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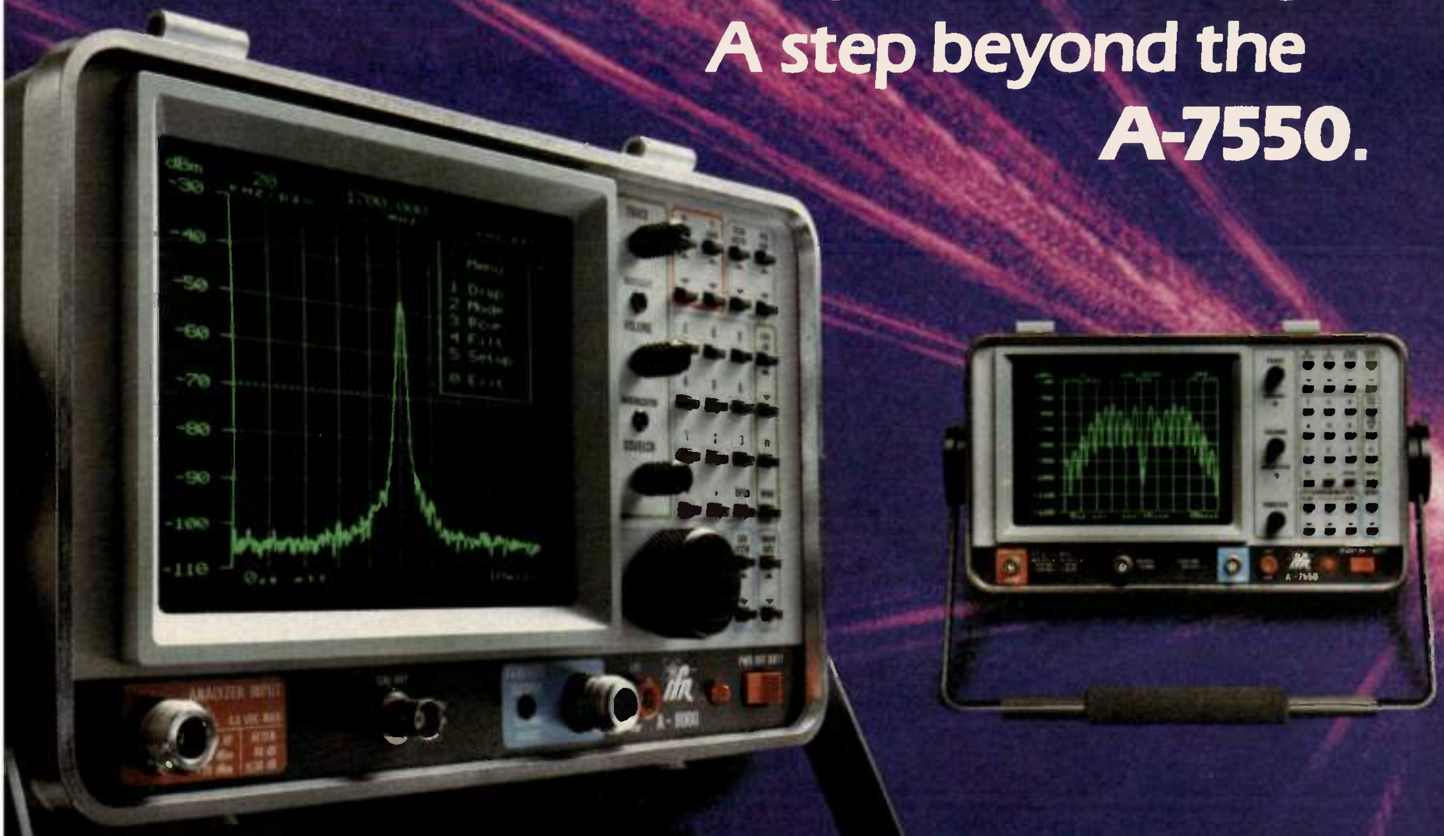

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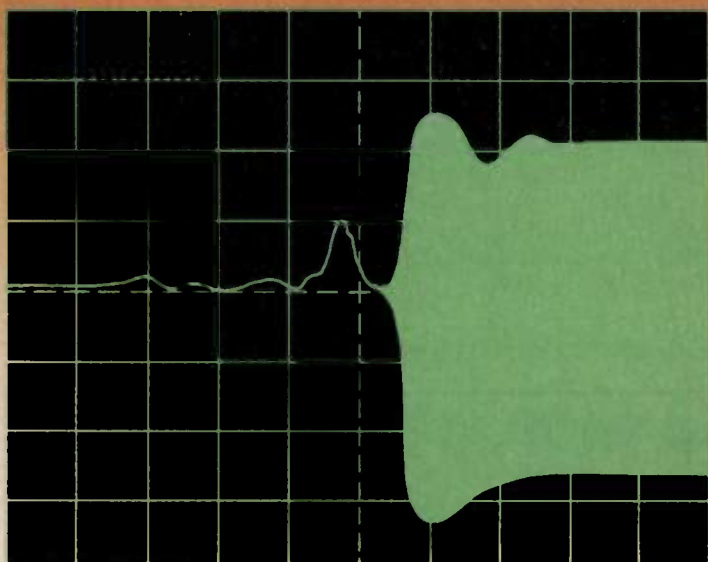

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RF design

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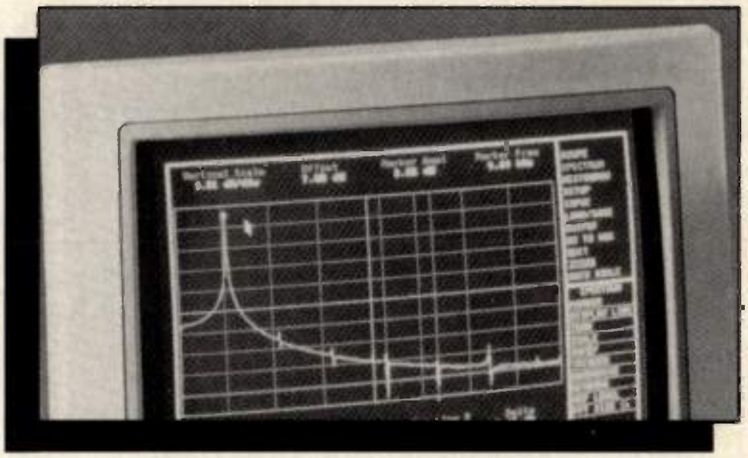

Our cover, provided by Burr-Brown, represents increasing use of digital technology in RF applications. In the photo is the ZPA1000 Dynamic Signal Analysis software for digitized signals. Available digitizers and buffers will handle signals up to 10 MHz.

R.F. DESIGN (ISSN: 0163-321X USPS: 453-490) is published monthly plus one extra issue in September. April 1991. Vol. 14, No. 4. Copyright 1991 by Cardiff Publishing Company, a subsidiary of Argus Press Holdings, Inc., 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111 (303) 220-0600. Contents may not be reproduced in any form without written permission. Second-Class Postage paid at Englewood, CO and at additional mailing offices. Subscription office: *RF Design*, 5615 W. Cermak Rd., Cicero, IL 60650. Domestic subscriptions are sent free to qualified individuals responsible for the design and development of communications equipment. Other subscriptions are: $38 per year in the United States; $48 per year in Canada and Mexico; $52 (surface mail) per year for foreign countries. Additional cost for first class mailing. Payment must be made in U.S. funds and accompany request. If available, single copies and back issues are $5.00 each (in the U.S.). This publication is available on microfilm/fiche from University Microfilms International, 300 Zeeb Road, Ann Arbor, MI 48106 USA (313) 761-4700.

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RF editorial

Engineering Productivity Part I: The Team Concept

By Gary A. Breed
Editor

What does it take to be successful in an increasingly competitive marketplace? Usually, it means getting more done with less — fewer employees, lower manufacturing costs, lower overhead, plus fewer returned goods and service calls. In the next few months, I'll be commenting on some of the techniques that have been developed to meet these challenges, identifying the role of the engineer in each of them.

To meet the overall goal of increased productivity, many companies are adopting new operating methods. You have probably heard some of the terminology these methods use: world class, just in time, design for manufacture, integrated management, and total quality management. All these concepts have the single goal of creating a quality product without wasted time or material.

What these methods feature is involvement from start to finish by participants in the product's design, manufacture, costing, sale and service. Perhaps a management committee has been formed with representatives from the different departments. Or, an individual project manager might oversee cross-checking and review by the various departments. In either situation, this process should allow operating features and performance to be optimized for customer requirements, serviceability, and the most economical manufacturing processes. Putting this concept into practice can be extremely difficult, because it usually means a major change in the way a company operates.

I have to note that these aren't really new ideas. Consider a very small company made up of a few partners who want to develop an idea into a product. Every decision is made by the partners, who represent every company function. Design tradeoffs are examined, costs evaluated, manufacturing methods explored, and plans are developed based on the group's consensus.

Implementing this cooperative process on a larger scale is the new part, and it presents some problems for those involved (such as design engineers). First, in a large company the participants do not have the urgency of a financial commitment like entrepreneurial partners. Also, there may be difficulty adjusting after many years of working on pieces of projects with no attention to overall goals. Another common problem is that each department feels it has to protect its "turf."

These problem areas point out how hard it is to make fundamental changes in the way a company operates. However, these changes will continue to take place, mandated by the need to achieve the highest possible efficiency. We can't (and in most cases, shouldn't) fight these changes, so we must adapt. In the next couple of editorials, I'll comment on some of the areas where engineers will find changes. I will also point out some areas where engineers can benefit significantly from their employers' commitments to improve competitiveness.

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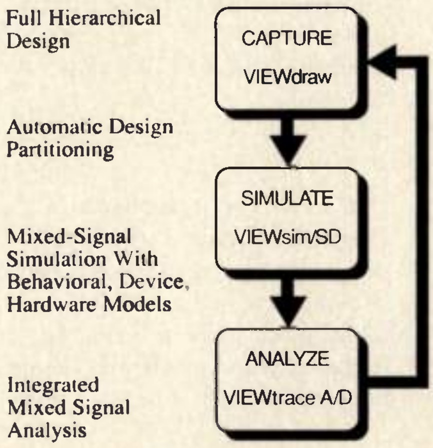

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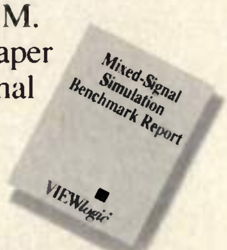

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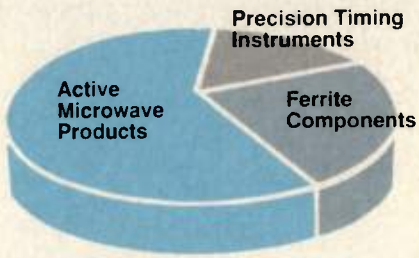

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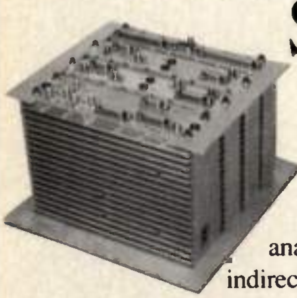

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RF letters

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

Driftless VCO Comments

Editor:

The "Linear Driftless VCO" in the January 1991 issue of *RF Design* has a fatal flaw. The "frequency comparator" is a linear circuit and cannot actually compare frequency. Instead it compares the average DC levels of two variable duty cycle pulse streams.

If the 74HC74 flip-flops used as one shots and 4016s connected to the Q outputs are replaced with a true digital frequency comparator, such as phase comparator 2 in the 74HC4046, the circuit should work as described. This should bring the frequency drift down from the 5.2 kHz measured by the author to within calculated limits.

Jon Nelson
Uniplex Corporation
St. Paul, MN

The Author's Reply

Editor:

The frequency comparator used, as said by Mr. Nelson, "compares the average DC levels of two variable duty cycle pulse streams," but each pulse stream has as many pulses as the input frequency, so its duty cycle is proportional to the input frequency and the output will have an average perfectly proportional to frequency difference (with an offset of $V_{cc}/2$).

Actually, it does compare frequency:

$$AVG(V_{out}) = k(f_1 - f_2) + V_{cc}/2$$

As stated in the article, the design needs a frequency detector that keeps within the linear region of operation at any input frequency, to produce an error signal proportional to the difference of frequencies. Thus, a mean can be extracted and the VCO will remain locked on that mean. The suggestion to use a type 2 phase comparator such as the 4046 will fail to work.

At any instant the frequency comparator is dealing with two different frequencies, because the reference is switching between two different values and the prescaled VCO (assumed as stable) will be in between. A type 2 phase comparator is within the linear region of operation only at equal input frequencies. At different input frequencies, the phase comparator will usually saturate.

In a conventional PLL this will allow to search for the lock. In our case it produces a square wave that is a replica (but not exactly equal) of the multiplexing signal, because we are switching between one frequency above and one below the prescaled one. This way the circuit seems to work but in fact it is not locked. This circuit with a 4046 was tested in early stages of development, without success.

Luis Cupido
C&TC
Aveiro, Portugal

Comb Generator Source

Editor:

I was very interested in the article which appeared in the January issue's EMC Corner, "RFI Measurements Using a Harmonic Comb Generator."

EMCO has been manufacturing an almost identical product for two years. The product idea came from a paper delivered by Scott Roleson at the 1987 IEEE Symposium, describing a device designed by Bill Royce. As were the authors of your article, Scott and Bill are employees of Hewlett-Packard. The EMCO device is designated the Model 4610 Royce Field Site Source.

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Glen Watkins
The Electro-Mechanics Co. (EMCO)
Austin, TX

Two-Channel Receiver Correction

An alpha (α) was omitted from Figure 1 in "A Simple Two-Channel Receiver for 8-PSK," February, 1991, p. 29. The corrected figure is shown below. In addition, in Figure 3, an epsilon (Σ) was omitted from the summation circle on the right side of the figure.

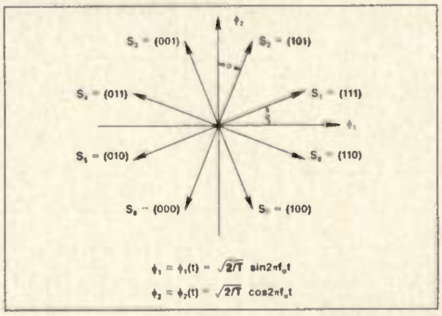

Figure 1. Signal space diagram showing gray level coding of symbols.

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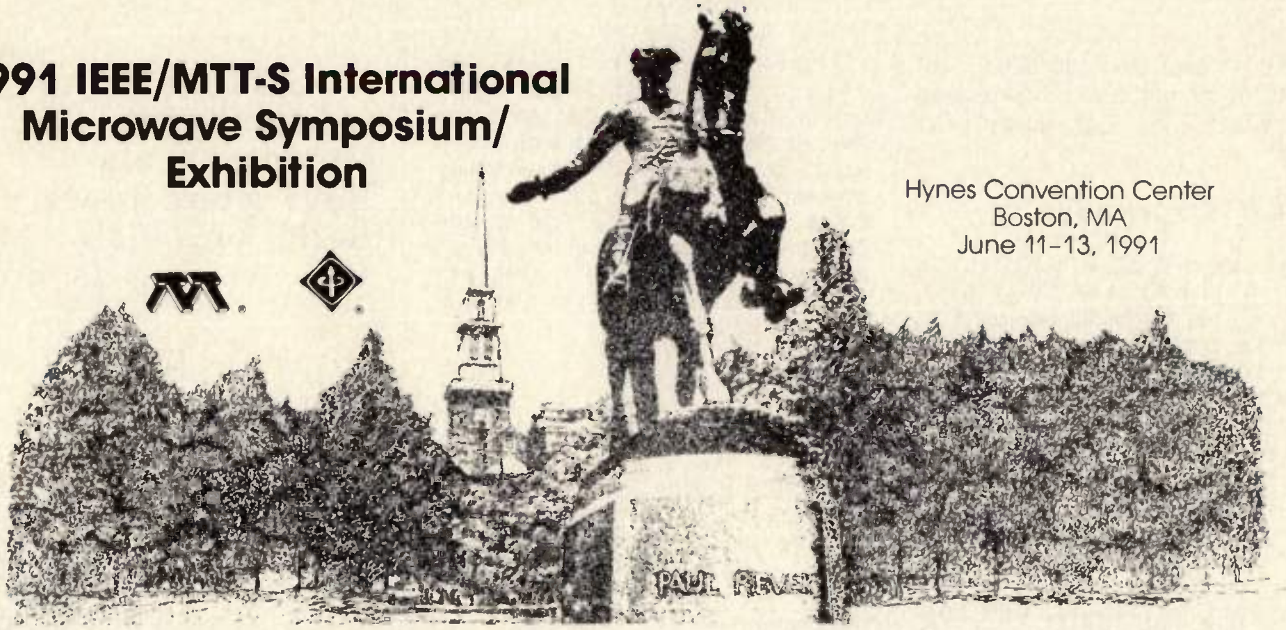

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Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616.

15-18 **Seventh International Conference on Antennas and Propagation**
University of York, UK
Information: Conference Services, IEE, Savoy Place, London, England WC2R 0BL. Tel: 44 071 240-2871 ext 222.

15-18 **NAB '91**
Las Vegas Convention Center, Las Vegas, NV
Information: NAB, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (202) 429-5350.

16-18 **Electro/International**
Jacob K. Javits Convention Center, New York, NY
Information: Electronic Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045-3194. Tel: (213) 772-2965.

28-1 **The 1991 MIC Workshop**
Kingsmill Resort & Conference Center, Williamsburg, VA
Information: Arlon, Microwave Materials Division, 1100 Governor Lea Road, Bear, DE 19701. Tel: (800) 635-9333.

May

13-15 **41st Electronic Components and Technology Conference**
Westin Peachtree Plaza, Atlanta, GA
Information: Electronic Industries Association. Tel: (202) 457-4930.

14-16 **IEEE Instrumentation and Measurement Technology Conference**
Omni Hotel, Atlanta, GA
Information: Robert Myers, 3685 Motor Avenue, Ste. 240, Los Angeles, CA 90034. Tel: (213)-1463. Fax (213) 287-1851.

29-31 **45th Annual Symposium on Frequency Control**
Los Angeles Airport Marriott, Los Angeles, CA
Information: Annual Symposium on Frequency Control, PO Box 826, Belmar, NJ 07719.

June

3-5 **First Virginia Tech Symposium on Wireless Personal Communications**
Virginia Tech, Blacksburg, VA
Information: Prof. Theodore S. Rappaport, Director, Mobile & Portable Radio Research Group, Virginia Polytechnic & State University, Blacksburg, VA 24061-0111.

11-13 **MTTS International Microwave Symposium and Expo**
Boston, MA
Information: Tel: (617) 769-9750.

13-14 **ARFTG**
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RF courses

ESD - Electrostatic Discharge
April 17-19, 1991, Madison, WI
Information: University of Wisconsin - Madison, Department of Engineering Professional Development. Tel: (608) 262-2061. Fax: (608) 263-3160.

Fundamentals of Radar Cross Section
May 20-24, 1991, San Diego, CA
Information: Kelly Brown, SCEEE, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146.

Microwave/Millimeter Wave Monolithic Integrated Circuits
June 4-7, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Infrared/Visible Signature Suppression
April 16-19, 1991, Atlanta, GA
Information: Education Extension, Georgia Institute of Technology. Tel: (404) 894-2547.

Analog/RF Fiber-Optic Communications
April 24-26, 1991, Washington, DC
Microwave Systems Engineering
April 29-May 3, 1991, Washington, DC
Vulnerability of Spread Spectrum AJ and LPE Communications Systems
May 6-9, 1991, Washington, DC
Nonlinear Digital Signal Processing and Applications
May 6-10, 1991, Washington, DC
Electromagnetic Interference and Control
May 6-10, 1991, Washington, DC
Frequency Hopping Signals and Systems
May 20-22, 1991, Washington, DC
New HF Communications Technology: Advanced Techniques
June 3-7, 1991, Washington, DC
Spread Spectrum Communications Systems
June 10-14, 1991, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Fiber Optic Communications
June 10-12, 1991, Sunnyvale, CA
Information: Center for Professional Development, Arizona State University. Tel: (602) 965-1740.

Digital Signal Processing Workshop
June 12-14, 1991, Norwood, MA
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Design Seminar: Principles of EMC
May 7-9, 1991, Mariposa, CA
Information: CKC Laboratories, Registrar. Tel: (209) 966-5240. Fax: (209) 742-6133.

Worst Case Circuit Analysis
May 13-15, 1991, Washington, DC
Information: Design and Evaluation, Inc. Tel: (609) 228-3800.

Academy: Schematic/Layout
April 16-19, 1991, Westlake Village, CA
May 7-10, 1991, Baltimore, MD
Touchstone/Academy
April 15-19, 1991, Baltimore, MD
Libra/Academy
April 22-26, 1991, Westlake Village, CA
April 22-26, 1991, Gilching, Germany
Information: EEsof. Tel: (818) 991-7530. Fax: (818) 991-7109.

RF and Microwave Design II: Non-Linear Circuits
April 22-26, 1991, Pisa, Italy
Digital Microwave Systems: Theory and Applications
April 22-25, 1991, Pisa, Italy
Satellite Communication and Broadcasting
April 22-26, 1991, Italy
Frequency-Time and Spatial-Time Signal Processing
June 10-14, 1991, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

RF/MW Linear/Nonlinear Circuits and Applications
May 2-8, 1991, Washington, DC
Information: Besser Associates. Tel: (415) 949-3300, Fax: (415) 949-4400.

Modern Power Conversion Design Techniques
April 29-May 3, 1991, Phoenix, AZ
May 20-21, 1991, San Rafael, CA
Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

EW Receivers
May 7-9, 1991, Syracuse, NY
ELINT Analysis
May 7-9, 1991, Syracuse, NY
ELINT Interception
May 14-16, 1991, Syracuse, NY
ELINT/EW Applications of Digital Signal Processing
May 14-16, 1991, Syracuse, NY
Integrated EW
May 21-22, 1991, Syracuse, NY
ELINT/EW Data Bases
May 21-23, 1991, Syracuse, NY
Radar Vulnerability to Jamming
June 4-5, 1991, Syracuse, NY
Electromagnetic Propagation
June 4-6, 1991, Syracuse, NY
Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Basic Network Measurements Using the 8510B Network Analyzer
April 30-May 2, 1991, Los Angeles, CA
May 14-16, 1991, Boston, MA
May 29-31, 1991, Los Angeles, CA
Microwave Fundamentals
May 14-17, 1991, Los Angeles, CA
Information: Hewlett-Packard Company. Tel: (714) 999-6700.

Design for EMC
April 22-23, 1991, Philadelphia, PA
Practical EMC Retrofits
April 24-26, 1991, Philadelphia, PA
Information: R&B Enterprises. Tel: (215) 825-1960.

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INFO/CARD 12

45th Annual Symposium on Frequency Control

This year's Frequency Control Symposium, co-sponsored by the IEEE Ultrasonics, Ferroelectronics and Frequency Control Society and the U.S. Army Laboratory Command, Electronics Technology and Devices Laboratory, is being held May 29-31 at the Los Angeles Airport Marriott, Calif. The symposium serves as a leading international forum on all aspects of frequency control and precision timekeeping.

Major areas of interest include fundamental properties of piezoelectric crystals, theory, design and processing of piezoelectric resonators and filters, stable oscillators and synthesizers, atomic and molecular frequency standards, frequency and time coordination and distribution. Special sessions have been scheduled this year for the increasing role of optics and lasers in frequency standards, and on high temperature superconductors.

The day before the start of the symposium, seven tutorials will presented including: Introduction to Phase Noise, Introductions to Quartz Frequency Standards, and Introduction to Frequency Synthesis. For information contact: Clark Wardrip, Bendix Field Engineering Corporation, PO Box 6147, Vandenberg, AFB, CA 93437. Tel: (805) 865-3214.

International Telecommunications Union Starts On-Line Computer Based Communications Services — The ITU recently announced the implementation of an on-line telecommunications related information exchange. The system will offer such facilities as electronic mail, bulletin boards, document interchange, computer conferencing, distributed access to ITU databases and notification of telecommunication information including a terminology infobase of 30,000 telecommunications terms in English, French and Spanish. The use of TIES (the ITU Telecom Information Exchange Services) is basically free of charge. The system is accessible through public telecommunications networks and is available 24 hours a day, seven days a week. For further information on TIES contact: Mr. L. Goelzer, Chief, Computer Department, ITU. Tel: (41) 22 730 5333. Fax: (41) 22 730 5337.

Wavetek to Enter Merger — Wavetek Corporation has announced that its Board of Directors has entered into an agreement in principle with Torrey Investments Inc., whereby Torrey Investments would acquire Wavetek by merger. in which the stockholders of Wavetek would receive $3.15 per share in cash. The value of the transaction is approximately $28,300,000. The merger is subject to execution of a definitive agreement and its approval by the Board of Directors of both corporations. In addition, the proposed merger is conditional upon Torrey Investments obtaining certain financing.

ASEP Guides Electroless Plating on Plastics Standards — The American Society of Electroplated Plastics (ASEP) will hold two half-day electroless plating of plastics roundtable discussions April 23 and 25 respectively, in Boston and San Jose. Specifiers, applicators and industry suppliers will want to attend the discussions to establish standards for evaluating the performance of electroless plating on plastics for EMI shielding. Topics will include: electrical performance, shielding testing, adhesion tests, accelerated environmental aging methods, durability tests and appearance guidelines. For more information call ASEP headquarters at (202) 371-1323 or Fax (202) 371-1090.

Microstrip Patch Antenna Developed — NIST has recently come up with a new development that makes certain antenna measurements and electromagnetic interference/compatibility tests more convenient. NIST researchers have developed a small (20 cm square) microstrip patch antenna that can be used as a standard transmitting and receiving antenna at frequencies below 500 MHz in an anechoic chamber. Researchers proved that the resonant frequency, driving point impedance, antenna radiation pattern and radiated field strength of the microstrip patch antenna could be calculated theoretically from its geometry and are accurate to within three percent. Paper no. 2-91 describes the antenna in detail and is available from Jo Emery, Div. 104, NIST, Boulder, CO 80303. Tel: (303) 497-3237.

Piezoelectric Devices Call for Papers — The 13th Piezoelectric Devices Conference to be held October 9-11, 1991 has issued a call for papers. Papers may be tutorial or application-oriented, dealing with recent progress in design, development, processing, or manufacturing control in areas represented by the following topics: properties of natural and cultured quartz, design of quartz resonators, oscillator and filter design, CAD/CAM, manufacturing and process control, measuring and test techniques, packaging and hybrid techniques, quality and reliability assurance, and SAW devices and applications. An abstract of the proposed paper clearly describing the content, scope, organization key points and presentation time is required for paper selection and session assignment. Mail three copies of the abstract with a phone number by May 6, 1991 to: Components Group, Electronic Industries Association, 2001 Pennsylvania Ave., N.W., Washington, DC 20006.

Scholarships Available for Radio Amateurs — The Foundation for Amateur Radio, Inc., plans to awards thirty-six scholarships for the academic year 1991-1992 to assist licensed Radio Amateurs. Licensed Radio Amateurs may compete for these awards if they plan to pursue a full-time course of studies beyond high school and are enrolled in or have been accepted for enrollment at an accredited university, college or technical school. The awards range from \$500 to \$2000 with preference given in some cases to residents of specified geographical areas or the pursuit of certain study programs. Additional information and an application form can be requested by letter or QSL card, postmarked prior to May 31, 1991 from: FAR Scholarships, 6903 Rhode Island Avenue, College Park, MD 20740.

Exhibitors Set for NAB '91 Convention — The National Association of Broadcasters has announced that more than 700 exhibitors will participate at its annual convention in Las Vegas, April 15-18. The NAB convention will offer exhibitors more than 430,000 square feet of show space at the Las Vegas Convention Center. The Engineering Conference is scheduled for April 14-18. In addition, NAB is sponsoring the HDTV World Conference & Exhibition during the same dates.

Motorola and DSC to Develop Open Cellular System Standards — Motorola, Inc. and DSC Communications Corp. have announced that they have reached an agreement in principle to cooperate in developing, in connection with appropriate standards bodies, open cellular system interface standards for the worldwide cellular market. The standards will be based on existing international switching and cellular standards such as CCS7, GSM, ISO and ISDN. Under the proposed agreement, Motorola and DSC will continue, on a non-exclusive basis, to market, enhance and support cellular systems worldwide. The agreement in principle, which is subject to preparation of definitive agreements, contemplates modifying the existing agreement between the two companies in the area of exclusivity, cross licensing of certain technology, marketing and manufacturing rights, and pricing and payment terms.

TriQuint Semiconductor and GigaBit Logic Merge — TriQuint Semiconductor and GigaBit Logic recently announced the signing of a letter of intent that will allow the two companies to be merged into one corporation. The new company will be jointly owned by the existing investors. The name of the new company will be TriQuint Semiconductor Inc. A distinct digital product line, which will include digital standard products and ASICs, will be marketed under the GigaBit Logic name. Until the definitive agreement is signed and conditions for the merger are met, the two companies will continue to operate independently. The merger is expected to be completed by the end of the first quarter of 1991.

Report Available on HDTV and Related Systems — A report released by the Office of Technology Assessment (OTA), *The Big Picture: HDTV & High Resolution Systems*, examines the scope of recent major technological advances and how the United States might benefit. The report evaluates the potential influence high definition television (HDTV) and related high resolution systems (HRS) may have in the consumer entertainment industry, and how they may affect communications, national security, research and education. The report also focuses on how HDTV and HRS are causing the U.S. to re-examine current national policies dealing with manufacturing, educational and training standardization, communications, military command, structural economic problems and the relationship between government and business. Finally, the report discusses the importance HDTV holds for the U.S. in rebuilding its leadership role in global and domestic electronic technology markets. The report, stock number 052-003-

01193-0 is available for $5.00. To order, send prepayment to Dept. 36-IP, Superintendent of Documents, Washington, DC 20402-9325. To order with VISA or MasterCard phone (202) 783-3238.

NEC and CEL Expanding Design Center — The NEC Corporation and California Eastern Labs have announced plans to expand the NEC/CEL Design Center to include development of silicon MMIC devices. The program will concentrate on the development of both standard products and customer specific devices based on NEC's proprietary DNP III direct nitride passivation process.

Burr-Brown Offering Direct Marketing Services — In an effort to streamline the entire product selection and order placement process, Burr-Brown has established a direct marketing organization within its Customer Service Center. Customers can now call toll-free and receive real-time applications assistance, request and receive one-hour fax service on product literature, receive price and delivery quotations, and place their orders over the phone charging their purchases to VISA, MasterCard or Discover Cards. To implement the process, Burr-Brown has installed a toll-free number (800) 548-6132 and a telefax capability (602) 741-3895.

Symposium Dates Changed — The Third International Symposium on Recent Advances in Microwave Technology is now scheduled from August 18-21, 1991. The symposium will cover all the topics in Microwave Technology and its applications including components, circuits, antenna and radar, MICs and MMICs, remote sensing, biological effects, communication systems, CAD techniques, and propagation and measurements among others. For information on the symposium contact: Prof. Banmali Rawat, Dept. of Electrical Engineering, University of Nevada - Reno, Reno, Nevada 89557-0030. Tel: (702) 784-6927. Fax: (702) 784-1300.

Viewsonics Announces Move — Viewsonics recently announced that they will be moving to Florida from their present New York location. Their new address will be: Viewsonics, Inc., 6454 E. Rogers Circle, Boca Raton, FL 33487. Tel: (407) 998-9594. Fax: (407) 998-3712. The toll free number will remain (800) 645-7600.

13th Annual Antenna Measurement Techniques Symposium Call for Papers — The AMTA has issued a call for papers for the October 7-11 symposium. Topics include, advanced antenna measurement techniques, practical aspects of measurement equipment, systems and equipment interfacing, theory and applications of measurement techniques, range, design, automation and evaluation, near-field techniques, anechoic chamber and absorber design and evaluation, phased array testing, and compact range design and evaluation. Authors should submit a 200 word abstract along with the contact author's phone and address by May 10, 1991 to: Prof. Roger C. Rudduck, AMTA Technical Coordinator, Electroscience Laboratory, Ohio State University, 1320 Kinnear Road, Columbus, OH 43212. Tel: (614) 292-6113.

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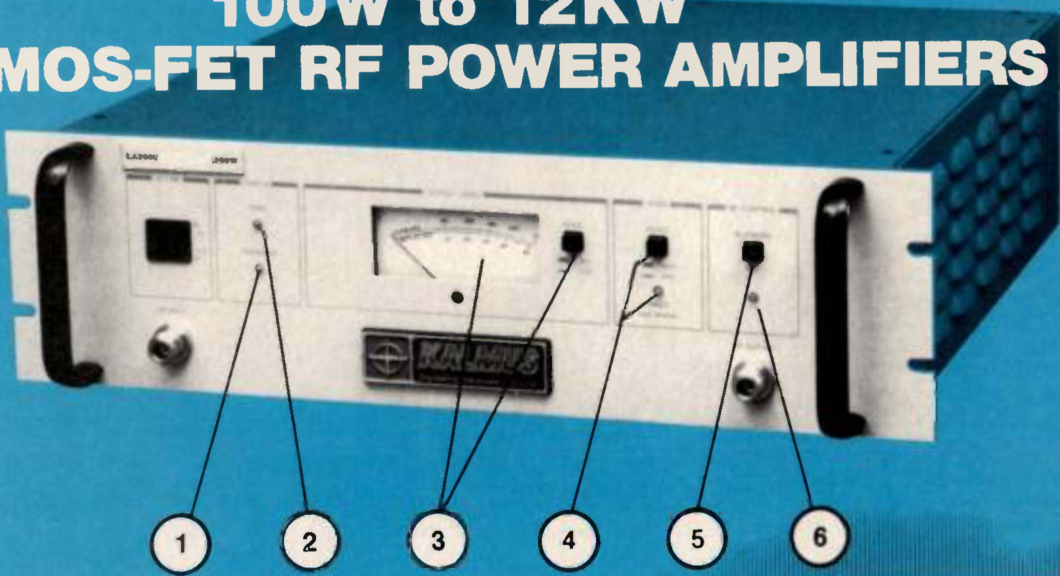

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INFO/CARD 17

RF Technology in Broadcasting and CATV

By Gary A. Breed
Editor

Broadcasting was one of the first commercial applications of RF technology, and radio, television and cable television continue to have a substantial influence on the development of new technologies. Studio, head-end, and transmitting facilities are still strong markets for advanced RF technology. This short report can only touch on some of the major developments.

In television broadcasting, Varian has had two major projects underway. The first is development of the Klystrode, as its name implies, a hybrid of the klystron and tetrode. This tube was created to improve the efficiency of the linear power amplifiers required for UHF transmitters. It doesn't take much math to estimate the cost of operating a 110 kilowatt transmitter using a 30 percent efficiency klystron, and then come up with the cost savings when that is improved to 40 or 45 percent. With development and test cooperation from TV transmitter manufacturer Comark, the Varian Klystrode has been proven to provide lower operating costs through higher efficiency.

The second project, also at Varian, is a higher efficiency klystron using a depressed-collector configuration. Because this tube is a true klystron, it has applications at much high frequencies than are available with the Klystrode. Both this and the Klystrode have applications beyond broadcasting, in radar, EW and ECM, and satellite communications. This is why NASA is a participant in depressed-collector klystron development, as well as the broadcasting industry.

It would be unfair not to mention work on solid-state television power amplifiers. Several companies offer VHF transmitters with solid-state aural amplifiers (Class C) operating up to several kilowatts. The video amplifiers remain a challenge, mainly because television requires linear amplification with its necessary lower efficiency and higher heat dissipation. Still, all current high power transmitters have solid state amplification through the driver stages.

Television production equipment utilizes new wideband linear integrated circuits, and has a major stake in high speed digital circuitry for image processing. Studio video for current video standards requires from 10 to 20 MHz bandwidth depending on the specific application. New operational amplifiers, buffers and analog multiplier ICs from companies like Elantec, Comlinear, Harris Semiconductor, Analog Devices and Burr-Brown are found in video distribution equipment, and in the analog input and output stages of digital video processing equipment. Analog switches from Maxim, Siliconix, and Harris are used for routing video and audio signals. In high performance applications, switches built with discrete FETs can still be found.

High speed digital technology had some of its first applications in television, with digital time base correctors, standards converters, and frame synchronizers. All of the special effects and graphics that television viewers take for granted are truly technological masterpieces. Pioneering these applications were companies like the Grass Valley Group, 3M, Ampex and others. Advanced image manipulation requires high speed analog-to-digital and digital-to-analog conversion plus high speed processors to perform the vector mathematics required for manipulating each pixel that makes up the image. Components for these applications are manufactured by Sony, TRW, Burr-Brown, Analog Devices, Datel and Comlinear.

Cable television continues to require medium power, low distortion amplifiers for signal distribution, with Motorola (ex-TRW) and Philips the main suppliers. High performance head end modulators are needed to minimize adjacent-channel interference, using SAW IF filters, and low distortion modulator and amplifier circuitry. Test equipment for CATV has seen considerable improvement in recent years, primarily for real-time system performance monitoring, and fast troubleshooting of equipment failures.

Encryption technology is another area where high-speed digital processes are required. From simple remotely switched and monitored converters to high-security scrambler/descrambler units that continually change the encryption code, high speed video processing reaches many cable subscribers. Even simple set-top decoders require excellent dynamic range and intermodulation performance, as well as programmability over the wide bandwidth of CATV signals (as much as 500 MHz).

Radio broadcasting continues to change from vacuum tube power amplifiers to solid state devices. One transmitter manufacturer estimates that nearly 50 percent of the local AM radio stations (1 kW power or less) are now using solid state transmitters. At very high power, companies like Harris Corp., Continental, Nautel, and others have developed unique combining techniques for generating up to 50 kW at medium wave frequencies using multiple MOSFET amplifier modules. FM transmitters at the 3.5 kW level and under are available in solid state designs, but have only been sold for three years or so.

If you include the radios, stereo systems, television sets and VCRs used for the reception of broadcast and CATV programming, broadcasting is probably be the largest RF industry segment, worldwide. In the U.S., where consumer equipment is a more modest part of RF manufacturing, the transmission equipment for radio and television is still a considerable market for RF components and a source of significant technological advancements. ***RF***

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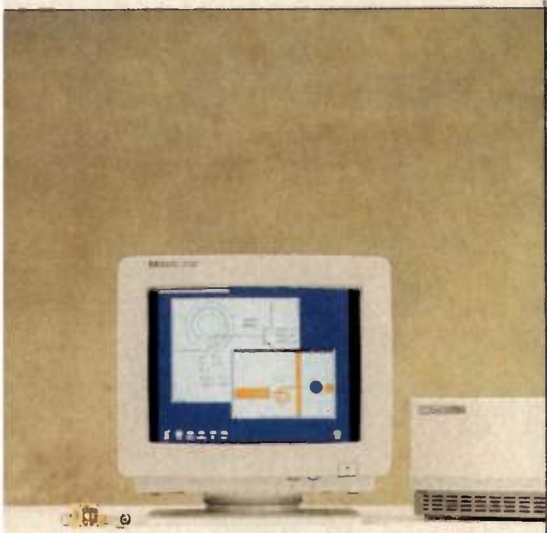

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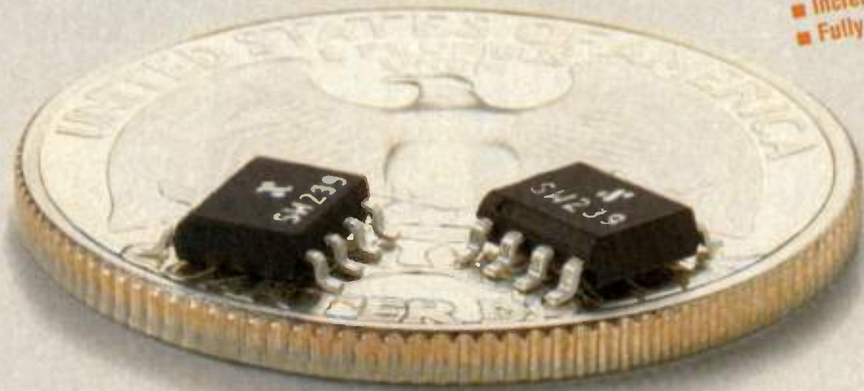

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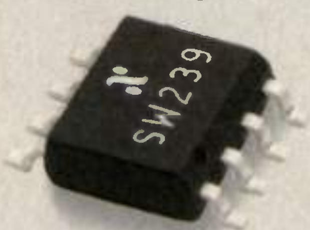

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1 dB Compression	+27 dB
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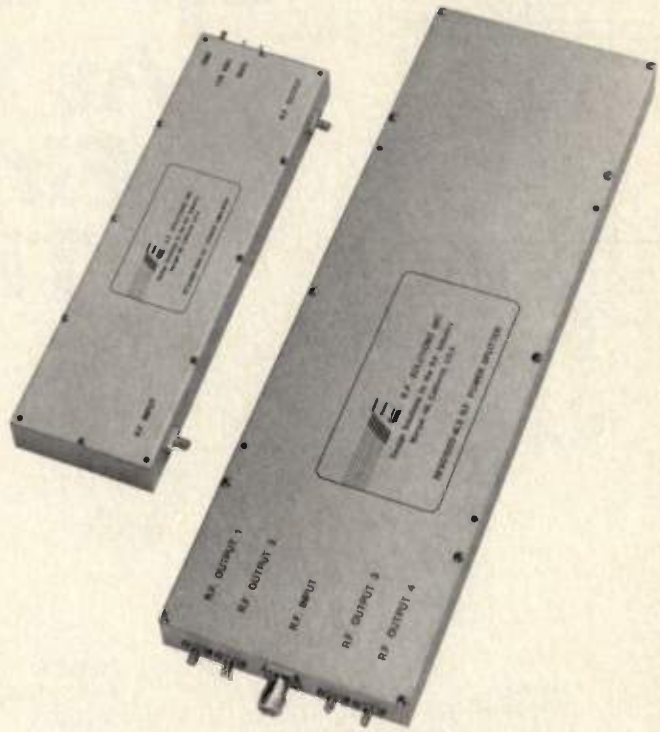
INFO/CARD 21

OFF THE SHELF

RF AMPLIFIERS

MODEL NUMBER	POWER Watts	GAIN dB	SUPPLY Volts	PRICE
FREQUENCY RANGE 5 – 50 MHz				
RFP0550-100	100	44	50	$2,100.00
RFP0550-1000	1000	16	50	$5,040.00
FREQUENCY RANGE 50 – 100 MHz				
RFP0800-100P50	50	30	50	$1,485.00
RFP0800-100P100	100	30	50	$1,660.00
RFP0800-100P200	200	30	50	$2,200.00
RFP800-100	600	16	50	$2,424.00
FREQUENCY RANGE 1 – 100 MHz				
RFP01100-300	300	46	50	$3,150.00
FREQUENCY RANGE 76 – 108 MHz				
RFP0810-600	600	16	50	$1,780.00
FREQUENCY RANGE 75 – 150 MHz				
RFP0800-150P50	50	30	50	$1,485.00
RFP0800-150P100	100	30	50	$1,660.00
RFP0800-150P200	200	30	50	$2,200.00
RFP800-150	500	14	50	$2,424.00
FREQUENCY RANGE 100 – 200 MHz				
RFP0800-200P50	50	30	50	$1,660.00
RFP0800-200P100	100	30	50	$2,900.00
RFP800-200	400	13	50	$3,636.00
FREQUENCY RANGE 225 – 400 MHz				
RFP0204-4	4	20	28	$ 484.00
RFP0204-10	10	30	28	$ 685.00
RFP0204-25	25	30	28	$1,140.00
RFP0204-50	50	40	28	$1,695.00
RFP0204-100	100	40	28	$2,200.00
FREQUENCY RANGE 400 – 500 MHz				
RFP0405-4	4	20	28	$ 435.00
RFP0405-10	10	30	28	$ 616.00
RFP0405-25	25	30	28	$1,026.00
RFP0405-50	50	40	28	$1,525.50
RFP0405-100	100	40	28	$1,980.00
FREQUENCY RANGE 1 – 500 MHz				
RFP00105-4	10	30	28	$1,450.00
RFP00105-25	25	30	28	$2,800.00
RFP00105-50	50	40	28	$3,752.00
RFP00105-100	100	40	28	$5,600.00



COUPLER/SPLITTER

MODEL NUMBER	INPUT/ OUTPUT PORTS	INSER-TION LOSS dB	POWER Watts	PRICE
FREQUENCY RANGE 5 – 50 MHz				
RFS100-5-50-4	4	0.3	100	$ 360.00
RFS1000-5-50-4	4	0.3	1000	$ 610.00
FREQUENCY RANGE 50 – 1000 MHz				
RFS100-FL-FU-2	2	0.15	100	$ 500.00
RFS100-FL-FU-3	3	0.15	100	$ 600.00
RFS100-FL-FU-4	4	0.3	100	$ 800.00
RFS100-FL-FU-6	6	0.3	100	$1,000.00
RFS1000-FL-FU-2	2	0.15	1000	$ 750.00
RFS1000-FL-FU-3	3	0.15	1000	$ 900.00
RFS1000-FL-FU-4	4	0.3	1000	$1,200.00
RFS1000-FL-FU-6	6	0.3	1000	$1,500.00

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The STEL-2110A is a digital bit synchronizer and PSK demodulator, providing high performance in spread spectrum BPSK and QPSK applications.

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The STEL-2210 Block Phase Estimator allows burst signals to be acquired rapidly and demodulated coherently when used in conjunction with the STEL-2110A.

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Designing a Direct Sequence Spread Spectrum Secure Communication System

By Soo Kwan Eo, Chingwo Ma, Incheol Jang
Comdisco Systems

Spread spectrum signals in communications are used for the purpose of combatting the detrimental effects of interference due to jamming, interference arising from other users of the channel and self-interference due to multipath propagation. The spread spectrum signal can be hidden from unintended listeners by transmitting it at low power, thereby achieving message privacy in the presence of other listeners.

One form of spread spectrum, direct sequence spread spectrum (DS-SS) modulation, can be achieved by superimposing a pseudo-random number (PN) sequence upon the information bits. In the receiver, the received signal is cross correlated with the same PN sequence. The effect of this cross correlation is the reduction of the information spectrum bandwidth and, at the same time, the widening of the unwanted jamming signal and receiver noise in order to effectively increase the signal-to-noise ratio. A merit of this scheme is that power spectral dispersal can significantly improve the bit error rate of the information pulse train.

The performance of the PN sequence spread-spectrum system can be further increased, with respect to its immunity to the interfering signal, by incorporating a linear adaptive LMS filter before the de-spreading operation takes place. Some algorithms (such as that of Theodoridis (1)) employ LS linear phase FIR filters to suppress the interference. While they provide adequate performance, if the interference is a slowly time-varying signal these algorithms are ineffective. A better approach would be to adopt a time-recursive linear phase scheme such as the Widrow LMS algorithm (2,3).

The design of such a system requires careful analysis of three main concerns:

1) The reduction of cross-correlation of different pseudo-random sequences due to the lack of absolute randomness in the different sequences generated. (Cross correlation close to zero is needed to generate a sequence that approximates white noise.)

2) The receiver's ability to synchronize the generation of the pseudo-random sequences in order to properly convolve the transmitted data.

3) The attributes of the interference suppression filter, (i.e., its ability to suppress interference of some predefined type as the relative strength of the data signal and the interfering signals vary.).

In this article we will present a design methodology that allows the designer of such a system to rapidly simulate the effects of changing system topology and parameters in a DS-SS system. The design software used is the Signal Processing WorkSystem, from Comdisco Systems. The overall system architecture is presented and the operation of the transmitter and receiver are discussed. For this example we will assume that the interfering signal is a high-power narrow band signal that may be swept across the bandwidth of the data signal. The filter chosen for this example is a linear adaptive LMS filter that will be described and illustrated.

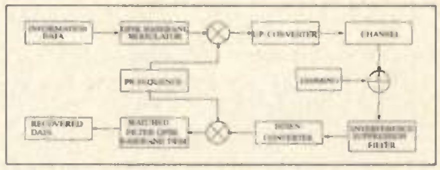

Figure 1. Direct spectrum spread signal system overview. A major design goal of this system is to recover the data signal in an environment with a jammer-to-signal ratio (JSR) of up to 30 dB.

To demonstrate the ease of analysis provided by this design methodology, the injected narrow band jamming signal and the corresponding interference rejection filter response are shown. Designers of such systems want to know

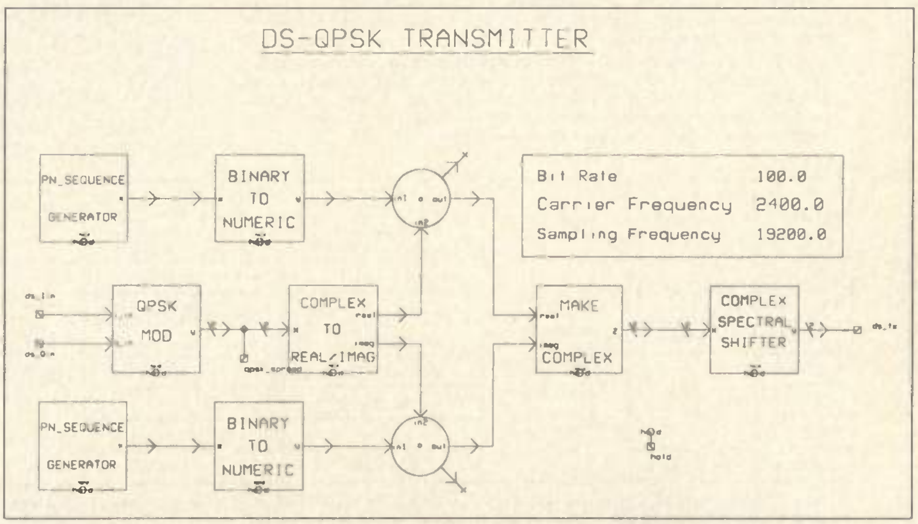

Figure 2. Direct sequence spread spectrum QPSK transmitter (as displayed on Comdisco System's Signal Processing WorkSystem). The QPSK output is at a carrier frequency of 2400 Hz with a sampling rate of 19200 samples per second and a bit rate of 100 bps.

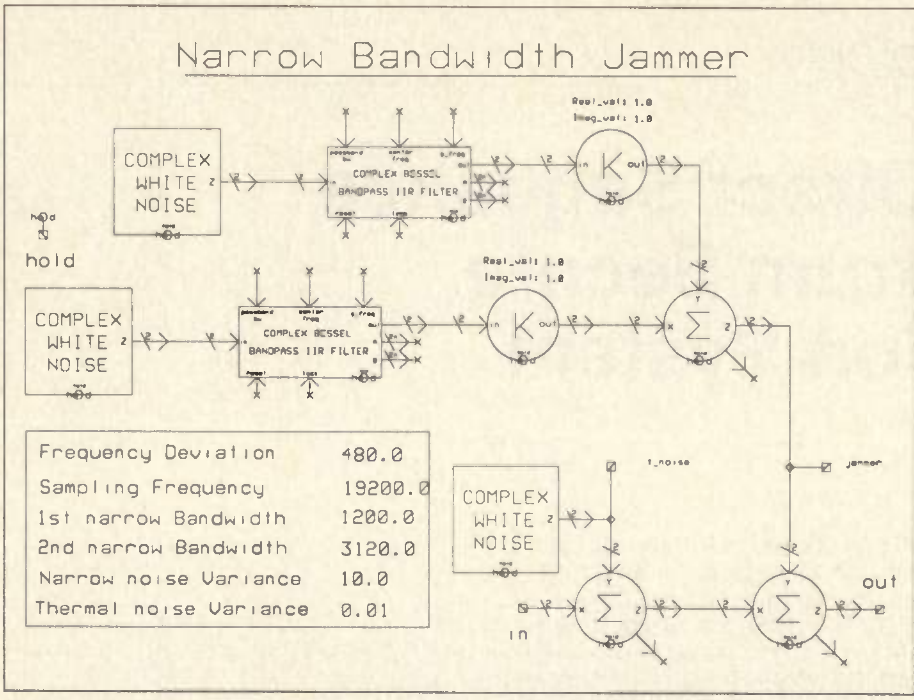

Figure 3. Narrow bandwidth jammer. This model is used to simulate interference that centers around two carrier frequencies (1200 Hz and 3120 Hz) with a frequency deviation of 480 Hz.

how the received symbol error rate varies as the jammer to signal ratio/chip varies. The techniques for performing this analysis are also demonstrated.

The Design Environment

The Signal Processing WorkSystem (SPW) provides simulation and analysis tools that automate digital signal processing system design. The user can select from a rich library of communications functional blocks, each of which can be represented hierarchically. The blocks are placed on the workstation display screen and then interconnected. To facilitate the simulation of the captured design, the user can either generate his own stimuli signals using a powerful signal editor, or, he can use real-world signals that have been digitized. The designer does not have to write any simulation code since the simulation program is automatically generated and compiled by SPW. The results of the simulation can be analyzed and displayed using FFTs to produce frequency-domain plots, eye diagrams, scatter plots, etc. Even the bit-error-rate (BER) of a digital communications system can be automatically calculated and displayed.

The software allows the designer to test the design at a high level of abstraction and to easily perform what-if trade-off analysis to optimize the design. It also lets the designer move to lower levels of abstraction so that an easy transition to hardware implementation

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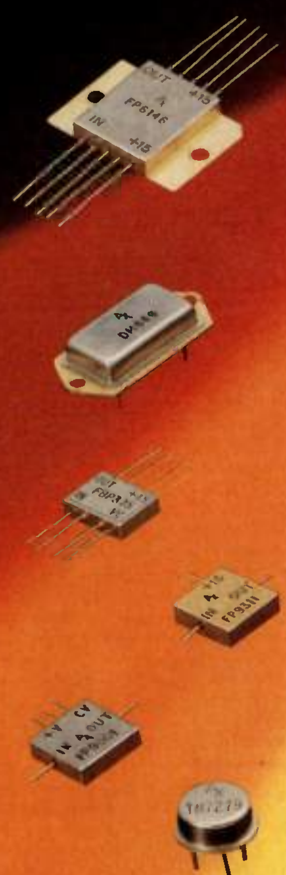

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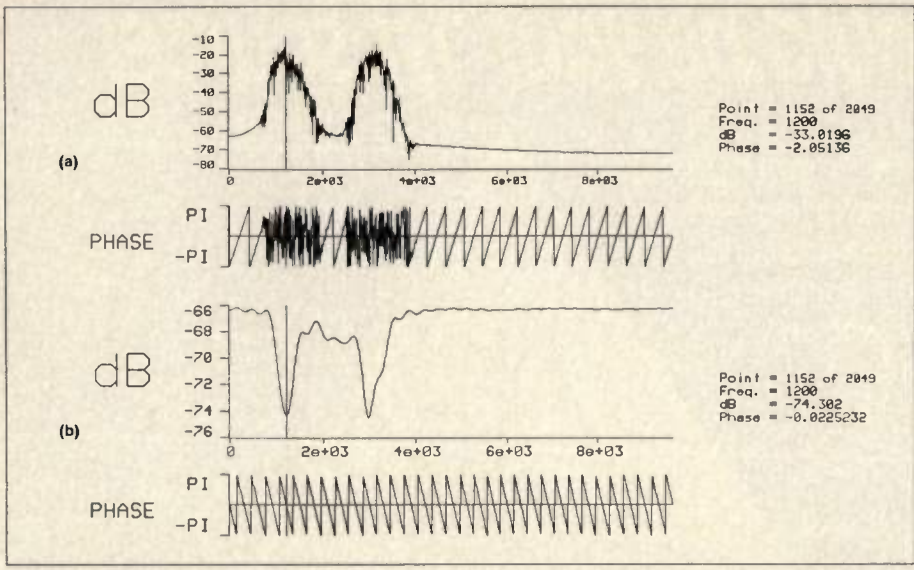

Figure 4. Interference suppression adaptive filter response. In (a) the two narrow band interference signals center at carrier frequencies of 1200 Hz and 3120 Hz with a 480 Hz frequency deviation. The response of the linear phase adaptive LMS filter is shown in (b).

is possible. Tools are provided that allow the user to implement his design in discrete integrated circuit standard parts, ASICs, or within many of the currently available digital signal processors utilizing Comdisco's Hardware Design System (HDS) for implementation in VHDL or the Code Generation System (CGS) for direct generation of processor code.

The communications library includes a majority of the blocks a designer is likely to need. In fact, all of the blocks used in this design are contained in the library. For those designs that require blocks not already in the SPW library, custom block building facilities are provided.

DS-SS QPSK System Overview

The direct sequence spread spectrum system presented here uses an adaptive suppression filter to minimize interference caused by time-varying jamming or interfering signals. A major design goal of this system is to recover the data signal in an environment with a jammer-to-signal ratio (JSR) of up to 30 dB. In the following sections, the direct-sequence spread spectrum (DS-SS) transmitter, jamming model, interference suppress filter model, and DS-SS receiver model are discussed.

The DS-SS QPSK system overview is illustrated in Figure 1. Sample information data is generated (or captured) within the software and a QPSK signal is output by the QPSK baseband modulator. The QPSK signal is multiplied by a binary sequence generated by the pseudo-random number (PN) sequence generator. The spectrum of the signal is further shifted to the carrier frequency region by an up-conversion module and is transmitted into the communications channel. To receive the signal it is first processed by an interference suppression filter, then down-converted to the baseband. Then the signal is multiplied by the same PN sequence used in the transmitter. Finally a matched filter and QPSK demodulator reproduce the digital information.

For this example (see Figure 2), the transmitter uses a random data generator as its information source. Two sets of random data with equiprobable 0 and 1 bits are converted from binary to numeric form and are input to the QPSK modulator. Before the phase modulation takes place, the in-phase (I) and quadrature (Q) inputs are QAM (quadrature amplitude modulated) and EXORed with the pseudo-number sequence. The EXORing is performed with multiplication using the "1" or "-1" amplitude of the PN sequence. The spreaded information bit stream is not phase modulated before it is transmitted via the channel. In this example the bit rate is 100 bps, the carrier frequency is 2400 Hz, and the sampling rate is 19,200 samples per second. The spreading is done by chip pulse and there are 24 chips per information symbol.

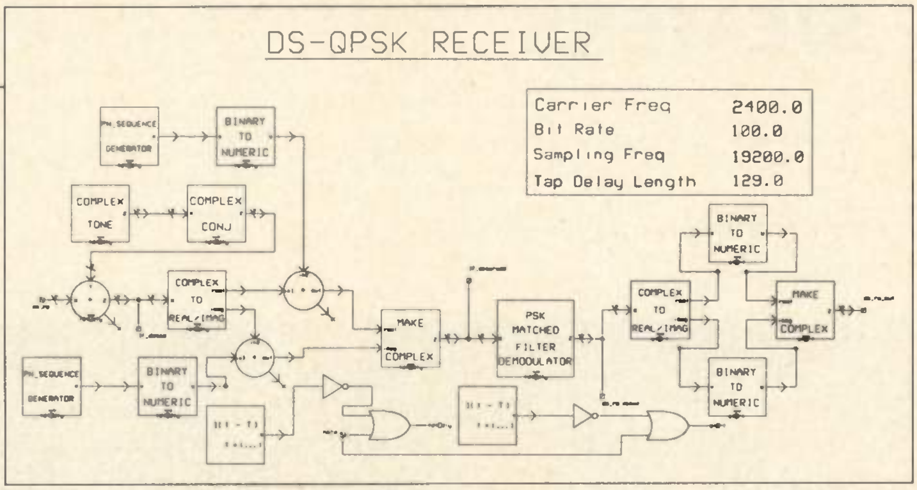

Figure 5. DS-QPSK Receiver.

Narrow Bandwidth Jammer Model

Figure 3 shows the modeling of a subsystem that simulates a narrow bandwidth jammer. Other jamming modes, such as high-power CW jamming, or random power-variant jamming, are of course, possible. The system designer must decide which jamming modes will be modeled. He is also faced with the need to determine whether or not a jamming signal is present (not a very difficult task) as well as the more challenging problem of identifying the jamming mode in order to activate the appropriate anti-jamming algorithm.

It is assumed that the jammer concentrates the jamming signal on the baseband data spectral region. In this example, a dual-band interfering signal is injected into the transmission channel (at 1200 Hz and at 3120 Hz) with a frequency deviation of 480 Hz using a

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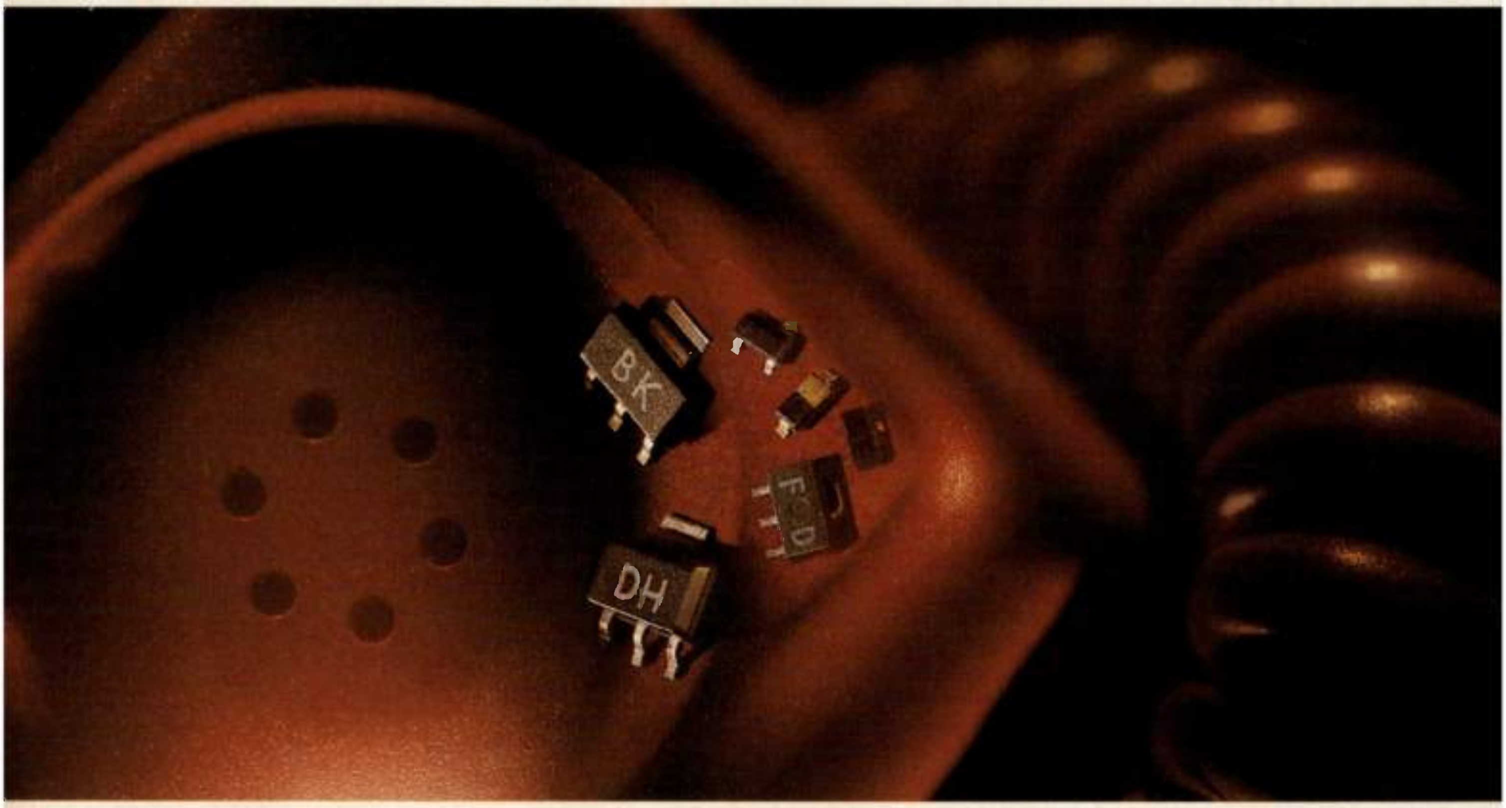

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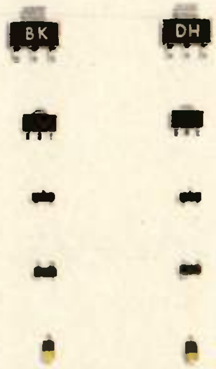

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BFP 280	8	10	1.4	18	1 to 5/0.2 to 8 low-noise pager
BFP 181	12	20	1.3	19	1 to 8/0.5 to 10 low-noise pre-amplifier
BFP 182	12	35	1.2	19	1 to 8/1 to 20 low-noise amplifier
BFP 183	12	65	1.2	19	1 to 8/2 to 28 low-noise amplifier
BFP 193	12	80	1.2	19	3 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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JSR/chip (dB)	−8.5	−1.5	3.7	13.3	15.7	18.6	21.5	23.2
W/O DS-SS	9.8×10^{-5}	.033	.05	0.3	N/M	N/M	N/M	N/M
W/O Filter	N/D	N/D	0.21	0.082	0.15	0.25	0.32	N/M
DS-SS, filter ($\times 10^{-4}$)	N/D	N/D	1.85	1.9	2.27	2.5	4.2	5.8

N/M : Not Measured
N/D : Not Detectable

Table 1. The symbol error rate for various JSR/chip ratios.

sampling rate of 19,200 with a level variance of 10 dB. Receiver thermal noise is modeled as white gaussian noise with a variance of 0.01 dB.

Interference Suppression Adaptive Filter

A linear phase adaptive LMS filter can be used to suppress the interference signal before the de-spreading operation is resumed at the receiver. In this example, an Equalizer Tap Coefficient block with backward error predicting scheme is used to implement the filter. Since the pseudo-random sequence is assumed to be almost white, the predictor predicts the high-power interference signal and the corresponding prediction error filter rejects the interference. It is assumed that the jamming signal is not synchronized to the signal's symbol rate but interferes on a bit-by-bit basis. The tap coefficients are appropriately updated to reject the interference. In Figure 4, the injected narrow band jamming signal and corresponding interference rejection filter response are shown.

DS-SS QPSK Receiver

The receiver operation is the exact inverse of the transmitter (see Figure 5). The DS-QPSK receiver recovers the data signal from the jamming signal by using a combination of linear adaptive filtering and spread spectrum techniques.

The transmitter sends a burst of synchronizing bits before it sends real data and the receiver uses this burst signal to synchronize the receiver (not shown in this current model).

Once interference is rejected, the received signal is phase demodulated and then de-spread using cross correlation between the received signal and the exact replication of the PN sequence that was used for transmission. This cross correlation further collapses the information signal to its original bandwidth. It also reduces the corresponding level of the narrow-band interference by spreading it across the bandwidth occupied by the PN sequence.

After the received signal is collapsed to its original bandwidth and interference signal power is minimized, it is fed into the matched filter and PSK detector.

Simulation

The ultimate goal of this system is to recover a signal that is being jammed by narrowband interference. The best

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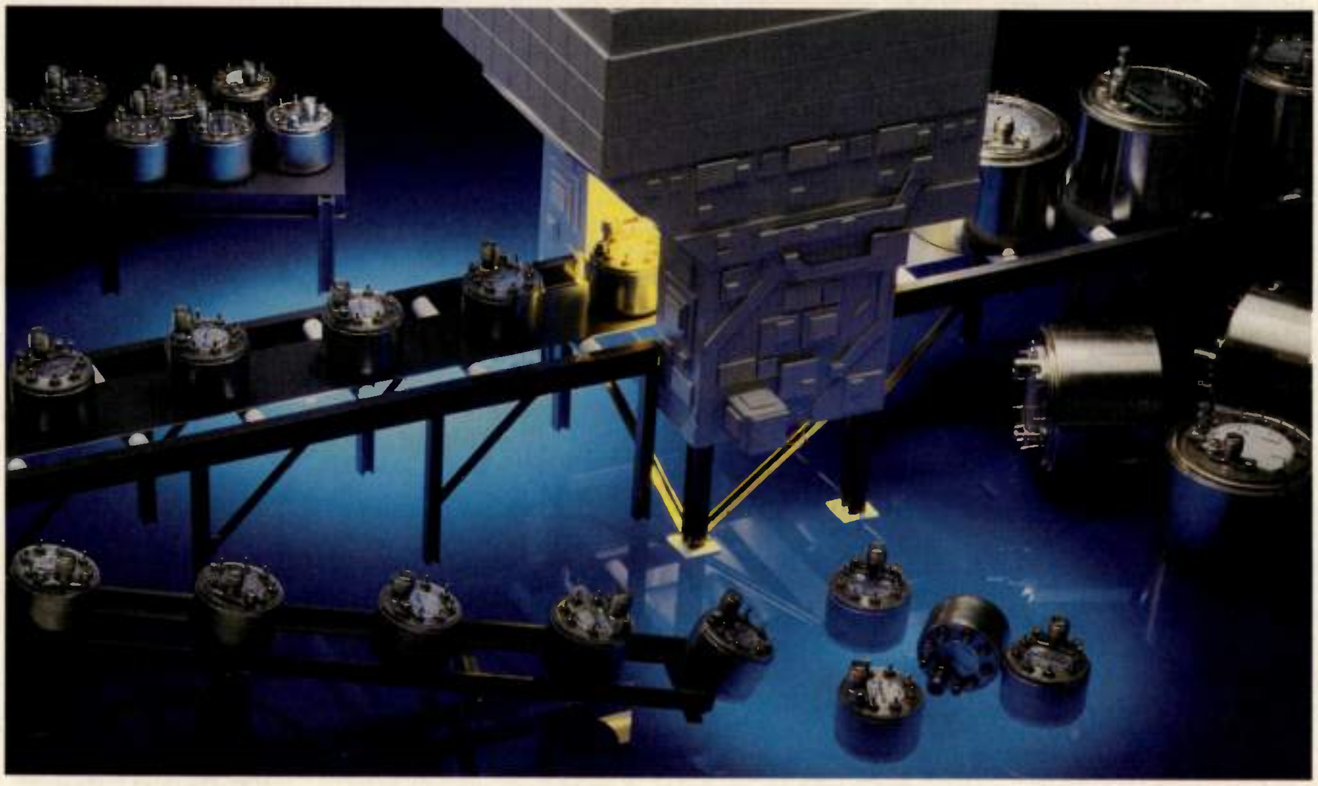

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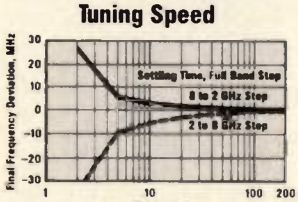

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measure of the effectiveness of the design is an examination of the received symbol error rate. Due to the complexity of this DS-SS QPSK system, determining this symbol error rate using hand coded simulation and analysis is nearly impossible. The Comdisco Signal Processing WorkSystem, however, allows the designer to accurately simulate the design, and automatically calculate the bit-error-rate to judge its effectiveness.

The simulation was configured with the assumption of a 129 tap filter, a jammer-to-signal ratio of 23 dB, and the two bands of narrow band interference described earlier in the article. Table 1 shows the symbol error rate for various JSR/chip ratios derived from the simulation. Each case was run with 2 to 3 million iterations. Once the symbol error is detected, the matched filter in the receiver loses phase synchronization with the incoming signal and tends to produce burst symbol errors. The symbol error rate is, therefore, calculated before the receiver experiences this burst error. Three scenarios were modeled and simulated. First, the system was simulated using no direct sequence spread spectrum techniques ("W/O DS-SS"). Then DS-SS was used but with no LMS adaptive filtering at the receive end ("W/O Filter"). finally, the system was simulated using both DS-SS and filtering ("DS-SS, filter").

Note the multiplier of 10^{-4} for the DS-SS, filter case. As can be seen from this data, the combination of DS-SS and LMS adaptive filter produce greatly improved symbol error rates.

Conclusion

This design methodology offers the designer of a direct sequence spread spectrum system dramatically new options. He can now experiment with different filter schemes, interference patterns, PN generators, etc. to optimize system performance, minimize complexity, reduce cost, etc. before building any prototype. Within a workstation environment, the designer need only swap functional blocks, change interconnections, vary parameters within those blocks, and then analyze the data after the simulation is performed with the new specifications. ***RF***

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3. L.B. Milstein, L.A. Iltis, "Signal Processing for Interference Rejection in Spread Spectrum Communications," *IEEE ASSP Mag.*, pp. 18-31.

About the Authors

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IQ Sampling Yields Flexible Demodulators

By Thomas Hack
Comlinear Corporation

I/Q demodulation is commonly found in Doppler Radar and communications systems including AM, DSB, and QAM formats. With additional circuitry FM, PM, SSB, ISSB and other formats can also be added to the list. I/Q demodulators are quite flexible, but this flexibility comes with some problems. The use of modern high speed analog-to-digital convertors (ADCs) allow traditional "analog" processing of I/Q demodulators to be carried out digitally, greatly reducing systematic errors. This article covers the practical implementation of a high-performance I/Q demodulator using modern ADCs.

A basic I/Q demodulator block diagram is given in Figure 1. An arbitrary IF signal enters the power divider, where it is split and fed to two mixers. Quadrature-phased local oscillators also feed these mixers. Low-pass filters remove the "sum" mixer products leaving the I and Q signals. When demodulating some signals (such as QAM) this is all that is needed. In others, ADCs and digital signal processing may follow.

There are several disadvantages to this approach. When implemented with conventional RF components, DC offsets, quadrature-phase errors, gain matching, carrier leakage, and impedance matching (among others) degrade demodulator performance. With advanced analog-to-digital converters such as Comlinear's CLC926, we can actually build a better demodulator.

Building a Better Demodulator

To illustrate the advantages of our approach, we'll start with a basic I/Q demodulator and modify it in stages showing the advantages after each modification. Since ADCs are very efficient harmonic mixers (see appendix), we can eliminate the mixers by replacing them with high-speed ADCs (Figure 2). All of the functions that follow the ADCs are now digitally implemented which eliminates matching errors in this part of the implementation and reduces most of the other errors.

Unfortunately ADC matching errors remain. Quadrature errors result from effective aperture delay mismatching between the two ADCs. ADC gain-matching errors are also present.

We could eliminate these errors if we somehow used only one ADC. Figure 3 provides us with a solution. Figure 3 shows the traditional local oscillator signals required in I/Q demodulators using analog mixers. For the approach in Figure 2, all signals after the ADC are in discrete-time form. As a result, the local oscillator signals can be represented as impulses of strength -1 and 1. The samples from the ADCs coincide

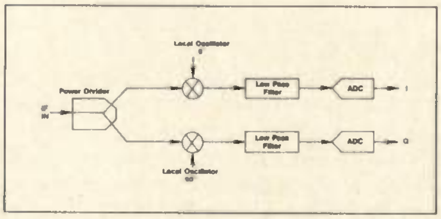

Figure 1. A basic I/Q demodulator.

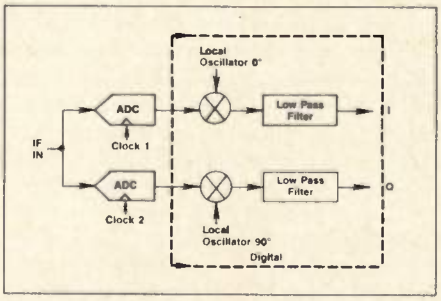

Figure 2. Another approach to I/Q demodulation.

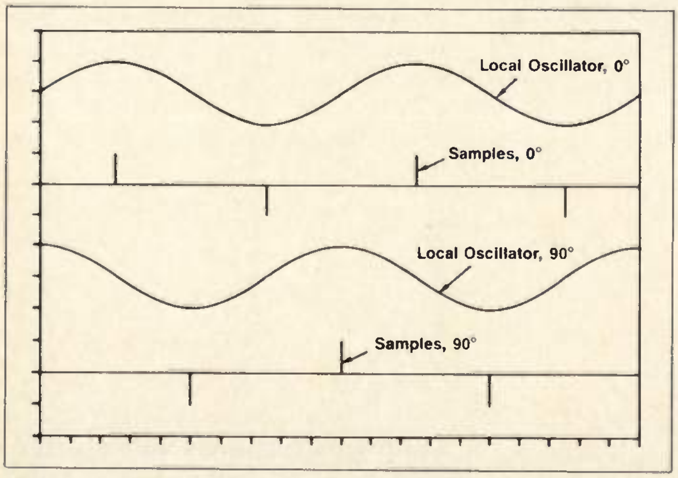

Figure 3. The upper 2 waveforms depict in-phase and the lower waveforms quadrature-phase local oscillators for the analog (Figure 1) and digital (Figure 2) approaches to I/Q demodulation.

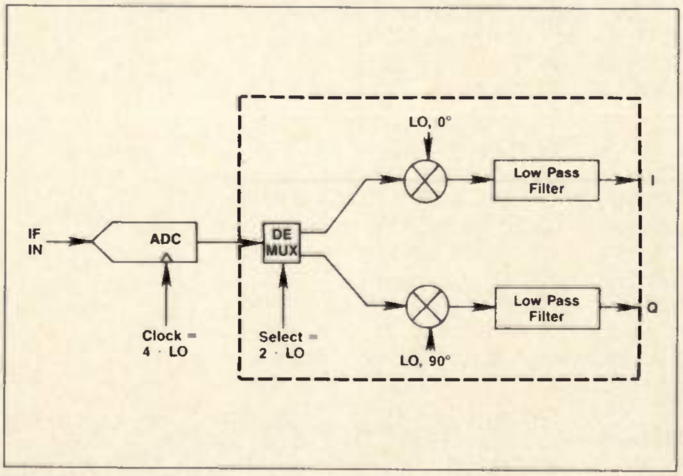

Figure 4. An improved I/Q demodulator.

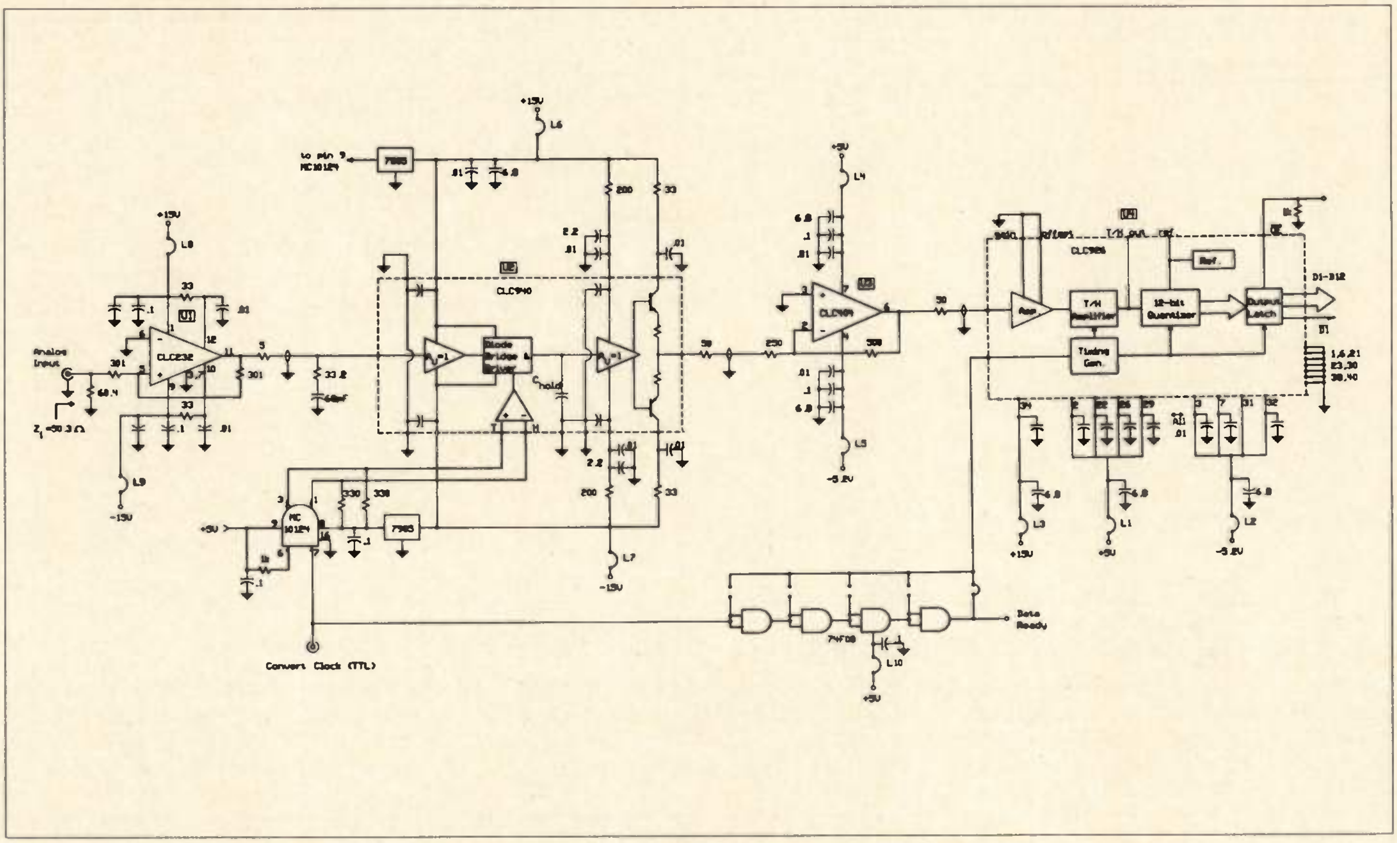

Figure 5. High performance IF sampling analog-to-digital convertor subsystem.

with the locations of these impulses.

From the diagram we can see that the in-phase ADC's samples occur exactly midway between the quadrature-phase ADC's samples. By using one ADC, doubling the sampling rate and sending alternate samples to the I and Q signal processing chain we can get the same effect. With this arrangement our sample clock will operate at four times the original local oscillator frequency. With both the I and Q samples coming from the same ADC, ADC quadrature and gain-matching errors are eliminated. The final block diagram is given in Figure 4.

Depending on speed requirements, discrete logic programmable DSP chips, or a mix of discrete logic and DSP chips could be used. For example, one DSP chip could be used after the ADC providing all demultiplexing, mixing, filtering, as well as any additional downstream signal processing. If more processing power is required, discrete logic can split the ADC data stream into separate I and Q signals with a DSP processor connected to each to speed things up. The same approach might also allow 2 cheaper DSP chips to replace one faster but more expensive chip. Mixing is also well suited to dedicated hardware, if desired. Since mixing is done by multiplying the ADC

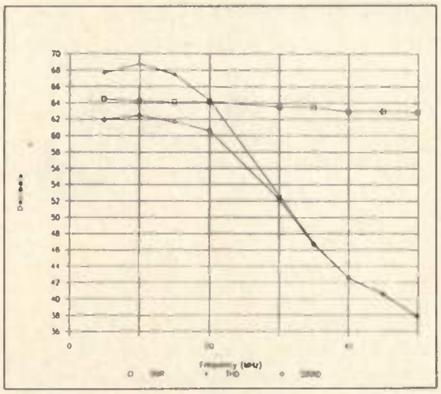

Figure 6. IF sampling performance.

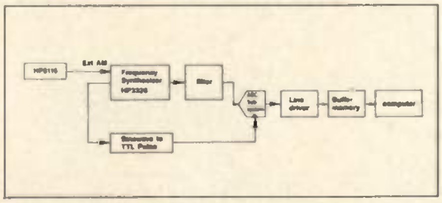

Figure 7. I/Q demodulation demonstration.

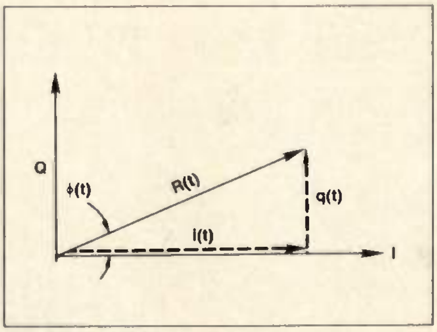

Figure 8. Quadrature carrier description of an IF signal.

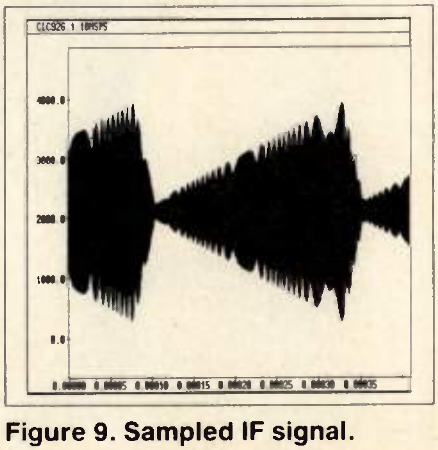

Figure 9. Sampled IF signal.

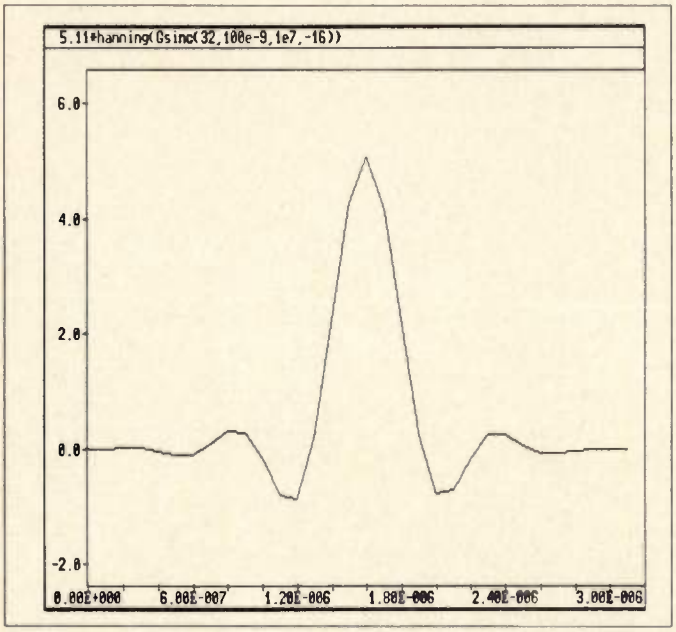

Figure 10a. Filter time domain response (impulse).

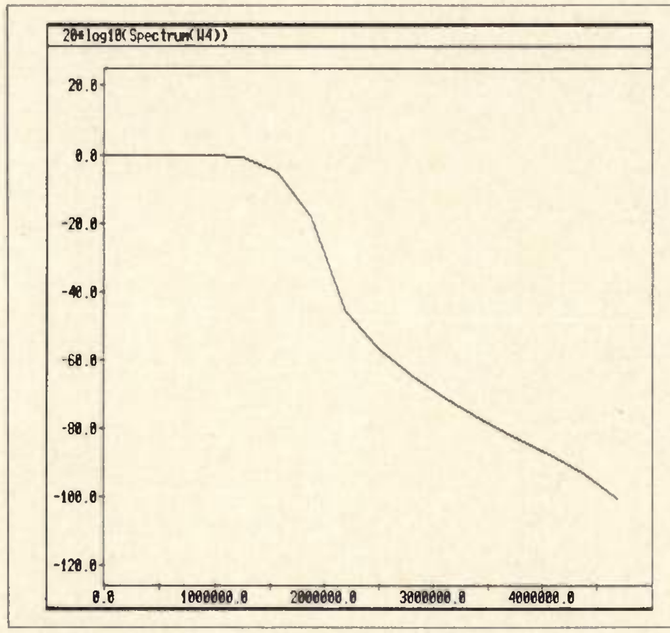

Figure 10b. Filter frequency response.

samples by alternating ones and minus ones (our local oscillator), simple hardware (not a full-blown multiplier) will suffice. The rest of the signal processing can be investigated in similar fashion.

While the technique outlined isn't new, it hasn't had wide appeal because data converters didn't offer enough performance except at very low IF frequencies. When state-of-the-art ADCs and track-hold amplifiers such as the CLC926 and CLC940 are combined together, however, IFs up to 21.4 MHz and beyond become practical. Figure 5 is a schematic for a high-performance IF sampling analog to digital convertor subsystem. Analog-to-digital conversion is provided by U4, a CLC926 (a 10 MSPS, 12-bit ADC offering 70 MHz analog input bandwidth). The CLC926 is a good choice with its state-of-the-art spurious-free signal range, signal to noise ratio, SINAD, and THD — and all of these specifications are tested on every part.

Performance is extended at higher IF frequencies by placing track-hold amplifier U2 (a CLC940) in front of the ADC. Since the ADC does its conversion on a held signal its dynamic errors are reduced. U2 sets the aperture jitter to 1 ps rms typical so signal to noise ratio of the incoming signal is not degraded by the track-hold even at high IF frequencies.

To reduce distortion, the CLC940 track-hold is intentionally operated at reduced amplitude. With typical second and third harmonic distortion performance of -57 and -60 dBc at 20 MHz and 2 V_{pp} output amplitude, distortion isn't low enough for our purposes. But since the CLC940 has classic intercept-point behavior, dropping the amplitude 6 dB to 1 V_{pp} should reduce HD2 by 6 dB (or -63 dBc) and HD3 should drop 12 dB to -72 dB. A CLC404 wideband, low distortion amplifier (U3) boosts the 1 V_{pp} CLC940 output to 2 V_{pp} — the full-scale range of the CLC926 ADC (This insures highest ADC dynamic range). Since the CLC404 is after the track-hold, dynamic performance requirements on it are somewhat relaxed.

Several other tricks are used to maximize performance. U1 (a CLC232) buffers the input signal providing less than 10 ohms driving-point impedance over frequency-minimizing track-hold amplifier distortion products. Distortion is also slightly improved by using differential ECL track-hold control signals.

Looking at the results from single-tone tests (Figures 6), it is easy to see that our efforts have been rewarded. At 21.4 MHz IF (-1 dB full scale), SINAD (signal to noise + distortion) is better than 60 dB (or 9.8 effective bits). Total harmonic distortion is -64 dBc and in-band harmonics are less than -66 dBc or better. Signal-to-noise ratio when measured over the full Nyquist bandwidth is 64 dB.

A Simple Demonstration

I/Q demodulation has been demonstrated with the circuit in Figure 5 using a PC-based data acquisition system (Figure 7). An HP 3326A synthesizer provides modulated signals which are filtered prior to being sent to the ADC circuit. Filtering removes synthesizer harmonics and wideband noise. A 10 MHz clock is also provided by the synthesizer by level shifting the 10 MHz reference output. This gives us complete control over the LO/carrier frequency ratio (phase is not as easy to control, however).

ADC samples are sent through a line driver, down a cable into a buffer memory. The personal computer accesses the buffer and performs the digital signal processing algorithms using an off-the-shelf DSP package.

For our demonstration we externally amplitude modulated the synthesizer with a 4 kHz sawtooth wave from an

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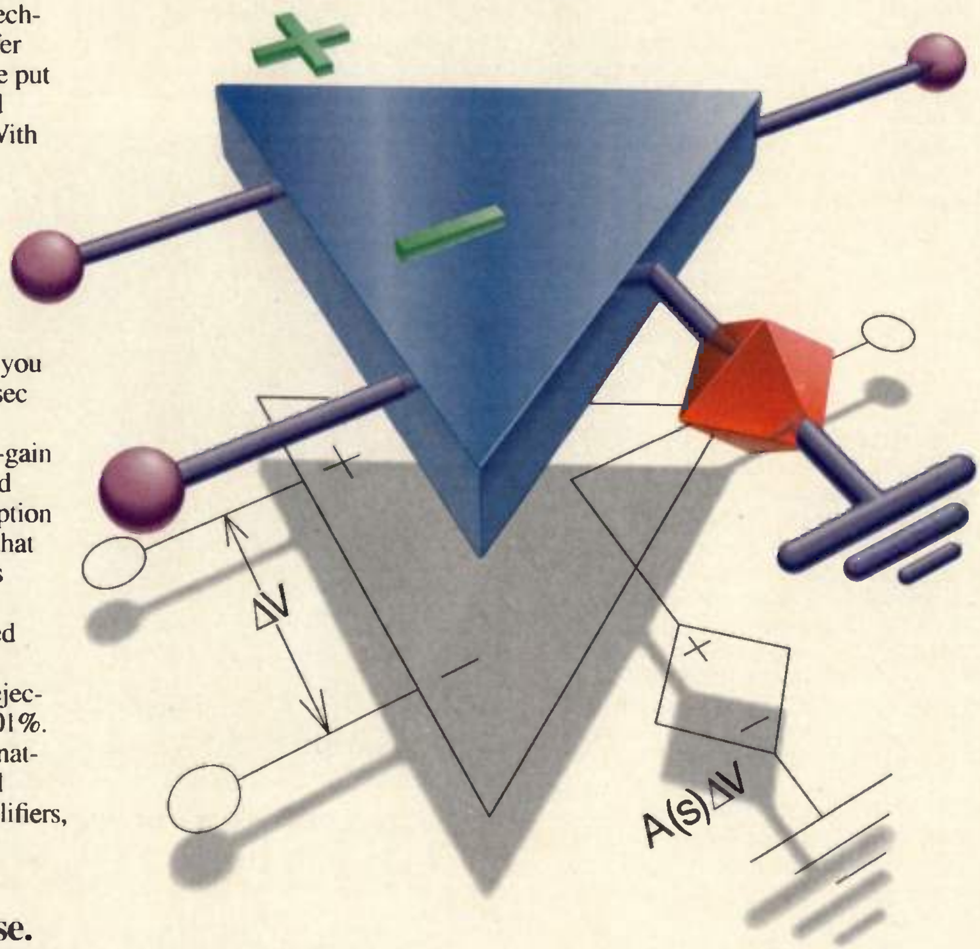

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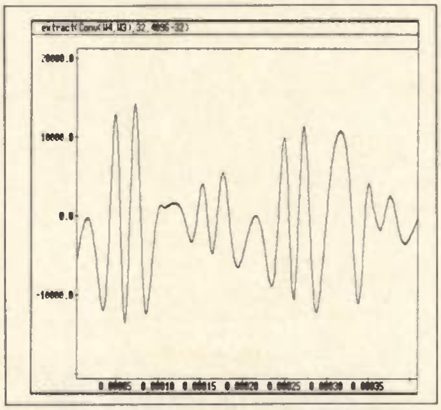

Figure 11a. I and Q signals.

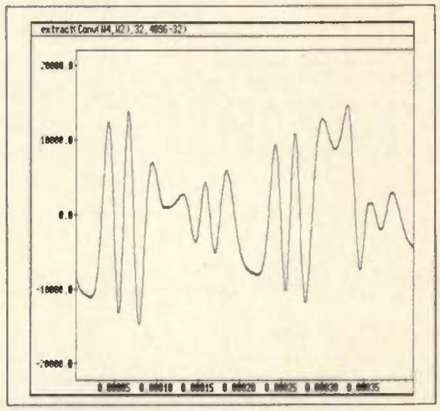

Figure 11b. I and Q signals.

HP8116 function generator and internally phase modulated with a 5 kHz sinewave (peak deviation 2π radians). The center frequency was set to 12.5 MHz.

To understand how to demodulate these signals, let's look at quadrature-carrier description of an arbitrary IF signal (Figure 8). We start by defining axes which represent our in-phase and quadrature local oscillators. An incoming IF signal can be resolved into components which lie along these axes — in fact this is what our mixers will do. The incoming signal has instantaneous amplitude R(t) and phase $\phi(t)$.

To recover the AM signal, recognize that R(t) represents the demodulated AM signal. We will compute R(t) from:

$$R(t) = \sqrt{i(t)^2 + q(t)^2} \tag{1}$$

An important point to make is that we do not need to phase lock our sample clock (thus our local oscillator frequency) to the IF carrier frequency. If the IF carrier and local oscillator frequencies are at different frequencies, the vector representing the IF signal will be spinning relative to the I and Q reference axes — that is the instantaneous phase will be changing. However the equation for R(t) is true no matter what value $\mu(t)$ assumes.

To recover the phase modulation, we could apply the relation:

$$\phi(t) = \arctan\left(\frac{q(t)}{i(i)}\right) \tag{2}$$

but this falls apart for large phase deviations.

A way out of this problem is to develop an FM detector:

$$\phi'(t) = \left(\frac{1}{1+\left(\frac{q(t)}{i(t)}\right)^2}\right)\left(\frac{i(t)q'(t) - q(t)i'(t)}{i^2(t)}\right) \tag{3}$$

$$= \frac{i(t)q'(t) - q(t)i'(t)}{R^2(t)}$$

The FM detector output is integrated to produce the PM signal. When implemented with DSP, a discrete-time approximation for the derivatives will be used, as will a discrete-time approximation for integration.

The results are given in Figures 9-14. The IF signal is sampled (Figure 9), and mixed as described earlier (by sending alternate samples to the I and Q arrays and alternately digitally multiplying the I and Q samples by 1 and -1). The results are filtered using a 32 order finite-impulse response filter whose response is given in Figure 10. The I and Q signals are shown in Figure 11. Finally, the amplitude modulation is recovered by using Equation 1 (Figure 12) and the phase modulation is recovered by FM detection using Equation 3 (Figure 13) and integrating (Figure 14).

In this example, signal to noise ratio could have been increased by lowering the bandwidth of the low pass filters after the mixers. Assuming a 10 MSPS sample rate, the I and Q channel each receive 5 MSPS for a Nyquist frequency of 2.5 MHz. If we had reduced the low pass filter bandwidths so that the noise equivalent bandwidth was approximately 25 kHz, signal-to-noise ratio would increase by approximately 10log(100) = 20 dB for a net signal to noise ratio of over 80 dB. This is based on the fact that noise power spectral density of ADCs are approximately uniform. For other modulation formats, signal to noise ratio improvements (through processing gain) may be available.

The example shows only a fraction of the potential of this technique.

Summary

When high speed ADCs, such as the CLC926, as combined with state of the

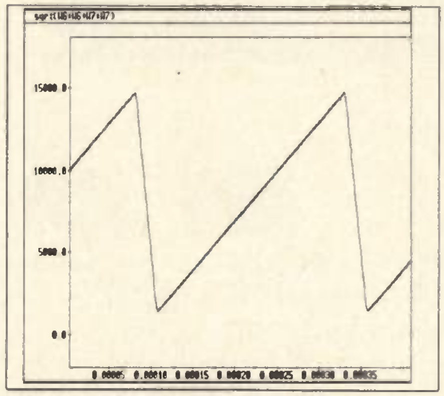

Figure 12. Recovered AM.

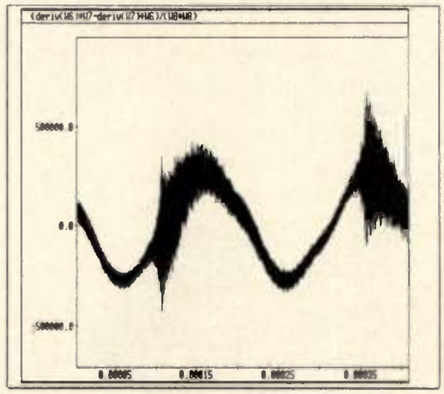

Figure 13. FM detection.

art track-hold amplifiers, such as the CLC940, previously unattainable ADC performance is reached. With this new potential comes the ability to digitally implement I/Q demodulators at common IF frequencies (such as 21.4 MHz).

CLC926-based I/Q sampling when teamed up with modern DSP has the potential for accurate, stable, repeatable, and flexible performance.

Acknowledgement

Thomas Hack would like to acknowledge DSP Development Corporation, Cambridge, Mass., for use of their DADiSP Software in this article. ***RF***

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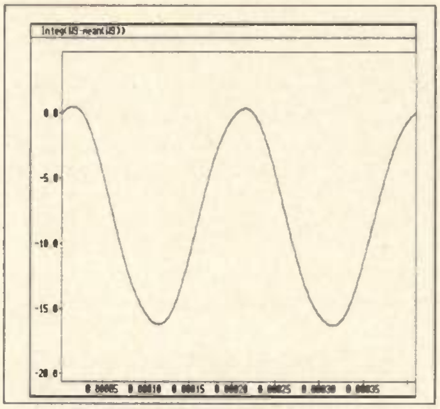

Figure 14. PM detection.

About the Author

Thomas Hack is an Applications Engineer with the Data Conversion Group of Comlinear Corporation.. He received his BSEE from Cooper Union, his MSEE from Renesselaer Polytechnic Institute, and his MBA from the University of Colorado. He can be reached at (303) 226-0500 and his address is 4800 Wheaton Drive, Fort Collins, CO 80525.

Appendix: A Quick Tour of IF Sampling

Figure 15 diagrams a common way of looking at the sampling behavior of an ADC. x(t), a bandlimited analog signal (from DC to some arbitrary frequency), is multiplied by sampling function s(t) simulating ADC sampling. At an assumed sample rate of 10 MSPS, the ADC aperture is substantially smaller than the sample period so s(t) can be approximated by a series of unit impulses spaced 100 ns apart (Figure 16). The multiplier produces a series of weighted impulses corresponding to discrete-time samples of the analog input. The actual observable samples at the ADC output are of course digital, not analog.

Figure 17 shows the same signals, but in the frequency domain. We will initially assume that the bandwidth of the analog input X(f) is less than 5 MHz (shown by the solid line). The sampling function described earlier produces a comb spectrum S(f) with spikes at DC and integer multiples of 10 MHz. Since the samples' output is produced by multiplying x(t) by s(t), the frequency domain representation of the output is obtained by convolving X(f) with S(f). The result is the repeating spectrum shown in Figure 17c — a result of harmonic mixing.

Nyquist's sampling theorem recognizes that alias-free sampling occurs when the sidebands at each of the harmonics of the sampling frequency don't share the same frequencies (overlap). For example, if the analog input bandwidth were higher as depicted by the dotted lines in Figure 17, aliasing occurs due to sideband overlapping (shown by the cross hatched

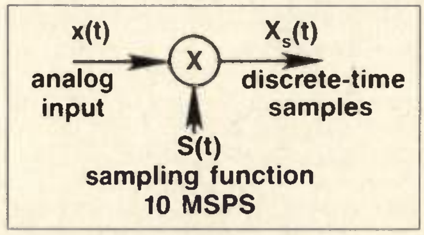

Figure 15. Sampling model.

regions in Figure 17c).

From this concept comes Nyquist's sampling theorem which states that the sampling frequency must be twice the bandwidth of the signal to be sampled. This is often interpreted to mean that the highest signal frequency must be less than half the sampling frequency or 5 MHz in this example. For video sampling —where the signal extends from virtually zero frequency to some upper frequency, as it has in our example, this would be correct.

But what happens for narrow-band signals, such as we might find in a receiver IF? Figure 18 shows the results from passing a narrow-band IF signal into the same ADC. We start off with an IF signal centered at 17.5 MHz — well above half the sampling frequency. To make things general, the signal is assumed to have independent lower and upper sidebands. As a result of the same sampling action as before, the sampled spectrum has these sidebands repeated at 5 MHz intervals. One of these sidebands will always occur below Nyquist frequency. (Hence IF sampling can replace analog mixers in some instances.)

To prevent overlapping sidebands we must prevent the sidebands from extending greater than ± 2.5 MHz from each center frequency. Since each sideband is independent of the other, the net bandwidth available is the same as before, 5 MHz! In order to attain the alias-free bandwidth claimed by Nyquist's sampling theorem, carefully choose the sampling clock frequency based on the center frequency of the IF.

In the example, we never completely bring the sidebands down to baseband — the lowest mixer products are centered at 2.5 MHz (resulting from mixing of the 2nd harmonic of the sampling frequency with the IF signal). In many instances, there may not be a need to mix to baseband, or it may not be desirable.

A few final points: 1) we could have just as easily placed our signal at another frequency and produced the same output from the ADC without changing the clock frequency. 2) For the interested, a similar analysis can be used to explain sequential repetitive sampling in oscilloscopes.

Our conclusions:

1) IF sampling doesn't violate Nyquist's sampling theorem. The required sample rate is determined from the bandwidth of the IF signal, not the center frequency.

2) To get the most from the technique, sampling clocks should be carefully selected.

3) To select suitable sampling frequencies determine the lowest sampling frequency based on Nyquist's sampling theorem and the bandwidth of the IF signal. Then look at how the IF signal mixes with harmonics of the sampling clock and adjust the sampling frequency relative to the IF frequency so that sidebands don't overlap.

4) The results are slightly different in the case where the upper and lower sidebands are not independent and our conclusions are altered somewhat.

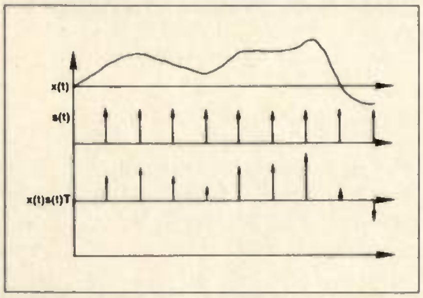

Figure 16. Sampling in the time domain.

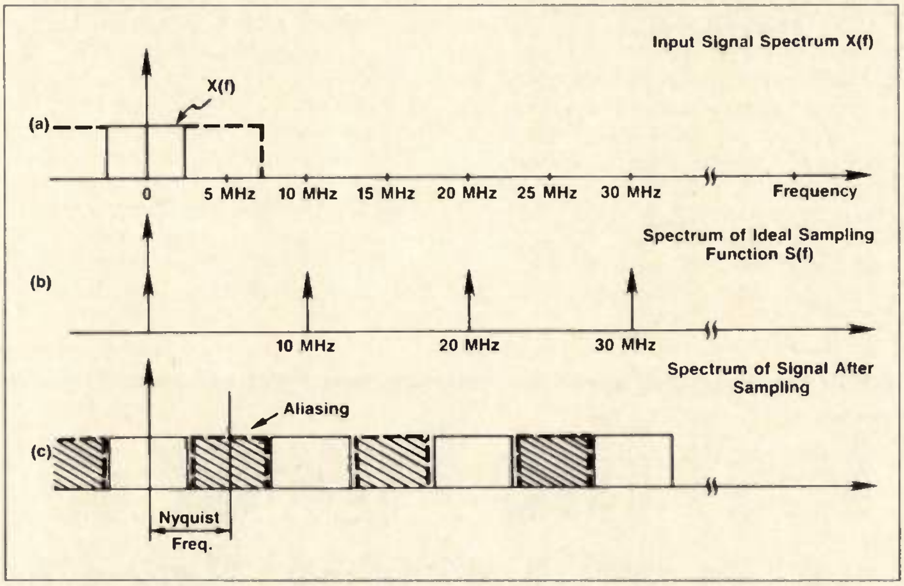

Figure 17. Video sampling: frequency domain representation.

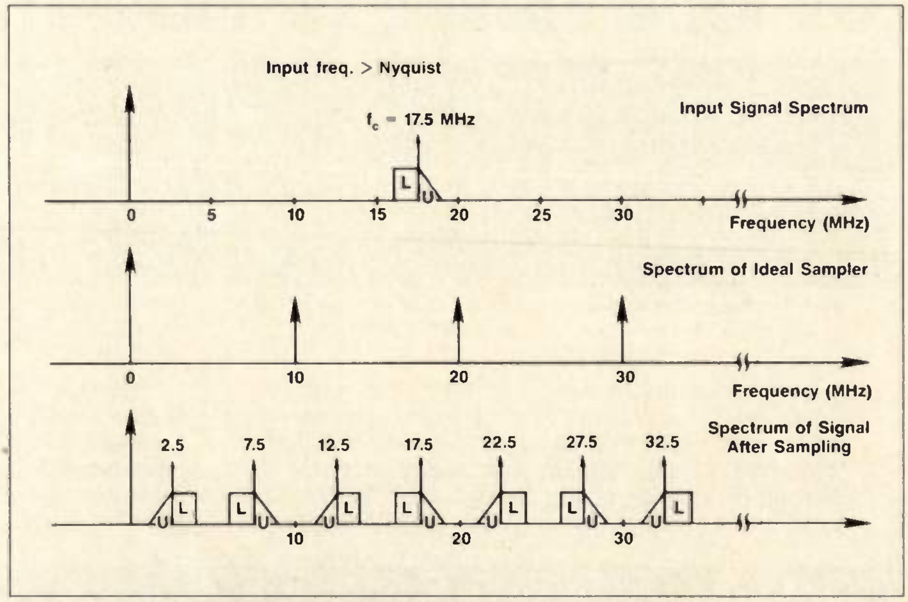

Figure 18. IF sampling: frequency domain representation.

PIN Diodes Handle Multi-kilowatts at HF/VHF

M/A-COM introduces new KILOVOLT™ PIN diodes, designed for use as high power switching elements in multi-kilowatt HF and VHF applications. MRI switches, filter switches, antenna couplers and power amplifier bypass switching are some of the uses targeted for these components. Long carrier lifetime of 10 usec for the 2000-volt version and 20 usec for the 3000-volt model assures operation at frequencies as low as 1 MHz. Maximum series resistance (4 MHz, with 0.5 A forward current) for each version is .20 ohms and .25 ohms, respectively. The diodes will handle 25 amps forward current, DC and RF. M/A-COM uses ultra high resistivity silicon material for the intrinsic region, on which P+ and N+ regions are deposited using an epitaxial process. Glass passivation results in a hermetically sealed chip that has been accepted in many hi-rel military programs. The diodes are available in pill, stud, or insulated stud packages with or without a solder lug on the anode.

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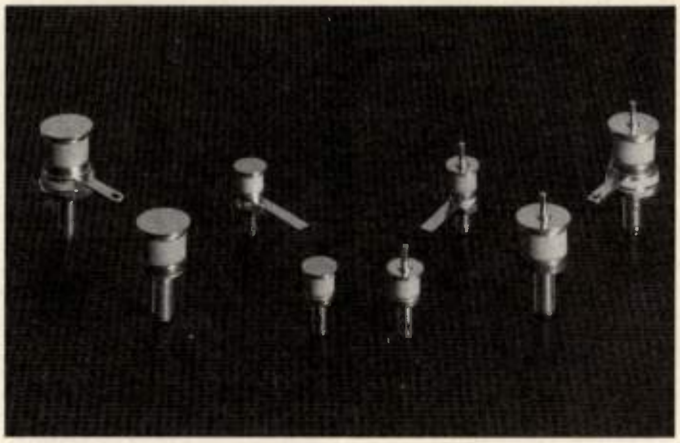

Personal Radiation Safety Monitor

A non-ionizing radiation monitor, the Nardalert™ Personal Monitor, Model 8840B, has been introduced by Loral Microwave-Narda. With this monitor, workers in potentially hazardous areas can have a hands-free, full-time sensor to warn of illumination by high

power radar, communications systems or EW systems. The small, wearable, pocket-sized unit allows accurate monitoring from 2 to 18 GHz for use in areas where conventional meter and probe are not viable solutions for protection. Thermocouple-based detection is used for true RMS average field sensing, avoiding multiple signal or modulation errors which might generate false alarms. Built-In-Test circuitry verifies operation each time it is turned on, and a low battery indicator warns when replacement is required.

Loral Microwave-Narda
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Function Generator Synthesizes Arbitrary Signals

The DS345 Synthesized Function Generator from Stanford Research Systems offers digitally synthesized waveforms to 30 MHz with 1 uHz resolution. Outputs can be simple sine, triangle, ramp or square waves, or arbitrary signals with up to 16,300 (12 bits) and sampling times to 25 nsec can be generated. Internally synthesized modulation capabilities include phase continuous linear and logarithmic frequency sweeps, plus amplitude, frequency, phase and burst modulation. Complex applications include binary shift keying, phase shift keying and arbitrary waveform generation. Rear panel trigger outputs, external input for amplitude modulation, sweep and span markers, IEEE-488 and RS-232 interfaces and external timebase inputs are additional features. Spurious components are better than −55 dBc. Output level is 10 mV to 10 V_{p-p}. List price is $1895, computer interface with Arbitrary Waveform Composer software is $495, and a high stability timebase option is $650.

Stanford Research Systems
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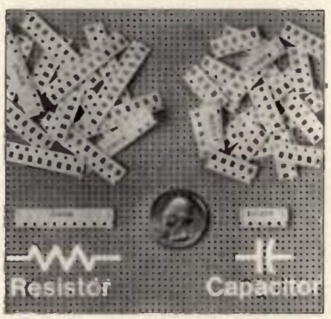
SMT Resistors and Capacitors Available in Small Quantities

Communications Specialists now offers surface mount resistors and components in small quantities and individual values. These components are offered to service shops and design labs which typically cannot meet the minimum purchase requirements of major component distributors. The unit of sale is per strip. Resistors are furnished in strips of 10 and sell for $2.50 per strip. Capacitors are furnished in strips of 5 and sell for $1.25 per strip.



Values are clearly marked on each strip. Minimum order is $10.00. Available resistors and capacitors are the same as those provided in the company's CC-1 and CR-1 prototyping kits, which generally are standard 5 percent resistor values, and 10 percent ceramic capacitor values.

Communications Specialists, Inc.
INFO/CARD #247

PC-Based Counter Operates With Windows 3.0

A new 10 Hz to 2.4 GHz counter has been introduced by Optoelectronics. The PC-10 Universal Frequency Counter-Timer is in the form of a 9-inch card for personal and laptop computers, using Windows 3.0 as a control panel and display window. Input signals of 10 mV or greater are accepted and displayed with up to 10 digit resolution. Input gating is continuously variable from 1 msec to 28 seconds for optimum measurement time and resolution. Measurement modes include discrete and average frequency, pulse width, time interval, period, and ratio between two frequencies. The counter can also control a radio receiver, such as the ICOM R7000, for operation faster than the typical scanner, or automatically tuning the radio to a detected signal frequency for surveillance applications. The PC-10 is priced at $335 each.

Optoelectronics Inc.
INFO/CARD #246

Frequency Multipliers

KW Microwave announces a series of active and passive frequency multipliers with integrated amplifiers, featuring multiplication factors from 2 to 96, with minimum input frequency of 1 MHz and maximum output frequency of 12 GHz. An example multiplier with a multiplication factor of 16 has an input frequency of 450 MHz and an output of 7.2 GHz. Typical input power is +23 dBm and typical output power is 0 dBm. Spurious and harmonic signals are −50 dBc. Multipliers are available in plug-in, drop-in or connectorized versions.

KW Microwave Corp.
INFO/CARD #245

Large Area Diamond Heat Sinks

Thick, large area chemical vapor deposited (CVD) diamond heat sinks are now being produced for power device and high speed applications. In dimensions up to 25 mm square and 0.5 mm thick, diamond heat sinks offer thermal conductivity about 2.5 times that of copper and 5 times that of beryllia at 100 C. For price sensitive applications, diamond-coated silicon substrates are also available.

Crystallume
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120 MHz Synthesizer IC

GEC Plessey Semiconductors introduces the NJ88C33 synthesizer, designed for operation from 2.5 to 5.5 volts with 3 mA current consumption. The device may be operated as a single-chip 120 MHz synthesizer, or with an external prescaler for operation to

frequencies beyond 2 GHz. On-chip are programmable 16-bit 'R', 12-bit 'N' and 7-bit 'A' counters which are addressed by an I2C bus. The bus will operate up to 2 MHz and can achieve channel loading in 20 us. In 100s, the price is $7.87.

GEC Plessey Semiconductors
INFO/CARD #243

GaAsFET Switch Matrix

A 5-in, 2-out GaAsFET switch matrix is introduced by K&L Microwave, covering DC-2 GHz, with TTL compatible control inputs. Switching speed is typically 40 nsec, insertion loss is 4.5 dB maximum, isolation is 60 dB, and VSWR is a maximum of 1.5:1. The matrix is enclosed in a hermetically sealed 3 x 1.5 x 0.4 inch enclosure.

K&L Microwave, Inc.
INFO/CARD #242

Communications Test Set

Motorola announces the R-2600A Communications System Analyzer, with cellular testing capability. Available cellular formats include EAMPS, ETACS, JTACS and NTACS. Test functions include sweep generator, spectrum analyzer, receiver, digital storage oscilloscope, terminated wattmeter, counter, signalling generator and decoder, SINAD meter, distortion meter, AC/DC voltmeter, and audio generator. The display includes softkeys, windowing and fast autoranging scales for ease of operation.

Motorola Government Electronics Group
INFO/CARD #241

DC- 4 GHz Differential Amplifier

The AM-4001, a stand-alone differential amplifier has been introduced by Tektronix. This pulse amplifier has a DC-4 GHz bandwidth with 18 dB gain and gain flatness of less than 1 dB from 1 MHz to 2.5 GHz. Available output level is ±2 V into 50 ohms from each output. The amplifier is designed for applications requiring accurate time domain characteristics such as preamplification

for real-time or sampling oscilloscopes and spectrum analyzers. Applications with ill-defined ground can benefit from using this unit, such as high speed logic, gigabit data transmission, and advanced radar systems. Price of the AM-4001 is $4000.
Tektronix, Inc.
INFO/CARD #240

Nonvolatile, Programmable Components

A family of serially programmable, nonvolatile integrated circuits has been introduced by Hughes' Semiconductor Products Center. These parts are designed for circuits requiring precision tuning, trimming, or reconfiguring. The first nine types of devices include: an eight-output alternative to DIP switches, a quad analog switch, eight-channel analog data selector, two trimmer potentiometers, a trimmer capacitor, sine/cosine generator, security code generator/detector, and a 16-bit divide-by-N counter.
Hughes Aircraft Company
INFO/CARD #239

Eight-Line EMI Filter

Better than 15 dB attenuation of common mode noise in up to eight data lines can be obtained with the Coilcraft CCDLF-8000 surface mount filter. By passing all signals through a single magnetic structure, these filters remove common and differential mode noise. Typical inductance is 5 uH per winding, maximum DC resistance is 250 milliohms, and current capacity is 100 mA. In 10,000 production quantities, the price is $2.00 each.
Coilcraft
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Mini Pak DBM

Vari-L introduces a double balanced mixer in the smallest available hermetic package, the model DBM-145. Frequency coverage is 10-1500 MHz with 5 dB conversion loss and VSWR of 1.5:1 at mid-band. LO drive is +7 dBm, and the third order intercept point is +12 dBm. In addition to frequency conversion, applications include bi-phase modulators, pulse modulators, phase detectors and frequency doublers.
Vari-L Company, Inc.
INFO/CARD #237

PC-based Counter

The GT100 from Guide Technology is a low cost frequency counter for automated test systems. The counter includes a full-size PC card and a DOS-based software package which provides a virtual front panel on any XT, AT or compatible PC. 100 MHz or 1.3 GHz models are available, with selectable gate time, fast frequency measurement, time interval, pulse width, period, ratio, and totalize functions. Price of the GT100 is $995 for the 100 MHz model and $1745 for the 1.3 GHz version.
Guide Technology Inc.
INFO/CARD #236

SMT Crystal

Fox Electronics offers a ceramic surface mounted crystal in a package compatible with existing SMT designs. The FC model is available in 10 MHz - 50 MHz frequency with standard tolerance of ± 50 PPM at 25 C and aging of 5 PPM/year, maximum. Shunt capacitance is 7 pF maximum, and equivalent series resistance (ESR) is 50 ohms max., 10-25 MHz, and 100 ohms max., 25-50 MHz.
Fox Electronics
INFO/CARD #235

LC Filters for HDTV

The HBF line of video filters and delay lines for high definition video and precision imaging is available from Toko America. Designed for very precise amplitude and group delay characteristics, lowpass filters are available with cutoff frequencies from 5 to 36 MHz, and delay lines are available with 300 to 600 ns delay times. Prices range from $202 in single piece sample quantities.
Toko America, Inc.
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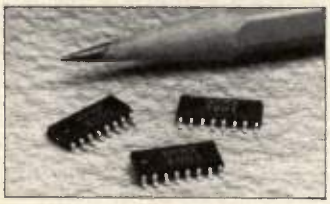
Plug-In Attenuator

JFW Industries has added the model P50-021 to their line of hybrid programmable attenuators. Attenuation range is 0-31 dB in 1 dB steps, frequency range is 50-250 MHz, and switching speed is 5 usec. Maximum VSWR

is 1.5:1, maximum input power is +15 dBm, and attenuation accuracy is ±0.5 dB or 2 percent. Control is CMOS compatible. Other models and custom configurations are available.
JFW Industries, Inc.
INFO/CARD #233

Monolithic QPSK Modulator

The first of a new series of RF ICs, the HPMX-2001 QPSK modulator has been introduced by Hewlett-Packard Co. The device incorporates twin double balanced modulators cross-coupled into a summing amplifier, plus a standby circuit for controlling current sources. Applications include digital cellular and digital cordless telephone equipment, as well as other applications requiring



amplitude modulation, single-sideband transmission or vector generation. Pricing is $7.25 for 1-99 units.
Hewlett-Packard Company
INFO/CARD #232

1 MV Pulser System

A pulser system generating 250 kV to 1 MV pulses with a risetime of 65 nsec and duration of up to 10 ms is announced by Maxwell Laboratories. The Model 40404 Marx Generator can be used as in high power microwave sources, for accelerating electron beams, in pulsed lasers, flash X-ray units or for general high-voltage laboratory use. Low jitter and droop, and low overshoot are additional features of this design.
Maxwell Laboratories, Inc.
INFO/CARD #231

Low Loss SAW Filter

Thomson-ICS has developed a 71 MHz SAW filter featuring less than 10 dB insertion loss and single component tuning circuits, intended for mobile radio applications. Bandwidth is 170 kHz with a shape factor of approximately

4:1. Packaging is a 1.4 × 0.5 × 0.2 inch DIP.
Thomson-ICS Corp.
INFO/CARD #230

ECL Compatible Oscillators

A new VHF miniature oscillator covering 200-500 MHz is available from Murata Erie. Outputs include ECLinPS differential, 100K logic, and sine wave (optional). Stability is ± 50 PPM to ± 100 PPM from 0 to +70 C. Other features include low jitter and excellent waveform symmetry. Applications include high resolution graphics, CPUs, fiber optics and communications.
Murata Erie North America
INFO/CARD #229

Tunable Notch Filters

A new line of tunable notch filters from Trilithic covers a range of 25 to 1000 MHz with octave band units, offering notch depths up to 50 dB. 3 dB bandwidth varies from 1 to 16 MHz, depending on the octave band selected, while passband loss is 1 dB maximum. Models are available in 50 or 75 ohm impedances, with type N, SMA, TNC or BNC connectors.
Trilithic
INFO/CARD #228

Mica Chip Capacitors

Type MC mica chip capacitors are now available from Cornell Dubilier. Natural mica dielectric retains high Q up to self-resonant frequencies exceeding 1 GHz. Capacitance ranging from 1 pF to 2000 pF (100 V) or 1200 pF (500 V) and tolerances as tight as ±0.25 pF or ±0.25 percent are offered. Bulk or reel packaging is available. OEM quantity pricing starts a $0.20 each.
Cornell Dubilier Electronics
INFO/CARD #227

RF Prototyping Board

Ivanboard is a universal circuit board for quick, reliable construction of all types of electronic circuits. Using a grid of pads with 0.1 inch spacing on one side, and a solid low-impedance ground plane on the opposite side, stable

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RF, analog and digital circuits can be built.
Europa Marketing
INFO/CARD #226

Attenuator Uses ACT Drivers

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KDI/Triangle Electronics
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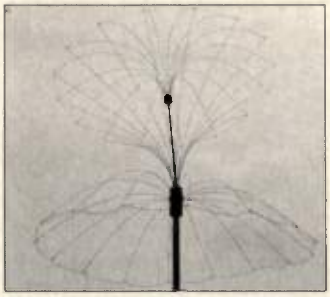

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Temperature Range .... -54 to + 85°C
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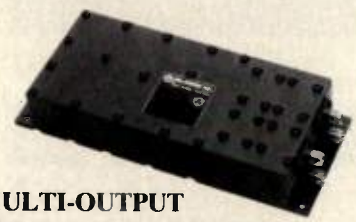

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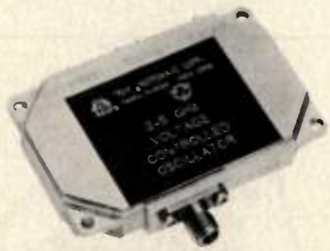
Connectors
RF Output .... SMA (Female)
Power Supply .... Solder Terminal
Power Supply .... +15 V DC ±5%



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Specifications

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Harmonics .... -15 dBc
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INFO/CARD 50

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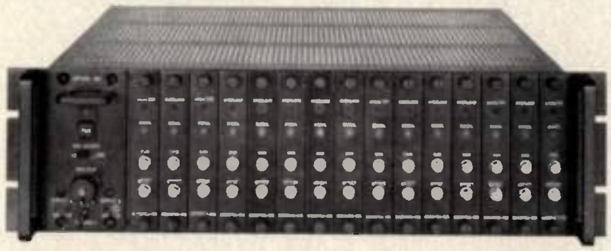

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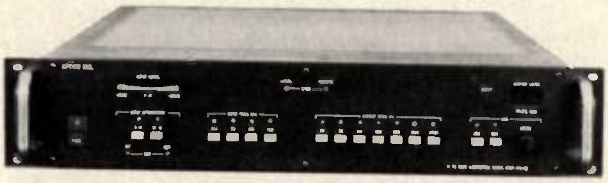

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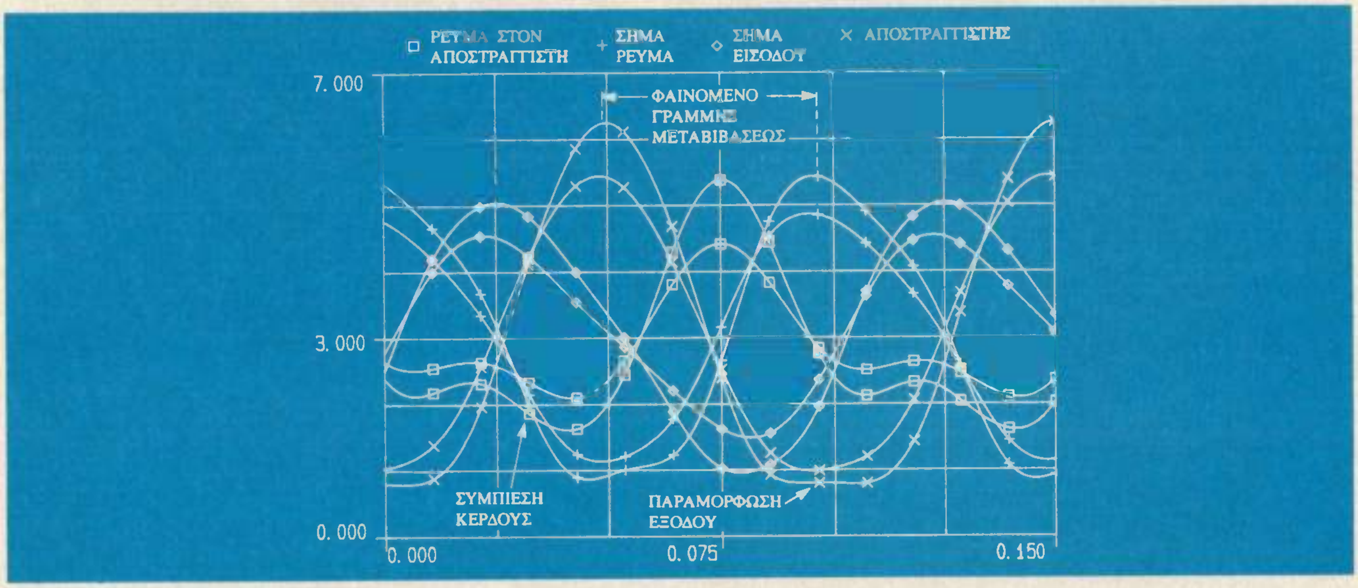

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INFO/CARD 22

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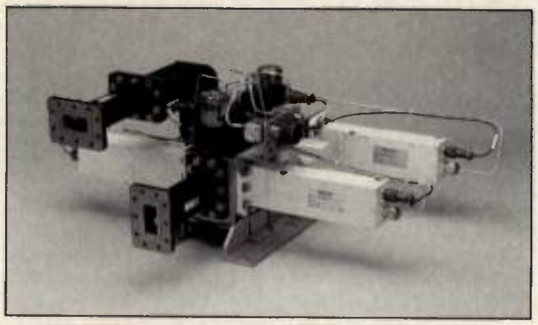

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Communications Range and Reliability of Part 15 Devices

By Bernard Kasmir
ADEMCO

Low power communication devices authorized under FCC Part 15 have signficant perfomance limitations. Some of these limitations are a result of regulations, and others are part of the mechanical and electrical design, cost, and complexity requirements of the manufacturer. The key performance concerns from the user's perspective are range and reliability. Part 15 device designers have the task of dealing with the limitations while maximizing range and reliability.

When we speak of radio range, we talk in terms of how far we can have communications. The definition becomes ambiguous unless we define initial conditions. There are, for example, best case, worst case, and typical performance; or the more generalized probability of communications under defined conditions.

When using Part 15 low power communication devices for the protection of life and property, reliable range, however it is defined, becomes a significant specification. The typical transmitter is a small PC board housed in some type of plastic case. Usually, the transmitter does not have an external antenna. Indeed, the antenna can often not be completely defined since it consists of a printed loop plus radiation from all components including the battery, and sometimes any external sensor wires.

The FCC specifies "best case" signal level, which is measured by finding the maximum level of radiation considering all possible orientation positions. This signal is expressed in field intensity in microvolts per meter at some specified distance (usually 3 or 10 meters). The FCC is not concerned what type of antenna is used (if any) as long as the best case signal level is maintained.

Radio receivers are usually measured for sensitivity in a 50 ohm system. However, the ability of a receiver to pick up a signal is also determined by the receiver antenna. The true measure of the receiver sensitivity is the ability to intercept a given field intensity. The actual receiver sensitivity can then be expressed in microvolts/meter. The antenna factor of the receiving antenna expresses the ability of the antenna to generate a voltage into 50 ohms for a particular field intensity. In other words:

AF = E/V

where, E is the field intensity in uV/m and V is voltage into 50 ohms. The lower the antenna factor the more voltage is delivered to the receiver for a particular field intensity.

Since the transmitter is specified in microvolts per meter (at some defined distance) and the receiver sensitivity can be defined in terms of field intensity in uV/m, a theoretical best case range can be defined by the use of linear interpolation. Thus, the field intensity (E2) at distance of S2 would be:

$$E2 = E1 \times (S1/S2) \quad (1)$$

In this case, S1 would be the distance where the transmitter field intensity E1 is specified, usually 3 meters. This shown in Figure 1 as the "1/X" curve where signal level (voltage) is halved as the distance is doubled.

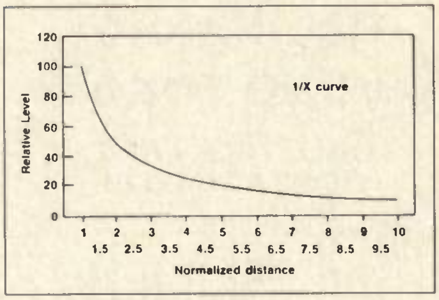

Figure 1. Signal level vs. distance (1/X curve).

The transmitter signal level as specified by the FCC is dependent upon frequency and duty cycle. Part 15.202 lists signal levels as follows: from 260 to 470 MHz, 1500-5000 uV/m. The field intensity at any operating frequency can be found by linear interpolation of this range. The field intensity E (in uV/m at 3 meters) at frequency F is equal to:

$$E = 16.67\,(F-260) + 1500$$

Duty cycle

This is determined by the average-to-peak value of signal level. The FCC

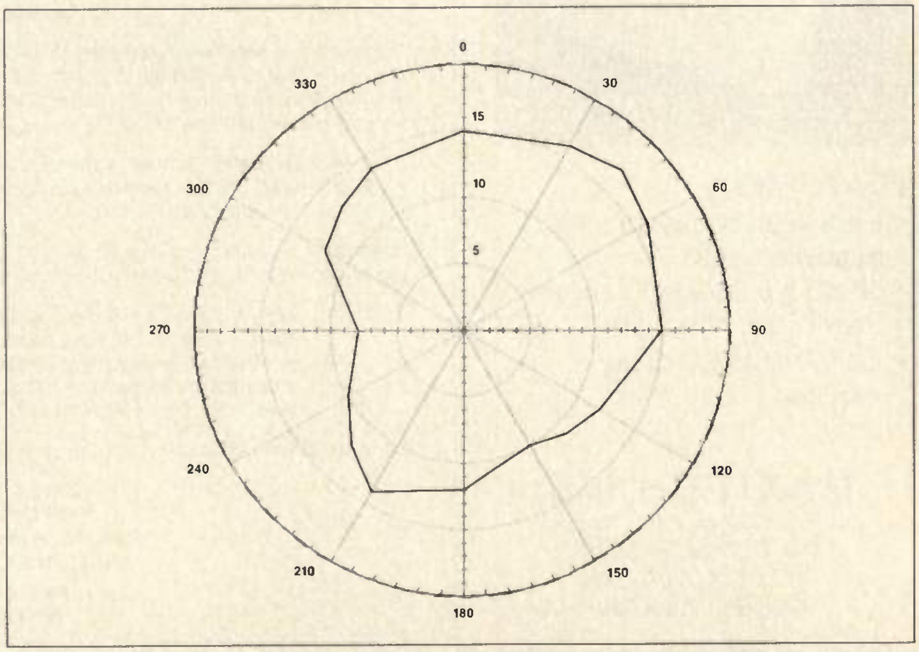

Figure 2. Signal level vs. orientation for a Part 15 transmitter.

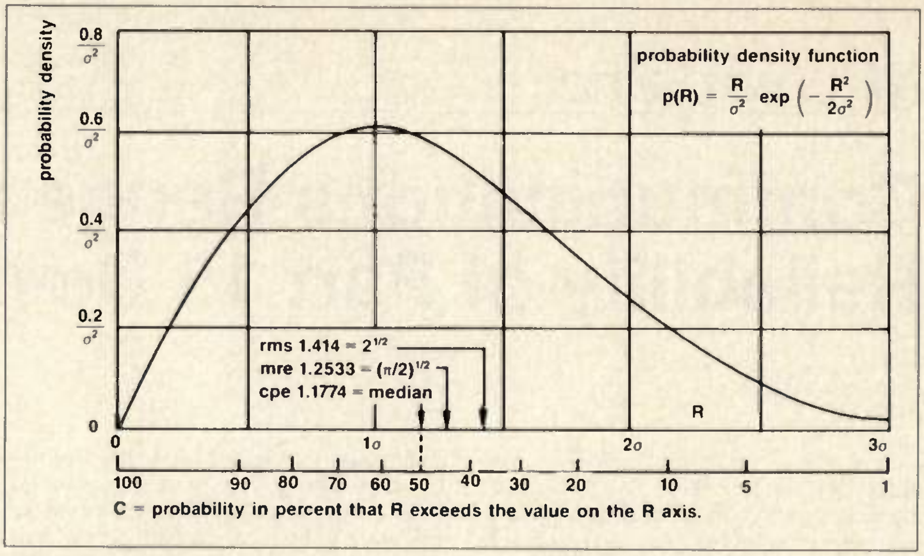

Figure 3. Rayleigh probability distribution model.

allows up to 20 dB (10x) peak signal when pulse modulated and averaged over 100 ms. In the following example we shall calculate range based upon some initial values. The values chosen do not represent any specific product, but rather typical values used in this tutorial.

For example, at 300 MHz, the average signal level allowed would be 2166 uV/m. Assume a duty cycle of .2 (the signal is on 20% of the time averaged over 100 ms). This yields a duty cycle advantage of 5, bringing the peak signal level to 10,830 uV/m at 3 meters.

Also, for this example, assume a receiver with a 30 dB antenna factor (numerically 31.6) and a receiver with a 2 microvolt sensitivity into 50 ohms. The receiver sensitivity in terms of microvolts per meter would be;

$31.6 \times 2 = 63.2$ uV/m

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The "best case" range would then be, from (1):

10,820 × 3 / 63.2 = 513 meters, or 1650 feet.

A typical distance between the control panel and a remote transmitter in a house may be 50 feet. It certainly appears as if we have plenty of signal to spare. In this example, we have about 30 dB signal margin, that is, a best case signal level 30 dB about threshold (20 Log 1650/50).

However, there is a big difference between theoretical best case and actual operation. As we shall see, there are many hazards that tend to reduce the range of a system.

Orientation

Figure 2 shows relative radiated field intensity of a transmitter as a function of orientation or position. The ideal antenna pattern would be omnidirectional (equal signal level in all directions), but this is not practical to achieve. As a result, the field intensity at the receiver is dependent upon the position of the transmitter. This field pattern is also influenced by building materials where the transmitter is mounted.

As with the transmitter, the receiver antenna can also exhibit directional behavior, with varying sensitivity depending on its orientation. In cases where the receiver location is fixed and the transmitter does not move over a wide azimuth relative to the receiver, fairly constant performance can be expected. If neither transmitter nor receiver are fixed, the variation can exceed that shown in Figure 2.

Attenuation

Building materials are translucent (not transparent) to radio waves at the frequencies typically used by Part 15 devices. A signal from a remote transmitter to a receiver will experience some attenuation greater than free space path loss. Studies have been conducted, but some controversy remains on levels of attenuation.

Interference

More and more electronic devices are used in homes and businesses. Devices such as digital clocks, computers, VCRs and even other receivers can cause interference. Although emissions are regulated, the close proximity of an electronic device to a radio receiver can cause significant interference. This results in reduced receiver sensitivity as well as reduced range.

We usually think in terms of deterministic models where performance can be precisely calculated. In the field of communications, there is some randomness of events (apparent randomness due the complexity of the signal propagation) and we cannot obtain a precise calculation. Rather, it is more useful to work with the probability of an event over a defined interval. The signal attenuation caused by the three factors discussed above cannot be given precisely. However, unless we obtain a probabilistic model, we can only operate using intelligent guesses.

Saving the best for last, the most significant specific phenomena that produces variability in received signal level is multiple reflections. The received signal level is a vector of all direct and reflected signals. This has been mathematically modeled and is represented by the Rayleigh probability distribution function. This model is comprised of an infinite number of small reflectors. Figure 3 shows the shape of this curve, which is called a log-normal curve. The mean represents zero signal attenuation. There is a 50 percent probability that the signal is enhanced anywhere up to 6 dB (left side of the curve). If reflections are all completely in phase the signal can be enhanced up a maximum of 6 dB. However, complete cancellation can theoretically result in infinite signal attenuation.

Table 1 summarizes the probability of additional attenuation caused by multiple reflections.

What does this mean? First, there is a 50 percent probability that the additional attenuation caused by multiple reflections will be 0 dB or less. There is a 90 percent probability that the additional attenuation caused by multiple reflections will be 10 dB or less. Similarly, the chart covers 20, 30 and 40 dB attenuation.

Attenuation	Probability
0 DB	50%
10 DB	90%
20 DB	99%
30 DB	99.9%
40 DB	99.99%

Table 1.

Take any one of the above chart, 20 dB for example. Since there is a 99 percent probability that the additional attenuation is 20 dB or less, then, if we have a 20 dB margin, there is a 99 percent probability that the signal will get through. By knowing the signal margin of a system, we can predict the reliability.

Now, back to our example; since we do not have a good model for items such as attenuation, orientation and interference, let us assign a "best guess" of 10 dB for these above items. This reduces our previously determined signal margin from 30 dB to 20 dB. According to the above table, we have a 20 dB signal margin and a 99 percent reliable system.

Conclusion

To assure maximum systems reliability, more than basic transmitter and receiver performance is required. It is important to use good installation practices to minimize the added attenuation due to the factors covered in this article. There are many techniques used to assure maximum received signal. Advantageous orientation, a clear signal path, no interference sources should be combined with predictable transmitter and receiver performance, including duty cycle and peak power, receiver sensitivity, and antenna factor. Subsequent articles will address these techniques. ***RF***

About the Author

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Parasitic Positive Feedback Frequency Acquisition in a PLL

By Jonathon Y.C. Cheah
Hughes Network Systems

When the use of a double balanced mixer as a phase comparator is necessary, frequency acquisition is an inherent problem. Double balanced mixers are commonly used in PLLs with a very high detector frequency or circuits such as analog Costas' loops where digital frequency/phase comparators are not suitable.

The parasitic positive feedback frequency acquisition scheme is not well known (1). A probable reason is that the direct application of this concept has the problems of achieving reliable sweep oscillation and controlling the sweep waveform. Therefore, its use is limited (2). A practical implementation that overcomes these problems is described here. This implementation fully demonstrates the automatic scan and stop action quality of this concept as well as its simplicity. As the feedback circuit is independent of the PLL's parameters other than the sweep rate, this circuit can, in many cases, be a "boiler plate" add-on to where it is needed. In areas such as Costas' loop carrier recovery circuits, where complex frequency scanning is needed to achieve frequency acquisition, this circuit can be a good candidate.

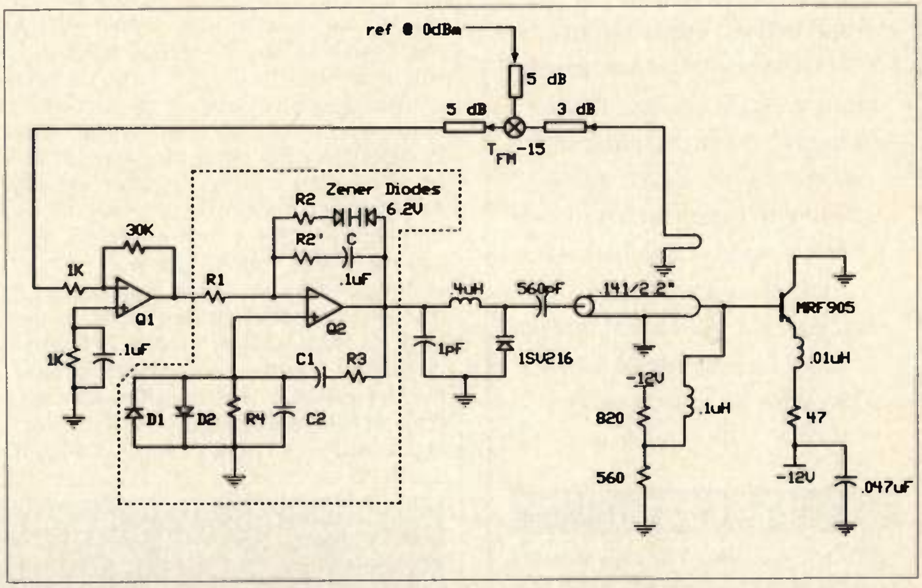

Figure 1. The box indicated by dotted lines shows the 2nd order loop filter with added positive feedback circuitry. The component designators are as used in the test and their values are provided.

The frequency acquisition design can be described as a Wein-Bridge oscillator superimposed on a 2nd order loop filter. When the loop is unlocked, the positive feedback oscillator is activated to provide frequency sweeping action. When the PLL attains the lock condition, the required negative gain feedback to sustain a continual oscillation condition is

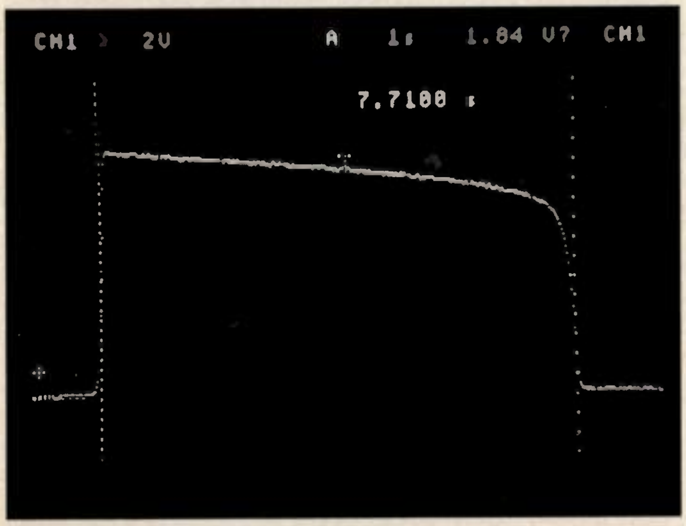

Figure 2. The sweep voltage at the output of the 2nd op-amp when the PLL is unlocked.

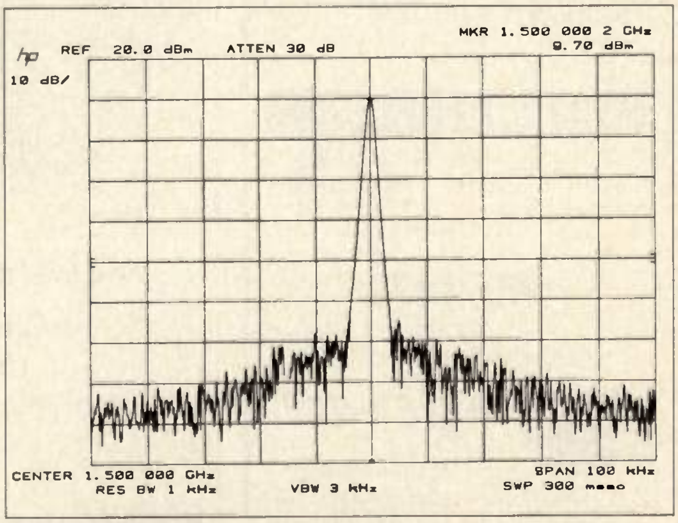

Figure 3. The frequency spectrum of the VCO when the loop is locked.

discontinued, while the infinite DC gain required by the phase lock loop is maintained.

Consider the transfer function of an active 2nd order PLL:

$$F(s) = \frac{s\tau_2 + 1}{s\tau_1} \quad \text{where} \quad \left\{ \begin{array}{l} \tau_1 = R_1C \\ \tau_2 = R_2C \end{array} \right. \qquad (1)$$

as shown in Figure 1.

When the frequency offset is large, the low pass loop filter effectively reduces the gain of the op-amp. For microwave VCOs, the offset frequency will most likely be beyond the 2nd pole of the op-amp's frequency response. However, at DC, the amplification gain is essentially the open loop gain of the op-amp. The output voltage of the amplifier will therefore reach the positive or negative rails of its maximum DC voltage swing. Referring to Figure 1, C is designed to be a high impedance at the sweep oscillation frequency. If the back to back Zener diodes have the combined Zener and forward diode biasing voltages less than the rail voltages of the op-amp and series resistor, R_2, provides a negative feedback gain of 2 or higher, the Wein-Bridge oscillator gain condition of $G(j\omega) \geq 1$ is satisfied where:

$G(j\omega) =$

$$\frac{1 + \frac{R_2}{R_1}}{1 + \frac{R_3}{R_4} + \frac{C_2}{C_1} + j\left(\omega R_3 C_2 - \frac{1}{\omega R_4 C_1}\right)} \qquad (2)$$

The purpose of using the Zener diodes is two-fold. First, they are in place to obtain a low frequency negative feedback path for the oscillation condition to be met.

Second, when the amplifier output voltage falls below the Zener voltage, they provide a virtual open circuit. In this way, the PLL function is isolated from the sweep oscillator.

The shunt small-signal diodes D_1 and D_2 provide a means of ensuring that the output voltage swing is clamped by the Zener diodes at the correct level. When an excessively large value of R_2 is used, the effectiveness of the clamping function will be lost.

For the PLL to acquire lock under a frequency sweep condition (1) with bet-

Zener Diode Voltage	Δ V^{-1}
3.3 V	160 ± 20 msec/V
3.6 V	160 ± 10 msec/V
3.9 V	160 ± 20 msec/V
4.3 V	120 ± 30 msec/V
4.7 V	110 ± 20 msec/V
5.1 V	75 ± 20 msec/V
6.2 V	40 ± 10 msec/V

Table 1. ΔV was measured at 1 to 0 Volt crossing.

ter than 90 percent successful lock probability, the rate of change of the output VCO frequency ω_{VCO} is an approximate function of the loop bandwidth ω_n.

$$\frac{d\omega_{vco}}{dt} \leqslant 0.5\omega_n^2 \qquad (3)$$

By definition, the rate of change of the output VCO frequency can be written as,

$$\frac{d\omega_{vco}}{dt} = \Delta V K_{vco}$$

$$\Delta V = 2f_{so}V_{pk} \qquad (4)$$

$$= \frac{\omega_{so}}{\pi} V_{pk}$$

where,

V_{pk} is the maximum voltage swing potential of the sweep oscillator

$\omega_{so} = 2\pi f_{so}$ is the angular frequency of the sweep oscillator

K_{VCO} is the gain constant of the output VCO.

Thus the maximum ΔV, and therefore the frequency of the sweep oscillator can be calculated.

It is also clear that the lock condition should occur at a point where the output voltage V_l of the loop filter is less than $|V_z|$, where V_z is the Zener voltage of the back to back Zener diodes. Under the lock condition, the negative feedback condition required by the Wein-Bridge oscillator is almost non-existent due to the fact that very little or no current flows through R_1. In addition, the open-circuit action of the Zener diodes prevent any further positive feedback tendency.

There is a small penalty suffered in the PLL loop dynamics. Since it is inconvenient to limit $R_4 = R_1 // R_2'$, the op-amp output bias current will create a small amount of DC offset. The selection of a low input bias current op-amp will mitigate this problem. In most cases, this problem is not important when compared with other sources of DC offset phenomena such as that of a mixer.

It is essential to ensure that the input of R_1 does not have a significant amount of fixed DC offset. This offset would prevent sweep oscillation from taking place.

A test circuit was designed to illustrate the usefulness of this method. A 1.5 GHz VCO was designed to provide ± 55 MHz deviation for the control input voltage of 0 to 10 V. The relatively wide tuning range of this VCO is intentional, so that the frequency acquisition action can be more clearly demonstrated. A second order loop is constructed as shown in Figure 1 where $\omega_n = 2\pi \times 10{,}000$ radians/sec and $\zeta = 2$. Thus the maximum rate of voltage change required at the DC crossing is $\Delta V < 28.6$ V/sec.

A heuristic estimate of the V_{pk} to V_z of about 25 is used, as V_{pk} is not always immediately obvious. With this initial value, all the circuit components can now be determined.

Table 1 shows the measured maximum ΔV with respect to the choice of Zener diodes. Obviously, the smaller the Zener voltage, the smaller the VCO tuning range is allowed. Figure 2 shows the sweep voltage waveform and Figure 3 shows the quality of the lock output spectrum of the VCO.

The requirements on all the circuit components are not critical and are simple to implement. In fact, it is convenient to adopt a slow sweep rate configuration, such as the one shown in the example, as general framework for PLLs of this kind which can tolerate equal or faster ΔV. ***RF***

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Biasing Solid State Amplifiers to Linear Operation

By Helge Granberg
Motorola, Inc., SPS

All solid state devices intended for linear operation must have a certain amount of "forward bias" (idle current) in order to place their operating points in the linear region of the transfer curve. Bipolar devices require a constant voltage source, whereas MOSFETs can be biased with simple resistor divider networks. Both will get more complex however, if temperature stability is required. Examples of applications requiring amplifier linearity include all amplitude modulated systems for communications and broadcast, nuclear magnetic resonance, magnetic resonance imaging, digital cellular telephone and signal sources for instrumentation. Circuit examples are presented from the most simple ones to sophisticated closed loop systems.

Since the base current of a bipolar transistor is equal to I_c (peak)/h_{FE}, the base bias supply must be able to supply this current without considerable excursions in the base-emitter voltage between the no-signal and the maximum signal conditions. This requires a constant voltage source, as variations of a few millivolts represent a large portion of the nominal 0.63 - 0.67 volt typical value. Depending on the specification of a specific application, various degrees of requirements are set for the base bias voltage source. In some instances a large value capacitor can be seen connected across the bias voltage supply to further reduce its AC impedance. However, this makes the impedance dependent on the frequency of modulation and is a good and practical solution only in applications where the modulating frequency is in the medium to high audio frequency range. One of the simplest biasing circuits for bipolar transistors (4,5,6), is shown in Figure 1a. It uses a clamping diode to provide a low impedance voltage source. The diode forward current must be greater than the peak base current of the transistor. This current is adjusted with R2 and the resistance of RFC1 and R1 combination is used to reduce the actual base voltage to a slightly lower value than the forward voltage of D1. The diode can be mechanically connected to the heat sink or the transistor housing to perform a temperature compensating function to Q1. This technique works adequately, although for perfect temperature tracking, Q1 and D1 should have similar DC parameters. One disadvantage with the circuit shown in Figure 1a is its inefficiency especially in biasing high power devices since (V_{cc} -V_b)I_b(max) will always be dissipated in the dropping resistors.

This deficiency of the circuit shown in Figure 1a can be overcome by amplifying the clamping diode current with an emitter follower (1,2) as shown in Figure 1b. Two series diodes (D1-D2) are required since one has to compensate for the V_{BE}(f) drop in Q1. In this case, low current signal diodes can be used and their forward current is equal to I(bias)/h_{FE}(Q1). For best results, Q1 should have a linear h_{FE} up to the peak bias current required and in higher power systems it must be cooled by some means. Ideally, Q1 and one of the series diodes should remain at ambient temperature, whereas the other diode (D1 or D2) can be used for temperature compensation of the RF device. An effective and fast responsive system is obtained if the diode (having long leads) is located near the RF transistor. The leads can be suitably formed allowing the body of the diode to be pressed against the ceramic lid of the RF transistor and fastened in place with thermal conductive epoxy. R1 is used to set the bias idle current and R2 limits its range of adjustment. The value of R2 depends on the supply voltage employed. The function of C1 and the RFC is simple to prevent the RF signal from getting into Q1.

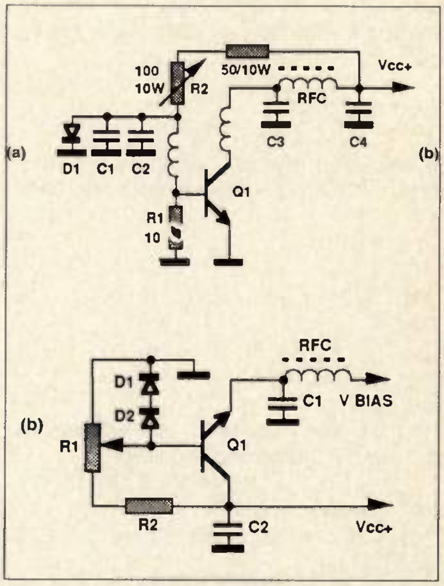


Figure 1. Two simple bias sources for bipolar transistors. (a) Uses all passive components, but is inefficient. This drawback is not present in (b), although it is slightly more complex.

Another fairly simple bipolar bias source (3) is shown in Figure 2. Its output voltage equals the base-emitter junction drop of Q1 plus the drop across R3. R1 must be selected to provide sufficient base drive current for Q2, set by its h_{FE}. Normally this current is in the range of a few milliamperes, and Q1 can be any small signal transistor in a package configuration that can be easily mounted to the heat sink or RF transistor housing for temperature compensation. The only requirement is that its V_{BE}(f) at that

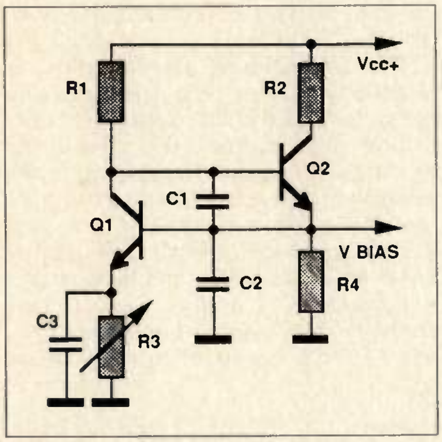


Figure 2. This circuit features the lowest source impedance of the less complex ones. Therefore it is recommended for high power device biasing and for demand applications.

current must be lower than that of the RF transistor at its bias current level. The maximum current capability depends on Q2 and R2. The power dissipation of Q2 can be up to a few watts and in most cases should be heat sunk, but must be electrically isolated from ground. The value of R2 can be calculated as: $(V_{CE}\text{-}V_{CE}(sat))/I_b$. C1 through C3 are a precaution to suppress high frequency oscillations, but may not be necessary depending on the transistors used and the physical circuit layout. Output source impedances for this circuit, when used in conjunction with a 300 W amplifier, have been calculated as low as 200-300 mOhms.

More sophisticated bias sources can include an integrated circuit voltage regulator (2). In most cases a pass transistor is required for current boost and to lower the source impedance. There are high current regulators available today, such as the LM317, LM337, etc., but the author has not investigated whether they are suitable for applications such as this. The circuit in Figure 3 uses a 723 regulator, which is available from several manufacturers with a variety of prefixes. It has been used for bipolar bias sources since the early 70s and more recently for MOSFET biasing as well. The 723 is specified for a minimum V_{out} of 2 volts, but with certain circuit modifications can be lowered to less than 0.5 V. The main advantages of this type bias source are: 1) It provides the lowest source impedance at a relatively low cost. 2) The bias voltage remains independent of variations in the power supply voltage. 3) Temperature compensation is easy to implement. In Figure 3 D1 performs this function and should be in thermal contact with the heat source. The same technique discussed with the circuit shown in Figure 1b can also be adapted here. Depending on the current requirement and the pass transistor used, Q1 may have to be cooled. It has a positive temperature coefficient to the bias voltage, but is negligible compared to the negative coefficient of D1. This permits Q1 to be attached to the main heat sink. R1 and D2 are only necessary if the RF amplifier is operated at a supply voltage higher than 40V, which is the maximum rating for the regulator.

Biasing of MOSFETs

Since MOSFETs have gate threshold voltages up to 5-6 volts, they require some base bias voltage in most applications. They can be operated in class C, (zero gate bias) but at a cost of low power gain. In such cases the input

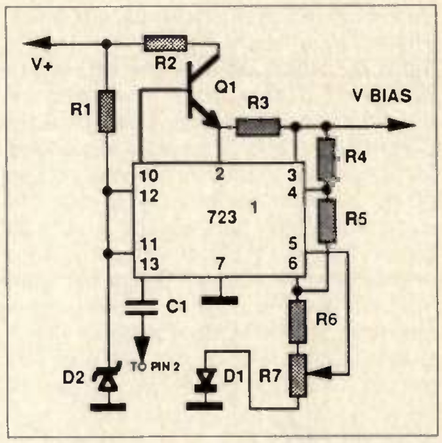


Figure 3. An integrated circuit bias source. The idle current remains constant regardless of the supply voltage. The source impedance mainly depends on the h_{FE} of Q1.

voltage swing must have an amplitude sufficient to overcome the gate voltage from zero to over the threshold level. The drain efficiency is usually higher than in other classes of operation. Especially if overdriven, the class of operation can approach class D. Zero bias is often used in amplifiers intended for FM or CW; and efficiencies in excess of 80 percent are not uncommon. In class B the gate bias voltage is set just below the threshold resulting in zero drain idle current flow. The power gain is higher than in class C, but the drain efficiency is 10-15 percent lower. Class B is suitable only for the FM and CW modes. Between these two classes of operation, one must decide whether the system has power gain to spare and how important is efficiency. At higher frequencies, such as UHF, a good compromise may be class B or even class AB. In class AB the gate bias voltage is somewhat higher than the device threshold, resulting in drain idle current flow. The idle current required to place the device into the linear mode of operation is usually given in a data sheet. In this respect MOSFETs are much more sensitive to the level of idle current than are bipolar transistors. They also require somewhat higher current levels compared to bipolars of comparable electrical size.

The temperature compensation of MOSFETs can be most readily accomplished with networks consisting of thermistors and resistors. The ratio of the two must be adjusted according to the thermistor characteristics and the g_{fs} of the FET. The changes in the gate threshold voltage are inversely proportional to temperature and amounts to approximately 1 mV/degree C. These changes have a larger effect on the I_{DQ} of a FET with high g_{fs} than one with low g_{fs}.

Unfortunately the situation is complicated by the fact that g_{fs} is also reduced at elevated temperatures, making the drain idle current dependent on two variables. In spite of this, this method of temperature compensation can be designed to operate satisfactorily and is repeatable for production. The thermistor is thermally connected into a convenient location in the heat source in a manner similar to that described for the compensating diodes with bipolar units discussed earlier. An example of a simple MOSFET biasing circuit (2) as described here is shown in Figure 4.

Most MOSFET device data sheets give V_{GS}(th) versus I_D data, but the values are only typical, and in some cases the g_{fs} can vary as much as 100 percent from unit to unit. Thus, in production the devices should have g_{fs} values that are matched on the parameter above to at least 20 percent. Otherwise each amplifier must be individually checked for temperature tracking. Some manufacturers such as Motorola supply RF power FETs with color coded g_{fs} matching.

The circuit in Figure 5 shows a typical MOSFET bias voltage source using the 723 IC regulator (2), which was earlier presented for bipolar transistor biasing. Since a MOSFET draws no gate bias current, except in the form of leakage, the pass transistor (Q1) has been omitted and D1 replaced by R5-R6 combination. The values of other passive components have also been modified to produce a maximum output of 8 volts. The temperature slope is adjusted by the ratio of the series resistor (R5) and the thermistor (R6). In addition to maintaining a constant bias voltage, this circuit also features bias voltage regulation against changes in the power supply voltage.

Figure 6 shows a closed loop system for MOSFET biasing. It provides an automatic and precise temperature compensation to any MOSFET regardless of its electrical size and g_{fs}. No temperature sensing elements need be connected to the heat sink or to the device housing. In fact, FETs with different gate threshold voltages can be changed in the amplifier without affecting the level of the idle current. The gate threshold voltage range is about ± 0.5 V with a single initial setting of the idle current. This means that the gate threshold voltage can vary within these limits over short or long periods of time for whatever reasons. In addition to temperature, other factors affecting V_{gs}(th) might be moisture, atmospheric pressure, etc.

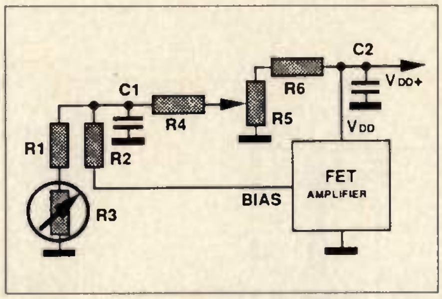


Figure 4. A simple MOSFET bias circuit using a thermistor-resistor network for temperature compensation.

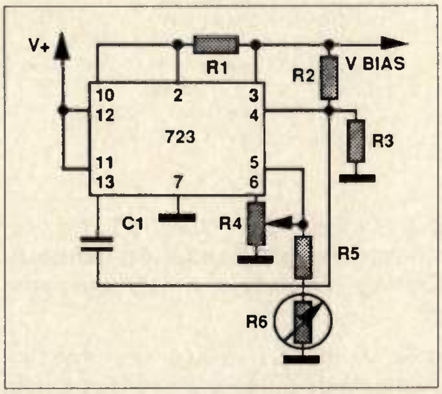
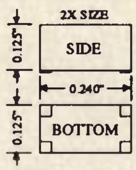
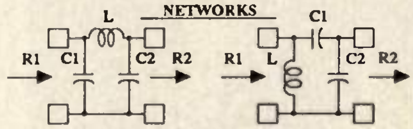


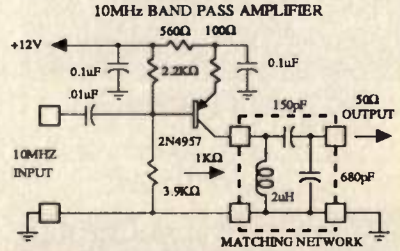
Figure 5. A more sophisticated MOSFET bias system with an integrated circuit voltage regulator. It also employs a thermistor for temperature tracking.

The principle of operation of circuit in Figure 6 is as follows: The idle current of the MOSFET amplifier is initially set to class A, AB or anywhere in between these bias limits by R8, which also provides a stable voltage reference to the negative input of the operational amplifier U1. This results in a current flow through R1 with a consequent voltage generated across it. This voltage is fed to the positive input of U1, which results in the output of U1 following it in polarity, but not in amplitude. Due to the voltage gain in U1, which operates in an open loop mode, its output voltage excursions are much higher than those generated across R1. Thus, if the current through R1 tends to increase for any reason, part of the output voltage of U1 fed to the amplifier gate bias input will adjust to a lower level, holding the current through R1 at its original value. A similar self adjustment will take place in the opposite direction as well.

The values for the resistive voltage divider R5-R6 have been selected for a range of ± 0.5 Volts at the amplifier FET







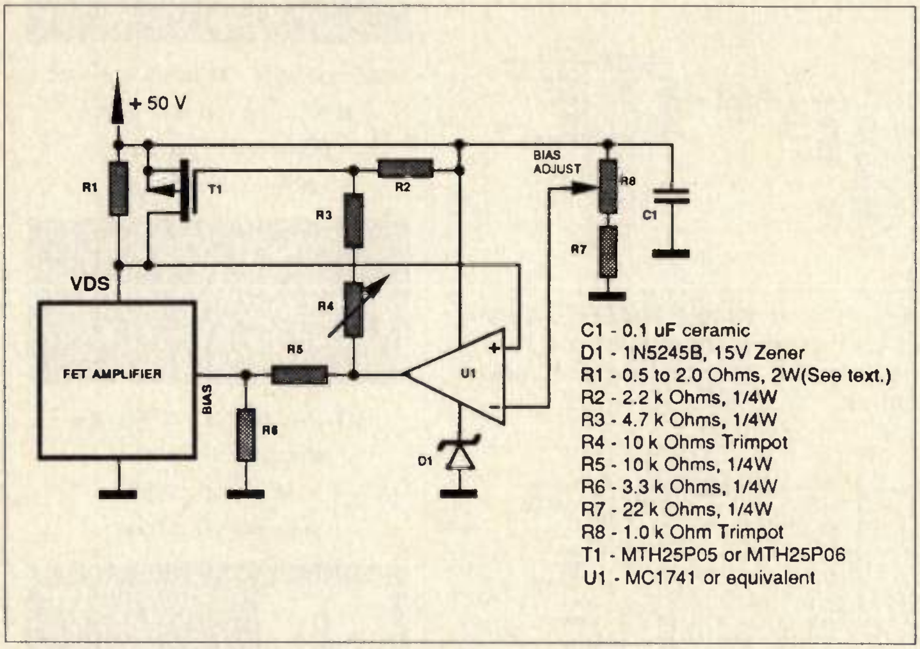


Figure 6. An automatic bias tracking system for MOSFET power amplifiers. It provides automatic temperature compensation without sensors as well as versatility for substituting a variety of electrical sizes of FETs operating at any supply voltage.

gate for the full voltage swing at the output of U1. This larger voltage swing is required to provide negative bias to the gate of the P-channel FET Q1, turning it on harder the higher the current is through R1. The r_{DS}(on) of Q1, which can decrease as much as 50 percent with 1V increase in gate voltage, depending on where the operating point is set by R4, is used in parallel with R1 to function as a voltage variable resistor. r_{DS}(on)(Q1) in parallel can be less than 0.2 ohms at 10 amperes. The value of R1 must be selected according to the desired level of idle current and R2 through R4 combination. R4 is adjusted only once (starting with its maximum value) according to the V_{GS}(th) and other parameters of Q1 to the point where the idle current previously set starts to increase. With correct component values, the gate bias voltage can be made to remain constant, or even increase (if desired) from the idle current level to the maximum drain current drawn under RF drive conditions. A bipolar Darlington transistor such as MJH6085 can be also used in place of the P-channel FET (Q1), but its h_{FE} must be 3000-4000 minimum in order for U1 to be able to drive it.

Summary

The circuits presented in this paper are of a basic nature, and may require refinement or modification according to specific applications. Equations to calculate the component values for most circuits described herein are given in the references and have not been duplicated here. No known description exists for the circuit shown in Figure 6. Circuit analysis and calculation of component values will be presented in a forthcoming paper by the author.

Many of these circuits have been in use by the industry in various forms for years, but the designers are constantly looking for simpler and better performing bias sources for solid state amplifiers. One possibility might be an Application Specific Integrated Circuit (ASIC) in a form of SMART power. These integrated circuits are gaining popularity and are manufactured by several companies including Motorola. Most are intended for automotive applications.

An integrated circuit can certainly be designed that satisfies the requirements for bipolar transistor and MOSFET biasing combining all the features presented in this paper. It is not certain though, that the existing market can justify such an IC development. However, new applications requiring linear amplification, such as digital cellular telephones, are being created and older vacuum tube designs are being converted to solid state designs at an ever increasing rate. An estimated increase in solid state linear amplifiers for all applications combined is at least a factor of 4-5 within the past ten years and is expected to double by the end of the decade. ***RF***

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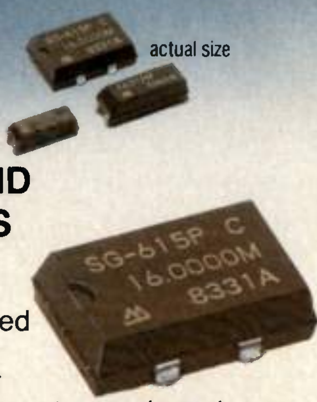

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A Stable Future for the Oscillator Market

By Liane G. Pomfret
Associate Editor

It comes as no surprise that the SAW and crystal oscillator business is doing well in the face of economic uncertainty. Their already widespread use, coupled with emerging technologies has provided a large market for manufacturers. SAW and crystal oscillators have not undergone any recent technological breakthroughs. Rather, changes continue to be small and cumulative.

The technology behind SAW and crystal oscillators has not changed much recently, nor is it expected to undergo any major changes in the near future. However, a number of small changes are occurring. One visible area of improvement is that component suppliers are improving their own quality control processes and passing these improvements on to the oscillator manufacturers. "We're getting nicer crystals," notes Liz Ronchetti, Vice President at Wenzel Associates, "the crystal manufacturers are doing better." Frank Perkins, Vice President of Marketing at RF Monolithics commented on the same trend in the MMICs and transistors that his company buys. In both cases, the suppliers, in trying to improve their own product, have helped improve SAW and crystal oscillator products.

Manufacturers are also paying closer attention to their own design processes. The use of computer aided design techniques in the design of circuits has improved reliability and performance and reduced the time and costs associated with developing new products. Problems such as phase noise under vibration, stability, low power, temperature performance, and special frequencies are all problems that are easier to solve because of the versatility of CAD programs. In the high performance area, the advent of microprocessor controlled oscillators has caused a bit of a stir within the industry. The technology first became available about four years ago and it's only been in the last two that it has become popular.

Another push lately has been towards surface mount packaging and miniaturi-

zation. While surface mount has been in use for years within the RF industry, it is just beginning to appear in the oscillator industry. One example of this, is from Wenzel Associates who recently introduced three new products, all in miniature packages. Marty Finklestein, President of MF Electronics observes, "We feel that surface mount is starting to catch on. Not only high volume parts, but more difficult products as well." Often the people who are pushing for surface mount are not the large, commercial end-users but the high-reliability, tight specification military users. Bob Murphy, Director of Aerospace Business at Frequency and Time Systems indicated that their switch to surface mount for some of their products was due to requirements for smaller sizes and lower power for Lightest programs. Since many of the companies in this report manufacture military, space or high reliability products, their attention to specifications and standards is much stricter than a high volume, popcorn part manufacturer. The military appears to be a driving force these days in development of smaller and better oscillators. Their need for small, high tech defense systems has forced manufacturers develop the necessary components on a much smaller scale than previously required. Tom Mihalek, Contract Administrator at Anderson Labs noted that improving quality is always an ongoing process — an idea few people would argue with. As Brian Rose, Vice President of Engineering at Q-Tech observes, "There's a responsibility to the customer."

These days, customers are using oscillators in just about every imaginable technology. With the new wave in communications technology, SAW and crystal oscillators are being used in everything from military spread spectrum to children's toys. Wenzel Associates has noticed that many of their new orders are coming from high performance products using spread spectrum and frequency hopping. Others have found that digital communications is the area with the most activity. Marty Finklestein indicated that standard telecommunications, or telephones are very strong for his company. RF Monolithics on the other hand, has seen their orders coming from areas such as air traffic control, aircraft and microwave radios. The list of products that use SAW and crystal oscillators is long and includes other areas such as radar, computers, spacecraft, satellites, cellular telephone and many more.

Everything points to the fact that the SAW and crystal oscillator business is doing well. Comments range from "we're tickled to death" to "we're pleasantly surprised." There were indications of a slow period at the end of last year, but business picked up again in January and is forging forward. With constant refinements and fine tuning, the oscillator industry is creating better products — from high cost, low volume military all the way down to high volume, low cost commercial. **RF**

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RF software

Automated Smith Chart

This software is helps the user to design, modify, test, and optimize impedance matching and other ladder type network circuits. The menu-driven software performs all mathematical computations, locus manipulations, and graphical display interactions internal to the program. The program prompts the user for specifications and values, then plots a full-size Smith Chart grid. It is available for IBM PCs and compatibles and HP 9000s.
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Extended Test Manager

Wavetek recently introduced WaveTest[R]-XTM which runs under Digital VMS and ULTRIX operating systems. The XTM designation stands for "Extended Test Manager." The new software combines the current capabilities of their PC-based WaveTest product with new features for addressing integrated factory requirements. The single unit price is $7,995 with deliveries four to five weeks after receipt of order.

Wavetek has also released their latest version of WaveTest™. WaveTest 3.0 offers two new icons and improvements to existing icons such as a dispatch icon, execute icon, graphical operator prompts, faster execution and better program control. The Test Program Generator module is available for $1495 or $495 for those who wish to upgrade from version 2.6.
Wavetek
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Analysis-only Version of SANA Available

T-SANA, the analysis-only version of SANA, is now available from Webb Laboratories. T-SANA offers nodal network definition, unrestricted noise analysis, output graphics postprocessing, complete manufacturer's active device libraries, PC-DOS 3.0+ and 4.0+ compatibility. The software is available from stock for $575.
Webb Laboratories
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Conversion Software

Units + Conversion Factors™ is a new software that automates conversions between units of any type. If the conversion units needed are not available, they may be added by the user. The software is IBM PC, XT, and AT compatible, requires DOS 2.0 or higher, 512 K available memory, one 5 1/4" floppy disk drive, and a CGA or EGA card or compatible. Price is $89.95.
David F. Taylor, Jr.
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Advertising Index

RF literature

Conductive Coatings Literature

Spraylat's new brochure addresses the latest in conductive coatings technology. The brochure discusses the benefits of their Series 599 copper conductive coatings which are either water or solvent based. Cost savings, ease of application, greater shielding effectiveness, and mechanical and long term environmental properties are reviewed.
Spraylat Corporation
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SPICE Newsletter

Intusoft's latest newsletter describes how to use and create analog behavioral models with any SPICE program. The newsletter demonstrates how to solve a set of differential equations that describe a physical spring-mass system, how to model a mixed-mode rotation system, and shows a complete servo model. SPICE netlists are also provided. The newsletter also includes information on the company's latest EDA tools, new op-amp model libraries and training seminars.
Intusoft
INFO/CARD #199

ELINT/EW Catalog

The 1991 ELINT/EW catalog for courses and software is now available from Research Associates of Syracuse. Courses include propagation, EW receivers, ELINT interception, ELINT analysis, ELINT/EW data bases, adaptive ECCM signal processing, ELINT/EW applications of DSP, radar vulnerability to jamming and integrated EW. Software includes the radar vulnerability analysis system and PCEMPE propagation program.
Research Associates of Syracuse
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RF & Microwave Engineering Book and Software Catalog

A new 72 page catalog from Artech House describes over 200 basic and advanced electrical engineering references, including 25 new titles to be published in 1991. Areas covered include: RF/Microwave applications, radar, antenna, optoelectronics, telecommunications, acoustics, and electronic materials.
Artech House
INFO/CARD #197

New and Used Test Equipment

RAG Electronics' Winter Update of available new and used electronic test equipment is now available. The used equipment section highlights recent arrivals, best values, and one-of-a-kind clearance items. The new equipment section includes Tektronix oscilloscopes, Fluke DMMs and more.
RAG Electronics
INFO/CARD #196

Microwave and Millimeter Wave GaAs MMICs

Alpha Industries has released its new RF, Microwave, and Millimeter Wave GaAs MMIC catalog. The catalog contains specifications and application notes for control, receiving and amplifier products. Extensive specifications are given for switches, attenuators, amplifiers, mixers, combiners and power dividers.
Alpha Industries, Inc.
INFO/CARD #195

Microprocessor Quartz Crystal Catalog

BOMAR has introduced their 1991 Microprocessor Quartz Crystal Catalog. The crystals featured are the standard frequencies most requested within the industry for microprocessor applications. Data is also supplied on custom cut crystals. Toll-free phone and fax numbers are available.
BOMAR Crystal Company
INFO/CARD #194

Tracking, Telemetry and Control

AML's latest catalog update contains specifications, outline dimensions and photos of AML's latest additions to its line of standard and custom integrated tracking components. Products covered include Dual Pseudo Monopulse Scanners, Switch Combiners, Monoscan Converters, and Monopulse Comparator Networks.
AML, Inc.
INFO/CARD #193

Test Equipment Catalog

The new Test Equipment Catalog from Lectronic Research Labs lists over 5,000 new and pre-owned products. More than 180 manufacturers are represented including Hewlett Packard, Fluke, Gen Rad, Tek, Honeywell, L&N, Lambda, Kepco, Sorenson and others. All reconditioned products are guaranteed to meet or exceed original manufacturer's specs.
Lectronic Research Laboratories
INFO/CARD #192

Capacitor Catalogs

Murata Erie has released two new catalogs on capacitors. Catalog No. C-06-A describes their complete line of trimming capacitors for HF, VHF, UHF and Microwave applications. It includes complete technical specifications including both electrical and mechanical characteristics. Catalog No. 62-09 covers their line of high voltage/high power ceramic capacitors and high voltage resistors. The capacitor line includes units ranging from small disc capacitors, rated at 7.5 kV to high power, water cooled units for RF transmitter applications.
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RF design

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Log slope, nom.	(mV/dB)	20	20	20
Linearity, nom.	(dB)	±1.0	±1.5	±1.75
Volts out, nom., into 93Ω@0dBm	(mV)	1450	1450	1350
IF output, nom.	(dBm)	0	−2	−4
Video rise time	(nsec)	≤20		
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