less frequency drift than allowed by the DECT standard.

Receiver Parameters

Most of the requirements for the receiver can be derived from type approval specifications. A noise figure, gain and linearity budget can be made for the receiver using data from available components.

Receiver sensitivity — Receiver sensitivity is normally defined as the signal level needed to produce a certain signal to noise ratio, or, for digital systems, the level needed to produce a certain Bit Error Rate (BER) at the output of the receiver. In the case of DECT, the BER at the sensitivity level is 10^{-3}, at nominal temperature and supply voltage and for frequency offsets of –50 kHz, 0 kHz, +50 kHz of the received signal. The required sensitivity for all DECT equipment is –83 dBm. For equipment meant for public access use, like the U.K.'s Telepoint, the requirement is –86 dBm.

In a receiver for digital modulation, the sensitivity is composed of two figures of merit: the receiver front-end sensitivity, normally expressed by the noise figure, and the demodulator performance. The latter depends on the demodulator chosen. Circuit simulations can be used to get an initial idea of the demodulator performance. Figure 5 shows a simplified block schematic for a possible DECT receiver. The IF channel selection filter is a Murata 110.592 MHz SAW filter SAFC110.6MA50T; the limiter-discriminator circuit is a National Semiconductor LMX2240; the lowpass filter is a discrete filter; finally, the baseband processing function (i.e. bit slicing, symbol timing recovery and bit slice threshold setting circuit) is contained in a National Semiconductor LMX2410.

The BER performance for this specific

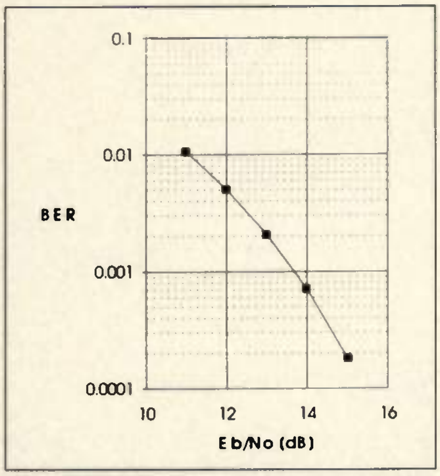


Figure 6 - Simulated BER for the DECT receiver.

circuit is shown in Figure 6. The implication of frequency offset is roughly indicated in Figure 7. As shown by these figures, an E_b/N_o ratio of approximately 13.6 dB is needed for a BER of 10^{-3} and needs to be increased by approximately 1 dB to compensate for a frequency offset of 50 kHz. The receiver (demodulator) could compensate for the frequency offset/error using a DAC to set the threshold level for the bit slicer (data comparator), assuming that the discriminator is sufficiently wideband. The latter will be the case in a DECT receiver, since offsets are on the order of 50 kHz, which is only a fraction of the required discriminator bandwidth.

The following equations assume ideal compensation. Not compensating for frequency offsets increases the requirements in receiver sensitivity by approximately 1 dB. The required noise figure for the receiver front-end can be derived by knowing the demodulator's E_b/N_o performance and assuming conducted measurements, as in the case of DECT. The E_b/N_o ratio equals the signal to noise ratio on the output of a filter with a noise bandwidth equal to the bit rate, r_b:

$$\frac{E_b}{N_o} = \frac{S}{r_b N_o} \qquad (1)$$

Since

$$N_o = kT_oF_n \qquad (2)$$

where k=1.38054 x 10^{-23} K/J, the Bolzmann constant; T_o=290K, the standard temperature; and F_n is the noise factor. Expressing the ratio E_b/N_o in decibels (dB), we get:

$$\left(\frac{E_b}{N_o}\right)dB = S - F_{dB} - 10\log_{10}(r_b kT_o) \qquad (3)$$

$$F_{dB} = S_{dBm} - \left(\frac{E_b}{N_o}\right)dB + 113.4 \qquad (4)$$

where S_{dBm} is the signal in dBm

Inserting the specified sensitivity for all DECT equipment, S_{dBm} = –83 dBm, and the needed E_b/N_o ≥13.6 dB, yields

$$\begin{aligned} F_{dB} &= -83 - 13.6 + 113.4 \text{ [dB]} \\ &= 16.8 \text{ [dB]} \end{aligned} \qquad (5)$$

The requirement is therefore a maximum noise figure of 16.8 dB for all DECT equipment, and similarly NF = 13.8 dB for equipment meant for Public Access.

Receiver linearity — Non-linearities in a receiver, such as gain compression, will cause signals from several transmitters to mix with each other in the receiver. The result is intermodulation products. In the worst case, these intermodulation products can end up having the same frequency as the desired RF signal. The intermodulation products can either be products from the desired signal or unwanted signals. Thus, the receiver intermodulation performance is a measure of the receiver's ability to avoid being disturbed by other DECT-like signals on other DECT channels. An ideal receiver is completely linear and would not suffer from intermodulation.

Receiver design is a trade-off between sensitivity, linearity, power consumption and component cost. Receiver linearity is normally expressed as the receiver's third-order Input Intercept Point, (IIP_3). Requirements for IIP_3 can be derived from the type approval measurement methods and specifications.

The receiver intermodulation performance is measured using three signals: a desired DECT signal and two undesired signals. The three signals are on three different DECT RF channels. The frequencies of the undesired signals are chosen such that one of their third-order intermodulation products becomes equal in frequency to the desired signal. Of the undesired signals, one is modulated with a DECT-like signal and the other is not. The intermodulation product of interest is the one that carries DECT modulation with nominal frequency deviation. The frequencies of the two undesired signals are chosen such that the intermodulation product of interest appears at the receive frequency. Hence the intermodulation product will appear as an undesired DECT interferer with the nominal frequency deviation and on the same channel as the desired signal.

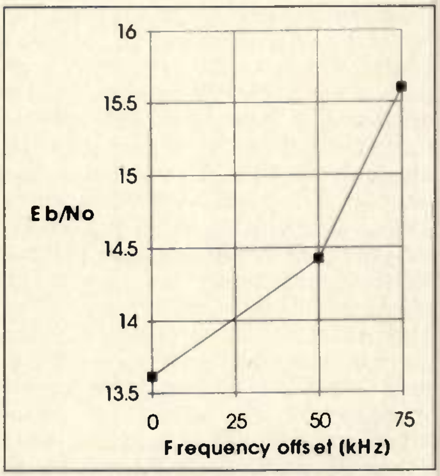


Figure 7 - Required E_b/N_o as a function of the frequency offset on the received signal.

The level of the desired signal is specified to be –73 dBm and the level for each of the two undesired signals is –46 dBm. The third order intercept point for a non-linear 2-port input (IIP_3) is defined as the point at which the third order intermodulation product equals the ideal linear, uncompressed output.

The intercept point can be found by :

$$P_{IIP3} = \frac{3P_{uw} - P_{Im3}}{2} \qquad (6)$$

with:

P_{IIP3} is the third order input intercept point (in dBm).

P_{im3} is the related intermodulation product for the two unwanted signals (in dBm).

P_{uw} is the power of each of the unwanted signals (in dBm).

P_{uw} is known from the DECT type approval documents. P_{im3} can be derived, remembering that the intermodulation product appears as a co-channel interfering signal, and DECT documents specify the co-channel interference rejection performance.

Co-channel interference rejection is defined as the ability of the receiver to cope with DECT-like interfering signals appearing at the same RF frequency as the desired signal. The co-channel rejection ratio is the ratio of the desired signal to the undesired that produces a certain BER. For a desired signal of –73 dBm, the co-channel rejection ratio is specified to be 10 dB maximum. This 10 dB ratio can be directly used, as the intermodulation performance is measured at the same level of the desired signal. Thus P_{im3}≥–83 dBm. Inserting for P_{uw}=–46 dBm and P_{im3}=–83 dBm, yields a receiver input intercept point of P_{IIP3}=–27.5 dBm.

Table 2 shows an example of a typical RF chain budget for a DECT receiver. The design issues are to have as low a noise figure as possible, a high input intercept point and a high overall receiver gain. The latter is in order to have the limiter saturated all the way down to the sensitivity level. From the table it can be concluded that (in this particular example) the mixer sets the limit for the overall noise figure, and that the LMX2240 in combination with the IF filter sets the limit for the overall input intercept point. Both figures of merit for the receiver are met by a reasonable margin, i.e. receiver sensitivity is met by a margin of 4.2 dB and input intercept point by a margin of 4.1 dB.

				Cumulative Data				
# Component	Gain	NF	OIP3	#	Gain	NF	IIP3	OIP3
1 Filter	–1.0	1.0	100.0	1	–1	1.0	97.5	96.5
2 Circulator /Switch	–1.0	1.0	40.0	2	–2	2.0	42.0	40.0
3 LNA/ LMX2216B	10.0	4.7	7.0	3	8	6.7	–1.0	7.0
4 Filter	–2.0	2.0	100.0	4	6	6.8	–1.0	5.0
5 Mixer/ LMX2216B	6.0	17.0	0.0	5	12	12.3	–12.3	–0.3
6 IF SAW-Filter	–4.0	4.0	100.0	6	8	12.4	–12.3	–4.3
7 LMX2240	70.0	8.0	55.0	7	78	12.6	–23.4	54.6

System Cumulative Values

Required E_b/N_o	13.60 [dB]	Gain	78.0 dB
Sensitivity	–87.20 [dBm]	N Fig	12.6 dB
IIP3	–23.4 dBm	OIP_3	54.6 dBm

Table 2 - A typical RF chain budget for a DECT receiver.

Baseband Issues

Also facing the radio designer are baseband issues such as bit or symbol timing recovery (STR), compensation for DC drift due to carrier frequency drift through a discriminator, and the baseband filtering requirements for the transmitter. National's LMX2410 addresses all of these concerns. Symbol timing recovery is achieved through the use of an all-digital phase locked loop and a non-linear element at the input. Timing is recovered by over-sampling the input data stream (after a threshold comparator) and finding the rising or falling edge. The symbol clock phase is then adjusted to lock it to the incoming data stream's phase. The STR circuit in National's DECT solution can lock the internal clock to the received data in as few as five edges of the incoming data stream.

DC compensation of the incoming data stream is important when using a discriminator because carrier frequency drift will manifest itself as DC drift after an FM discriminator. The DECT preamble is 32 bits in length with a balance of 16 1s and 16 0s. A good way to track the DC level of the incoming signal is to monitor the duty cycle of the preamble and adjust the LMX2410's threshold DAC for the data comparator.

Baseband pre-modulation filtering is important in achieving the proper shaping of the transmit spectrum. In DECT, the pre-modulation filtering is specified as a Gaussian filter characteristic with a –3 dB bandwidth of half the bit rate (B_bT = 0.5). National achieves this pre-modulation filtering through the use of a ROM look-up table and an 8 bit DAC on the LMX2410. This allows an ideal Gaussian characteristic to be saved in the ROM. National's chip is also mask-programmable, which lends itself to easy adaptation for other standards.

System Issues

Three issues which face DECT phone designers are range, multipath fading (or delay spread), and voice quality. With the high bit rate, a DECT phone will have more susceptibility to multipath reflections. A normal delay spread in indoor environments is on the order of 100-200 ns. The bit time of a DECT bit is 880 ns. This means that the potential delay spread due to multipath reflections is 10-20 percent of a bit time. Typically, some method of diversity is used to combat such relatively small fractional delay spreads.

Switched antenna diversity is the simplest to implement. It involves using two antennas (usually in the base station) and measuring the received signal strengths of both antennas. The larger signal is then received for the duration of the burst. Full receiver diversity involves two complete receivers from antenna to the cyclic redundancy checker (CRC) in the baseband section. Whichever received signal passes the CRC is the one that is used for the duration of the burst.

Another, even simpler method is dual slot diversity. In this method, only one receiver is used, but two (usually successive) timeslots are used for receiving one data burst by each of the antennas. Identical information is thus transmitted in both timeslots. In the receiver, the timeslot that was received with the fewest errors is used. This method adds the performance enhancement of the dual receiver diversity principle to a single receiver architecture. However, it decreases system capacity due to the extra timeslots being used.

The range and voice quality of a DECT phone must be minimally wireline quality. The range of the phone must be long enough so that the phone is not always in hand-off mode, but short enough to allow for multiple cell (base station) sites. In the indoor office environment, a good range is about 100 meters and about 200-300 meters outdoors.

Summary

A new chip set for DECT applications is now available from National Semiconductor. Using a single-conversion receiver and direct- modulation transmitter, these components offer a low cost, low power consumption solution for DECT equipment. *For more information, circle Info/Card #260.* ***RF***

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Benny Madsen has a Ph.D in Electrical Engineering from Aalborg University in Denmark. Recently, he has worked at Dancall Radio, Denmark, developing analog and digital cordless and cellular phones. Since January 1993, he has worked at National Semiconductor Corporation as an engineering project manager in the Wireless Communications Group.

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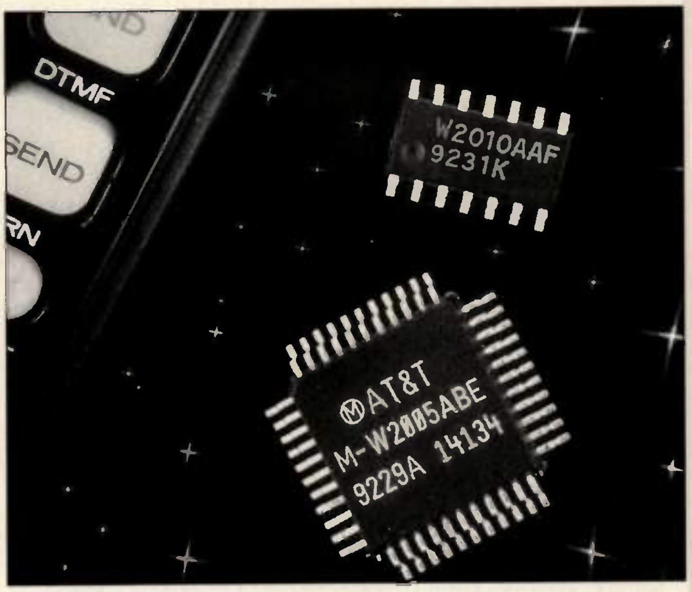
They can be reached at National Semiconductor, M/S A1500, P.O. Box 58090, Santa Clara, CA 95052-8090.

Dual-Mode Wireless Chip Set



AT&T Microelectronics now offers a pair of integrated circuits specifically designed for the IS-54 cellular standard. The 900 MHz transceiver chip set incorporates virtually all receive and transmit functions needed for detecting and transmitting both analog and digitally modulated signals. The AT&T W2005 receiver is the first single IC to address the unique requirements of the dual mode IS-54 standard and is the only device to perform both FM (analog) and DQPSK (digital) demodulation on the same chip. The W2005 uses a superheterodyne architecture to shift the frequency of the 824-849 MHz signals to an intermediate frequency of 455 kHz, and then offers a dual path for detecting either the analog or digitally modulated signals. The companion W2010 transmitter uses direct-up conversion to modulate the voice signal at the 900 MHz transmission frequency. The quadrature modulator features an on-chip phase splitter that reduces phase imbalance to less than 0.95 degrees and amplitude imbalance to less than 0.15 dB. Samples of both ICs are currently available, with production quantities expected in mid-1993. In production quantities, the AT&T W2005 Cellular Receiver IC costs $6.00 each, and the W2010 Quadrature Modulator sells for $4.50 each.

AT&T Microelectronics
INFO/CARD #250

Medium Power Amplifier

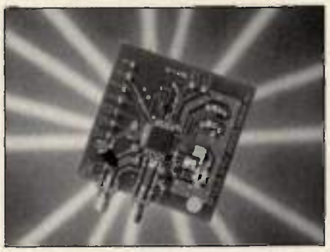
RF Micro Devices has announced the availability of the first commercial linear power amplifier fabricated using TRW's Heterojunction Bipolar Transistor (HBT)

technology. The RF2103 is designed for use as the final linear RF amplifier in UHF radio transmitters operating between 800 MHz and 1 GHz. The device is packaged in a 14-pin ceramic SOIC package. The device is self contained with the exception of the output matching network and power supply feed line. It produces a peak output power level of 800 mW (CW), or 40 mW average for a two tone input, at a supply voltage of 6.3 VDC. The device can also be used in 3 V battery applications, where the maximum CW output is 135 mW. Total gain is 25 to 30 dB, with low VSWR across the operating frequency. The RF2103 is now available in a ceramic SO-14 package and will be available in a plastic package later this year. Pricing for the RF2103 in the ceramic SO-14 package is $15 in quantities of 100K. Projected prices for the plastic version are less than $10.

RF Micro Devices, Inc.
INFO/CARD #249

GPS "Front-End" Evaluation Card

The GP1010-1 Evaluation Card from GEC Plessey Semiconductors uses the company's GP1010 RF Downconverter to produce sampled digital signals from a global positioning system antenna/low noise RF pre-amplifier, or from simulated signals. The board provides an integrated solution with a minimum of premounted components around the GP1010 IC. Provision is made for RF input at 1.57 GHz and for the



10 MHz reference signal via SMA jacks. The first IF is designed to use low cost LC filter components with a 3 dB bandwidth of 20 MHz. The second IF at 35 MHz uses a GEC Plessey DW9230 SAW filter for better than 40 dB out-of-band noise rejection. The board is supplied assembled and tested. The 60 x 60mm board requires a 5 V power supply, sample clock source, and 10 MHz local oscillator reference source. The board is free of charge to engineers involved in actively designing GPS receivers.

GEC Plessey Semiconductors
INFO/CARD #248

Spectrum Analyzer for Digital Cellular

Anritsu Wiltron introduces the MS2602A spectrum analyzer designed for digital cellular radio

measurements. The spectrum analyzer has a frequency range of 100 Hz to 8.5 GHz, and a level accuracy up to ±1.1 dB. The MS2602A is a high performance spectrum analyzer, with a high signal purity of –120 dBc/Hz at a 100 kHz offset. This characteristic, plus low intermodulation distortion, allows measurements such as adjacent channel leakage power and linearity measurements of power amplifiers to be made relatively easily. Simultaneous display of frequency and time domain is provided by the MS2602A. Other capabilities include AM/FM demodulation, auto tuning, signal search, zone sweep, 10 multiple markers, GPIB, and an FM waveform display. The MS2602A is priced at $23,950.

Anritsu Wiltron Sales Co.
INFO/CARD #247

UHF, Broadband Power MOSFETs

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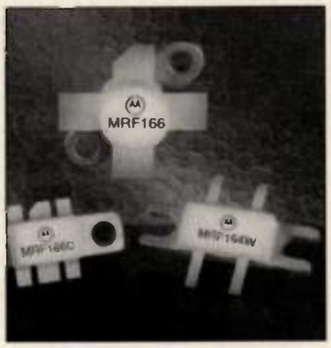

2 W and 20 W in different package options. Example devices are: MRF158, 2 W output power, 20 dB gain (typ.), 0.45 pF C_{rss} (typ.), $18.10; MRF164W, 20 W output power, 17 dB gain (typ.), 2.5 pF C_{rss} (typ.), $79.00; MRF166C, 20 W output power, 17 dB gain (typ.), 4.5 pF C_{rss} (typ.), $31.45. All prices are for U.S. delivery in low quantities.

Motorola Semiconductor Products Sector
INFO/CARD #246

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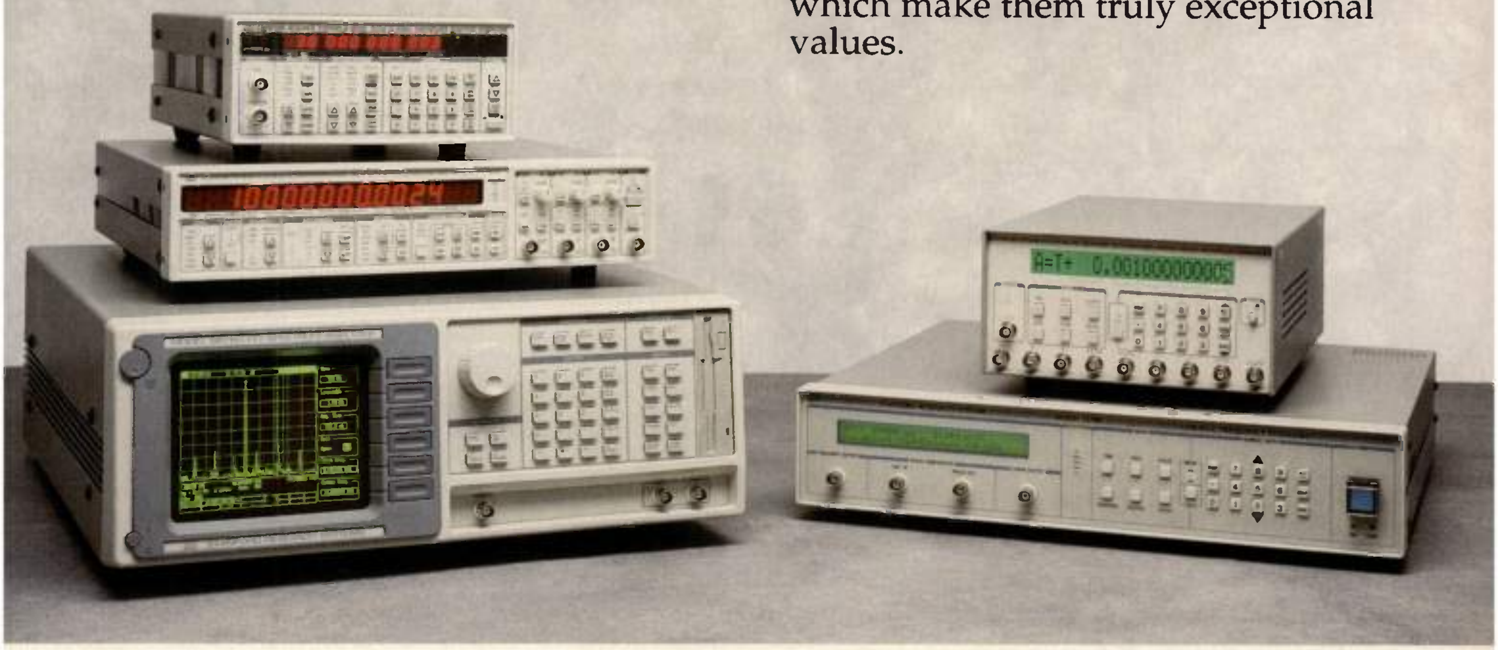

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Metelics Corp.
INFO/CARD #245

Space Qualified

A full line of Hi-Rel MELF devices screened to JANTX or JANTXV specifications for space level military projects are available from Compensated Devices. Surface mount semiconductors include Schottky, 1 Amp Schottky, Zener and current regulating diodes.
Compensated Devices, Inc.
INFO/CARD #244

Simulate Spread Spectrum Communication

Spread spectrum radios can operate on frequencies where many other, sometimes more powerful, transmitters are simultaneously operating. An accurate evaluation of the performance of a spread spectrum radio in such environment would require many transmitters and would be very costly and awkward to perform. Using a UFX-BER Series noise generating instrument to simulate interference from other radios operating within the same frequency band is a much more practical alternative.

Direct-sequence spread spectrum communication techniques such as Code Division Multiple Access (CDMA) are simulated well by white noise. This is because direct sequence spread spectrum looks much like white noise. Frequency-hopping spread spectrum systems, in which several radios are transmitting at the same time, are also well simulated by white Gaussian noise because of the central-limit-theorem. This theorem states that when many non-correlated stochastic variables are added, the result approaches a Gaussian distribution function, provided that none of the variables dominates the sum. Simulation of many non-correlated, frequency-hopping radios can be made by use of white Gaussian noise.

The relation between the number of interferers, their power, and the necessary noise power density of the simulation is:

$$N = 10_{\log} M + \langle P \rangle$$

where:

N = the average white noise power (in dBm)
M = the number of interferers
$\langle P \rangle$ = the average power of interferer (in dBm)

It is therefore necessary to be able to generate carrier-to-noise ratios (C/N) from −40 through +40 dB. The UFX-BER Series does this with an accuracy of greater than 0.21 dB RMS.

The UFX-BER instruments include precision step attenuators for the signal and the noise source, plus a signal and noise combiner, built-in carrier-to-noise-ratio power meter, and carrier power monitor (Figure 1).

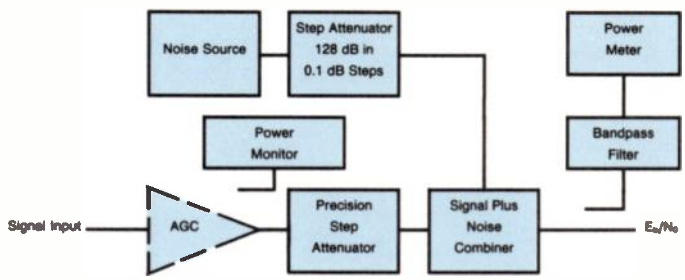

Figure 1. Block Diagram of UFX-BER Series Instrument.

The built-in power meter measures the carrier input power. The carrier output power is then set to the desired level by use of the instrument's precision step attenuator. The signal carrier power may be kept constant by passing the signal through an optional linear automatic gain control (AGC) amplifier. The instrument then generates the desired amount of noise power (also measured with the power meter) and combines the input signal with noise.

The output power and the average signal input power are alternately monitored. The carrier output power, and the peak deviation of the carrier power are then displayed simultaneously on the front panel along with the resultant accurately generated C/N.

The selectable, customer specified filters are all inserted in the power meter and in the noise path only, so as not to introduce any group delay distortion in the carrier path. The filters in the power meter path improve the signal-to-noise-ratio, and thereby the accuracy of the carrier power measurements. The filters in the noise path assure adequate dynamic range to maintain a 5:1 (14 dB) crest factor.

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E_b/N_o Noise-generating Instruments for Bit Error Rate Testing

UFX-BER Series

General Specifications

Carrier Path

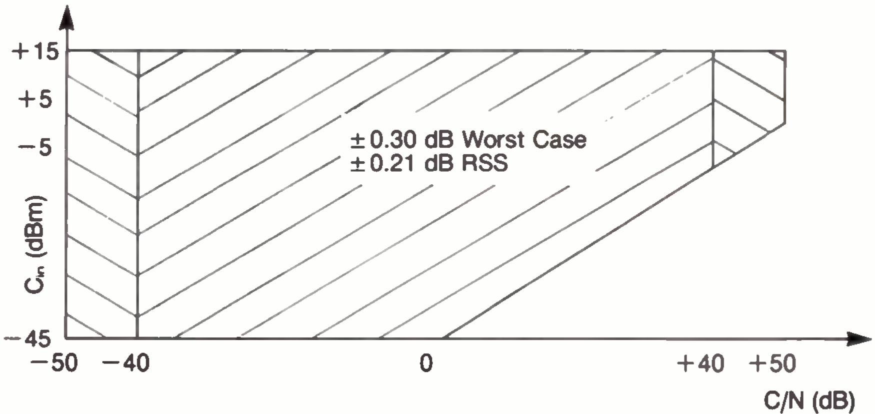
Attenuation	0 to 60 dB, automatically programmed
Range of carrier input power (C_{in})	−45 to +15 dBm
Flatness	±0.20 dB for 70 MHz (±20 MHz) ±0.30 dB for 140 MHz (±40 MHz)
Group delay	0.20 ns or less

Output Path

Range of carrier output power	−45 dBm to carrier input level
Range of noise output power (N)	−45 dBm to +5 dBm
Noise flatness	±0.50 dB or less from 100 kHz to 200 MHz
Accuracy of C/N	See graph
Third-order intercept point	+29 dBm typical
Crest factor (peak to average noise voltage)	14 dB or more

UFX-BER SERIES

MODEL	CENTER FREQUENCY	BANDWIDTHS
UFX-BER-455	455 kHz	Filters are customer specified
UFX-BER-10	10.7 MHz	
UFX-BER-21	21.4 MHz	
UFX-BER-45	45 MHz	Maximum bandwidth 200 MHz
UFX-BER-70	70 MHz	
UFX-BER-140	140 MHz	
UFX-BER-IF1	70 MHz, 140 MHz	
UFX-BER-_ _	User Specified	



C/N Accuracy Versus Carrier Input Level at 23±3°C

Features:

- Direct display of E_b/N_o or C/N
- Accuracy of 0.21 dB RSS
- 75 ohm impedance BNC connectors
- Variable output power

The UFX-BER Series of noise-generating instruments display the actual E_b/N_o or C/N for testing of satellite communications or digital radio equipment. The UFX-BER generates and measures E_b/N_o for IF back-to-back or loop-back testing with extreme precision over a broad range of input or output power, steps, increments, decrements, and desired E_b/N_o values can be entered directly in dB in 0.25 dB increments from the front panel or by IEEE-488 bus.

Settings are stored in non-volatile memory and the user can create up to eight test routines (including delay time and loops) that will run automatically under program control. The instrument measures 5 1/4 in. high, 17 in. wide, and 20 in. deep and comes with brackets for 19-in. rack mounting.

Options:

1. Automatic gain control amplifier to maintain constant carrier power level
2. Other custom configurations
3. 50-ohm input and output impedance instead of standard 75-ohm impedance
4. RS-232C, RS-422, or RS-423 interface in addition to IEEE-488 interface
5. 220 VAC, 50 Hz, instead of standard 110 VAC, 60 Hz
6. Interferor input
7. Switched filter bank with up to six customer-specified filters

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RF products

SUBSYSTEMS

Radio Modem Board

The System 200, from Monicor Electronic Corp., communicates any RS-232 data, input and output, at any output level from 2 W to 1 mW. System 200 comes with Monicor's highly efficient TurboLink 2.0 operating system, supporting the X.30 protocol. Transceiver operation is narrowband FM between 450 and 470 MHz. System 200 is priced at $465 per set in OEM quantities.
Monicor Electronic Corp.
INFO/CARD #243

GPS Synchronization

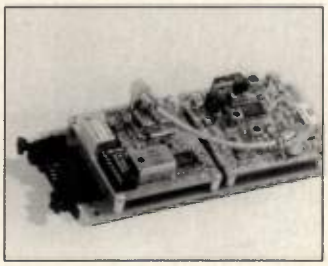
Datum Timing has added a GPS time input option to their Model 9700 Programmable Time System. The option is easily installed in any Series 9700 product. Once enabled, the GPS time input option determines position and transfers time to the time code generator with sub-microsecond precision.
Datum Timing, Inc.
INFO/CARD #242

RF Modem

The FM 96 Data Radio from IWL Communications provides fast, reliable data transfer using a UDS half-duplex wireless modem. Asynchronous serial data can be transmitted at 9600 bps over commercial frequencies at 450 - 470 MHz.
IWL Communications, Inc.
INFO/CARD #241

900 MHz RF Modules

RF modules from RF Industries use a synthesizer PROM control interface with an optional serial or parallel microprocessor control. These OEM modules measure 2 x 2 inches and can be configured for multiple channels, 12.5 or 25 kHz channel spacing and wide frequency range from 850 to 960 MHz.
RF Industries, Ltd.
INFO/CARD #240



AMPLIFIERS

Ultra Low Noise

AML announces model ALV02202408 S-band low noise amplifier with internal signal level monitor and automatic gain control loop. Maximum gain is 37 dB. Gain with full AGS attenuation is 17 dB. Maximum noise figure at maximum gain is 0.7 dB.
AML, Inc.
INFO/CARD #239

10 kHz to 100 MHz, 1 kW

Amplifier Research has extended the power range of its solid-state Series "A" line to 1000 watts minimum, linear output, with the introduction of the model 1000A100 RF amplifier. Price is $74,000, with delivery in 30 days ARO.
Amplifier Research
INFO/CARD #238

50 W, Solid State

The model AR1929-50 solid state RF amplifier operates over the 1000 to 2000 MHz frequency range. The amplifier from Power Systems Technology provides 50 W at 1 dB compression, 55 watts at saturation, and operates Class A with an instantaneous bandwidth of 1000 MHz. The rack or bench mountable amplifier includes a built-in power supply.
Power Systems Technology, Inc.
INFO/CARD #237

Power Module Boards

Tactical Electronics offers printed circuit boards to facilitate the usage of the following power modules: MHW series (except 714-04 case) from Motorola, M series from Mitsubishi, BGY series from Philips and the SA series from Toshiba.
Tactical Electronics Corp.
INFO/CARD #236

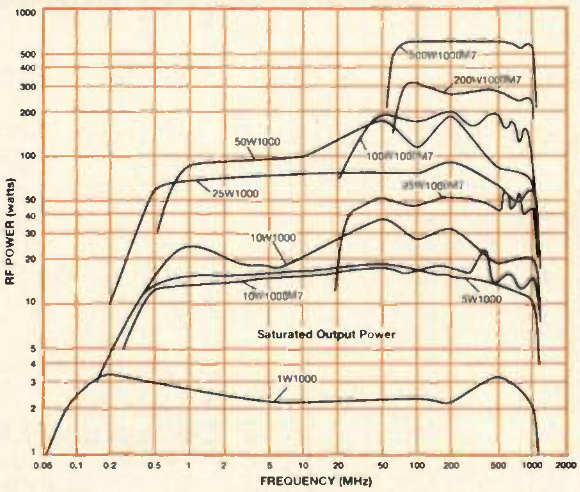
When it gets lonely out there at 1 GHz, you'll be glad you have all this power.

The only amplifier that can deliver 500 watts through that balky decade from 100 to 1,000 MHz is our new Model 500W1000M7. That's the majestic one on the left, with the front-panel bi-directional power meter, ALC, and a bagful of other controls that will let you actually *enjoy* automatic sweep testing in your lab.

The 500W1000M7 presently heads our well-known "W" Series of all-solid-state linear amplifiers that cover four crucial decades of bandwidth from 100 kHz to 1 GHz. Today, you may need only 1 watt (the little portable on the top), or 5, or 10, or 25, or 50, or even 100 or 200 watts *minimum*—all with that fantastic bandwidth instantly available without tuning or bandswitching—a combination of power and bandwidth that's comforting when you work the outer reaches of the rf spectrum. Wherever your power requirements are today, they're sure to go up. And one of these days you're going to want a 500W1000M7.

The "W" Series is part of a complete line of AR amplifiers offering rf power up to 10 kilowatts for rf susceptibility testing, nmr spectroscopy, plasma/fusion research, and a host of other test situations that demand rf power of unconditional stability—immunity to even the worst-case load mismatch, shorted or open cable, wild swings of VSWR—with no fear of oscillation, foldback, or system shutdown.

You might learn a lot from our free booklet, "Guide to Broadband Power Amplifiers." Send for a copy. Or call our toll-free number (800-933-8181), which will connect you directly to one of our applications engineers.



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RF products

SEMI-CONDUCTORS

Radio Data Demodulator

Featuring on-chip anti-alias, bandpass and signal reconstruction filters, the SAA6579 Radio Data System (RDS) demodulator from Philips Semiconductors requires the addition of only six passive components and a crystal to recover RDS data from FM radio broadcasts. Cost is in the region of $2.20 in volume.
Philips Semiconductors
INFO/CARD #235

MMIC Amplifier

Model P35-4104 GaAs MMIC amplifier features 18 dB of gain. The operating frequency for the amplifier from Daico Industries is 0.05 to 3.0 GHz. Typical power output is +13 dBm, with only a 4.5 dB noise figure. Input and output are matched to 50 ohms. The part is available in chip or surface mount form.
Daico Industries, Inc.
INFO/CARD #234

Speech Compression IC

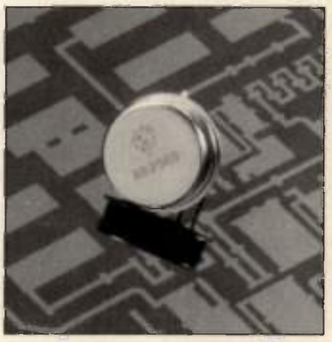
Qualcomm today announces the Q4400 variable-rate vocoder, the world's first VLSI circuit implementing the QCELP (Qualcomm Codebook Excited Linear Predictive code). Compression rates of 16:1 are possible with the 100-pin plastic quad flat pack device. Power consumption is less than 1 W. Pricing is as low as $69 each in thousands.
Qualcomm, Inc.,
VLSI Products
INFO/CARD #233

45 MHz Op Amps

Linear Technology's LT1208 and LT1209 dual and quad op amps have a 50 degree phase margin and are stable driving any capacitive load. Slew rate is 400 V/us. The LT1208 and LT1209 are available in 8-pin and 16-pin packages, respectively. Both SO and DIP packages are available. Pricing for the LT1208 dual is $4.45 in plastic in quantities of 100.
Linear Technology Corp.
INFO/CARD #232

High Power Amplifier

The model AR2569 thin-film cascadable amplifier from Cougar Components operates over 50 to



2500 MHz. Typical gain is 16.8 dB. Typical output power at the 1 dB compression point is +27.5 dBm. The device is packaged in a 4-pin TO-8B housing, with SMA and flatpack available.
Cougar Components
INFO/CARD #231

Amplifier

Amplifonix announces a low-cost, high output amplifier operating from 30 to 500 MHz. Model RZ9604 has typical output compression of +21.0 dBm, with gain of 23 dB. The device is packaged in a TO-8B package and costs under $30.00 in large quantities.
Amplifonix, Inc.
INFO/CARD #230

DISCRETE COMPONENTS

RF Coils

Adjustable RF coils — both shielded and unshielded — are available from J.W. Miller. Series 4900 adjustable coils provide choices within the 30 to 250 MHz range. Series 48A and 49A adjustable coils provide stability and uniformity within the 10 to 250 MHz range. Inductance values from 0.037 to 0.788 uH are available from stock.
J.W. Miller Division of
Bell Industries
INFO/CARD #229

Chip Inductors

AVX Corp. has introduced the ACCU-L series of surface mount RF inductors. Available as an 0805 chip, ACCU-L inductors exhibit high SRF, high Q and low resistance. They are ideal for applications between 100 MHz and above 2 GHz. This series covers the range from 2.7 nH to 15 nH with tolerances of ±0.5 nH and ±5 percent.
AVX Corp.
INFO/CARD #228

TEST EQUIPMENT

Calorimeter

Dielectric Communications' model 7200 calorimeter may be utilized with any existing fluid cooled load, offering an easy method of accurate RF measurement. Virtually any cooling fluid can be utilized, and power can be displayed in kilowatts, kcal/h or MJ/h.
Dielectric Communications
INFO/CARD #227

Tracking Generator

The 2707 external tracking generator from Tektronix provides swept-frequency coverage from 100 kHz to 1.8 GHz. Output level adjustments can be made between 0 to –48 dBm, in 0.1 dB steps. The 2707 works with Tektronix 271X spectrum analyzers, and is priced at $3950.
Tektronix
INFO/CARD #226

Noise Figure Measurement

Noise Com has introduced the NCT 6000 series of instruments for noise figure measurement. The series includes a calibrated white noise source, low noise amplifier, impedance matching circuitry, step attenuator with 0.1 dB resolution, power supplies and switches necessary to perform accurate noise figure measurements when using a spectrum analyzer.
Noise Com, Inc.
INFO/CARD #225

Frequency Counter

A portable 300 MHz frequency counter has been introduced by John Fluke Mfg. The PM 6685 has a resolution of 10 digits per second of measuring time. The counter has ovenized oscillator time-base options and stability as

high as 5×10^{-10}/day. Optional RF inputs extend the upper frequency range to 1.3, 2.7 or 4.5 GHz. The PM 6685 U.S. list price is $1860.
John Fluke Mfg. Co., Inc.
INFO/CARD #224

Radar Signal Simulator

The RA-100AM can simulate up to ten asynchronous pulsed emitters, each with individual pulse widths, pulse repetition intervals, pulse groups, jitter and signal-to-noise ratios. The RA-200 is a rack-mounted, modular RF Radar signal simulation system.
S.T. Research Corp.
INFO/CARD #223

SIGNAL SOURCES

Tiny OCXO

Vectron Laboratories has developed the CO-738S series of oven controlled crystal oscillators with aging rates of 5×10^{-8}/year and temperature stabilities of $\pm 1 \times 10^{-9}$ over 0 to 50 degrees C. Available in frequencies from 32 kHz to 32 MHz, the oscillators measure 1.5 x 1.25 x 0.86 inches. Price is $373 each for 100 pieces.
Vectron Laboratories, Inc.
INFO/CARD #222

Coaxial Resonator VCO

T and M Microwave model VCR 8001 is a high Q, coaxial resonator oscillator operating at a center frequency of 1525 MHz. The surface mount devices can be tuned ±10 MHz with a tuning voltage of 0 to 12 V. Typical output power is +13 dBm.
T and M Microwave, Inc.
INFO/CARD #221

A Basic Review of Feedback

By Gary A. Breed
Editor

Feedback is the principle that makes it possible to have accurate linear systems. Amplifiers, oscillators, automatic gain control and phase locked loops are among the common RF applications of feedback. Although the design equations for these circuits are well known and most undergraduate linear systems courses provide a mathematical derivation, this short note reexamines the most basic concepts of feedback from an intuitive perspective.

There are two essential concepts to remember regarding feedback: 1) feedback trades gain for control, and 2) uncontrolled feedback can corrupt the desired results and cause oscillation. The first represents the power of feedback for designers; the second is a warning to the same designers.

The generalized feedback system is shown in Figure 1. The input passes through the forward signal path, an amplifier with gain A. The output is subjected to some transfer function, β, and summed in an inverted form with the input. The net input is now the difference between the original input and a modified portion of the output. This is the classic "negative feedback" concept.

This linear feedback system can be represented by the equation:

$$V_{OUT} = A(V_{IN} - \beta V_{OUT}) \qquad (1)$$

or, rearranging:

$$V_{OUT}(1 + A\beta) = AV_{IN} \qquad (2)$$

To obtain the overall gain (closed loop gain):

$$A_{CL} = \frac{V_{OUT}}{V_{IN}} = \frac{A}{1 + A\beta} \qquad (3)$$

In a perfectly-designed feedback system, we can think of the transfer function β as detecting any error in the input/output relationship. The error-correction signal is then subtracted from the input, resulting in an error-free signal at the output of the system.

The simplest illustration of the principle is an operational amplifier. Figure 2 shows a circuit with the input applied to the non-inverting input, and two equal-value (10kohm) resistors making a voltage divider in the feedback loop to the inverting input. According to equation (3) above, β = 1/2, since the voltage divider presents 1/2 of V_{OUT} to the inverting input. The gain of this system will be:

$$A_{CL} = \frac{A}{1 + A/2} \qquad (4)$$

In an ideal system, A is infinite, but we can live with a very large A of perhaps 100,000. In this case:

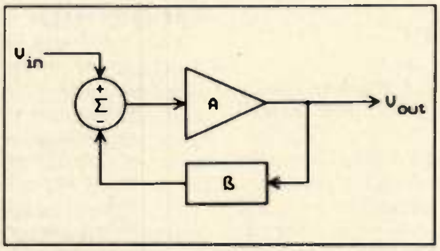


Figure 1. The classic representation of negative feedback.

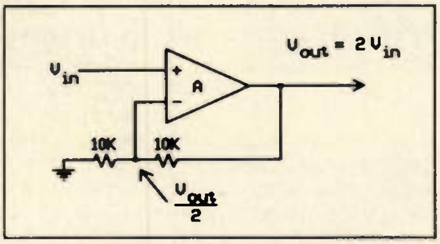


Figure 2. A simple gain-of-2 operational amplifier example.

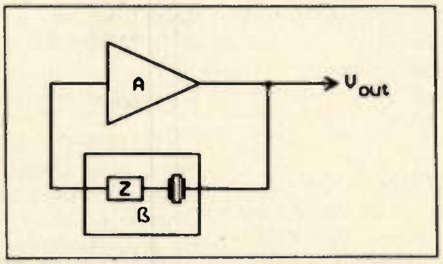


Figure 3. The classic representation of negative feedback.

$$A_{CL} = \frac{100{,}000}{1 + 50{,}000} = 1.99996 \qquad (5)$$

As A goes to infinity, the closed loop gain will become exactly 2, or 1/β.

Let's follow this process around the op amp loop. First, an input V_{IN} is applied, which is amplified by gain A, which gives us some output V_{OUT}. However, half of the output is then subtracted from the input, so the input is now $V_{IN} - V_{OUT}/2$. If we use the "error function" concept noted above, then $V_{IN} - V_{OUT}/2 = 0$ when the system is operating error-free. This condition is satisfied only if $V_{OUT} = 2V_{IN}$; or $A_{CL} = 2$, as shown by equation (5).

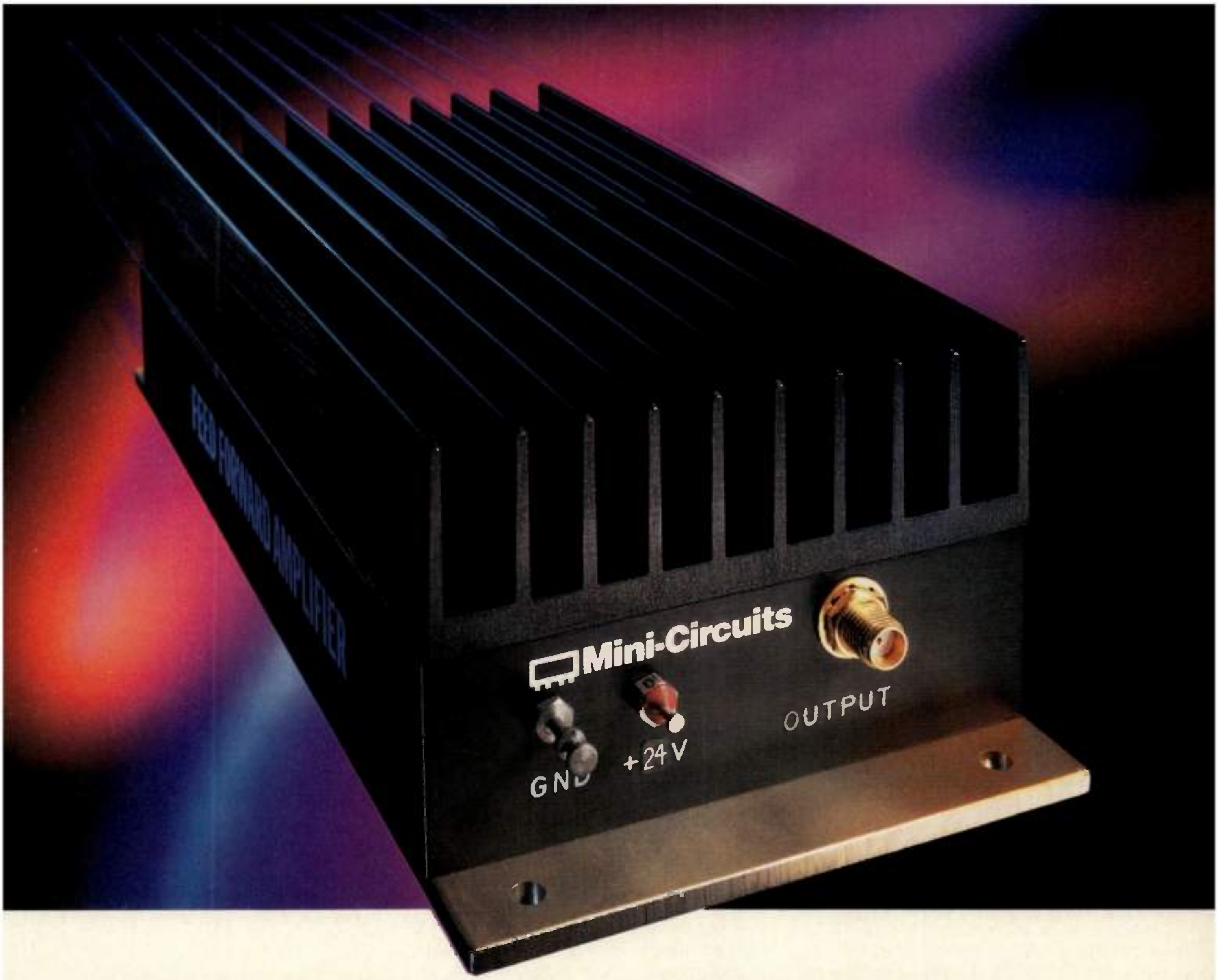
This example demonstrates principle 1) above: we can trade gain for control. Passive feedback components alone define an exact gain-of-2 system. The more gain we have to trade (i.e., A approaching infinity), the better the accuracy.

Non-ideal Systems

Of course, feedback is not limited to ideal amplifiers and resistive feedback networks. The feedback can be any impedance, including frequency-and time-dependent networks of resistors, inductors and capacitors. The network can also be non-linear, using diodes or transistors.

The amplifier need not be ideal, either. Frequency response and phase shift must be included in the forward transfer function, which can then be applied to the feedback (β) transfer function. As long as the operational parameters of the circuit are known, feedback networks for the circuit can be computed.

An oscillator is a common RF circuit requiring application of feedback principles. Let's use the simplified oscillator circuit in Figure 3 as an example. Let's assume that the amplifier A is a packaged device which has a gain of 10 at the frequency of interest, and is an inverting amplifier with a phase shift of 180 degrees at that frequency. For β, we will use a crystal along with additional circuity (Z) to obtain the necessary feed-



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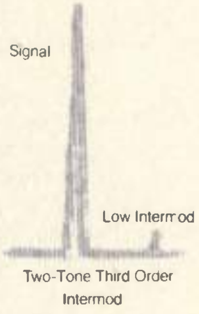

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Two models, 50-500MHz and 50-1000MHz, unconditionally stable, with more than 20dB gain and 45dB reverse isolation will satisfy most system needs for low intermodulation distortion. Even amplifier second order intercept point is greater than +63dBm!

So, check the specifications listed and re-evaluate your amplifier design decisions for improved system performance at lower cost.

One-week shipment guaranteed. Only from Mini-Circuits.

SPECIFICATIONS	ZHL-1-50P3	ZHL-2-50P3
Frequency, MHz	50-500	50-1000
Gain, Min, dB	22	21
Power Output(1), dBm	27(2)	25
Intercept Point, dBm		
IP2	70	63
IP3	50	43
DC Power		
Volts	24	24
Current, Amps	1.3	0.65
VSWR, 50 ohms, In/Out	1.5:1	2:1
Price (1-9)	**$695.00**	**$895.00**
Size(3) L x W x H, inch	7 x 3.25 x 2.13	

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F151 REV ORIG

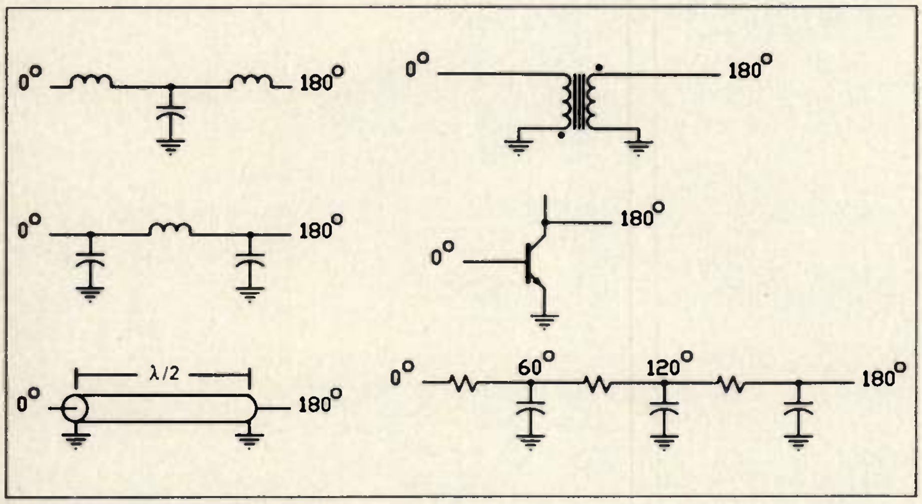

Figure 4. Methods of obtaining the necessary 180-degree phase shift.

back. We'll assume the crystal resonator is operating in its series-resonant mode, so that it is lossless and has no phase shift at the operating frequency.

The criteria for oscillation require that the total phase shift through the amplifier and feedback circuit be (n)2π. Also, when steady-state conditions are reached, the gain of the system must be 1; or more correctly, the output must remain at a fixed level. The components we have chosen must work together with the feedback network to meet these criteria. Note that in the circuit described, the feedback is not inverted, but presented directly to the input.

Since the amplifier has a phase shift defined as 180 degrees, Z must supply the additional phase shift of 180 degrees to meet the requirement of 2π total phase shift. Figure 4 shows some design options. This part of the feedback network can be a lumped element network with a 180-degree phase shift, or it could be a transmission line section with a 180-degree, or 1/2 wavelength time delay. At lower RF frequencies, a transformer might also be used to provide the phase shift, since 180 degrees is a simple inversion. Other possibilities include introducing the required phase shift using cascaded R-C networks or an inverting amplifier.

The other part of the oscillation criteria, a gain of 1 (output at a constant level), requires a non-linear function. As shown in Figure 5, one method is to use opposed diodes to limit the maximum amplitude of the signal. The forward voltage drop of the diodes will determine the actual amplitude, which must be high enough to let the amplifier operate in its optimum power range. Mainly, we want to assure sufficient excess gain when the oscillator is starting, and then maintain that the output level well above the inherent noise level. Another method of limiting the maximum amplitude is to allow the amplifier itself reach saturation. The power supply voltage and bias levels can be selected to perform the same function as the diodes, as long as the ratings of the active devices used in the amplifier are not exceeded.

With a feedback network of diodes for limiting and a transmission network for phase shift, the oscillator should operate (Figure 6).

Again, we can follow the process of operation through the system in an intuitive way. In an oscillator, the system noise is the initial input, which is amplified by A. The entire output voltage is returned to the input via the crystal and phase shift network. At the crystal's resonant frequency, the fed-back output is exactly in phase with the input, boosting the level again and again until it reaches the limiting level of the diodes or the saturation voltage of the amplifier. When this point is reached, only the frequency that corresponds to the resonance of the crystal is contained in the signal — In the process of re-amplification, energy at other frequencies is not in phase after feedback, and is amplified less and less in successive loops.

This is an intentional oscillator, but sometimes an undesirable feedback path meets the oscillation criteria. If a circuit has feedback due to stray inductance and capacitance, poor power supply decoupling, or improper p.c. board layout, we have the situation warned about in principle 2): uncontrolled feedback can result in unwanted oscillation. This is especially important with the new generation of MMIC, video and operational amplifiers, where lots of gain is available that can make a lossy, indirect feedback path exceed the gain of 1 criterion!

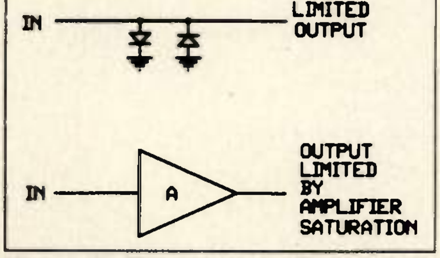

Figure 5. Two ways to limit amplitude.

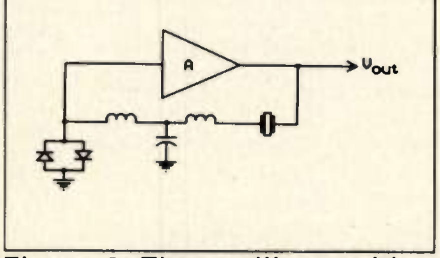

Figure 6. The oscillator with a complete feedback circuit.

Conclusion

Controlling a circuit through feedback is a powerful tool, allowing us to obtain precision linear (and non-linear) responses for amplification, filtering, and other signal processing applications. We can get very good performance from an imperfect active device (as long it has sufficient gain) by applying accurate passive components in a feedback circuit.

This powerful design tool can also work against us, since it can also be used to induce oscillations, intentional and unwanted. Especially in RF circuits, we must always remember that feedback can exist in ways that do not appear on the circuit diagram. ***RF***

References

1. T.M. Frederickson, *Intuitive Operational Amplifiers*, McGraw-Hill, 1988, Chapters 2, 3, 4.
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An RF Swept Filter for a Medical Ultrasound System

By Alexander Pummer
P.C.S. Consultants

This design is part of a medical ultrasound imaging system. It was designed to eliminate the shortcomings of an existing circuit. The attenuation of ultrasound waves in the body is depth and frequency dependant. The rate of attenuation is approximately 1 dB/cm/MHz. Because of this, the spectrum of the reflected energy is different from that of the transmitted energy. The higher frequencies are more attenuated. This causes a relative increase of noise at higher frequencies.

The normal way of solving this problem is to use a bandpass filter which is tuned to cut out the noisy part of the returning spectrum. Since we want to observe different depths in the body, we have to adjust the passband of this filter. This adjustment should happen continuously since we want to see a continuous picture. The filter has to be tuned during the receiving period.

The presently used solution utilizes a tuned LC circuit where the tuning element is a varicap diode. This circuit has two major problems. The first problem is that the relative high level signal will add to the tuning voltage and contribute to the tuning of the circuit and will therefore cause intermodulation. The second problem is a result of the first one. The filter cannot be tuned exactly to a certain frequency because of the uncertainty in the tuning voltage. The new design separates the tuning element from the signal path. Therefore, there is no parasitic tuning effect, no intermodulation, and the center frequency of the filter is exact at any time.

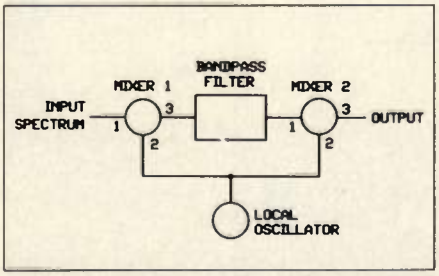

Figure 1. Block diagram of swept filter

The circuit (shown in Figure 1) works as follows: the input spectrum, f(s), feeds into the first mixer port #1. The local oscillator, f(LO), is driving the mixer at port #2. The combination frequencies f(s)±f(LO) are available at port #3. The bandpass filter could select f(s)+f(LO) or f(s)–f(LO), but for our application, we will select f(s)+f(LO). The selected spectrum f(s)+f(LO) feeds into the second mixer at port #1. The local oscillator f(LO) is also driving the second mixer at its port #2. The output spectrum of the second mixer is: f(s)+f(LO)±f(LO)=f(s)+2f(LO). A lowpass filter will separate the component f(s) from f(s)+2f(LO). If we sweep the local oscillator, the output spectrum of the system will be shaped by the amplitude frequency response of the bandpass filter and the virtual center frequency of the system will be equal to the frequency of the local oscillator f(LO). This system will act exactly like the ideal filter we need. The tuning of the local oscillator is totally separated from the signal path. The frequency of the local oscillator is controlled by a frequency closed loop. This loop will exactly tune the oscillator frequency to the required value. The bandpass filter is a fixed filter built with amplitude independent components.

Some Details of the Circuit

The mixers are single balanced only, but they are designed to carry high level signals without producing noticeable intermodulation. Figure 2 shows the mixer. The two switches S1 and S2 are conducting alternately. The sequence is controlled by the local oscillator. When S1 is on, S2 is off, and vice versa. These two switches are JFETs. A mixer will generate intermodulation products if the input level is high enough to influence the conduction angle of the switching elements. The conduction angle for a good mixer is determined by the oscillator only. Using the JFET as a switching device in this configuration will allow us to use relatively high signal levels. The signal level of the system at the input of the mixer is approximately 1 Vpp. The control signal for the JFETs comes from a fast CMOS output. The input capacitance of the devices is low, (3 pF max.), therefore the oscillator power is low. For a diode ring mixer we would have to use much higher oscillator power, which could create interference problems for the rest of the system.

The local oscillator looks relatively complicated, but the two loops of the circuit will take care of all the component tolerances. One of the loops keeps the oscillator amplitude constant. The other is the frequency control loop.

The ultrasound system provides a ramp to tune the oscillator. The loop ensures that at an instantaneous value of this ramp the oscillator will tune to one specified frequency. The actual oscillator frequency is twice that required for the mixing process. A 2:1 frequency divider (D flip-flop) is used to divide the output frequency of the oscillator, and provide the 50 percent duty cycle signal required for the proper function of the mixers.

From the production point of view it works like a digital circuit. In other words, if the right parts are loaded into the board it works without the hand-selecting and tweaking normally associated with RF oscillator circuits. The frequency closed loop usually has a frequency to voltage converter. I decided on a different method. One of the main system's microcontrollers provides a tuning voltage, via a D/A converter, for the oscillator. It then takes a sample of the oscillator frequency by using a ten bit binary counter.

The required frequency accuracy is

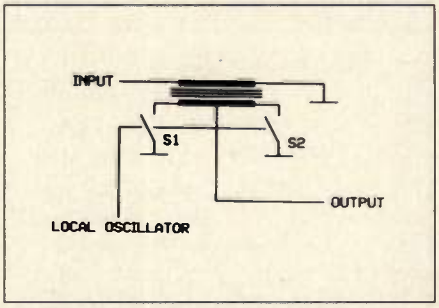

Figure 2. Illustration of mixer action used in the filter.

eight bits. By changing the tuning voltage in a stepwise fashion, the microcontroller will map the oscillator's control voltage/output frequency transfer function and will create a look up table. The control voltage to output frequency transfer function of the oscillator is a non- linear function. This look up table will be created during the initialization of the ultrasound system, and it will be refreshed whenever the filter is not in use and the microcontroller has the time. The loop will cancel any part tolerances, temperature effects, aging and drift. Since this loop is not an analog closed loop, it can be tuned quickly and without an expensive high-speed loop amplifier. This eliminates the problems associated with fast, non-linear closed loops.

The microcontroller also helps to set the overall gain by switching the feedback path of the intermediate frequency amplifier.

Details of Actual Circuit

The design specifications for the actual circuit are shown in Table 1. The schematic is shown in Figure 3.

Design specification:	
System gain:	0dB ±1.5dB
Max input level	1 V pp
Spurious-free dynamic range	at least 55 dB
Band width	±400 kHz for 0.5 dB roll off and min 18dB roll off for 1 MHz
Tuning range	2 MHz to 10 MHz
Tuning accuracy	8 bits = 0.25%
Tuning band width	50 kHz
Power consumption	+5V @ 100mA max −12V @ 50mA max
Parts cost	$50 max
Additional requirements:	
Low labor cost	
Minimal adjustment time	
Highly reliable, insensitive circuit	
No interference with the main system.	

Table 1. Design Specifications.

The input signal feeds from the main system into transformer T1. This should be a high level transformer in order to avoid nonlinearities due to saturation. U1 is a fast JFET array. U2 is the frequency divider and it should be located close to U1. The interconnection between U2 and U1 must be extremely short. The clock input line of U2 is not

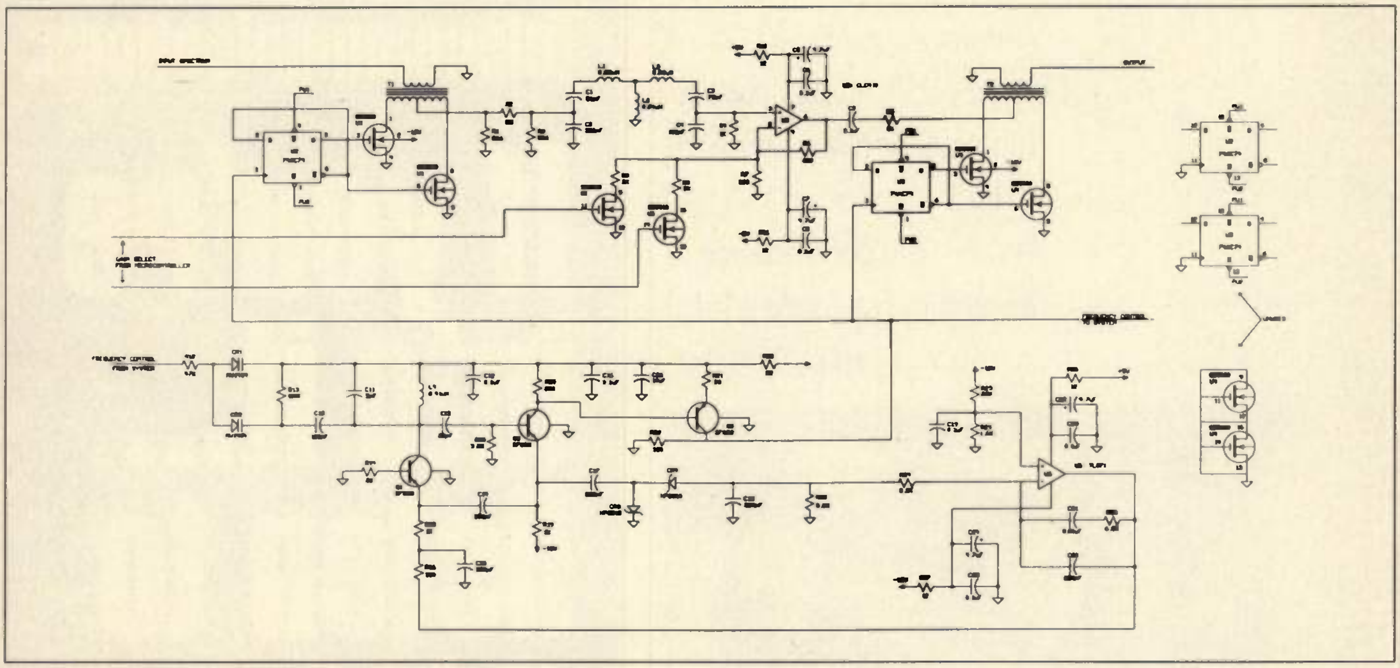

Figure 3. Schematic of the swept bandpass filter.

so critical. R1, R2 and R3 form a 4 dB path which separates the mixer output from the reactive input of the intermediate frequency filter. C1, C2, L1, L2, C3 and C4 constitute the intermediate frequency filter. The coupling is designed for maximum bandwidth and maximum flatness (so-called critical coupling). The coupling works via L2, which is a piece of stripline. The inductance is set by the PCB design, no adjustment is required. the only adjustable parts are L1 and L3. They are tuned for maximum output at 25 MHz. The input and the output of the filter are not equally tapped (i.e. the output loading is less than the input). Amplifier U3 compensates for the insertion loss of the filter, the attenuation of the 4 dB path, and for the conversion losses of both mixers. The self test routine of the main system provides a carrier to set the overall gain to 0 dB ±1 dB. The level of the input and output signal will be compared and the microcontroller will set the feedback path of the amplifier by turning on or off the switches Sa and Sb of U1. The output of the amplifier drives the input of the second mixer via the R5 matching resistor. The layout requirements for U4 and U5 are the same as for U1 and U2. The transformer T2 is identical to T1.

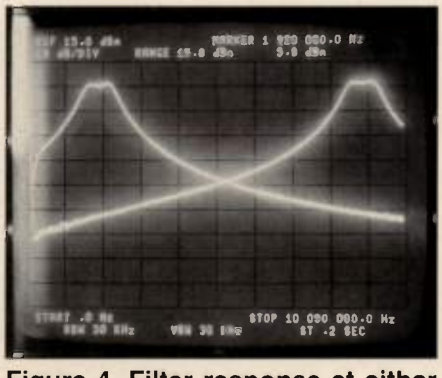

Figure 4. Filter response at either end of its tuning range.

Local Oscillator

The main requirement is simplicity for production. Coils should have two ends only, no tap, no secondary. Although a transistor array is available (RCA 3100), I used discrete parts. Q1 and Q2 are the active elements of the oscillator. The frequency is determined by L4 and the resultant value of C11, C10 and the two junction capacitances of the varicap diodes CR1 and CR2. The oscillator level is 4 Vpp across L4. This is a good compromise between noise and radiation. The 4 Vpp requires the use of two varicap diodes for first order cancellation. The oscillator amplitude is stabilized by the servo. The oscillator level is set by the voltage divider R25 and R29. The output of Q3 is a CMOS logic level signal which drives the two frequency dividers of the two mixers. It also provides a sample of the oscillator frequency for the control circuit. The utilized output frequency range of the oscillator is 27 MHz to 35 MHz. The lowpass filter mentioned earlier is not part of the circuit. It is already present as an anti-aliasing filter in front of the ultrasound system's A/D converter.

Results

This circuit was breadboarded and tested. It fulfilled the specifications with a healthy margin. The circuit was then integrated into the ultrasound system and worked according to our expectations. Figure 4 shows the filter's response. A patent application has been filed for this filtering scheme. *RF*

About the Author

Alexander Pummer received his Dipl.Ing degree at the Polytechnik University in Budapest Hungary. He built his first ham radio from scratch at age seven. Since then, he has designed equipment operating from DC to microwaves and in both analog and digital domains. He is a member of the IEEE Electromagnetic and Communication societies. He can be reached at P.C.S. Consultants, 4349 Krause St., Pleasanton, CA 94588 or by phone at (510) 426-8056 or by FAX at (510) 846-9058.

A Resistive Attenuator Calculation Program

By Jouni Verronen
Engineering Consultant

Attenuators are frequently needed in RF circuits. This program is aimed at quick calculation of resistive attenuators of the Pi- and T-type. Also, the resistors for an L-pad of minimum attenuation are easily calculated. In addition to the synthesis of basic attenuator circuits, RESATT also provides analysis. One can check either a T-circuit or Pi-circuit using standard resistor values. The analysis program calculates total loss and impedance match at both ports.

This is one of those small calculation programs that I have written for personal use to allow fast solutions to specific, often-encountered, technical design problems. The general requirements for small, quick programs, are discussed in Reference 1.

The program is divided into two parts, synthesis and analysis. Upon starting, the synthesis portion is activated. If the program was used before, there is a data file (RESATT1.DAT) residing in the same directory as RESATT.EXE. This file contains the latest input values, and the program is initialized with those values. New initial values can be selected by hitting an appropriate character. These are listed in a help file, which can be accessed from F1.

The user-supplied values are the resistive terminal impedances and the loss of the attenuator. Every time one of these values is changed the program calculates new resistances for both T and Pi topologies. Figure 1 shows a typical screen. If m is hit, the program calculates the minimum attenuation L-network which can match the two terminal resistances given.

One must calculate the real world losses and impedances when standard resistor values are used. This is done in the analysis part of RESATT. The analysis is activated by hitting the right arrow at the keyboard. Hitting the left arrow returns operation to the synthesis half.

To perform an analysis, first either a Pi- or T-circuit must be chosen by striking p or t. After that, the resistor values can be entered. The analysis program calculates the following items of interest.

- impedance seen towards the attenuator at both ports
- return loss at both ports
- insertion loss

```
Resistive attenuator design
---------------------------------------

Date: 1992-03-18          Time: 02:47:23

-----------------------------------------------------------------

Attenuator synthesis for L =   10.00 dB

T-circuit:

             R1 =  25.97 ohm            R2 =  25.97 ohm

Ro1 =  50.00 ohm          R3 =  35.14 ohm          Ro2 =  50.00 ohm

Pi-circuit:

Ro1 =  50.00 ohm          R3 =  71.15 ohm          Ro2 =  50.00 ohm

             R1 =  96.25 ohm            R2 =  96.25 ohm

-----------------------------------------------------------------

Attenuator analysis for a Pi-type resistor network:

Ro1 =  50.0 ohm           R3 =  68.0 ohm           Ro2 =  50.0 ohm

     ->      R1 = 100.0 ohm            R2 = 100.0 ohm    <-

RL1 = 49.6 dB                                      RL2 = 49.6 dB
Ri1 = 50.3 ohm                                     Ri2 = 50.3 ohm

Calculated loss:          L =   9.63 dB
```

Figure 2. A printed report.

```
           Resistive attenuator design  1.00  by JV

  Synthesis :         Ro1 ≥ Ro2           Analysis :
                                        ---------------------------------
 Ro1 =  75.0 ohm     Ro2 =  50.0 ohm      for T-attenuator:

          L =  5.72 dB                  — Rt1 =      ———————┬——————— Rt2 =  —
                                          39.00 ohm         |         0.00 ohm
 Stroke m for Lmin                                        Rt3 =
                                                          82.00 ohm

 — Rt1 =       ——————┬—————— Rt2 =  —     when:
   43.30 ohm         |        0.00 ohm    Ro1 = 75.0 ohm and Ro2 = 50.0 ohm
                   Rt3 =
                   86.60 ohm              ->                            <-

 ——————┬——————     Rp3 =     ——————┬——————  Ri =  70.06 ohm   Ri =  47.69 ohm
       |           43.30 ohm       |        RL =   29.4 dB    RL =   32.5 dB
   Rp1 =                         Rp2 =
 ******** ohm                    86.60 ohm  Total loss =   5.61 dB

 ——— F1: help ——— F7: exit ——————————————————————————————————————————
```

Figure 1. Typical screen showing both synthesis and analysis portions.

In all these calculations, the ports of the attenuator are terminated with the same resistances (Ro1 and Ro2) which were used in the synthesis part.

For both synthesis and analysis a printed report can be generated. F5 prints the common heading. F6 prints the calculation report for the active side. Figure 2 shows an example of a printed report.

Equations Used

Equations for attenuator synthesis are well known and are not rewritten here, (see References 2 and 3). For analysis, the terminated resistor network was first reduced to one input resistance, for which return loss was then determined. Insertion loss was calculated after that using basic concepts.

Program Versions

The program was written using

QuickBASIC, and the compiled version RESATT.EXE is directly executable under DOS. It requires an EGA or VGA monitor. The source code is also provided in ASCII form (RESATT.BAS). For other monitors it should be modified and then compiled for stand-alone operation.

Conclusions

In these calculations the equations involved are not particularly complicated, but compared to using tables or calculators, this program is quicker, especially as it does both analysis and synthesis; and one gets a printed report, too. ***RF***

This program is available on disk from the RF Design Software Service. See page 79 for ordering information.

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3. Gary A. Breed, "Attenuator Basics", *RF Design*, February 1992.

About the Author

Jouni Verronen is an independent consultant, specializing in RF and analog design. He was recently involved in the design of scanning supervisory equipment, which is intended to control cellular telephone traffic more optimally in the NMT cellular system. He holds a M.Sc. in electrical engineering. He can be reached at Kannokkotie 13, 90550 Oulu, Finland, or by telephone at 358 81 5546790.

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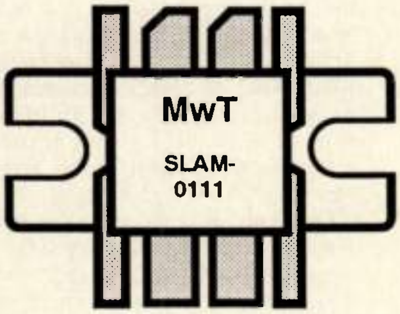

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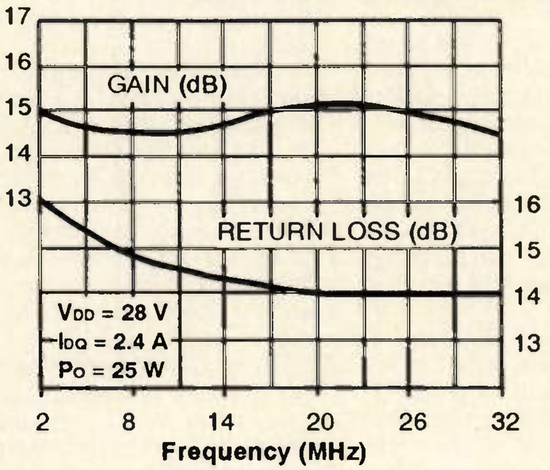

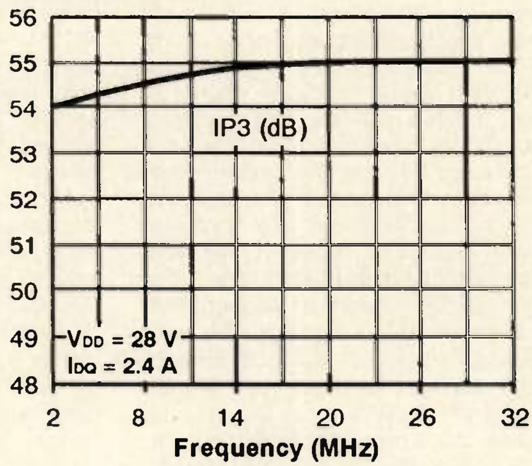

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Tallying Up Counters

By Andy Kellett
Technical Editor

What use are frequency counters? Do you want to tune an oscillator so it is spot on? The frequency counter would be a good choice except that your oscillator is synthesized and crystal referenced; you will be more interested in spurs than the oscillator's minute drifts. For this case, a spectrum analyzer is the tool to use. Suppose you want to look at the output of a VCO as it is quickly swept over its 10 percent bandwidth. You can get accurate readings at the endpoints where the frequency is stationary, but how did the frequency behave as it went between the endpoints? These examples may merely point out a maxim that applies to all test equipment; no instrument can measure everything. However, the question remains: what use are frequency counters? Each different type of counter sold is a different answer to that question.

Fast, precise, full-featured counters are used both in production work and as research and development instruments. In production applications, the most important specifications are normally precision and speed. "Usually when you buy a counter for production it's for a specific reason. It needs to measure this frequency with this many digits of resolution at this rate," says Doug Barker, product manager for John Fluke Manufacturing Co. Buyers for research and development applications look more for flexibility. "The R&D engineer is looking for more capability, because the counter is probably going to be a shared instrument. Somebody may need it for time interval analysis, and the next person will need it to characterize a VCO," says Barker. While performance and features are increasing, prices continue to decline. "It is true for a lot of our product lines that our older instruments cost more, while our newer models do more but cost less," notes Keithly Instruments applications engineer, Dale Cigoy.

The ability to produce a basic frequency counter for less money is part of the reason frequency counters are being included as a feature in other measuring instruments. "As more counter circuits are available on single chips, it becomes very simple and relatively inexpensive to add the functions. So it makes sense in many applications to add the function simply because it doesn't cost much," says B&K Precision product manager, Bob Kral. Audio and low HF measurements can be made with the counters integrated into today's multimeters. Frequency counters built into spectrum analyzers have advantages over stand-alone counters. "The advantage of the spectrum analyzer's counter is that you can be very selective about what you are looking at. The spectrum analyzer/counter forms a frequency selective system, so you can make sure you are counting the signal of interest," notes Tektronix applications engineer, Cliff Morgan.

VXI and PC based instruments represent another branch of the counter family tree. Automatic testing is a growing field which is beginning to incorporate RF measurements. "In a PC, the time it takes to get measurements to the computer, which is really where you want to analyze them, is much shorter than if you have to go over a GPIB," says Shalom Kattan, president of Guide Technology. The computer can supply this processed information on screen, acting as a virtual front panel, or can incorporate the data as part of a system feedback loop.

Increased emphasis on EMI requirements has created new uses for frequency counters. Counters from Optoelectronics are meant for in-field frequency verification of RF communications and part 15 signals, and for counter surveillance. The high quality handheld instruments directly count up to 230 MHz, and possess a high gain front-end for signal input from an antenna. A patented digital filter design prevents the self oscillation which can accompany a high gain design.

As counters have become faster, it has become possible to make frequency measurements of an essentially different type. Modulation domain analyzers measure frequency vs. time. "In a modulation domain analyzer you make rapid frequency or time interval measurements correlated in time, so that you can display frequency vs. time. It's like comparing a voltmeter to a digital scope. A digital scope is really a voltmeter that makes very fast measurements and displays them as a function of time," says Guide's Kattan. The VCO example at the beginning of this report would be handled well by a modulation domain analyzer. Announced at last month's RF Expo West, Guide's PC-card modulation domain analyzer joins an instrument from Hewlett-Packard, a counter and software from John Fluke and a VXI instrument from Racal-Dana.

Manufacturers of basic counters see the counter market slowing down for two reasons. First, other instruments now have frequency measurement capabilities. Second, crystal based circuits have reduced the need to tune circuits. "Because of that [the use of crystal locked oscillators] you see less use of frequency counters, but some of that is taken up by timer counters," says Arlene Meadows, product marketing manager for Racal Dana.

Portions of the market do show strength however. "At Racal-Dana, our VXI business is increasing," says Meadows. Fluke's Barker noted, "Our counter business hasn't been doing too badly. Fluke/Philips is building market share in the counter arena." For those manufacturers who have opened a new part of the market, such as Guide Technology and Optoelectronics, business is expanding.

The recession slowed the entire market, and recent upswings have affected everyone as well. The thing that several counter manufacturers expressed concern over is the changing tax structure and how that will affect their companies, and their companies' customers.

Counters continue to develop rapidly, and developments from the most expensive, state-of-the-art counter systems quickly find use in other segments of the counter industry. Hewlett-Packard product marketing engineer Dave Cunningham comments on how modulation domain analysis technology has found use in the counters he works with, "We've taken what we've learned from these more advanced products, and now we can take that same technology, and because we know how to use it, we can use it a lot less expensively in a product like our lower cost counters." ***RF***

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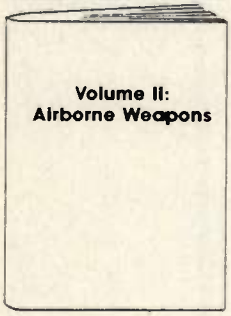

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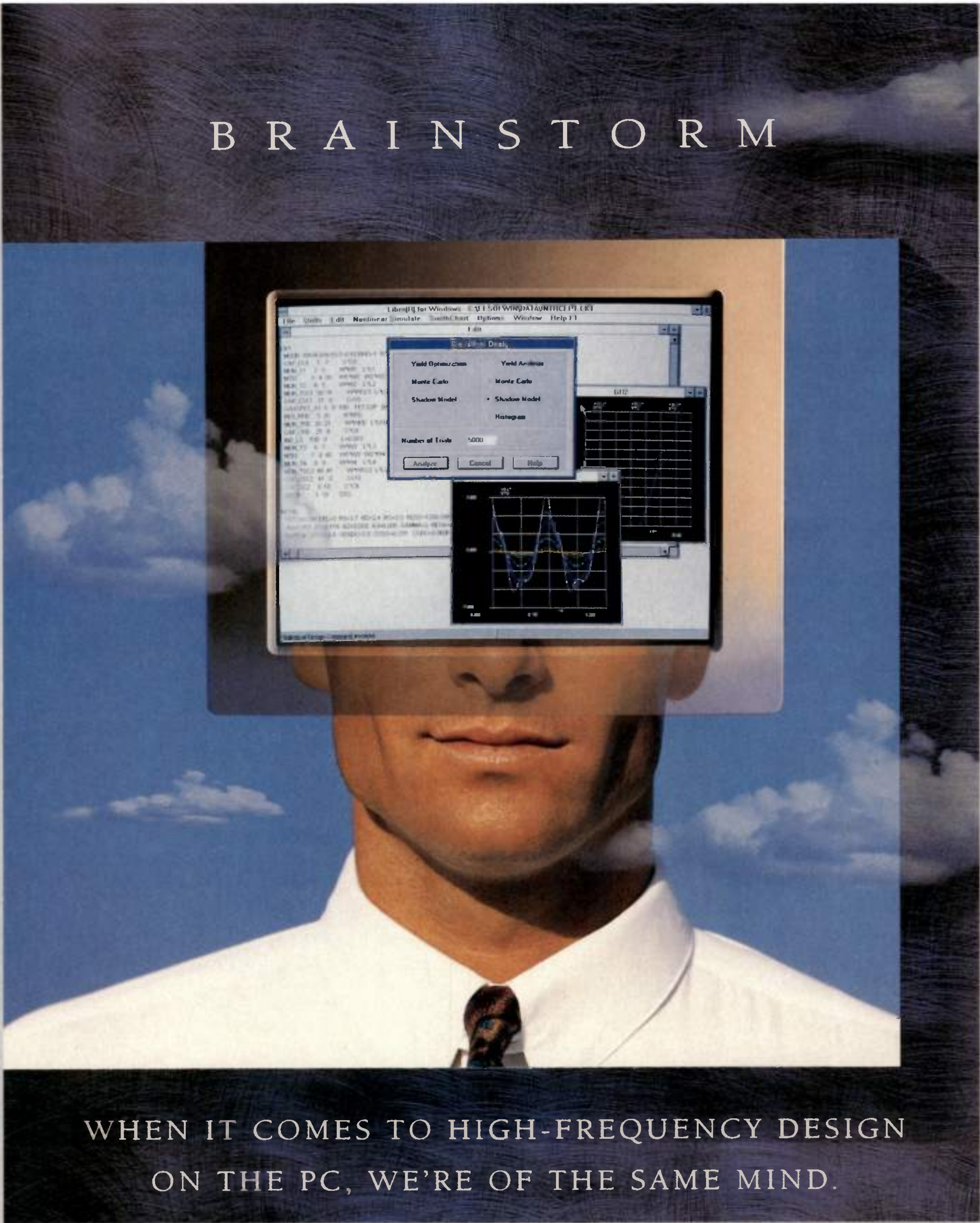

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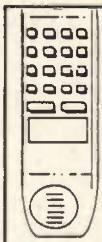
AudioLogic is an exciting new company designing innovative DSP based products for the hearing impaired, located near the Rocky Mountains. Our products require creative solutions in the areas of low power, miniature, battery operated RF communications. Appropriate experience would be 5 years working with consumer or miniature RF products, such as pagers or cellular. If you seek a technical challenge, competitive salary, equity participation, a stimulating work environment and would like to help society, send your resume and salary history to John Melanson, *AudioLogic*, 1355 S. Boulder Rd #F137, Louisville, CO 80027. EOE.



...YOUR CAREER

Cellular Engineers: Design/develop RF and analog circuits for high capacity cellular systems. Requires minimum of 2 years experience in any of the following: DSP, ASIC Design, CAE Development, Digital Modulation, Digital Mobile Communications, Channel Equalizers, Transmitter-Receiver-Synthesizer or Audio Design, Digital Signal Processing.

Sr. RF Design Engineer: You will be responsible for the design and development of 900 MHz RF circuitry. Also, you will provide support throughout the development cycle. A BSEE with emphasis on analog/RF design techniques and at least 6 years of experience designing RF subsystems are essential. A background in RF circuit design, including RF amplifiers (low and high power), synthesizers, VCO, VCXO and PLL is necessary. In addition, you need knowledge of modern communication systems theory.

Whatever you are seeking in it...

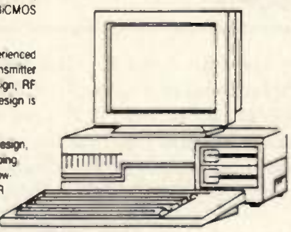
MMIC Engineer: Develop L/S band GAA's MMIC power amplifiers for commercial wireless communications. Requires M.S. or BSEE, + 2 years experience with GAA's MMIC design, simulation, packaging and test.

Amplifier Designers: Both Senior and Junior positions exist for designer with experience in Microstrip/Stripline, Hybrid Mic, MMIC Amplifiers. Markets involved will be cellular, wireless and mobile communications. BS/MSEE.

RF Design Engineer: Responsible for design of analog and RF systems and circuits for consumer and commercial digital wireless products. Five to ten years experience in RF systems analysis and design. Experience with low-cost design techniques for frequency synthesizers, power amplifiers, up/down converters and baseband circuits for digital communications systems. Must be able to derive RF systems and module requirements to meet overall performance and cost goals. Familiarity with time division duplex or CDMA a plus.

Staff/Principal Engineer: This individual will be responsible for the design/development of UHF/VHF amplifiers. Will be required to control the design on a stand-alone basis while meeting cost and schedule requirements with the support from junior engineers and experienced technicians. Hands-on engineer will also be required to lead IR&D efforts in UHF/VHF RF Power Amplifier design to enhance transmitter technology position. BSEE/MSEE.

RFIC Design: MS or PhD in Electrical Engineering with minimum 5 years related experience is preferred. The candidate should have a good knowledge and experience in Linear Bipolar High Frequency IC design and measurement techniques to design IC's like Amplifiers, Mixers, Oscillators, VCO's, Prescalers, Synthesizers, Limiting Amplifiers, etc. operating up to 2 GHz in Bipolar or BiCMOS technologies.



RF/Synthesier Design: Positions available for candidates experienced in the design and development of communication receiver/transmitter systems and subsystems. Experience in RF synthesizer design, RF circuitry, low power circuit design and GaAs/Silicon MMIC design is desirable. At least a BS degree in Engineering is required.

Wireless LANS: Responsible for the electrical and mechanical design, manufacture and production of a 2.4-2.5 GHz frequency hopping spread spectrum radio transceiver. Designs include power & low-noise amplifiers, filters, mixers, VCO, synthesizer, doubler, T/R switch and antennas. BS/MSEE.

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CREATING A WIRELESS WORLD

RF DESIGN ENGINEER

Motorola is the source of exciting innovation in cellular and wireless communications. Our engineers are utilizing the latest RF technologies, working on tomorrow's products; redefining the way the world communicates.

As a team member of a group of engineers, you will work on product design and development of portable cellular telephones for the Western European Market. You will be involved in multiple projects at various stages of completion-from initial concept through production. Responsibilities include: project engineering, cost estimation, proposal development, RF and analog design, as well as circuit design and testing.

The qualified candidate will possess a BSEE, MS preferred; and have 2+ years experience in RF transceiver design, design tools and development. Knowledge of receivers, transmitters or synthesizers, 450 to 900 MHz is strongly preferred. Familiarity with high frequency filter, RF power amplifier, antenna design, surface mount devices and microstrip/stripline applications is required. This position is based in Libertyville, Illinois; 40 miles northwest of Chicago.

We offer an excellent salary, comprehensive benefits package and opportunities for professional growth. For immediate consideration, please send your resume and salary history to: Supervisor, Professional Recruitment, Dept. RF/493-TH, Motorola, Cellular, 1501 W. Shure Drive, Arlington Heights, IL 60004. Or FAX your resume to our Resumix FAX line: (708) 632-7382.

Motorola is an equal opportunity/affirmative action employer. We welcome and encourage diversity in our workforce.

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RF literature

EMI/RFI Filters
Murata Erie North America's line of EMI/RFI filter devices is covered in detail in their new EMI/RFI Filter Catalog No. E-06-A. The 97-page catalog is free of charge.
Murata Erie North America
INFO/CARD #180

Coils and Filters
Toko America's 164-page catalog of coils and filters covers Toko's complete lines of coils and balun transformers, as well as LC, dielectric, SAW, ceramic, helical and active filters.
Toko America, Inc.
INFO/CARD #179

Sigma-Delta A/D
The design of a 16-bit sigma-delta analog to digital converter using the Comdisco Systems' Signal Processing WorkSystem (SPW) is described in a six-page application note available free of charge from the company.
Comdisco Systems, Inc.
INFO/CARD #178

Digital Signal Processing
The latest issue of DSPatch from Analog Devices includes features about the ADSP-2115 DSP chip, use of the ADSP-2101 chip in a video compression application, questions and answers, tips, and other information concerning the use of Analog Devices' DSP devices.
Analog Devices
INFO/CARD #177

Quartz Products
A full-color brochure from Tele Quarz describes the company's design and manufacturing capabilities for the production of quartz crystals, filters and oscillators.
Tele Quarz USA
INFO/CARD #176

Telecom Reference
Butterworth-Heinemann has recently published the *Telecommunications Engineer's Reference Book*. The 62-chapter book was edited by F.F. Mazda, and covers mathematical techniques, physical phenomena, communications principles, and applications. Cost of the 1,000 page book is $125.00.
Butterworth-Heinemann
INFO/CARD #175

VXI Catalog
Hewlett-Packard's 1993 VXI catalog includes a high-power 486-embedded controller running MS-DOS and supporting compiled SCPI, a family of high performance A to D converters, a Baseband Signal Analyzer System using HP's new VXI Digital Signal Processor, and a High Speed Data Acquisition System.
Hewlett-Packard Co.
INFO/CARD #174

Receiving Multicouplers
AML announces a fully revised and expanded edition of its Receiving Multicoupler Handbook. The new edition contains applications notes and specs for 50 multicouplers from 10 kHz to 2.4 GHz.
AML, Inc.
INFO/CARD #173

ISO 9000 Training Materials
Six manuals and three videos which cover ISO 9000 related topics are offered by PPM Associates. Guidelines for PC design and assembly, surface mount techniques and design for manufacturability are among the topics covered in the manuals and videos.
PPM Associates, Inc.
INFO/CARD #172

Microwave Oscillators
Forty-six microwave oscillators are described in a information packet from Electronic Surveillance Components. Included are voltage controlled oscillators in the 100 to 3500 MHz range that feature low-cost pin output and surface mount versions.
Electronic Surveillance Components, Inc.
INFO/CARD #171

Shielding Data
Now available is a compilation of shielding quality data on Spira's entire line of EMI gaskets and honeycomb filters.
Spira Mfg. Corp.
INFO/CARD #170

Cable Assembly Guide
C.E. Precision's guide to specifying flexible, semi-rigid, silicon dioxide, and other coaxial cable assemblies defines what the design engineer needs to consider in specifying coaxial cables for their application.
C.E. Precision Assemblies
INFO/CARD #169

VXI/VME Guide
Analogic announces a free, 74-page "Designer's Guide to High Performance VXI/VME Systems," featuring Analogic's precision VXI instruments, software support, technical notes, and a family of VME-based signal processing boards.
Analogic Corp.
INFO/CARD #168

Mobile Data News
Dataradio has initiated the quarterly publication of its *VIS TIMES*. *VIS TIMES* is a newsletter focusing on products, applications and related information on private mobile data networks using Dataradio's VIS (Vehicular Information Series) open architecture radio modem products.
Dataradio Corp.
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