

High Level Integration Comes to DDS Devices

Technical Features — Baseband Signal Processing SS Sliding Correlator Grow Art inars



When you're looking for the leader of the pack in CDMA testing, there's only one name to call: Noise Com.

The WIS Series (Wireless Impairment System) — A complete, automated solution for testing CDMA base and mobile stations.

To completely characterize the performance of CDMA base and mobile stations, a hodgepodge of general-purpose test equipment just

won't do. You need an integrated solution designed to satisfy all of the demands of test standards such as IS-97 and IS-98 for cellular, and ANSI J-STD-019 and ANSI J-STD-018 for PCS applications. That's precisely the job Noise Com's WIS Series is designed to do.

The WIS Series is the only test station to combine so many automated CDMA measurement capabilities in a single, rack-mount system. They can provide emulation of wireless channel impairments such as additive white Gaussian noise (AWGN), multipath fading, as well as interference.

Multipath fading, AWGN and interference...

To emulate multipath fading, the WIS Series can model wireless communication channels between base stations and mobile transceivers using Rayleigh, Rician, Log-Normal, Suzuki, and Nakagami fading statistics. The WIS Series is also the only CDMA test solution with an

INFO/CARD 1

The Wireless Telecom Group is a publicly traded company and is listed on the American Stock Exchange. Symbol: WTT

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AWGN generator that can precisely set Eb/No, C/N, and C/I ratios. The WIS series can be configured with either base or mobile station interference sets or both, providing a complete impairment solution.

Combine the WIS Series with either a base station or mobile transmitter, or a base

or mobile station simulator, and you have the most accurate, comprehensive CDMA measurement solution available.

The configuration of the WIS Series is flexible, too, so you can pick and choose the capabilities you want right now, with the option to modify or upgrade them later. So, whether your measurement needs are in product development and design verification, production testing, or quality control, the WIS Series is the system youcan rely on today...and tomorrow.

Call us about our entire line of wireless test solutions.

When you want to deal with the leader of the pack in CDMA, there's only one name to remember: Noise Com. For more details, callus today at (201) 261-8797, fax us at (201) 261-8339, e-mail: noisecom@haven.ios.com, or write us at E. 49 Midland Ave., Paramus, NJ 07652. Noise Com. We're your global partner for wireless, telecommunications, and noise testing solutions.



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Mini-Circuits ushers-in a new era of technology and economy with ERA monolithic GaAs amplifiers. Just check the specs! These surface mount and drop-in amplifiers cover your applications to 8GHz with higher gain, more output, and flatter response. Characterized with S-parameter data, these amplifiers are very easy to use. Simply sketch an interconnect layout, and the design is done. And ERA's are engineered with wider bandwidths to eliminate your need for costly compensation networks and extra gain stages. So, review your present design and replace with Mini-Circuits new ERA technology. Lower overall cost, wide bandwidth stability, and lots to ... gain!

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	ERA-1	ERA-2	ERA-3
	ERA-1SM	ERA-2SM	ERA-3SM
Frequency	DC-8000	DC-6000	DC-3000
(MHz) f _L -f _U	DC-8000	DC-6000	DC J000
Gain, (dB)	11.6	14.9	20.2
	11.0	13.1	19.4
Max. Power Out	13	14	11
(dBm, @1dB comp.)	13	13	
Dynamic Range	NF IP3	NF IP3	NF IP3
	7dB 26dBm	6dB 27dBm	4.5dB 23dBm
	7dB 26dBm	6dB 27dBm	4.5dB 23dBm
¹ Price (\$ea., Qty.10)	1.80	1.95	2.10
	1.85	2.00	2.15

Note: All specifications typical at 2GHz, 25°C.

*Low frequency cutoff determined by external coupling capacitors I Price (ea.) Oty. 1000: ERA-1 \$1.16, ERA-2 \$1.31, ERA-3 \$1.46 Add \$.05 for SM option.

Designer's Amplifier Kits

K1-ERA: 10 of each ERA model (30 pieces) only \$49.95



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ERA-1SM

SIZE

ERA-1 ACTUAL

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F 214 Rev Orig

RFdesign

contents

April 1996

featured technology

32 Sideband Suppression Considerations for Complex Baseband Processing of FDM Waveforms

Accuracy in I/Q modulation and demodulation affect system performance. With zero-IF techniques, frequency translation is direct to baseband. This article discusses the types of errors and their effects.

- Rock Roberts

40 A Spread Spectrum Sliding Correlator for Propagation Measurements

Analyzing wireless communications channels often requires channel sounding mea surements to establish propagation char-

acteristics. Pulse techniques and frequency domain measurements are often used, but best results are achieved with a spread spectrum system using a sliding correlator such as the one described in this article.

— William G. Newhall, Theodore S. Rappaport, Dennis G. Sweeney

Cover story 56 High Level Integration and Performance Comes to DDS

The new AD9850 DDS + DAC integrated circuit offers a complete high frequency synthesis solution.

- Dave Crook, Jim Surber

tutorial

74 Active Filter and Other Op Amp Circuits for RF Applications

High speed op amps allow traditional active circuit design to be used for RF functions. This article reviews circuits for active filters, a precision rectifier and an isolator/circulator.

- Gary A. Breed

95 Structured Design of Third-Order, Type-2 and Type-3 PLLs

A review of PLL characteristics with an emphasis on graphical representation for quick analysis. Mathematical background is also presented. — Fu Nian Ku

engineer's notebook

102 Simple Direct-Conversion Receiver Checks Frequency Counters

This circuit uses the timebase of a counter as the local oscillator. — Michael A. Covington



cover story - p. 56

departments

- 8 Editorial
- 18 Calendar
- 20 Courses
- 22 News
- 28 Industry Insight
- 65 New Products
- 69 Product Focus
- 108 New Software
- 109 New Literature
- 110 Marketplace
- 122 Advertiser Index
- 122 Company Index
- 126 Info/Card



EMC coverage starts on page 81 —

News, Products, and A P.C. board design feature article...



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RF editorial

I will be leaving *RF Design*, but it's not "Good-bye"

By Gary A. Breed Editor

After more than ten years at RFDesign, I am moving on to those proverbial "bigger and better things." (No, this isn't an April Fool's joke) My new job will be President of Noble Publishing, which some of you may recognize as a technical book publishing company. Instead of seeking out the most appropriate technical articles for you to read, I will be finding authors and developing the most appropriate textbooks and reference books to put in your library.

RF Design will be in good hands. No, we haven't yet found the next Editor, but he or she is certainly out there among the many excellent RF engineers who have a broader-than-usual outlook on business and technology. In the mean time, I will help the present staff in any way necessary to maintain *RF Design*'s quality.

This magazine has prospered while I've been here, growing steadily even through the dark years of military cutbacks. I suppose I should take some of the credit for that success, but only a part. The concept of providing useful, timely engineering information in *RF Design* was established long before I became Editor. With only a few dissenters, you have repeatedly told me that I've done a good job fulfilling that mission. Your appreciation of my efforts is humbly acknowledged. Thanks for your support.

As the title above suggests, you won't be rid of me! Noble Publishing's books emphasize RF design techniques and other communications-related topics. I will be at many of the same engineering conferences and trade shows where you've seen me before, talking with you about all things related to RF. I will probably continue to contribute articles to *RF Design*, maybe even some editorial comments like you've found in this column every month. I will still be easy to find!

See You at IWCE and the

RF Design Seminar Series Don't forget to make the trip to Las Vegas April 23-26! I will be there, hosting the *RF Design*-sponsored events. The RF Design Seminar Series offers valuable short courses that are still open for registration (if you hurry). In addition, the giant International Wireless Communications Expo (IWCE) will have hundreds of radio equipment suppliers and more than 70 exhibit booths with the RF suppliers you need to see. They will be present-

and software for you to evaluate. IWCE presentations cover all aspects of the radio communications business — regulations, operations, standards, business concerns, and of course, a Thursday afternoon session of RF-related technical papers. This is the place to catch up on the issues surrounding mobile communications!

ing the latest components, equipment

IWCE provides a unique environment, combining end-user products (hand-held, mobile and base station radios and accessories), with systemlevel hardware (antennas, control systems, transmission lines and towers), and now including components, software and instruments for the design of those finished products. The synergy of users, system operators and design engineers all in one one place should create lots of new ideas, and a new appreciation for each level of technology.



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LIFD-3010-080	30	10	-70 to +5	110	1
LIFD-3010P-80	30	10	-80 to 0	100	0.5
LIFD-6010-70	60	10	-70 to 0	65	1
LIFD-6020P-80	60	20	-80 to 0	50	0.5
LIFD-12020-80	120	20	-80 to 0	50	1
LIFD-16040P-70	160	40	-65 to +5	25	0.5
LIFD-16040-70	160	40	-70 to 0	30	1
LIFD-300100-60	300	100	-60 to 0	10	1

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MLIF-500/100-70	500	100	-70 to 0	10	1
MLIF-750/500-62	750	500	-65 to 0	2	1
MLIF-1000/250-60	1000	250	-60 to 0	2	1
MLIF-1000/500-65	1000	500	-65 to 3	5	1.5
MLIF-1500/1000-60	1500	1000	-60 to 0	1.25	1
MLIF-1575/20-40*	1575	20	-38 to 3	N/A	1.5

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FMDM-21.4/4-5	21	4	1000	150	2
FMDM-30/6-8	30	6	1000	120	2
FMDM60/10-15	60	10	250	75	2
FMDM-160/35-15	160	35	100	30	3
FMDM-160/50-25	160	50	20	17.5	5
FMDM-300/100-20	300	100	20	20	2.5
FMDM-1000/300-70	1000	300	10	5	5

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	Model Number	Center Frequency (MHz)	Bandwidth (3 dB) (MHz, Min.)	Dynamic Range (dB, Min.)	Phase Change (±Deg., Max.)	Output Power (dBm, Min.)	guide
1	LCPM-30/10-70	30	10	-70 to 0	4	10	
	LCPM-60/20-70	60	20	-70 to 0	5	10	All and a second se
	LCPM-60/14-65	60	14	-70 to -5	2.5	10 👝	
	LCPM-160/40-65	160	40	-70 to -5	5	10	Baar
	LCPM-160/40-70	160	40	-65 to 5	3	10	M/N LCDM
	LCPM-300/50-55	300	50	-55 to 0	5	3	FREQ 30 MH
í	LCPM-500/100-45	500	100	-45 to 0	5	3	751346

Variable Gain Control Linear Amplifiers

- Operating frequencies from 21 to 850 MHz
- Standard models up to 75 dB of gain control range

			Contraction of the local sector of the local s	and the second second	and the second se
Model Number	Center Frequency (MHz)	Bandwidth (3 dB) (MHz, Min.)	Dynamic Range (dB, Min.)	Noïse Figure (dB, Max.)	P. Out @ 1dB Compr. (dBm, Min.)
VGC-7-30/10	30	10	-80 to -10	4	10
VGC-7-60/20	60	20	-85 to -10	6	10
VGC-7-70/10	70	10	-60 to -10	4	10
VGC-6-70/20	70	20	-50 to +10	5	10
VGC-7S-140/40	140	40	-75 to -10	7	10
VGC-6-160/40	160	40	-65 to -5	6	1
VGC-7-250/100	250	100	-70 to -10	6	10
VGC-4-720/100	720	100	-40 to 0	6	0
VGC-4-850/100	850	100	-40 to 0	6	0

utomatic Gain Control Linear Amplifiers Operating frequencies from 21 to 850 MHz Standard models up to 75 dB of gain control range

Model Number	Center Frequency (MHz)	Bandwidth (3 dB) (MHz, Min.)	Output Power (dBm, Min.)	Power Variation (dB, Max.)	Dynamic Range (dBm, Min.)
AGC-7P-30/15	30	15	10	2	-60 to +5
AGC-8-70/20	70	20	10	2	-75 to 0
AGC-6-70/30	70	30	3	2	-65 to -5
AGC-6-140/30	140	30	4	2	-65 to -5
AGC-4S-140/55	140	55	5	2	-40 to 0
AGC-5S-370/100	370	100	5	2	-60 to -10
AGC-5-387/175	387.5	175	-3	1	-60 to -15

For available options, custom design information, technical questions, or any additional information, please contact Boris Benger at [516] 436-7400, extension 140.

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RF calendar

April

16-17 Minnesota EMC Event Bloomington, MN Information: Judie Anderson, 3001 Habor Lane, Suite 150, Plymouth, MN 55447. Tel: (612) 559-0332; Fax: (612) 553-9326. 23-26 International Wireless Communications Expo (including the RF Design Seminar Series) Las Vegas, NV Information: Intertec Presentations, 6151 Powers Ferry Road, N.W., Atlanta, GA 30339. Tel.: (800) 828-0420 or (770) 618-0499. Fax: (770) 618-0441. 28-1 IEEE Vehicular Technology Conference-VTC '96 Atlanta, GA Information: Wendy Rochelle, IEEE/VTC'96 Registrar, P.O. Box: 1331, Piscataway, NJ 08855-1331. Tel: (908) 562-3870; Fax: (908) 981-1769. E-mail: VTC96@IEEE.com June 5-7 Frequency Control Symposium Honolulu, HI Information: Michael Mirarchi, Synergistic Management, 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 280-2024. Fax: (908) 681-9314. 16-21 **MTT-S International Microwave Symposium** San Francisco, CA Information: Derry Hornbuckle, Hewlett-Packard; Tel: (707) 577-3658; Fax: (707) 577-2036, or Jerry Fiedziusko, Space Systems/Loral Corp. Tel: (415) 852-6868. Fax: (415) 852-5068. 24-25 **Time and Frequency Seminars: Introduction-Level 1** Boulder, CO 26-28

Time and Frequency Seminars: Fundamentals-Level 2

Boulder, CO

Information: Wendy Ortega Henderson, National Institute of Standards and Technology, 325 Broadway, Boulder, CO 80303-3328. Tel: (303) 497-3593; Fax: (303)497-6461. E-mail: ortegaw@boulder.nist.gov

July

21-26

1996 IEEE AP-S International Symposium and URSI

Radio Science Meeting

Baltimore, MD

Information: Mr. Jon Moellers, Steering Committee Chair, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331. Tel: (410) 993-6774; Fax: (410) 993-7432.

August

21-23

Wireless Communications Workshop Boulder, CO

Information: Dr. Roger Marks, National Institute of Standards and Technology, 325 Broadway, MC 813.06, Boulder, Co 80303. Tel: (303) 497-3037; Fax: (303) 497-7828; E-mail: marks@nist.gov

18

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RF courses

Antennas: Principles, Design and Measurements

May 14-17, 1996, St. Cloud (Orlando), FL Information: Kelly Brown, Northeast Consortium for Engineering Education, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146. Fax: (407) 892-0406.

DSP Without Tears

April 17-18, 1996, Fort Lauderdale, FL May 1-3, 1996, Boston, MA Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (770) 587-4812; Fax: (770) 518-8368.

RF and Wireless Engineering

April 23-25, 1996, Las Vegas, NV Practical High Frequency Filter Design April 23, 1996, Las Vegas, NV Oscillator Design Principles April 24, 1996, 1996, Las Vegas, NV Digital Modulation and Spread Spectrum for Wireless Communications April 23, 1996, RF Power Transistors and Amplifiers

April 24, 1996, Las Vegas, NV

Information: RF Design Seminar Series, Argus Trade Shows, 6151 Powers Ferry Rd., N.W. Atlanta, GA 30339. Tel: (800) 828-0420.

Maximize Performance of Transmission Lines

April 9-11, Orland Park, IL May 7-9, 1996, Orland Park, IL June 4-6, 1996, Orland Park, IL Information: Andrew Corporation, Dept. 355, P.O. Box: 9000, San Fernando, CA 91341-9978. Tel: (800) 255-1479 ext. 117.

Modern Receiver Design

April 15-19, 1996, London, U.K. Wireless Infrastructure Network Engineering for Cellular, PCS, LEO and WPBX April 15-19, 1996, Washington, DC Microwave Tubes, High-Power Transmitters, and Microwave Systems: Basic Principles April 29-May 3,1996, Washington, DC Grounding, Bonding, Shielding and Transient Protection April 29-May 3, 1996, San Diego, CA Satellite Communications Engineering Principles May 7-10, 1996, Washington, DC **Digital Image Processing** May 13-14, 1996, Washington, DC Mobile Satellite Communications Systems May 13-15, 1996, Washington, DC Modern Digital Modulation Techniques May 13-17, 1996, Washington, DC Modern Digital Video Processing May 15-17, 1996, Washington, DC **Global Positioning System: Principles and Practice** May 20-23, 1996, Washington, DC Electromagnetic Interference and Control in Modern Communications Systems May 20-24, 1996, Washington, DC Mobile Communications Engineering May 22-24, 1996, Washington, DC

Hazardous RF Electromagnetic Radiation: Evaluation, Control, Effects, and Standards

June 12-14, 1996, Washington, DC

Information: The George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd Street, N.W., Washington, DC 20052. Tel: (202) 994–6106 or (800) 424–9773. Fax: (202) 872–0645.

Grounding & Shielding Electronic Systems, and Circuit Board Layout

June, 1996 (dates TBA), Chicago, IL

August 14-16, 1996, San Jose, CA Information: Continuing Education, University of Missouri-Rolla, 103 ME Annex, Rolla, MO 65409-1560. Tel: (314) 341-4132; Fax: (314) 341-4992.

Digital Video Technology

April 10-12, 1996, Los Angeles, CA

Wireless Voice and Data Communications April 16-19, 1996, Los Angeles, CA

Advanced Communications Systems Using Digital Signal Processing

April 22-24, 1996, Los Angeles, CA

Wired and Wireless Telecommunications Networking May 20-24, 1996, Los Angeles, CA

Information: UCLA Extension, 10995 Le Conte Ave., Suite 542, Los Angeles, CA 90024. Tel: (310) 825-1047; Fax: (310) 206-2815. E-mail: mhenness@unex.ucla.edu

RF IC Design of Wireless Communication Systems

April 23-26, 1996, San Francisco, CA Information: Mead, Microelectronics, Inc., 7100 NW Grand-

view Dr., Corvallis, OR 97330. Tel: (541) 758-0828; Fax: (541) 752-1405.

Training Program for Cellular, PCS Staff

Independent Learning Program Information: Virginia Polytechnic Institute and State University, Mobile and Portable Radio Research Group, 840 University City Blvd., Pointe West Commons, Suite 1, Blacksburg, VA 24061-0350. Tel: (540) 231-2970; Fax: (540) 231-2968.

Principles of Electronic Counter-Countermeaures April 9-11, 1996, Atlanta, GA Phased-Array Radar System Design

April 29-May 2, 1996, Atlanta, GA Information: Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel: (404) 894-2547.

RF and Wireless Made Simple April 29-30, 1996, Los Altos, CA July 8-9, 1996, Los Altos, CA Applied RF Techniques I April 29- May 3, 1996, Los Altos, CA July 15-19, 1996, Los Altos, CA Wireless RF System Design May 6-10, 1996, Los Altos, CA Applied RF Techniques II May 6-10, 1996, Los Altos, CA Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel:(415) 949-3300; Fax: (415) 949-4400.

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LRFMS-1A-13-SM6	+13	2-500	DC-500	6.0	44	36	\$ 7.95
LRFMS-1A-17-SM6	+17	2-500	DC-500	7.0	44	35	\$ 9.95
LRFMS-2L-SM6	+3	500-1000	DC-1000	5.6	30	20	\$ 5.95
LRFMS-2-SM6	+7	5-1000	DC-1000	7.0	35	30	\$ 3.95
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LRFMS-2-17-SM6	+17	5-1000	DC-1000	8.0	28	30	\$ 10.95
LRFMS-11X-SM6	+7	5-1900	5-1000	7.0	32	36	\$ 3.95
LRFMS-4-SM6	+7	5-1500	DC-1000	7.5	40	30	\$ 6.95
LRFMS-5-SM6	+7	10-2000	10-1000	6.5	40	30	\$ 7.95
LRFMS-5-13-SM6	+13	10-1500	DC-1000	7.5	40	35	\$ 13.95
LRFMS-5-17-SM6	+17	10-1500	DC-1000	8.0	35	30	\$ 15.95

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RF news

Forecast Predicts Small Increase in R & D Expenditures for 1996

R & D expenditures in 1996 are expected to increase to about \$174 billion according to a Battelle - R & D Magazine forecast. This is a 1.7 percent increase over the \$171 billion that the National Science Foundation estimates was spent in 1995. Although R & D investment stalled in the early 1990s, this period of stagnation is concluding and R & D spending will increase in the near future. The gradual recovery in R & D expenditures follows a pattern that Battelle predicted five years ago. Noting the growth of R & D expenditures in the 1980s, it was apparent that such growth would not be sustained. Battelle predicted a slowdown in growth in the early 1990s, followed by an increase in R & D spending by middecade. Major predictions for 1996 include:

- Federal R & D Spending will decrease, about a half-percent in 1995.
- Industry will increase R & D Spending, about 3 percent over last year.
- The remainder of R & D expenditures — \$8.8 billion — will be supported by universities and non-profit organizations.
- A key to industrial investment is that long-anticipated real growth, outdistancing inflation, will occur in 1996.

- Private industry will increasingly look for opportunities to outsource internal R & D functions.
- An increasing share of U.S. Industry's R & D will be performed off shore, primarily in facilities owned by the same industry.
- The federal government's efforts at budget-cutting, coupled with philosophical changes toward national science policy, will affect R & D support and the federal laboratories.

Video Module Interface Proposed

A video interface specification, initiated by SGS-THOMSON Microelectronics and offered as an open standard, is winning support from the PC

Business Briefs

Densitron Moves to Larger Facility — Densitron Corporation America has moved to a larger facility in Santa Fe Springs, California. The new facility will accommodate Densitron's Engineering, Sales, Marketing, Finance, and manufacturing staff for its North American operations. Effective March 1,1996, Densitron's address is: 10430-2 Pioneer Blvd., Santa Fe Springs, CA 90670. Tel: (310) 941-5000; Fax: (310) 941-5757.

Rogers Corporation Acquires Product Lines From ARLON, Inc. — Rogers Corporation has acquired the relatively small business of high dielectric constant PTFE based non-reinforced circuit laminate products from ARLON, Inc., including the products known as DICLAD®810 and EPSILAM®10 circuit board materials.

JEFA Acquires COMSAT RSI — JEFA International, Inc. acquired by COMSAT RSI, Inc. on October 1,1995, to become "COMSAT RSI JEFA Wireless Systems." This acquisition includes a wider range of equipment, services, and resources to design and implement large, complex communications systems. JEFA Wireless has also moved up the road to: 2100 Couch Drive, McKinney, TX 75069; Tel: (214) 547-4455; Fax: (214) 542-4557.

Group Technologies Corporation Announces Sale of Metrum Unit — F. W. Bell, Inc. has purchased substantially all the assets of Metrum, Inc., of Littleton, Colorado, a wholly-owned subsidiary of Group Technologies Corporation of Tampa Florida. Metrum designs, manufactures and markets high performance, digital and analog recording technologies for data acquisition applications.

Interad, Ltd. has moved — Their new address is: 8020 Queenair Drive, Gaithersburg, MD 20879. Tel: (301) 948-1626; Fax: (301)977-7559.

Fujitsu Towa Electron Opens U.S. Office — Fujitsu Towa Electron Limited, a Japanese manufacturer of electronic components and assemblies, announced its expansion into the U.S. with the opening of Fujitsu Towa Electron U.S.A., Inc. in San Jose, California. The new office will manage product sales, marketing and technical support for North America. Fujitsu Towa Electron U.S.A., is located at 1641 N. First St., Sutie 155, San Jose, CA 9512. Tel: (408) 437-8900; Fax: (408) 437-0700.

Comtech Consolidates SPS Division into PST Division — Comtech Microwave Products Corp., a whollyowned subsidiary of Comtech Telecommunications Corp., announced that effective immediately, its scientific Power Systems Division (SPS) will be consolidated into Comtech's Power Systems Technology Division (PST). All current SPS products and technology will be fully supported by PST.

Cadence Picks Systems Science to Provide Waveform Display Technology — Systems Science, Inc., a developer of design verification, test and visualization tools, and Cadence Design Systems, Inc., a major provider of EDA software and services, today announced that they have entered into an agreement to bundle Systems Science's SimWave® waveform display with Cadence's digital and analog simulators.

Cubic Communications Acquires Direction-Finding Product Line — Cubic Communications, Inc., has acquired the direction-finding product line assets of Ocean Applied Research (OAR) Corp.

Harris Semiconductor Announces Major Expansion to High Speed Converter Portfolio — Harris Semiconductor has added 14 new high speed converters acquired through a buy-resell agreement with Sony; more Sony converters will be added over the next six months, in addition to new Harris developed converters.

Weinschel relocates — Weinschel Corporation has relocated. Their new address is 5305 Spectrum Dr., Frederick, MD 217-7362. Tel: (800) 638-2048; (301) 831-4701; Fax: (301) 831-4570.



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INFO/CARD 18

RF news continued

Multimedia industry, including leading manufacturers such as STB Systems, Cirrus Logic, Trident, Sierra Semiconductor, Oak Technology, SiS, Alliance Semiconductor and ULSI systems. Known as the Video Module Interface (VMI), the proposal addresses the connection of video modules such as MPEG, video phone and video decoder, to GUI chip. Developed in conjunction with other manufacturers, including Cirrus Logic and Sierra Semiconductor, the key benefits of VMI are its flexibility and low-cost.

Report on Telecommunications Industry

With the passage of the Telecommunications Act of 1996, the spoils of

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INFO/CARD 19

wireless and wireline connectivity have been put on the table. At stake is the \$45 billion in revenues generated annually from local telecommunications service in the US, a market Allied Business Intelligence, Inc. (ABI) forecasts to grow to \$56.2 billion by 2001. This not including the CATV market, accounting for \$33.4 billion in subscriber revenues and expected to reach \$44.5 billion by 2001. Forecasting, historical data and a detailed breakdown of the component markets stemming from local loop telecommunications and CATV service are set forth in "The New Local Loop: Wireless & Wireline Telecommunications - 1996 to 2001 Strategic Analysis." The report was prepared over 10 months, from June, 1995, through March, 1996, by ABI's Senior Research Consultant, Francis X. Duffy. The qualitative section of the report offers details and insight about the local loop industry, regulation, established technologies, emerging technologies, and opportunities. The quantitative section takes a careful look at the US market by subscribership revenues and equipment sales. ABI is a high-tech strategic research firm based in Oyster Bay, New York. The technological developments now unfolding could change the shape of the wireless industry forever. For more details or to speak with the analyst regarding this study or a specific area of this study, please contact Tim Archdeacon at 516-624-3113.

"At the Asterisk, the Time Will be..."

NIST recently upgraded the equipment that provides its Automated Computer Time Service, or ACTS, for setting clocks in computers. Computers (and other equipment) can be programmed to call (303) 494-4774 for the time of day. The service has 12 incoming lines (in rotary), and PC based servers can accept calls at speeds up to 9,600 baud. These changes should provide smoother service to a broader variety of user modems. Software to use ACTS is available for \$45 by ordering Standard Reference Material 8101b from the NIST SRM Program, Tel: (301) 975-6776; Fax: (301) 948-3730; E-mail: srminfo@enh.nist.gov. World Wide Web surfers can see a printout of the exact time at http://time.nist. gov:13/, but this will not reset your computers clock. And what about that "asterisk" in the title?

We designed and manufactured an antenna to find a polar bear in a snowstorm. And a rattlesnake in a sand dune. We can develop the right antenna for your application, too.



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INFO/CARD 20

RF news continued

The NIST computer systems use and asterisk as the on-time marker, analogous to the tone transmitted on shortwave radio and on the telephone at (303) 499-7111.

Jobs May Be Getting Scarcer For PhDs

Landing the job of their choice may

be getting harder for science and engineering doctoral students, a recent national study suggests. Prospects are reported as particularly bad for those seeking a university research position. As few as one out of every three retiring university professors are replaced as of 1992. The number of job applicants outweighs the number of jobs available. Dr. Mary Frank Fox, a



Georgia Institute of Technology Sociologist, and Georgia State University Economist Dr. Paula Stephan of the Policy Research Center, looked at five fields of study: chemistry, computer science, electrical engineering, microbiology and physics. In a survey of 3,800 student, 2,400 students responded about their perceptions of their career fields. According to Fox, today's science and engineering doctoral students are facing three major economic changes that may hamper their career prospects, including government deficits and their effects on federal funding for research and research training; the end of the Cold War and its effect on funding for scientific research tied to defense; and the lifting of the mandatory age for retirement and its effect on the replacement of scientific personnel.

Contracts

Baltimore Selects Motorola — The City of Baltimore Board of Estimates has awarded Motorola a \$38 million contract to manufacture and install a new 800 MHz radio communications system for the city's public safety and public service agencies. The new Motorola ASTRO[™] digital system initially will service City of Baltimore Fire Department. The system replaces a number of individual radio systems whose capabilities couldn't meet growing department communications needs.

Cancom Completes \$20 Million Network — Canadian Satellite Communications (Cancom) put into service an MPEG-based digital video compression system from Scientific-Atlanta to make enhancements to its programming lineup without increasing costs for satellite transponder space. The US \$20 million network includes 25 uplinks and approximately 14,000 Scientific-Atlanta digital satellite receivers.

Motorola Division Chooses ADRA

- The Advanced Messaging Systems Division (AMSD) of Motorola, Inc., in Fort Worth, Texas, has selected Matrix, Adra System's unique PDM system, to manage and control information related to the product design process. Matrix will enable employees in AMSD to manage information and control the processes that govern the lifecycles of that information.

The Giga-tronics 8540B is the World's Best Power Meter for Communications Testing. But Don't Take Our Word for It.

Power meters generally fare well when characterizing simple continuous wave (CW) or pulsed RF/microwave signals. But modern communications signals, with digital information on phase-, frequency-, and amplitude-modulated carriers, pose problems for most power meters. The exception is the 8540B from Giga-tronics. The latest version of this high-speed measurement tool performs true average power measurements even on multitone signals with digital modulation.

> Microwaves & RF June 1995



Call 800-726-GIGA (4442) or fax 510-328-4700 for more information for a demonstration.



Giga-tronics Incorporated = 4650 Norris Canyon Road = San Ramon, California 94583 = Telephone: 800-726-4442 or 510-328-4650 = Telefax: 510-328-4700

RF industry insight

DSP Grows in Importance for New Communications Technologies

By Gary A. Breed Editor

On the north side of the Dallas, Texas metropolitan area, in Richardson, the Texas Instruments campus has five construction cranes and hundreds of hard hats, turning former parking lots into wafer fabs. The main reason for this huge investment is to supply digital signal processing (DSP) ICs for the millions of new wireless, cable, and satellite communications products that will come to market in the next few years.

Like Texas Instruments, other DSP companies like AT&T Microelectronics, Motorola Semiconductor, Harris and Analog Devices are making similar investments in the future. They, and dozens of other companies with interface and data handling ICs in their product lines, are anticipating dramatic growth in DSP.

Because most new communications technologies are data-based, rather than analog-based, DSP is a natural fit for "back-end" signal filtering, error correction, echo-cancelling, and other signal processing functions. With data transmission requiring smaller signalto-noise ratios than analog methods (but with wider bandwidth requirements), 10-bit or 12-bit digital words are sufficient, and easily handled in 16-bit DSP devices. For the dynamic range associated with analog transmission, 20 bits or more would be required.

Analog is Still Part of the Process

Analog techniques are, of course, essential for the amplification and detection of the RF transmissions that carry the data. In the transmitter, an RF carrier signal is modulated by the data, up-converted to the operating frequency, amplified, switched and filtered before being sent to the antenna. In the receiver, the signal is also filtered, switched, down-converted, filtered again, and demodulated to present the recovered data to the DSP



Like other DSP IC makers, AT&T Microelectronics is seeking new, large markets in wireless communications. Their DSP1627 has 50 MIPS computing power when operating from a 3 volt supply; 70 MIPS at 5 volts.

part of the system.

Harris and Stanford Telecom have pioneered the development integrated digital down- and up-conversion devices, which can eliminate one part of the analog RF circuitry. However, the speed/frequency range of these



MPEG 2 video compression is a major technology in new communications applications. The VLSI Technology VES2020 transport demultiplexer is part of the data management system used in digital video set-top boxes.

devices is not yet high enough to replace the intermediate frequency (IF) circuitry in most communications equipment operating at 800 MHz or higher. The accuracy of digital processing can improve performance in those applications where digital conversion and demodulation are appropriate.

DSP is here to stay, in conjunction with analog techniques. It is safe to suggest the the real "revolution" in communications technology today is the integration of many different techniques in a single product — or on a single chip. Radios aren't just radios any more. They are microprocessorcontrolled, power-managed, analog and digitally processed devices. The separation of RF and digital into different disciplines is disappearing.

Digital signal processing is powerful, precise, and cheap when compared to equivalent analog processing. DSP's place in the next generation of radiobased products is assured. *RF*

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

	NO.	TEMEX	PAS	SBAND		STOP	BANE)	LOSS	RIPPLE	ULT. REJ.	TERM.(Rp//Cp)
	POLES	P/N	dB	±KHz	dB	±KHz	dB	±KHz	dB	dB-MAX	dB-MIN.	OHM/PF
	2	TE5180	3	3.75	15	12.5	-		2	1.0	50	850//+6
	4	TE5190	3	3.75	30	12.5	-	-	3	2.0	70	850//+5
	6	TE5200	6	3.75	60	12.5		-	4	2.0	90	850//+5
N	8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850//+5
I	2	TE5220	3	6.50	15	20.0	-	-	2	1.0	50	1300//+2
5	4	TE5230	3	6.50	30	22.5		-	3	2.0	70	1400//0
-	6	TE5240	6	6.50	60	22.5		-	4	2.0	90	1400//0
4	8	TE5250	6	6.50	60	17.5	80	22.5	4	2.0	100	1400//0
-	2	TE5260	3	7.50	15	25.0	-	-	2	1.0	50	1500//0
N.	4	TE5270	3	7.50	30	25.0	-		3	2.0	70	1600//0
	6	TE5280	6	7.50	60	25.0	-		4	2.0	90	1600//0
	8	TE5290	6	7.50	60	20.0	80	25.0	4	2.0	100	1600//0
	2	TE5300	3	15.0	15	50.0	-		2	1.0	45	3000//0
	4	TE5310	3	15.0	30	45.0	-		3	2.0	60	3000//-1
	6	TE5320	6	15.0	60	45.0	-		3	2.0	90	3000//-1
	8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000//-1

	NO.	TEMEX	MODE	PASS	BAND	STO	PBAND	LOSS	RIPPLE	ULT. REJ.	TERM.(Rp//Cp)
N	POLES	P/N		dB	±KHz	dB	±KHz	dB	dB-MAX	dB-MIN.	OHM/PF
I	2	TE9420	3-0T	3	3.75	18	16.0	3	1	40	2000//-1.0
5	4	TE9310	3-OT	3	3.75	30	12.5	3	1	70	2000//-1.0
	2	TE7420	3-0T	3	7.50	18	28.0	2	1	40	3000//-1.0
0	4	TE7430	3-OT	3	7.50	40	30.0	3	1	70	3000//-1.0
S	2	TE7440	3-OT	3	15.0	15	47.0	2	1	40	8000//-1.5
4	4	TE7450	3-OT	3	15.0	30	50.0	3	1	70	8000//-1.5
	2	TE7730	FUND	3	15.0	15	50.0	2	1	40	1100//+1.5
	4	TE7740	FUND	3	15.0	40	60.0	3	1	70	800// +1.0

N	NO.	MODE	MODE PASSBAND			STOPBAND				RIPPLE	TERM.(Rp//Cp)	
I	POLES	P/N		dB	*KHz	dB	±KHz	dB	KHz	dB	dB-MAX	OHM//PF
Σ	2	TE10400	3-0T	3	7.5	18	30	35	-910	2	1	2000//-1
0	4	TE10410	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
o	2	TE10420	3-OT	3	10	15	30	35	-910	2	1	2500//-1
7	4	TE10430	3-OT	3	10	35	40	80	-910	3	1	2500//-1

N	NO.	TEMEX	MODE	PASSBAND		STOPBAND				LOSS	RIPPLE	TERM.(Rp//Cp)
Î	POLES	P/N		dB	[±] KHz	dB	[±] KHz	dB	KHz	dB	dB-MAX	OHM//PF
M	2	TE10440	3-0T	3	7.5	18	30	35	-910	2	1	2000//-1
	4	TE10450	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
9	2	TE10460	3-OT	3	10	15	30	35	-910	2	1	2500//-1
0	4	TE10470	3-0T	3	10	35	40	80	-910	3	1	2500//-1
0,	4	TE10480	3-0T	3	15	30	50	80	-910	3	1	4000//-1



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INFO/CARD 24

RF signal processing

Sideband Suppression Considerations for Complex Baseband Processing of FDM Waveforms

By Rick Roberts Harris Corporation

Frequency Division Multiplexing (FDM), in its various forms [1], continues to be popular for data communications. With the advent of the digital processor, a common technique for generation and processing of FDM is the zero Hertz IF; that is, quadraphase complex baseband waveform processing. This paper shows that the quadraphase characteristics can be degraded during frequency translation to/from a real frequency by I/Q imbalance resulting in cross channel interference between pairs of carriers. The amount of interference is quantified as sideband suppression. It is suggested that the frequency translation be done via DSP to achieve superior sideband suppression.

With the advent of digital signal processing, a common technique for generation and processing of FDM is a zero hertz IF baseband waveform in conjunction with the complex IFFT/FFT. It is desired that this baseband waveform be complex quadraphase, however, the quadraphase characteristics can be degraded during frequency translations due to I/Q modulator/demodulator amplitude/phase imbalance. As shall be shown, this I/Q imbalance can result in cross channel interference between pairs of FDM carriers. If the interference is extreme, one may be forced to use cross channel interference cancellation techniques to correct the problem. In this paper, the mechanism that gives rise to less than perfect sideband suppression will be discussed and an expression for the resulting level of interference will be derived. A method for doing the frequency translation via DSP is shown, which results in excellent sideband suppression with insignificant levels of cross channel interference.

Sideband Suppression

The concept of sideband suppression is illustrated by considering the down conversion of a real [2] bandpass signal to a zero Hertz centered complex bandpass signal using the I/Q demodulator shown in Figure 1. The terms A and q in the I arm respectively represent LO amplitude and phase imbalance.

The goal is to generate a single sideband LO signal, via an I/Q demodulator, so as to implement the Fourier frequency translation property as illustrated in Figure 2. However, any phase or amplitude imbalance between the I and Q arms of the demodulator introduces an undesirable sideband component to the LO (i.e. leaky LO) that causes spectral aliasing from the negative frequency spectral envelope as shown in Figure 3. The level of the aliasing interference is proportional to the level of the undesirable leaky LO sideband component, hence the term "sideband suppression".

This concept of spectral aliasing due to poor sideband suppression also applies to the upconversion process,









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	RF		1/0		DULATOR	RS
-0-	FR	EQ.		ONV. DSS	CARRIER REJ.	S
MODEL	(M f	fu	_ (×	α dB)	(-dBc) Typ.	
N QA-10M MIQA-21M MIQA-70M MIQA-70ML MIQA-91M MIQA-100M N QA-109M MIQA-195M	9 20 66 86 95 103 185	11 23 73 73 95 105 113 205	58 62 5.7 55 55 55 56	0 20 0 14 0.10 0 10 0.10 0 10 0 10 0 10 0 10	41 50 38 38 38 38 38 38 38	
MIQC-38M MIQC-88M MIQC-176M MIQC-895M MIQC-1785M MIQC-1860M	34 52 104 868 1710 1805	38 88 176 895 1785 1880	5.6 5.7 5.5 8.0 9.0 9.0	0.10 0 10 0 10 0 10 0 30 0 30	48 41 38 40 35 35	
MIQY-70M MIQY-140M	67 137	73 14 3	5.8 5.8	0.20	40 34	dal
JCIQ-88M JCIQ-1,76M	52 104	88 176	56 56	01	40 35	Jel

l. I U	×	DNV. DSS dB) o	CARRIER REJ. (-dBc) Typ.	SIDEBAND REJ. (-dBc) Typ	HAF SUPPI (-dBc) 3xl/Q	RM. RESS Typ. 5x1/Q	PRICE \$ Qty. (1-9)
11 23 73 73 95 05 13 05	58 62 6.2 5.7 55 55 55 56	0.20 0.14 0.10 0.10 0.10 0.10 0.10 0.10 0.1	41 50 38 38 38 38 38 38 38 38	40 40 38 38 38 38 38 38 38 38	58 48 48 48 48 48 48 48 48	68 65 58 58 58 58 58 58 58 58	49 95 39 95 49 95 49 95 49 95 49 95 49 95
38 88 76 195 185 180	5.6 5.7 5.5 8.0 9.0 9.0	0.10 0 10 0 10 0 10 0 30 0 30	48 41 38 40 35 35	37 34 36 40 35 35	54 52 47 52 40 40	65 66 70 58 65 65	49.95 49.95 54.9 90.9 99.95
73 43	5.8 5.8	0.20	40 34	36 36	47 45	60 60	19 95 19 95
88 76	56 56	01	40 35	35 35	45 45	65 65	49 95

-			
	RF		I/Q D
	FF	REQ.	
MODEL	fL	fU	× n
MIQA-10D MIQA-21D NIQA-70D	9 20 66	11 23 73	60 61 62
MIQC-38D MIQC-60WD MIQC-895D	34 20 868	38 60 895	5.5 5.3 8.0

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(MHz)			_ (dB)		(dB)	(Deg.)	(-dBc) Typ.	Oty.
fL	fu		x	σ	Тур.	Тур.	3x1/Q	5x1/Q	(1-9)
9	11	6	60	0.10	0.15	10	50	65	49 95
20	23		61	0 15	0 15	07	64	67	49 95
66	73		62	0 10	0 15	07	56	58	49 95
34	38		5.5	0 10	0.10	0.5	60	65	49 95
20	60		5.3	0 10	0.15	1.0	55	67	79 95
868	895		8.0	0.20	0.15	1.5	40	55	99 95
1.15	1.35		50	0 10	0 15	10	59	67	29 95
67	73		55	0 25	0 10	05	52	66	19 95
137	143		55	0 25	0 10	05	47	70	19 95
			Su	face M	ount Mode	ls			
104 868	176 895		5.5	01	015	2	52 45	65 65	54.95 99.95
710	1785		8	0.2	02	2	50	65	99.5

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associated with a I/Q modulator, as shown in Figure 4.

The exact expression for sideband suppression in terms of the I/Q phase and amplitude imbalance is derived in the appendix; but for those cases where it is reasonably near quadrature, the level of sideband suppression is given as:

$$\gamma_{\rm dB} \approx -6 + 10 \log \left[{\rm A}^2 + \theta^2 \right]$$

where A is the amplitude error and q is the radian phase error. For example, given an amplitude imbalance of




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Appendix: Quadraphase LO Sideband Suppression Let the LO signal be expressed as s(t) = a(t)+jb(t) where $b(t) = sin(\omega_0 t)$ and $a(t) = (1-A)cos(\omega_0 t + \theta)$. Notice that the term a(t) contains both the phase and amplitude error. By straightforward substitution of the exponential expression for the sine and cosine, we can decompose s(t)into a left spectral shifting function and a right spectral shifting function as shown below.

$$s(t) = \left[\frac{1}{2} + \left(\frac{1+A}{2}\right)e^{-j\theta}\right]e^{-j\omega_0 t} + \left[\left(\frac{1+A}{2}\right)e^{-j\omega} - \frac{1}{2}\right]e^{j\omega_0 t}$$

The sideband suppression is given as the ratio of the magnitude squared of the arguments of the exponential terms and is expressed in dB as:

$$\gamma_{dB} = 10 \log \left(\frac{2 + A(A-2) - 2(1-A)\cos\theta}{2 + A(A-2) + 2(1-A)\cos\theta} \right)$$

If θ is small (i.e. $\theta <<1$ radian), we can approximate the cosine term as:

$$1-\frac{\theta^2}{2}$$

In addition, if A<<1 we can approximate γ as:

$$\gamma_{dB} = -6 + 10 \log |A^2 + \theta^2|$$

0.35 dB (A = 0.04) and a phase imbalance of 0.065 radians, the sideband suppression is -28 dB. As is explained below, this imbalance may not allow enough performance margin required for a particular application, especially if the digital modulation on each FDM carrier is high order QAM.

FDM Cross Channel Interference

The spectrum of Figure 2 emphasis the lack of spectral symmetry with respect to the bandpass center frequency. This is especially the case for FDM, where perhaps the level of each carrier in the FDM spectrum is the same but the information content of the modulation per carrier is completely uncorrelated. Thus, with the introduction of an aliasing sideband, two or more carriers will pair up with each other causing spectral overlap as shown in Figure 5.

Assuming that the modulation on each of the FDM carriers is digital in nature, the net result will be degraded bit error rate performance of the desired signal. Even if a reasonable level of correct decisions on the desired signal is maintained, cross





Figure 6. Digital down/up conversion technique using DSP processing.

interference cancellation will require complicated signal processing between the desired signal and the interfering signals. It is important to note that the case for FDM is different than that for QAM. In QAM, the spectral content of both the I and Q signals completely fall within the demodulator passband. In this case, trigonometric relationships can be derived that show how a simple lattice network can be inserted in the I/Q arms to undo the signal distortion caused by poor I/Q quadrature.

DSP Frequency Translation

An alternative technique to the traditional analog I/Q modulator and demodulator for doing the frequency translation is the NCOM (numerically controlled oscillator/modulator), an example of which is the Harris HSP45116, in conjunction with the appropriate interpolator or decimator. Figure 6 shows a scheme for both upconversion and down conversion. Notice that the analog to digital interface occurs on a bandpass waveform (i.e. direct IF sampling) which allows the quadraphase characteristics of the frequency conversion to be determined solely by the precision of the digital processing. A-to-D subsampling techniques [3] can be used to introduce a

digital IF prior to baseband down conversion to minimize sampling clock speeds and hence reduce power consumption. The precision of the overall process is arbitrarily high which results in determinable sideband rejection.

The Harris HSP50016 Digital Down Converter (DDC) is an example of a highly accurate single chip down converter that integrates the LO, the mixer and the decimating lowpass filters all in one package. Its LO serves as an example of the current commercial state-of-the-art in digital LO technology. Table 1 lists the measured sideband suppression for the Harris HSP50016 digital downconverter.

Carrier Frequency	Suppression dB		
Fs/8	118.88		
Fs/5	118.80		
Fs/4	116.04		
Fs/3	117.87		

 Table 1. HSP50016 Measured Sideband Suppression Carrier Frequency Suppression dB

Conclusion

Analog I/Q complex waveform modulators/demodulators have finite sideband suppression due to amplitude and/or phasing errors between the I/Q arms. These errors can cause unwanted sideband leakage which, for an FDM waveform, causes interference between carrier pairs with respect to zero Hertz. DSP techniques exhibit much higher ratios of sideband suppression and the DSP processing of the FDM waveform yields much lower carrier interference ratios. In many cases, the DSP approach will be more attractive than the traditional analog approach when high sideband sup-RF pression ratios are required.

References and Notes

1. Orthogonal FDM and/or DMT (Discrete Multitone)

2. Real simply means a "real-world" analog signal with resulting symmetry between the positive and negative frequency spectrum.

3. Groshong, R. and Stephen Ruscak, "Undersampling Techniques Simplify Digital Radio", *Electronic Design*, May 23,1991.

About the Author

Richard Roberts has a PhD and MSEE From the Florida Institute of Technology and a BSEE From the University of Wisconsin. For the last 16 years he has worked for Harris Corporation where he is currently a communications systems engineer in the Semiconductor Sector. He is involved with both analog and digital modem design and has several patents and publications in the area of signal processing. He can be reached at (407) 724-7022 or via E-mail at rrober02@harris.com.



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RF spread spectrum

A Spread Spectrum Sliding Correlator System for Propagation Measurements

By: William G. Newhall, Theodore S. Rappaport, Dennis G. Sweeney Mobile & Portable Radio Research Group Virginia Tech

The design of a wireless communication system is dependent upon the propagation environment in which the system is to be used. Factors such as the time delay spread and the path loss of a radio channel affect the performance and reliability of a wireless system. These factors can be accurately measured through RF propagation measurements in the environments in which anemerging wireless technology is to be deployed.

This article describes the design of an RF channel measurement system based on the spread spectrum sliding correlator technique. Presented first are discussions on propagation time delay measurements and three techniques which may be used to measure RF channel responses. The article then focuses on the sliding correlator technique, and explains system parameters and hardware which implements the sliding correlator system used at the Mobile and Portable Radio Research Group (MPRG) at Virginia Tech. Our goal is to make the theory and implementation better understood for RF engineers.

Channel sounding, the process of determining the radio propagation

delays and attenuation, has become an important part of the design of modems for new frequency bands or new operating environments. In the past, channel sounding has been carried out by scientists and research labs, but with the demand for wireless applications and systems, channel sounding is becoming important for engineers who work with the design, development, and installation of new wireless systems.

Time Delay Measurements

Typical radio channels suffer from time dispersion due to reflecting objects and scatterers in the propagation environment. Reflectors and scatterers create multiple propagation paths (multipath) between the transmitter and receiver, each path having a different physical length. Paths having different physical lengths induce different propagation delays between the transmitter and the receiver, causing intersymbol interference and limiting data rate. This time dispersion exhibited by typical RF channels can be measured by observing the propagation delays which a transmitted signal experiences.

Because of the importance of the

multipath structure in determining the propagation delay and the fading effects, a number of wideband (i.e., pulse) channel sounding techniques have been developed. The term "wideband" is used to denote the fact that the measurement system is able to resolve multipath by using a wider bandwidth than the intended radio system. Wideband channel sounding techniques may be classified as direct pulse measurements, swept frequency measurements, and sliding correlator measurements [1].

Direct Pulse System

A simple channel sounding approach is the direct RF pulse system. This technique allows one to determine the power delay profile of an RF channel [2] [3]. A direct pulse system is essentially a wideband pulsed bistatic radar, transmitting a pulse of width τ seconds, and uses a receiver with a wide bandpass filter (BW = $1/2\tau$ Hz). The pulse is repeated every second or fraction of a second so that it may be averaged at the receiver. The received signal is amplified, detected with an envelope detector, and displayed and stored on a high speed oscilloscope. This gives an immediate measurement



Figure 1. Direct pulse RF Channel measurement system.



Figure 2. Frequency domain channel sounder

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of the square of the channel impulse response convolved with the probing pulse. If the oscilloscope is set on averaging mode, then this system can provide a local average power delay profile. An attractive aspect of this system is the lack of complexity, since offthe-shelf equipment may be used. A block diagram of the direct pulse system is shown in Figure 1.

The main problem with this system is that it is subject to interference and noise, due to the wide filter passband required for multipath time resolution. Also, the pulse system relies on the ability to trigger the oscilloscope on the first arriving signal. If the first signal becomes blocked due to the channel, it is possible for the system to miss the first arriving multipath.

Frequency Domain Channel Sounding

Because of the dual relationship between the time domain and the frequency domain, it is possible to measure the channel impulse response using frequency domain techniques.

Figure 2 shows a frequency domain channel sounder. A vector network analyzer controls a synthesized frequency sweeper, and an S-parameter test set is used to monitor the frequency response of the channel. The sweeper scans a particular frequency band stepping through discrete frequencies. Due to the Fourier transform, the number and spacings of these frequency steps impact the time resolution of the impulse response measurement. For each frequency step, the S-parameter test set transmits a known signal level at port 1 and monitors the received signal level at port 2. These signals allow the analyzer to determine the complex response (i.e., transmissivity $S_{21}(\omega)$) of the channel over the measured frequency range. The transmissivity response is a frequency domain representation of the channel impulse response, and is converted to the time domain using inverse discrete Fourier transform (IDFT) processing, giving a band-limited version of the impulse response. In theory, this technique works well, providing amplitude



and phase information in the time domain. However, the system requires calibration and hard-wired synchronization between the transmitter and receiver, making it useful only for very close measurements (e.g., indoor work). Another limitation with this system is the non-real-time nature of the measurement. For time-varying channels, the channel frequency response can change rapidly during a sweep period, giving an erroneous impulse response measurement.

Spread Spectrum Sliding Correlator Channel Sounding

The most popular channel sounding technique is based on a spread spectrum sliding correlator (see Figures 3 through 6). The advantage of a spread spectrum system is that while the probing signal may be wideband, it is possible to detect the transmitted signal using a narrowband receiver preceded by a wideband mixer, thus improving the dynamic range of the system as compared to the direct RF pulse system.

In a spread spectrum channel sounder, a carrier signal is "spread" over a large bandwidth by mixing it with a wideband pseudo-noise (PN) sequence. The power spectrum envelope of the transmitted spread spectrum signal is the classic $\sin(f)/f$ shape and has a null-to-null bandwidth of $BW=2R_c$ where R_c is the PN sequence code clock rate (chip clock rate) [4]. Typical chip rates are several megabits per second, which results in a transmitted bandwidth several MHz wide.

The spread spectrum signal is then received, filtered, and "despread" using a PN sequence generator identical to that used at the transmitter. Although the two PN sequences are identical, the transmitter chip clock is run slightly faster than the receiver chip clock. Mixing the chip sequences then implements a "sliding correlator" [4]. When the PN code of the transmitter catches up with the PN code of the receiver, the two chip sequences are aligned, giving maximal correlation. When the two sequences are not maximally correlated, mixing the incoming spread spectrum signal with the unsynchronized receiver chip sequence spreads this signal into a bandwidth at least as large as the receiver's reference PN sequence. In this way, the narrowband filter that follows the correlator can reject almost all of the

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Figure 3. Sliding correlator measurement system transmitter configured for 5850 MHz (50 MHz chip clock)



Figure 4. Sliding correlator measurement system receiver c onfigured for 5850 MHz (50 MHz chip clock)

incoming signal power. This provides the processing gain realized in all spread spectrum systems and allows the receiver to reject passband interference, unlike the direct RF pulse sounding system or the swept frequency system.

When the incoming signal is correlated with the receiver sequence, the signal is collapsed back into the original bandwidth (i.e., "despread"), envelope detected, and displayed on an oscilloscope. Since different incoming multipaths have different time delays, they maximally correlate with the receiver PN sequence at different times. The energy of these individual paths pass through the correlator at different time depending on the propagation time delay. Therefore, after envelope detection, the channel impulse response convolved with the pulse shape of a single chip is displayed on the oscilloscope. When expressed in units of power, the pulse shape shown on the oscilloscope is called the power delay profile of the channel, from which the path loss and RMS delay spread can be calculated. Cox [5] first used this method to measure channel impulse responses in outdoor suburban environments at 910 MHz.

Devasirvatham [6, 7] used a direct sequence spread spectrum channel sounder to measure time delay spread of multipathcomponents and signal level in office and residential



Figure 5. Sliding correlator measurement system transmitter configured for 915 MHz (10 Mhz chip clock)

buildings at 850 MHz. Bultitude [8] has also used this technique for indoor and microcellular channel sounding work, and MPRG researchers have used this technique for a range of measurements in buildings, cities, and train yards.

Sliding Correlator System Parameters

Explained in this section are the parameters used for the sliding correlator measurement system. These parameters are set to achieve the desired time resolution, slide factor, processing gain, and maximum unambiguous range capabilities of the system.

The time resolution Δt of multipath components is defined as the amount of time by which the components must be separated in order to resolve individual components in a measured profile and is given by

$$\Delta \tau \approx 2T_{\rm c} = \frac{2}{R_{\rm c}}$$

where $T_{\rm c}$ is the chip period and $R_{\rm c}$ is the chip rate. In actuality, the time resolution is smaller than $2T_{\rm c}$ and can be estimated as $T_{\rm c}$. The slip rate R_{slip} is defined as the difference between the transmitter and receiver chip rates, and can be expressed as:

$$R_{slip} = \alpha - \beta$$

where α is the transmitter chip rate and β is the receiver chip rate [1]. (R_c is equal to α and is approximately equal to β for time resolution calculations.) An important quantity in sliding correlator measurements is γ , called the slide factor, which relates observed multipath delay time at the receiver to the actual propagation delay time. The slide factor γ is defined as:

$$\gamma = \frac{\alpha}{\alpha - \beta}$$

and can be simply expressed as the transmitter chip rate α divided by the slip rate R_{slip} . The slide factor relates actual propagation time $t_{propagation}$ to observed oscilloscope time $t_{observed}$ as:

$$t_{propagation} = \frac{t_{observed}}{\gamma}$$

This means that if the slide factor γ is 1000, then a 100 μ s delay observed between two multipaths at the receiver is actually a delay of 100 ns in propagation time between the multipaths.

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er by setting its resolution bandwidth equal to the desired bandwidth. For sliding correlator measurements, the choice of resolution bandwidth is determined by:

$$ResBW = 2(\alpha - \beta)$$

to obtain the processing gain realized by the narrowband filter. The *ResBW* parameter determines the detector bandwidth required for proper display of the received multipath.

The unambiguous range of the system is the maximum excess path length of a multipath component which can be measured by the system. Since the PN sequences at the transmitter and receiver repeat periodically, and have virtually the same period, a received multipath component must have an excess delay less than the period of the sequences in order to be correctly interpreted. In other words, the apparent delay of a multipath component may not be the true delay if the true delay exceeds the delay corresponding to the maximum unambiguous range. The period of the PN sequence is:

$$\tau_{PN} = T_c l$$

where l is the number of chips in the PN sequence. Multiplying this quantity by the speed of light yields the unambiguous range dr of the system:

$d_r = T_c lc$

For example, if the transmitter uses a chip rate of 50 MHz, and the PN sequence length is 2047 chips, then the maximum unambiguous range would be about 12.2 km.

The dynamic display range can be defined as the difference between the maximum measurable correlation peak and the noise floor displayed on the oscilloscope. The dynamic display range of the sliding correlator system is limited by the one or a combination of the following factors: the index of discrimination of the system, the LO-RF isolation of the mixer, and system thermal noise. The dynamic display range for the current MPRG system is estimated to be between 20 dB and 25



dB, and is limited by the isolation of the mixer.

The MPRG Spread Spectrum Channel Sounder

A sliding correlator measurement system has been in use at the Mobile and Portable Radio Research Group at Virginia Tech for several years and has been redesigned for greater flexibility and use at higher bands. It uses "plug-in" amplifier and filter stages to accommodate the desired carrier frequency and bandwidth. The system is used to sound RF channels at 900 MHz, 1900 MHz, 2450 MHz and 6 GHz; however, the capability of measuring 9 GHz, 18 GHz, and 26 GHz bands are planned.

Two typical configurations of the MPRG measurement system are shown in Figures 3 through 6. Figures 3 and 4 show the system used to sound channels at 6 GHz, and Figures 5 and 6 show the system configured to sound the 902 MHz to 928 MHz ISM band.

Baseband Hardware

It is clear from Figures 3 through 6 that the baseband processing remains the same for different RF configurations. An Efratom rubidium frequency standard is used as a 10 MHz frequency reference at both the transmitter and receiver. At the transmitter, the rubidium standard drives the 10 MHz reference input of the HP86360A sweeper, and the frequency multiplier derives the transmitter chip clock from the 10 MHz reference output of the sweeper. The frequency multiplier was built at MPRG and can select 10 Mchip/s, 20 Mchip/s, 50 Mchip/s, and 100 Mchip/s chip rates. The eleven stage ECL shift register clocked by the output of the frequency multiplier produces a PN sequence 2047 chips in length.

At the receiver, another rubidium frequency standard drives the reference input of the receiver signal generator, a Fluke 6062A signal generator. Two PN sequence generators are used at the receiver, both produce the same sequence but are clocked at slightly different rates. The 'RX PN' and 'TX PN' sequence generators are identical, except that the TX PN sequence is clocked at the same rate as the transmitter PN sequence generator, and the RX PN sequence is clocked at a slightly slower rate (the RX PN clock signal is produced by the Fluke 6062A, whereas the TX PN clock at the receivUncover small signals close to carriers with -105 dBc/Hz internal phase noise.

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er comes directly from the rubidium source using a frequency multiplier). The RX PN signal drives the IF port of the mixer to despread the received spread spectrum signal. The Tektronix spectrum analyzer, set to zero span at the carrier frequency, acts as a receiver and narrowband filter and displays received power versus time. The oscilloscope trace shows the vertical output of the spectrum analyzer, a voltage which is proportional to the log10 of received power, versus time. The waveform shown on the oscilloscope is the power delay profile of the RF channel being measured, and is recorded using the Toshiba PC via the GPIB bus. Figure 7 shows a sample power delay profile for a 915 MHz channel using a chip rate of 10 Mchip/s.

The TX PN generator at the receiver and the use of the rubidium frequency standards permit absolute time delay measurements. The RX PN sequence generator at the receiver drives a second mixer at the receiver which correlates the output of the RX PN sequence generator with the TX PN sequence generated locally at the receiver. The result is a correlation peak which is used as a trigger pulse for the oscilloscope displaying the power delay profile.

Before the system is used for measurements, the TX PN sequence generator at the transmitter and the TX PN sequence generator at the receiver are synchronized. If the receiver reference frequency is exactly the same as the one in the transmitter and the two TX PN generators (the one in the transmitter and the one in the receiver) are synchronized together, then the TX PN generator in the receiver will produce exactly the same code as the one in the transmitter at the same time. In this way, the TX PN sequence generator in the receiver replicates the time at the transmitter. When the transmitter and receiver are separated, the trigger pulse produced by the correlation of the RX PN sequence and the TX PN sequence corresponds to time zero, and propagation time for line-ofsight or multipath components can be referenced to the trigger pulse.



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In order to keep the TX PN sequence generated at the transmitter synchronized with the TX PN sequence generated locally at the receiver, a highly stable frequency standard is required. The Efratom LCRO rubidium frequency standards suit this need. Without the use of the rubidium oscillators, a reference cable connecting the transmitter and receiver would be required throughout measurements, and would make long-distance propagation measurements impractical. With the frequency standards, only the link budget limits the transmitter-receiver (T-R) separation of the system.

Sychronization Circuit

Figure 9 is a simplified circuit for the TX PN sequence generator which includes the synchronizing circuits. The sync circuit permits synchronizing two TX PN generators so they produce the same code at the same time. One TX PN generator is at the transmitter and one is at the receiver, and they rely on the transmitter and receiver rubidium oscillators to remain sufficiently stable so that the receiver TX PN generator replicates the exact state of the TX PN generator at the transmitter.

The sync circuit consists of U5 which is a pair of ECL D-type flip-flops (FFs). U5A is configured as a Set/Reset (R/S) FF. This R/S FF is connected to the stop/run switch. When the switch is placed in the "stop" position, the Q output of U5A is forced HI and the D input of U5B goes HI as well. At the next clock transition, this HI is transferred to U5B's Q output. Due to the wired-OR nature of ECL, pin 13 of U4 in the PN shift register will now remain HI independent of the inputs on pins 14 and 15 of U4. Thisforces the shift register to fill with "1s" and "1s" will continue to circulate as long as the switch is in the "stop" position. In addition, U5B's Q output goes LOW and the LED lights. The LED is on in the stop mode.

When the switch is placed in the "run" position, U5A's Q output is now forced LOW. This LOW is transferred to the U5B's Q output at the next clock transition. The U4 pin 13 output now controls the operation of the shift register and the PN sequence generator begins normal operation. The sync circuit provides the eleven stage PN sequence generator illustrated in Figure 9 a way to begin from the state: 1111111111.

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Figure 6. Sliding correlator measurement system receiver configured for 915 MHz (10 MHz chip clock)

The two TX PN sequence generators are synchronized by connecting the run/stop switch in each generator in parallel through a cable. The switches have a center-off position, and with one off, the other will control both PN generators. Now, when the controlling switch is set to "run" mode, the two TX PN sequence generators will start together from the same state. After this simple "back-to-back" calibration procedure, the transmitter and receiver may be separated to make measurements without a cable connecting them together.

RF Hardware

The transmitter (see Figures 3 and 5) uses the HP86360A synthesized sweeper to generate a CW carrier at the center of the band being measured. This carrier is BPSK modulated by the PN sequence using a mixer, and the output at the RF port of the mixer is amplified and transmitted into the channel. The attenuator placed after the mixer dampens reflections due to impedance mismatches in the system. The RF signal produced at the transmitter may be band-limited by a bandpass filter, depending on the channel being measured and the power levels used. An omnidirectional antenna is used at the transmitter, usually a biconical antenna or $\lambda/4$ monopole tuned to the band of interest.

The 915 MHz transmitter configuration shown in Figure 5 was used for a recent ISM band propagation measurement campaign, where a 10 MHz chip clock rate was used to sound RF channels. This configuration uses a Mini Circuits ZEM4300 mixer as a BPSK modulator, and a Mini Circuits ZHL4240 amplifier as a final amplifier. The transmitter antenna used for the 915 MHz configuration is a $\lambda/4$ monopole. Figure 8 shows the output of the transmitter.

The 6 GHz system in Figure 3 uses a Macom DCX4-26





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Figure 7. Sample power delay profile

mixer to BPSK modulate the 5850 MHz carrier. The BPSK signal is band-limited by an interdigital bandpass filter centered at 5850 MHz, which prevents transmitting energy below approximately 5700 MHz and above approximately 6000 MHz. The output of this filter is fed to a 20 dB gain amplifier which increase the power to approximately 5 cBm to drive the power amp. The power



Figure 8. (right) Spectrum of transmitter output with $f_c=915$ MHz and $R_c=10$ Mchip/s.

amplifier can supply up to 40 dBm.

For measurements above 2 GHz, the transmitter uses a KinTronic Laboratories wideband biconical antenna for omnidirectional measurements, and a different antennas are used at the receiver. For example, recent 6 GHz measurements have used a specially designed rectangular horn antenna with 17 dBi gain in conjunction with an antenna rotor to determine multipath angle of arrival.

Recent 915 MHz measurements

used a $\lambda/4$ monopole at the receiver. Attenuation at the front end of the 915 MHz receiver is increased for close measurements to reduce the PN noise floor caused by the transmitter PN sequence. Two five-pole interdigital bandpass filters are used at the input to reject cellular signals below 894 MHz and paging signals around 930 MHz, since narrowband signals within the pass band of the receiver reduce the dynamic range of the system. Two Mini Circuits ZHL series amplifiers provide 65 dB of gain prior to the ZEM4300 mixer, were the received signal is demodulated.

The 6 GHz receiver configuration (see Figure 4) uses a bandpass filter identical to that used at the transmitter, rejecting signals below 5700 MHz and above 6000 MHz so signals in the radar and communications bands do not overload the receiver. The isolator provides about 20 dB of attenuation in the reverse direction to reduce the effect of an impedance mismatch between the bandpass filter and the amplifier. The signal is amplified in a



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Figure 9. Schematic of the spread spectrum sliding correlator.

20 dB gain low noise amplifier and then passed through a step attenuator and an additional 6 GHz amplifier. The two 6 GHz amplifiers have approximately 40 dB gain. The gain is necessary to overcome the high noise figure of the spectrum analyzer that is used as the receiver and the loss in the balanced modulator used as the correlator. The total system noise figure is approximately 5 dB to 10 dB, depending upon the length and type of the antenna cable.

Other frequency bands are sounded by replacing filters, amplifiers, and mixers with components meeting the specifications required by the new band. System parameters, such as chip rates, slip rate, and resolution bandwidth, must also be set appropriately to sound the new band.

Favorable System Parameters

Many combinations of system parameters may produce satisfactory measurement data, MPRG researchers have found through experience that certain combinations obtain the most accurate representation of a channel response. The most recent 6 GHz measurements were performed using a transmitter (TX PN) chip rate α of 50 Mchip/s to obtain a time resolution $\Delta \tau$ of 20 ns, and a receiver (RX PN) chip rate b of 49.95 MHz. These chip rates imposed a slip rate R_{slip} of 50 kHz, a slide factor γ of 1000, and a maximum unambiguous range d_r of about 12.2 km. A spectrum analyzer resolution bandwidth and video bandwidth of 30 kHz was used.

Measurements made during the recent 915 MHz campaign used a transmitter (TX PN) chip rate α of 10

Mchip/s to achieve a time resolution of 100 ns while confining most of the transmitted energy within the 902 MHz to 928 MHz ISM band. This configuration used a receiver (RX PN) chip rate β of 9.990 MHz, imposing a slip rate R_{slip} of 10 kHz, and a slide factor of 1000. This yields a maximum unambiguous range d_r of approximately 61 km. The resolution and video bandwidth of the spectrum analyzer was 10 kHz.

Conclusion

The sliding correlator technique is a popular and effective method to measure RF channels, and has been used in the past by several researchers to characterize the delay spread of different environments. This article has described Virginia Tech's MPRG channel sounder, which is a sliding correlator system that can be easily configured for several different bands. The system described here has been used recently to study multipath as a function of receiver antenna angle in an urban environment at 6 GHz, and has been used to measure the 902 MHz to 928 MHz ISM band in a large train yard, a propagation environment which had not been well characterized for time dispersive characteristics. MPRG students and staff have provided propagation measurements for a wide range of sponsors, and are able to conduct field studies and propagation analysis in rapid fashion. A future article will discuss one such project, which characterized propagation conditions in an active train yard, and demonstrate how to analyze received data for modem design work. RF

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RF cover story

High Level Integration and Performance Comes to DDS

By Dave Crook and Jim Surber Analog Devices, Inc.

Analog Devices' AD9850 is the world's first complete monolithic 125 MHz high-performance Direct-Digital-Synthesis (DDS) system. This innovative CMOS device integrates a proprietary high-speed 32-bit DDS core, with a performance-optimized 10-bit D/A converter (DAC) and high-speed, low jitter comparator. The AD9850 shatters the barriers associated with existing DDS and DAC synthesizer solutions by providing major advancements in virtually every feature and specification category: level of integration, power dissipation, size, speed, dynamic performance, ease-of implementation and price.

DDS has long been recognized as a premier technology for generating frequency-agile waveforms (please refer to reference [1] for technical discussion on the basic operation of DDS). DDS technology excels at delivering extremely precise digital-control of its synthesized output frequency, fast frequency "hopping," and high frequency accuracy in its output. However, previous 100 MHz DDS solutions have been relatively expensive and necessitated the selection and implementation of a companion reconstruction DAC. This additional design burden, coupled with the undesirable aspects of previous discrete high-speed DDS solutions (i.e., complex interface requirements, high relative expense, and high power dissipation), has relegated high-speed DDS technology to applications in high-end industrial, instrumentation, and military systems. The AD9850 rewrites the DDS/DAC frequency synthesizer story.

What is the AD9850?

The individual functional blocks contained within the integrated AD9850 device, as shown in Figure 1, are innovations in their own right. The numerically-controlled oscillator (NCO) core uses a highly-efficient proprietary algorithm to calculate and implement the required phase-to-sine



conversion function "on the fly," reducing the additional processing steps and circuitry associated with the traditional RAM or ROM look-up table implementation of phase-to-sine conversion. This algorithm minimizes the power consumption and size required for the NCO function.

The D/A converter (DAC) block

within the AD9850 likewise sets a new standard. This 10-bit DAC core is optimized for wide-band spurious-free dynamic range (SFDR) and advances the state-of-the-art for 125 MHz CMOS DAC distortion performance. Innovative analog design techniques coupled with an advanced fabrication process have yielded a segmented



Figure 1. Functional block diagram of the AD9850.



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PROGRAMMED TEST SOURCES, INC. 9 Beaver Brook Road, Littleton, MA 01460 Tel: 508 486-3008 Fax: 508 486-4495 DAC architecture that delivers >50 dB of wideband SFDR at a 40 MHz output frequency (125 MHz clock rate). This optimized CMOS DAC core is singularly responsible for the high level of dynamic performance delivered by the AD9850. As a side note, this core is spawning a new Analog Devices product family of high-speed, low-cost, discrete CMOS DACs. These "transmitter-quality" DACs will be targeted specifically at communications applications where high dynamic performance and an economical price are required.

DDS Demystified

The AD9850 has eradicates one of the stigmas usually associated with discrete DDS and DAC solutions: complex implementation. Discrete DDS technology has been considered by many system designers to be saddled with a formidable learning curve more of an exotic art form than an exact science. The AD9850 will change the perception of DDS technology for the following reasons: • The AD9850's DDS core is an integrated functional block of a complete synthesizer solution

• The DDS is deliberately implemented in such a way that the user need not understand DDS theory to use this device to its fullest extent (no learning curve)

• The AD9850 completely eliminates the non-trivial, and somewhat overwhelming, task of successfully selecting and implementing a high-speed reconstruction DAC to support a discrete DDS solution

•The device contains a straightforward set of control functions and input words that greatly simplifies its operation as compared to previous DDS/DAC solutions

• The AD9850 is priced to make digital frequency synthesis a viable alternative to analog synthesis implementations

Easy-to-Use Digital Frequency Synthesis

The external support components and user control interface required to



implement the AD9850 are of minimal complexity. It requires only an accurate input reference clock, a word-load clock, and a 40-bit control/data word to achieve complete operation. The 40-bit load word function contains the 32-bit frequency tuning word, 5-bit phase modulation data, power-down enable, and the selection of the serial or parallel data load format. The parallel-load format requires five loads of an 8 bit data word. This 8-bit iterative load format reduces pin-count and allows the AD9850 to be packaged in its ultra-small 28-pin package (25 mil pin spacing) — the smallest footprint available for this function. The 8-bit data bus format is relatively easy to implement and it enables an output frequency change rate of up to 23 million changes-per-second. The reduced pin count, and resulting small package size, directly contribute to the AD9850's low cost.

AD9850 Applications

Frequency Synthesizer - The AD9850 is a highly flexible, high-performance device that will serve many of the traditional DDS markets and applications. However, its breakthrough attributes, including size, price, functional integration and performance, makes it uniquely positioned to bring DDS technology into high-volume, consumer-oriented communications applications. The two foremost applications for the AD9850 are: 1) low-distortion, frequency-agile synthesizer and, 2) frequency/phaseagile clock generator. These applications will be individually discussed in greater detail.

The AD9850 was designed to perform as a superior frequency-agile synthesizer. The device's 42 MHz output bandwidth is appropriate for many communications applications, including frequency-agile local oscillators, digital modulators, and carrier-generators. In these applications, the wideband distortion characteristics of the frequency output are most important; these characteristics are directly determined by the performance of the internal DAC. As was previously described, the internal DAC of the AD9850 was designed for optimum wideband dynamic performance over its Nyquist output range. The SFDR performance at several output frequencies is shown in Figures 2a, b, and c. The AD9850 is clocked at 125 MHz in all of these examples.

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Figure 2a. ADC9850 output SFDR measurement at 41 MHz.

The >50 dB of wideband SFDR performance at 40 MHz analog out delivered by the AD9850, sets a new benchmark for 10-bit CMOS DAC performance. Some frequency synthesizer applications, however, are more interested in the spectral characteristics close-in to the fundamental (narrowband). Figure 4 shows a plot of a 4.5 MHz output fundamental being clocked at 20.5 MHz. In this plot, the spectrum analyzer is displaying a 10



Figure 2b. ADC9850 output SFDR measurement at 15 MHz.

kHz frequency span and the SFDR of the output frequency is greater than 80 dB over this span.

Frequency-Agile Clock Generator

In the AD9850's clock generator application, the device's internal comparator is utilized. The NCO and DAC blocks generate a frequency-selectable 0-1 Vp-p output sinewave that is brought off-chip for external low-pass filtering. This filtering is necessary



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Figure 2c. ADC9850 output SFDR measurement at 1 MHz.

because the DAC output is a quantized (stepped) sinewave that contains significant energy at the aliases of the fundamental frequency. Aliased images occur at the sum and difference of the fundamental and multiples of the clock frequency. The lowpass filter helps to attenuate the amplitude of the aliased energy and smoothes the quantized steps. The filtered sinewave is then brought back on-chip and applied to the input of the internal comparator, which is configured as a +0.5 V threshold comparator (half of the DAC's fullscale output voltage). The comparator's +0.5 V reference level can be conveniently generated by averaging the complimentary DAC outputs across a resistor divider and filtering with a .01 µfd. capacitor (see Figure 4).

The comparator functions as a sinewave-to-squarewave converter and its squarewave output serves as a CMOS-compatible clock source, with a rise and fall times of 2.2 ns into a 15 pF load. The DAC outputs and comparator inputs are differential but they can be driven either in singleended or differential mode. As a clock source, the AD9850 can be digitallytuned to any frequency up to 42 MHz (approximately one-third of the 125



Figure 3. 4.5 MHz output viewed narrowband (1 kHz/div.).

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Figure 4. AD9850 in a clock generator application.

MHz clock rate), and it delivers an output frequency resolution of .029 Hz. Furthermore, the output clock can be phase-modulated in increments of $+180^{\circ}$, $+90^{\circ}$, $+45^{\circ}$, $+22.5^{\circ}$, $+11.25^{\circ}$, and any combination thereof.

Output switching characteristics and jitter are obviously important specifications for clock generator applications. The internal comparator of the AD9850 always sees a full-scale DAC output swing and it was designed with adequate hysterisis to insure clean output switching (no "chatter") with low frequency/low slewing input signals. Figure 5a shows the AD9850's comparator output switching characteristic with an input signal frequency of 1 kHz. Note the absence of output chatter as the input relatively slowly slews through the comparator's threshold level. Figure 5b shows the comparator's output switching characteristic with a 41 MHz input signal.

The parameters of the external filter play an important role in the amount of overall output clock jitter. Figure 6 shows the jitter plot for the output of the AD9850 under the following conditions: 40 MHz output pulse, 125 MHz reference clock, +3.3 V supply voltage, and a 5th order 45 MHz low-pass filter with 200 ohm input/output impedance. By storing consecutive samples of the output pulse on a storage oscilloscope, the output jitter is demonstrated to be approximately 300 psp-p which well supports its major application as a clock generator in a spread-spectrum digital communication systems.

Spread-Spectrum Application

An example of the AD9850 as a clock generator in a spread-spectrum receiver in shown in Figure 7. This is the

block diagram of the I&Q digitizer section of Loral Communications Systems' EB 200 spread-spectrum modem. The Loral system utilizes a DDS-based clock generator function to lock the receiver onto the chip rate of the transmitter. This allows the entire spread-spectrum system to quickly reoptimize its TX-RX data transfer delay and timing to compensate for fluctuating transmission media characteristics such as that presented under changing atmospheric conditions, or as posed by system operation within mobile/portable environment.

In a spread-spectrum receiver application, the EB-200 system accepts an RF input and generates I&Q IF signals via a quadrature down converter stage. A pair of high-speed A/D converters digitize the IF I&Q channels at the chip rate of the spread data. An on-board ASIC, Loral's PA-100 chip, performs the basic despreading and baseband demodulation of the I&Q data as well as decoding its chip rate. The PA-100 device generates a 32-bit frequency control word corresponding to the chip rate, which is applied as the frequency tuning word to the AD9850. The AD9850, in turn, generates an encode command pulse frequency for the dual A/D converters that is locked to the chip rate of the transmitted data. As transmission conditions change, the AD9850 clock generator function allows the receiver to be digitally tuned for continued optimum performance.

The features and attributes of the AD9850 complement and enable the full potential of the EB-200 spreadspectrum modem. The EB-200 is a highly flexible digital demodulator function that fully supports many different modulation schemes: BPSK,

62



Figure 5a. Comparator output waveform at 1 kHz.

QPSK, and OQPSK — at data rates from 9.6 k to 40 Msymbols/s. The AD9850, as the supporting clock generator function, allows the EB-200 to target many commercial spread-spectrum applications such as RFID, wireless telephony, wireless LAN, and high-speed data modems. For additional information on Loral's EB-200 spread-spectrum digital demodulator system, contact Kent Erickson at Loral Communications Systems, 640 N. 2200W, Salt Lake City, UT, 84116 (801/594-2044).

User-Friendly AD9850 Evaluation Board Available

An evaluation board for the AD9850 is available which carries the concept of simple DDS interface a step further. The evaluation board interfaces to the printer port of a PC and includes control software that operates under WindowsTM. This provides an extremely user-friendly "point-and-click" control format for operating the device. Two versions of the evaluation board are available: the AD9850/FSPCB and the AD9850/CGPCB. The FS version is configured as a frequency synthesizer with the AD9850's DAC output connected to a BNC connector and termiohms. The 50nated in AD9850/CGPCB evaluation board configures the AD9850 as a clock generator function and includes a 43 MHz low-pass filter between the DAC output and the comparator input. The AD9850 comparator's squarewave output is connected to an output BNC jack for timing analysis. Both versions are available from stock with a cost of \$98.50, which includes AD9850 device and operating software.

More DDS to Come...

The AD9850 is the flagship product





Figure 5b. Comparator output waveform at 41 MHz.

Figure 6. Output clock jitter plot.



Figure 7. AD9850 as a clock generator in a spread-spectrum receiver.

for a new family of DDS-based products from Analog Devices. In addition to forming the core of higher-resolution frequency synthesizer products, the AD9850's CMOS DDS and DAC blocks will be further integrated with mixers and digital filters to form complete digital modulator devices. One such immediate follow-on product is a QPSK digital modulator device that will provide an optimum low-cost solution for the 5-40 MHz return path TX function as required in the emerging HFC cable modem industry.

For additional information, contact Analog Devices at 1-800-ANALOGD or circle Info/Card #250. RF

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3. EB-200 Spread Spectrum Modem Preliminary Data Sheet, published by Loral Communications Systems, 1996.

About the Authors

David Crook is a Development Engineer in Analog Devices' High Speed Converter group located in Greensboro, N.C. He has been with the company for nine years and has most recently been involved with the development of BiCMOS A/D converters and high-speed CMOS DAC and DDS architectures.

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Noise Com introduces the WIS-98 and WIS-018 CDMA mobile Station Test Systems for cellular and PCS applications, respectively. The combined Wireless Impairment System Model WIS-98/018 is used for simultaneous cellular and PCS testing. Systems are integrated



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Expanding their HP 8648 line of signal generators, Hewlett-Packard announces the HP 8648D, a low-cost 4 GHz synthesized model designed for RF communications systems testing. New applications require higher frequency operation than was previously needed, and this model offers low cost in the 3 GHz to 4 GHz range, where available instruments have generally been high-end products.



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Advanced Control Components introduces the ACAM-7911, 7912 and 7914 RF power amplifiers designed as drivers in microcell base stations operating



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1-Watt RFIC Power Amplifier

Pacific Monolithics introduces the PM2105 1-watt power amplifier for operation over the 800-2000 MHz frequency range. With extrnal matching, high performance can be obtinaed within 30 to 100 MHz bandwidth anywhere in the specified frequency range. The PM2105 is designed for operation in either linear or saturated power applications. P_{1dB} is 29 dBm, while saturated power is 30.5 dBm (typical). Bias pins either Class AB or Class C operation. The device ratings are specified with a 5 VDC supply, but the device is usable over 3 to 6 VDC. **Pacific Monolithics** INFO/CARD #193

SSTs Target HF and CATV Return Uses

New silicon FET amplifiers from MicroWave Technology are designed for low/medium power

applications at frequencies from 0.2 to 100 MHz, including HF communications, test instrumentation, and CATV reverse channels. The LPS-0151 covers 1-50 MHz with more than 0.5 watt of linear power output in a 75 ohm system. The LPS-0150 is a 1-watt, 50 ohm device for 1-50 MHz, and the LPS-0110 is a broadband 0.2-110 MHz device providing 0.5 watts into 50 ohms. All models are internally matched and biased, requiring only power and decoupling circuitry, and are specified for operation at 12.5 VDC. MicroWave Technology INFO/CARD #194

3-Watt GSM Amplifier

Anadigics has introduced the AWT0904 GaAs monolithic power amplifier, delivering 3 watts from a single +5.8 VDC supply. Provided in a low-cost IC package, the device uses three amplification stages to achieve 34 dB gain, 35 dBm output, 50% efficiency, and low receive-band noise. Similar amplifiers are being developed for the U.S. PCS band. Anadigics Inc. INFO/CARD #195

IC Combines Mixer and I/Q Modulator

TEMIC has released the U2891B, providing a 2.5 GHz mixer and 30-500 MHz I/Q modulator in a single package operating at a supply as low as 2.7 volts. Accuracy of the I/Q phase shifter is 0.5°, providing sideband suppression of 45 dB and LO suppression of 40 dB. The mixer section has an output compression point of -7 dBm and a third order intercept of -6dBm with a 50 Ω load. Conversion gain is 9 dB. **TEMIC Telefunken INFO/CARD #196**

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TEST EQUIPMENT

Enhanced Model RF Voltmeter

Boonton Electronics has improved the Model 9200C RF voltmeter with a new front panel display. The 9200C offers complete microprocessor control, a voltage range of 200 μ V to 3V, from 10 Hz to 2.5 GHz. Features include auto-zero function, an additional channel for differential measurements, DC recorder output, IEE-488 interface and a 3-1/2 digit voltage or dB display with 0.01 dB resolution. Starting price of the 9200C is \$3,600.

Boonton Electronics Corp. INFO/CARD #198

Cellular Base Station Emulator

The TAS 6600 Wireless Communications Analyzer from Telecom Analysis Systems offers complete cellular base station emulation, including: a mainframe analyzer with



EAMPS emulation, the TAS-6600-CAP Cellular Audio Processor (CAP) module, and the TASKIT[®]/6600 software for Windows. The compact unit simplifies, integrates and enhances the testing of cellular communications products.

Telecom Analysis Systems, Inc. INFO/CARD #199

PC-Based Arbitrary Waveform Generator

Gage Applied Sciences has introduced the CompuGen 1100, a PC-based arbitrary waveform generator with 12-bit resolution DAC and conversion rates up to 80 MSPS. The ISA bus card instrument is provided with control software that allows userdefined mathematical functions or waveforms from the software library. Output signals with up to 20 MHz bandwidth can be generated. With standard 512k memory (expandable to 8 million samples), the price is \$4,995. Software drivers for popular compilers are \$250.

Gage Applied Sciences INFO/CARD #200

10 MHz Data Acquisition Board for Workstations

Ultraview Corporation has developed an SBus data acquisition board that can continuously acquire analog data at 10 MSPS and 12-bit precision, including baseband and low-frequency RF signals. Real-time analysis of scientific, medical and communications signals is the primary application. A special pipeline mode allows the

entire 16 MBytes of RAM to be used for outputting 16 MBytes of analog data, while siultaneously inputting 16 MBytes of data. This allows immediate measurement of response to stimulus, useful for testing high-speed RF, disk drives, and MMICs. Single-unit price is \$4495 for the combined A/D and dual D/A board. Other models range from \$2995 to \$3995. **Ultraview** Corporation INFO/CARD #201



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Sawtek now offers a line of standard 70 MHz SAW filters with insertion loss that's substantially lower than any 70 MHz SAW filters we've ever built. On average, our new filters offer 50% lower insertion loss than any of our existing 70 MHz filters, making them more appropriate for your low power wireless applications.

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Part	BW3	Loss
Number	(MHz min.)	(dB max
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354652	1.0	TBD
354653	1.5	TBD
354654	2.0	TBD
854655	2.5	TBD
854656	3.0	TBD
854657	3.5	TBD
854658	4.0	7.5
854659	4.5	8.0
854660	5.0	8.5
854661	6.0	9.0
854662	7.0	9.5
854663	8.0	10.0
854664	9.0	10.5
854665	10.0	11.0
854666	12.0	12.5
854667	14.0	13.0
854668	16.0	13.5
854669	18.0	14.5
854670	20.0	15.0
854671	22.0	16.0
854672	24.0	16.5
854673	26.0	17.5
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RF products continued

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devices are less prone to leaching during soldering than silver technology components. A complete line of attenuators, resistors and terminations is available using this technology. **Bird Component Products** INFO/CARD #202

Chip Lowpass Filters for 800-2600 MHz

The LTF Series of chip lowpass filters from Toko America can be used to reduce harmonics of a system PA or LO. Provided in a miniature 1206 footprint $(3.2 \times 1.6 \text{ mm})$ and only 1.4 mm high, these filters provide a minimu of 30 dB attenuation at $2 \times f_0$ and 18 dB at $3 \times f_0$. Standard frequencies are available for standard wireless communications band from 836.5 MHz through 2450 MHz, including cellular, 900 and 2400 MHz ISM bands, PDC-1.5, PCS-1900, DCS-1800, DECT and PHS. Toko America, Inc. INFO/CARD #203

SAW Filters for Wireless Systems

A new family of SAW filters for wireless applications is now available from Mitsubishi Electronics. The MF1009S is designed for the 869-894 MHz AMPS cellular receive band, and the MF1010S series is designed for the 824-849 MHz AMPS transmit band. The new SAW filter has a small footprint of $3.8 \times 3.8 \times 1.25$ mm, and are offered optimized for either insertion loss (3.1 to 3.5 dB) or for stopband rejection (24-55 dB). Low amplitude ripple and wide temperature range operation are additional features of the SAW filter line. **Mitsubishi Electronics** INFO/CARD #204

Dual Celluar Bandpass Filter

K&L Microwave offers a dual cellular bandpass filter, model X10FV50-920/X80-N/N with center frequencies of 897.5 MHz and 942.5 MHz. -3 dB bandwidth at each frequency is 36 MHz, with insertion loss of less than 1.5 dB. Rejection at 860 and 980 MHz is greater than 40 dB, and greater than 70 dB at 920 MHz.

K&L Microwave INFO/CARD #205



Chip Inductors

Pulse offers a new series of 1008 wirewound RF chip inductors, with a ceramic core that provides exceptional Q performance, thermal stability and self-resonant frequencies reaching up to 6 GHz. The inductors are available in industry-standard values from 4.7 nH to 4700 nH, and are also available in the 0805 size. Both are designed for pick-and-place assembly. Q and inductance and S-parameter information are available in disk format for easier circuit modeling. In 100,000-piece OEM quantities, pricing is typically \$0.13 each. Pulse

INFO/CARD #206

1 kV Ceramic-Plate Capacitors

Philips Components has added a new series of 1 kV ceramic-plate capacitors in a wide range of values from 0.47 pF to 3300 pF, available in tolerances from 5% to 20%. High temperature versions are available, as are low-loss models for minimum dissipation.

Philips Components INFO/CARD #207

Product Focus — **Crystal Oscillators**

Vacuum OCXO Increases Performance

A new generation of OCXO technology, the Evacuated Miniature Crystal Oscillator (EMXO) uses the insulating properties of vacuum, with a cold weld enclosure to provide improved performance over conventional OCXOs. Steadystate power consumption is <1.2 watts at -55°C, with warm-up time at that temperature of two minutes. Stability is available to $\pm 1 \times 10^{-8}$ over -55 to +85°C. Vectron Labs

INFO/CARD #208

High-Stability SATCOM TCXO

Model 6202 from Oscillatek is based on the poular 6024 and 6079 SATCOM TCXO series. The 6202 is a 10 MHz model that will maintain ± 1 ppm 0 to +50°C including aging for 10 years. Pricing is \$65.00 each at 1000 pieces, and other performance options are available. Oscillatek INFO/CARD #209

Microprocessor Crystals

Ecliptek now offers the ECCM3 and ECCM4 µP crystals, with high stability needed for use in clock and reference oscillators for wireless, satellite and telecommunications. They are available in frequencies from 11 to 120 MHz, with ±10 ppm tolerance and ±5 ppm stability over -20 to +80°C. The ECCM3 and ECCM4 are provided in SMT packages designed for automatic assembly. **Ecliptek Corporation** INFO/CARD #211

SUBSYSTEMS

Distribution Amplifier

Frequency Electronics has introduced the FE-7923A 10channel sinewave distribution

Programmable **Master Clock** Oscillator

Micro Networks introduces a parallel-programmed master clock oscillator for frequencies from 40 MHz to 1 GHz. The M115 Series has



eight models that can be programmed over a wide range. For example, the M115-1 covers 40-80 MHz in 100 kHz increments, while the M115-8 covers 800 MHz to 1 GHz in 200 kHz steps. Stability is 35 ppm (0-70°C), with other specifications available. Prices start at less than \$155 in sample quantities, OEM pricing less than \$100. **Micro Networks** INFO/CARD #210

amplifier, operating at 1 to 10 MHz. Designed to meet stringent satellite ground station requirements, the unit offers low phase noise, low amplitude noise bursts, and isolation of more than 100 dB. The FE-7923A is provided in a 19-inch rack mount case. **Frequency Electronics**

INFO/CARD #212

Wireless Data Transceiver/Modem

Advanced Digital Systems (ADS) and Tait Electronics have jointly developed a rugged, costeffective mobile data package that marries a data-ready Tait mobile transceiver with a highspeed ADS "Smart Packet Modem." The D2000/SPM modem system transmits and receives data from 1200 to 9600 bps. The D2000 mobile transceiver is a synthesized fourchannel radio, remotely con-







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Product Focus — Crystal Oscillators continued

Surface Mount VCXOs for Communications

Champion Technologies introduces the K1526 Series of voltage-controlled crystal oscillators (VCXOs), available in a compact ceramic SMT package. The VCXOs are compatible with today's phase-locked loop applications in wireless telecommunications. Available in frequencies from 2.0 to 55.0 MHz, the VCXOs offer overall frequency stability tolerance as tight as ± 25 ppm over 0 to 70°C, and ± 50 ppm from -40°C to +85°C. Deviation range is available from 60 ppm to 150 ppm with a control range of 0.5 to 4.5 V. Typical pricing is \$19.000 each in 1,000-piece quantity. **Champion Technologies, Inc. INFO/CARD #213**

OCXO Series Developed from Satellite Applications

Model 106 from Reeves-Hoffman is an oven controlled crystal oscillator (OCXO) that was originally designed for high reliability Satellite Rescue (SARSAT) applications, with low power consumption comparable to a TCXO. Temperature stabilities are as tight as ± 0.05 ppm over 0-70°C, and ± 0.1 ppm over -40 to ± 70 °C. Power consumption is approximately 200 mW at



-40°C, ideal for battery-power applications. The Model 106 is packaged in a standard CO-08 package, and aging rates as low as 0.3 ppm per year are avilable. **Reeves-Hoffman INFO/CARD #214**

SMD Clock Oscillator

Vectron Technologies offers a fixed frequency crystal oscillator in a leadless chip carrier (LCC) package, for TLL and CMOS applications. Tape and reel as well as tube packaging is available. Typical applications include providing clock signals for digital signal processing (DSP), microprocessors (μ P) and disk drive circuitry.

Vectron Technologies Inc. INFO/CARD #215

Super-Low-Profile OCXO

Oak Frequency Control Group's new 4853 OCXO is available for the 10-50 MHz range, and features a low-profile package with dimensions of $1.0 \times 1.0 \times$ 0.45 inches. Temperature stability is $\pm 8 \times$ 10^{-8} over -20 to +70°C. Warm-up power is 2.5 watts, and steady-state power consumption at +25°C is 1 watt. Typical phase noise at 100 MHz is -130 dBc/Hz. An SC-cut crystal version is available. Oak Frequency Control Group INFO/CARD #216

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POS-200	100-200	-102	-24	18	11.95
POS-300	150-280	-100	-30	18	13.95
POS-400	200-380	-98	-28	20	13.95
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trolled by the modem. VHF and UHF models are offered for 136-174 and 400-470 or 450-520 MHz, respectively. Tait Electronics — USA, Inc. INFO/CARD #217

Radio Data Transceiver

The Pathfinder Explorer 9600 from Metric Systems Corp. is an intelligent UHF 9600 bps radio data transceiver for industrial control, monitoring and data acquisition systems. The unit offers 3-wire or RS-232 interface, packetized or transparent mode, individual station addressing (up to 64,000 addresses), full error correction, and the CSMA point-to-multipoint networking protocol. The Pathfinder Explorer 9600 is available in seven frequency bands from 403 to 512 MHz.

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Microwave Microcells

3dbm has announced the availability of the MMC series of Microwave Microcells. The units are designed to improve the capacity and coverage of cellular systems. Typical uses for coverage imporvement include shopping malls, convention centers, office buildings, tunnels, subways and indoor garages. They can be combined with 3dbm's SZS Microcell Zone Selectors to form intelligent microcell clusters. **3dbm, Inc.**

INFO/CARD #219

CABLES & CONNECTORS

Flexible Microwave Cable Assemblies

Low-cost microwave cable assemblies for applications up to 14 GHz are available from Richardson Electronics. Made by Gore, the assemblies are offered in 4-inch to 60-inch lengths, with straight or rightangle SMA male connectors. The cable features low loss RF characteristics and performance similar to semi-rigid cable, but in a more flexible alternative.

Richardson Electronics Ltd. INFO/CARD #220

Minimum-Profile F Type Jack

Connect-Tech Products now offers the CTP-169 minimum-profile F-type jack which mounts directly to the edge of a printed circuit board, supported by a slot in the jack. The edge slot can be changed from the standard .062 p.c. board thickness to any other required dimension. Applications include cable TV, LAN networks, TV interface, satellite and other communications. Unit cost of the CTP-169 in high volume is under \$0.35 each. Connect-Tech Products

Connect-Tech Products INFO/CARD #221

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RF tutorial

Active Filter and Other Op Amp Circuits for RF Applications

By Gary A. Breed Editor

Active filters and other high-frequency and high-speed op amp circuits have only recently become practical for RF applications. The development of the current-feedback operational amplifier in the 1980s, and subsequent improvements in the design and fabrication of all types of op amps and buffers have combined to make active circuits viable replacements for some passive R-L-C circuits.

Aseries of example circuits are presented in this month's tutorial, taken from past articles published in *RF Design*. These circuits are good representatives of the types of active circuits that are practical for *RF* applications.

An Active Bandpass Filter

The first circuit is an active bandpass filter using a wideband currentfeedback op amp, the Comlinear CLC400 [1]. A 1987-era product, the CLC400 has a -3 dB bandwidth of 200 MHz, and is designed for stability in low gain applications ($A_y = \pm 1$ to 8). The topology of the amplifier in Figure 1 was chosen to avoid placing reactive components in the feedback path, thus maintaining stability. While not a high-Q topology, the design provides predictable "peaked" response that is useful for reducing broadband noise,



Figure 1. A practical 40 MHz active bandpass filter circuit with no active elements in the feedback path.

performing a spectral "clean-up" following an oscillator or amplifier, or providing a tuned response to detectors or other test circuitry.

When designing this type of filter, the op amp cannot be assumed to be ideal, unless the filter center frequency is well below the -3 dB bandwidth of the amplifier. In the case of the CLC400, the time delay of 1.6 ns through the device must be included in design calculations.

25 MHz Tunable Active Filter

Figure 2 is the functional block diagram of the Krohn-Hite Model 35 tunable active filter [2]. This product was state-of-the-art in 1991, offering 4-pole Butterworth lowpass response, tunable from 170 Hz to 25.6 MHz. This type of filter is useful for EMI testing, Part 68 telecommunications testing, prefiltering for various instruments, ultrasonics and high-speed tape or disk drive testing.



Figure 2. Functional block diagram of a commercial tunable active filter with 25 MHz bandwidth (Krohn-Hite Model 35).

74



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Figure 3. Wideband voltage-controlled amplifier (VCA) circuit.



Figure 4. A high-speed precision rectifier

Wideband VCA and Rectifier

Mathews and Reimer described a series of high-speed op amp circuits in their 1989 article [3]. One of those circuits is shown in Figure 3, a wideband voltage-controlled amplifier (VCA) using a high-speed analog multiplier in the feedback loop. The multiplier acts as an attenuator, controlling the gain without having the control element in the direct path. A gain control range of 60 dB can be achieved.

The other circuit (Figure 4) is a precision rectifier, which can be used in a fast AGC detector or other high-speed amplitude monitoring application. The 0.1 dB bandwidth of this rectifier circuit using the HFA 0002 is 2.3 MHz, with about 40 ns response time from input AC to output DC.

Circulator/Isolator Circuit

The 1991 winner of the RF Design Awards Contest [4] submitted the circuit of Figure 5, which uses the phase accuracy of high-speed op amps to make a three-stage "circulator" for low frequencies. The circuit provides about



100

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1	INA-51063	Gain block	DC-2400	5	12	3.0	20.5	+6
	INA-52063	Gain block	DC-1600	5	30	4.0	22	+15
	MGA-81563	Driver anip	100-6000	3	42	2.7	12	+27
	MGA-82563	Driver amp	100-6000	3	84	2.2	13	+31
	MGA-86563	LNA	1500-6000	5	14	1.6	22	+15
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40 dB isolation between ports using low-inductance precision resistors. This broadband performance level cannot be achieved by passive circuit.

The author's initial application of this circuit was to measure output return loss of an oscillator while in operation, which requires isolation of the circuit output from the measuring instrumentation. Low-level amplifier, mixer or multiplier testing are other useful applications for this isolator. RF

References

1. Scott Evans, David Potson, "High Speed Monolithic Op Amps: New RF Building Blocks," RF Design, January 1988, p. 27.

2. Bill Kulas, "Tunable Active Filter Reaches 25 MHz," RF Design, June 1991, p. 41.

3. Brian Mathews, Dave Reimer, "High Frequency Op Amp Makes Precise Detector and AGC," RF Design, November 1989, p. 76.

4. Charles Wenzel, "Low Frequency Circulator/Isolator Uses No Ferrite or Magnet," RF Design, July 1991, P. 39.



Figure 5. An active circulator/isolator for "low" frequencies (<100 MHz).



79

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New Products — page 91

EVALUATEST & DESIGN

FEATURE

Grounding and Radiation Reduction in PCBs

Radiation from printed circuit boards can be reduced significantly is proper grounding techniques are used. Of particular importance are the boundaries between analog and digital portions of the circuit. This article offers solutions for grounding that minimize currents that can cause excessive radiation.

- David A. Weston

86

DEPARTMENTS

Calendar	82
News	84
Info/Cards	126*
New Products	91

Product Showcase	93
EMC Mart	94
Company Index	122*
Advertiser Index	122*

*Info/Cards, Company Index, and Advertiser Index can be found on the designated pages in the issue of RF Design which contains this EMC Test & Design supplement.

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News

NIST Helps Power Grids Make 'Lightning Saves'

A NIST Technique has been incorporated into the latest version of an electrical standard used to test the ability of electric power grid high-voltage equipment to survive lightning strikes. Now a part of IEEE Standard 4-1995, "Standard Techniques for High-Voltage Testing," the technique improves the measurement of specific high-voltage impulse waveforms that simulate lightning. Such test waveforms subject the equipment to up to 10 times its normal operating voltage. To certify that power grid equipment is lightning-proof, engineers must evaluate the strength of the waveform. They use a "high-voltage divider" device to scale the voltage down to levels that can be measured. However, the process can produce a distorted reading. The NIST technique improves the measurement process by enabling testers to mathematically estimate the distortion. For more information, contact Gerald J. FitzPatrick, B344 Metrology Bldg., NIST, Gaithersburg, MD 20899-0001. Tel: (301) 975-2737; E-mail: fitzpa@eeel.nist.gov.

New European Automotive EMC Directive

The Automotive Directive 72/245/EEC as changed by 95/54/EC of October 31,1995, is about EMC Type Approval of vehicles and automotive electronic components including after-market accessories. The Automotive Directive is a Single Directive per 89/336/EEC and applies beginning January 1,1996. The transition period is through October 1,2002. Requests EEC Type Approvals for vehicles, units or components are now subject to the new Directive of January 1. The labeling requirement is "e plus Country Code etc." for "CE." Electric powered vehicles, fork-lifts etc. are exempt and remain under the EMC Directive 89/336/EEC. Dual-Voltage equipment is subject to both directives resulting in "CE" plus "e" marking. EMCC Dr. Rasek, located at Moggast, D-91320 Ebermannstadt, Germany, was nominated as the first "Competent Body" for the full scope of EMC and was accredited as the first EMC test laboratory covering the new Directive requirements. Tel: 49 9194 9016; Fax: 49 9194 8125.

Power Quality Fellowship at Texas A & M University

Current Technology, Inc., the leading manufacturer of electrical transient suppression filter systems and other power conditioning products, has announced that the company will provide Texas A&M University with a \$7,500 Power Quality Fellowship for the 1996-1997 academic year. The funding will support graduate student research in the university's Electrical Engineering Power Quality Laboratory, which is presently under construction on the Texas A&M main campus in College Station.

New North American Headquarters

Technology International Incorporated announces the formal opening of their new North American Headquarters. Their new address is: 609 Twin Ridge Lane, Richmond, VA 23235. Tel: (804) 560-5334.

New Magnetics Technology Center

Walt Benecki, President of The Arnold Engineering Company in Marengo, Illinois, has announced the completion of their new Magnetics Technology Center (MTC). This state-of-the-art, 16,800 sq. ft. magnetics development laboratory will be used for new product and improved process development by Arnold Engineering and their customers. The MTC was developed to assist Arnold's customers in minimizing their product-cycle time and costs. It will house a multitude of process and analytical equipment making it the most complete magnetics laboratory in the united States.

Berndt Associates Named New Representative

Lindgren RF Enclosures, Inc. has named Berndt Associates, Inc. of Arlington Heights, Illinois, as its representative managing the sales of Lindgren EMI/RFI shielding systems for the Midwest. These states include Illinois, Iowa, Indiana, Minnesota, North Dakota, South Dakota and Wisconsin. Berndt Associates will represent Lindgren's line of industrial shielding systems including double electrically isolated (DEI) Screen and solid rooms, cell type enclosures, welded enclosures, and anechoic chambers for EMC testing.

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INFO/CARD 74

Grounding and Radiation Reduction from PCBs

By David A. Weston EMC Consulting Inc.

David A. Weston EMC operates Consulting Inc. Canada. For the last fifteen years he has specialized in all aspects of EMC including analysis. EMI problem solving and PCB layout. He has designed commercially available output power line filters for switching power supplies which can make almost any switcher as quiet as a linear supply. Mr. Weston can be reached at EMC Consulting Inc, P.O. Box 496, Merrickville, Ontario, Canada, K0G 1NO. Tel: (613) 269-4247; Fax: (613) 269-2045.

The importance of grounding in reducing radiated emissions from PCBs has been demonstrated in practical terms in a number of EMI investigations. This article describes an experiment which was conducted to prove the point and also to quantify the effect in one specific example of poor grounding. The article also describes two PCB grounding EMI case studies.

The PCB grounding experiment was tried during measurements made to characterize the radiation from thirteen different PCB layout configurations using six logic types. The results of these measurements were published in Reference 1.

The thirteen different layouts included:

- Transmission line (signal and signal return traces located either side by side on one surface of the PCB or one trace on the upper surface and one trace on the lower surface of the PCB)
- Microstrip (signal trace located above a ground plane which is used to carry the signal return current)
- Image plane (a transmission line, i.e. signal and return traces, on one surface of the PCB with a conductive plane on the other side connected to the signal return at one location only)
- Interleaved image plane (a microstrip configuration with image plane interleaved between the signal trace and a lower ground plane)
- Partial stripline (a microstrip with a narrow upper ground plane connected to the lower full ground plane using vias either side, and down the length, of the signal trace. The radiation was measured with the distance between the vias at 3.3 mm, 6.6 mm and 13.2 mm.
- Large stripline (the same basic configuration as the partial stripline but with the upper ground plane extending almost to the edge of the lower ground plane)
- Full stripline (the same basic configuration as the partial stripline but with the upper ground plane extending over the same area as the lower ground plane)
- Full stripline with additional vias (The same as the full stripline but, in addition to the rows of vias either side of the stripline, with vias, connecting lower and upper ground planes together, at 13.2 mm intervals around the edge of the PCB)



Figure 1. Connection of the semi-rigid line, used with the ECLinPS device, to the microstrip ground plane.

• Microstrip with a right angle bend constructed from either one 90° bend or two 45° bends

As expected the experiments showed that the major source of radiation from the PCBs was common mode current flow with a large, 56 dB, difference between a high emission level type PCB and a low emission level type PCB. In addition to the different layouts the investigation included a look at circuit level reduction techniques such as RC and RLC networks at the signal source end of the trace to slow down clock edges and reduce the high frequency radiation from the PCB.

The signal source used to drive the PCB trace, terminated in a passive RC load, comprised a battery, regulator, clock oscillator and driver IC contained in a shielded enclosure. The shielded enclosure ensured that only the radiation from the PCB was measured and not that from the clock and driver IC. Two versions of the enclosure were built. One contained the 74XX04 type devices and the second was used to house the ECLinPS device. The difference between the two enclosures was in the method used to make the signal and signal return connections to the PCB. The ECLinPS device is designed for a 50 ohm load and should be connected to the load by the use of a structure with a 50 ohm characteristic impedance.

The ECLinPS device was tested only with the microstrip PCB layout and a piece of 50 ohm semi rigid cable was used to connect the output of the driver to the PCB under test.

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INFO/CARD 75

Radiation Reduction

The shield of the semi rigid was soldered to the enclosure at the location where the line penetrated the enclosure. At the PCB end the semi rigid was soldered to a copper plate and this plate was soldered to the ground plane of the microstrip PCB as shown in Figure 2. Although the ECLinPS device was only tested with the microstrip layout so were the remaining seven different logic types and a comparison between the logic types is possible. A color plot was made of the relative emission levels using the eight logic types with the microstrip PCB. Plots were also made of the thirteen different PCB layouts with one baseline type of 74XX04 device. Thus a comparison can be made between the emission profile of the baseline type of device with the microstrip and any other logic type. This comparison can then be used to correct the emission profile of any PCB layout for a different logic type, including the ECLinPS type.

In the construction of the shielded enclosure used to contain the 74XX04 devices the enclosure must

be connected to the ground planes or the image plane in the microstrip, stripline or image plane PCB layouts. Whereas in the transmission line PCB layout the signal return is isolated from the enclosure. Figure 2 illustrates the connection of the enclosure to the signal wire from the source and the ground/image plane. The signal return connection is made from the PCB plane to the feedthrough in the enclosure via a short length of wire, as is the signal connection to the signal trace. To minimize radiation from these short lengths of wire a small copper dog house enclosure covers both the feedthrough connections and the lengths of wire. This dog house extension can be seen in Figure 2 and is used to connect the enclosure to the ground/image plane of the PCB. In the connection of Figure 2 none of the signal return current flows in the connection between the enclosure and the PCB ground plane.

The grounding experiment involved connecting the enclosure to the PCB via a short length of wire which was connected to the feedthrough and then via the signal return wire to the PCB. This grounding technique is illustrated in Figure 3 and it can be seen that the signal return current does flow on the connection between the enclosure and the PCB ground plane.

The radiated emissions from the different types of PCB layout were measured with this alteration in grounding. The level of radiation from the low emission type PCBs increased from 56 dB below high emitter types to 6 dB below these types.

The reason for this dramatic increase is that an RF potential difference is developed between the enclosure and the ground plane/s or image plane. This potential difference is caused by the signal return current flow on the short (5 mm) length of wire used to connect the PCB to the enclosure. Although the wire is short it exhibits significant inductance and it is across this inductance that the voltage is developed.

Whereas in the grounding connection shown in Figure 2 the enclosure to ground plane connection is low impedance, it does not carry the signal



Are you using cables to interconnect your electronic equipment? These cables often behave like antennas by coupling ambient electromagnetic noise to and from your system. With new IEC regulations covering radiated and conducted immunity, it's more important than ever for designers to have the tools to stop EMI without resorting to costly, time consuming product revisions.

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Figure 2. Connection of the source to a microstrip, stripline or image plane PCB layout.



Figure 4. PCB with digital and RF ground planes connected together with a thin trace.



Figure 3. The signal return is used as the connection between the PCB ground plane and the enclosure.



Figure 5. The use of semi-rigid cable to connect the ground planes together.

return current and so no RF potential is developed.

Why is the same problem not seen with the semi rigid cable connection shown in Figure 1? The reason no potential difference is developed on the outside surface of the semi rigid shield is due to both the mutual inductance between the center conductor and shield and due to the skin depth effect. The skin depth ensures that above some relatively low frequency the majority of the return current flows on one surface of the shield and the mutual inductance ensures that this is the inside surface of the shield. This effect is illustrated in Figure 5.

The problem of RF voltages developed between PCB grounds and equipment grounds or the enclosure has been seen in a number of practical EMI cases and two of these are presented as follows:

The first case study involved radiation from a PCB, which contained digital and RF circuits, to a nearby PCB which contained sensitive RF circuits. The source PCB generated emissions which were, at certain frequencies, 40 dB above the FCC class "A" limits!

The board layout was however exemplary, it contained all the "good" features such as effective ground planes and stripline configurations for signals which were potential sources of emissions. The problem was due to a thin trace used to connect the RF and digital grounds together at a single point. The thin trace carried the return current of a digital signal driving a device on the RF board. The two PCB ground planes and single point connection are shown in Figure 4.

The return current generated a voltage drop in the inductance of the return trace and so an RF potential was developed between the ground planes. This is the same problem as seen in the grounding experiment, in which the potential was developed between the ground plane and the enclosure. Many modern PCB antenna designs use a similar layout, driving the antenna at a similar point, and achieve a high gain, that is efficient radiation.

A significant reduction in radiation was seen simply by widening the thin signal return trace. Widening the trace reduces the inductance and the voltage drop in the return path. The majority of the signal return current flows in the return path directly beneath the signal trace. This means that when the width of the return trace is five times or more the height of the signal trace above the return trace that an effective ground plane has been achieved, for that single signal connection. This concentration of current under the signal trace is again due to mutual inductance. The reason the digital ground plane is connected to the RF ground plane at a single point is to reduce digital noise current flow in grounds used by sensitive RF circuits. However the width of this single connection can be increased when sensitive circuits are located far away from the connection or when digital signal return current is routed away from the location of the single point ground. For example ensure that digital return current does not flow between devices located across the increased width of the ground plane connection.

An even more effective solution is the use of a semi rigid cable connection shown in Figure 5 or a stripline PCB structure installed across the gap. The stripline is not as effective as the semi rigid due to some leakage through the gaps between the vias used to connect the upper and lower ground planes together. Another potential solution is to disconnect the two ground planes and to use an opto-isolator to transfer the digital signal to the RF section. Separating the two ground planes is very effective in reducing current flow between them. However if an RF potential exists between the planes, as it almost certainly does. then the level of radiation may remain high or even increase with separated grounds. The digital return current is forced onto a trace, which is disconnected from the RF ground, and the digital and RF grounds are tied together at a single point. Negligible current flows in this common connection and the two ground

Radiation Reduction

planes are close together in potential.

Another common example of RF potential developed between PCB ground planes is when signal and signal return connections exist between two or more PCBs. These connections are typically made via pins in edge connectors and a mother board or back plane. It is important that as many parallel signal return connections as possible are made. Also that these connections be as short and as wide as possible to reduce the inductance of the signal return, and therefore the potential difference between grounds.

If external cables are connected to the two PCBs then the RF potential between the two ground planes appears on the signal connections as a common mode voltage, regardless of whether the signal is balanced or unbalanced with reference to ground.



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The two cables result in a very efficient antenna as the RF potential between them drives one cable against the other, similar to the two rods of a dipole antenna. In this example the use of finger stock or a similar contact to connect the two ground planes together will reduce the RF potential between them. Ideally this common connection should be made at the location where the two cables leave the enclosure, or at least where they leave the PCB. To reduce radiation from the cables due to a potential difference between the PCB ground plane and the enclosure, finger stock or a direct connection may be made between the ground and enclosure. If a DC connection is not allowed, an RF (capacitive) connection should be made.

Another practical example of EMI was partially due to an RF potential between a ground plane on an approximately $8 \text{ cm} \times 4 \text{ cm}$ digital PCB and an RF ground. The digital PCB was located approximately 0.5 cm away from a receiver contained in a small well shielded case. The small microstrip antenna connected to the receiver was approximately 2 cm from the digital board and was referenced to the receiver case. Due to radiated coupling between the digital board and the antenna the receiver exhibited desensitization and narrowband and broadband spurious response. Part of the problem was due to the RF potential between the digital ground plane and the RF ground (shielded case). By shorting the case to the digital ground plane at a number of points with finger stock material the radiated coupling reduced significantly, although not sufficiently to totally cure the problem.

Conclusion

In conclusion, the use of low emission type PCB layouts in conjunction with ICs which generate only low level emissions can reduce radiation to levels where shielding of the PCBs is not necessary to meet commercial radiated emission limits. However when PCB-to-PCB or PCB-to-enclosure/ground connections are high impedance and carry RF current then none of these reduction techniques may be sufficient without the use of an effective grounding scheme.

Reference

1. D.A. Weston, "PCB radiation due to high speed logic and emission reduction techniques," EMC Consulting, 1992.

ENCLOSURES, INC.

New Products

Metallized Fabric Shielding Tape

3M now has a flexible metallized fabric shielding tape, expanding their line of shileding products. 3M 1190 tape combines copper-plated



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Low-Cost EMI Test System

Dynamic Sciences anounces the DSI-2020 EMI Test System, optimized for FCC-EN precompliance and certification testing over a 1 kHz to 1.9 GHz frequency range. The DSI-2020 consists of a single receiver and a power EMI software package that runs on the user's GPIBequipped computer. Built-in spectral and oscilloscope displays eliminate the need for external equipment. Price of the system is \$29,950. Dynamic Sciences International INFO/CARD #232

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Rohde & Schwarz is now offering EMI software ESxS-K1, designed for reilaible, fast and reproducible EMC measurements to be performed at small- and medium-sized companies. The software supports the Rohde & Schwarz test receivers of the ESHS/ESVS/ESS family, as well as the ESVD/ESVB/ESN/ESVN instruments for European standards. Rohde & Schwarz GmbH & Co. INFO/CARD #230

EMC Shielding Fingers

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EMC/ESD Literature

Suppressor Product Data on Disk

Semtech Corp. announces the TVS Access Disk, an alternative to thumbing through catalogs and data sheets. The disk allows customers to locate the right transient voltage suppressor for their applications. The disk includes a tutorial on suppression of transient voltages, standards, a selection guide, and ordering information.

Semtech Corporation INFO/CARD #223

INFO/CARD #223

Silicone Rubber Packaging Catalog

A new catalog from Fujipoly describes four silicone rubber product lines: RFI/EMI shielding and ESD grounding; ZEBRA elastomeric components; SARCON thermal management components; and custom moldings and extrusions. Technical background and mechanical details of each product are included.

Fujipoly America Corp. INFO/CARD #224

Data Sheets on Shielded Doors

Lindgren RF Enclosures is offering a series of technical data sheets on its high performance EMI/RFI shielded doors. Among the offerings are the Doubler Knife Edge (DKE) Door, a manually operated door with performance superior to single knife edge doors, the Pneumatic Sliding Door (PSD) and Pneumatic Hinged Door (PHD).

Lindgren RF Enclosures, Inc. INFO/CARD #225

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New Products

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RF synthesizers

Structured Design of Third-Order, Type-2 and Type-3 PLLs

By Fu-Nian Ku

PLLs are ubiquitous, and many books and papers have dealt with the subject. In this paper, I present PLL characteristics in a graphical format so that people can readily see the characteristics in whole and get approximate results. The complicated mathematical formulae are in the Appendix, where they can be examined or used in the reader's choice of computer program.

Comprehensive study of PLLs is a Very complicated task. There are different types of phase detectors, different requirements of lock-range and lock-time, different added-on filters. In addition, the real performance of devices must be considered, as must the noise rejection of PLL. Secondorder PLLs have been thoroughly studied in past two decades [2], but there is still much to learn about the behavior of third-order and higher PLLs. Przedpelski's papers [3] have won wide acceptance in the last two decades, and they have made highorder PLL characteristics calculable.

This paper is an introduction. It does not touch non-linear phenomena, and does not try to predict capturetime, capture-range or other quantities. However, it does give a basic relation among all physical-interpretationoriented parameters. I have simplified these complex formulae and extracted as few as possible parameters to show the physical meaning for easy design.

Type 2 PLLs (Active Filter)

I use the conventional symbols and definitions. Let $\theta_e(t)$ be the phase difference of two input of phase detector, $\theta_e(t) = \theta_i(t) - \theta_o(t)/N$ Assume this is a linear system; the output of phase detector is $K_d \theta_e(t)$. If the initial $\theta_e(0) = 0$. The Laplace transform of this loop results in the following equation 1(a):





Figure 1. (a) Loop filter circuit and open-loop transfer function. (b) Block diagram of the loop.

where,

$$K = \frac{K_d K_a K_o}{N}$$
(1b)

For the purpose of expressing parameters in a form that has physical meaning, we change the denominator in equation 1 to:

$$(2 + \alpha)\left(s^2 + \frac{2s}{\tau} + \omega n^2\right)$$

There are three types of input reference signals:

- Case 1: The input reference is constant phase θ_i(s) = φ/s.
- Case 2: The input reference is constant frequency f_0 , $\theta_i(s) = 2\pi f_0/s^2$.
- Case 3: The input reference is constant frequency ramp f_0 , $\theta_i(s) = 2\pi f_0/s^3$

Case 1 is rarely used; sometimes peo-

ple use case 3, but most people use case 2.

Appendix equations (a1) and (a2) give the solution for case 2 and case 3, where, $b = 1/T_3$.

In equations (a1) and (a2), there are trigonometric terms that have a common angle argument.

Let the argument:

$$\sqrt{\omega n^2 \tau^2 - 1} \frac{t}{\tau} = 2\pi F_S \frac{t}{T_s}$$
(3)

where,

$$\tau = \frac{T_s}{2\pi} \qquad \omega n = \frac{\sqrt{1 + F_S^2}}{\tau} \qquad (4)$$

In this way, the argument has physical meaning. F_S is swing frequency. T_S is the time scaling factor. We still need another parameter. I suggest the maximum deflection angle of the open loop

transfer function, ϕ , is another one [4]. The quantity ϕ is defined by:

$$T_2 = T_3 \beta \tag{5}$$

and

$$\beta = \frac{\cos\phi}{1 - \sin\phi} \tag{6}$$

When we define equation (2) as the denominator of equation (1), there is an relation between these three parameters, F_S, T_s and ϕ . Using this relation, we get:

$$B = \omega n^2 \cdot (1 - \beta) + 4/\tau^2 \tag{7}$$

$$\alpha = \frac{\tau}{4} \left[-\mathbf{B} + \sqrt{\mathbf{B}^2 - \left(4\frac{\omega n}{\tau}\right)^2} \right]$$
(8)

$$\mathbf{b} = \alpha + 2/\tau \tag{9}$$

 α in equation (8) must be real, so ϕ has a limitation:

$$\phi_{\min} < \phi < \pi/2 \tag{10}$$

where (11):

$$\phi \min(\mathbf{F}_{S}) = \frac{\pi}{2} - 2 \operatorname{atn} \left(\frac{1}{1 + \frac{2}{\sqrt{1 + \mathbf{F}_{S}^{2}}}} \right)$$

When $F_S = 0$, ϕ_{min} has the largest value, 53.135°

Given F_s , T_s and ϕ (ϕ satisfying equation 10), we can calculate α , τ , ω_n and b using the above equations. Then equations (a1) and (a2) are calculable. The time constants of the active filter are given by the following equations:

$$T3 = 1/(a + 1/\tau)$$
 (12)

$$T2 = T3 \beta \tag{13}$$

$$T1 = K/(T3 \alpha \omega n^2)$$
(14)

where K is defined by equation (1b).

Analyzing the above three time constants, we discover that T1 is proportional to T_s^2 . T_2 and T_3 are proportion-al to T_s . T_1/T_s^2 , T_2/T_s and T_3/T_s are independent of T_s , leaving only the two parameters, F_s and ϕ . The following graphs show these relations.

First, F_S limits the range of ϕ by determining ϕ_{min} (equation 11). Figure 2 shows the relation between F_S and



Figure 2. ϕ_{min} as a function of F_s .

 ϕ_{\min} . The x-axis is F_S , and the y-axis is

 ϕ_{\min} in degrees. F_c is the free F_S is the frequency in the scaled time base. A large F_S means rapid variation, and fast settling to steady state (i.e. fast locking). Fast settling is the advantage of a large F_S. Its disadvantage is that a loop with large F_S is less stable, and is more likely to be unlocked due to noise, interference, or other causes. Figure 2 shows that a larger F_S has a smaller ϕ_{min} . This means we can select \$\$ from a bigger range. It does not give us more range of stability. Conventionally, engineers use phase margin to gauge stability. Large phase margin ensures good stability. Phase margin is defined as the difference between the phase angle of the open loop transfer function and π (or $-\pi$) when the amplitude of open loop transfer function is one. Using this defininition, phase margin is plotted in Figure 3.

The x-axis is F_s , and the y-axis is the phase margin in degrees. Different curves correspond to different values of ϕ . As mentioned before, the range of ϕ is limited by F_S . If $\phi = 20^\circ$, F_S can not be larger than 4.563. In Figure 3, the dotted curve corresponding to $\phi =$ 20° ends at $F_{S} = 4.563$. However, those curves corresponding to $\phi \ge 253.135^{\circ}$ do not break in the figure.

The y-axes of Figures 3(b,c,d) represent T_3/T_s , T_2/T_s and KT_s^{2}/T_1 , respectively. The x-axis of all figures is F_s . Figures 3a through 3d each have five curves corresponding to different values of \$. Sometimes, two curves are very close in the display and graphics only shows the second curve legend. When ϕ approaches 90°, the thirdorder case degenerates to the secondorder. In this case, $T_3 = 0$ (see Figure 3b). Using the $\phi = 89.99^{\circ}$ curve, we can investigate the properties of secondorder loops. Further analysis shows



Figure 3. Various quantities as a function of F_S and ϕ .

that when the input reference is constant, $\theta_e(t)$ Case 2, equation (a1), is proportional to $f_0^*T_s$, and $\theta_e(t)$ of Case (3), equation (a2), is proportional to $f_0T_s^2$. Now, we define $F_0 = f_0^*T_s$. If we define F_s , ϕ and F_0 , and use t/T_s as the time coordinate, equation (a1), denoted as $\theta_2(t)$, is independent of T_s . Meantime, the result of Case (3) (frequency ramp), denoted as $\theta'_2(t)$ is proportional to Ts. The curve $\theta'_2(t)/T_s$ is also independent on T_s , if we take t/T_s as the time axis.

In practicable design, selecting F_0 may be a problem. If F_0 is too big, the locking process will be slow, even it will not be locked (especially in analog phase detector). How to select F₀ is out of the scope of this paper. For safely quick locking, keep F_0 to a low value. The better way, but not necessary, is after selecting all F_S , ϕ and F_0 , plot $\theta_2(t)$ and $\theta_2(t)/Ts$ of equations a1 and a2 versus the t/T axis. These two curves are independent of T_s . $\theta_2(t)$ displays the result of the input as a frequency jumping F₀ in an infinitesimal time. While $\theta' 2(t)/T_s$ displays the result of input as a constant continuous frequency change, the rate of change is F_0 Hz in T_s second. When we design a real synthesizer, we use a crystal oscillator as the reference frequency generator. The reference frequency is not established instantly. So, the initial part of real description should resembles $\theta' 2(t)/Ts$ curve. Later, after reaching a certain frequency it should stop increasing and settle to zero. The real transient is below $\theta_2(t)$ and settles to a steady state zero. In a Type 2 PLL, $\theta'_2(t)$ does not tend to zero. Nevertheless, $\theta_2(t)$ tends to zero. Using a digital phase detector, the PLL should be locked finally. If we hope it has a fast cap-

200 150 50 0 -50 0 0.1 0.2 0.3 0.4 0.5 0.6 0.7 0.8 0.9 1.0

Figure 4. Phase error $\theta_e(t)$ of Case 2 and $\theta_e(t)/T_s$ of Case 3. The X-axis is $T_s(t/T_s)$; the Y-axis is in degrees.

ture, the maximum value of transient should below 180° which is the linear range of digital phase detector. If it is too big, lower $\theta_2(t)$ by decreasing F_0 and/or changing F_S and ϕ properly.

Example

A 3rd order PLL using a Type 2 active filter produces an output of 16.95 MHz from a 5-KHz reference, \mathbf{F}_0 . The phase detector, active filter and VCO have the following factors:

$$\begin{split} & K_{\rm d} = 0.19 \; \text{V/rad} \\ & K_{\rm a} = 1 \\ & K_{\rm o} = 10.6 \; 1 \; \text{o6 rad/slV} \\ & \text{Divider ratio N} = 16.95 / 0.005 = 3390 \\ & \text{then, K} = K_{\rm d} \; K_{\rm a} \; K_{\rm o} / \; \text{N} = 594.1 \end{split}$$



RF Design



Figure 5. For the Type 3 PLL — (a) Loop filter circuit and open loop transfer function; (b) Block diagram of the loop.



Assigning $F_S = 1.2$, from Figure 2 or equation 11, we have $\phi_{min} = 42.6^{\circ}$. Select $\phi = 42.7^{\circ}$ and $F_0 = f_0 T_s = 5$. The phase margin is 42.69° which can be found in Figure 3 (approximately). Next, assume T_s to be any value. Using above formulae, we can calculate ωn , τ , α and b. Substitute all parameters in equations (a1) and (a2). Draw $\theta_2(t)$, in solid line, and $\theta'_2(t)/T_s$, in the dashed line, on Figure 4. The Xaxis is in scale of T_s(t/T_s), and the Yaxis in degrees.

The maximum value of $\theta_{2}(t)$ is about 150°. Because it is smaller than 180°, it is a fast locking one. Be reminded that the curves on Figure 4 are independent of T_s . First, we select $T_s = 0.0165$. From Figures 3(b,c,d) or equations (3,4,6-8) and (12-14), we have:

Tl = 3.61×10^{-3} sec. $T2 = 3.66 \times 10^{-3}$ sec. $T3 = 7.030 \times 10^{-4}$ sec.

These data are equal to result of Przedpelski's [5] within a tolerance of 10%. Because, $f_0 = F_0/T_s = 5/T_s = 303$ Hz. That means the initial jump of reference frequency is not bigger than 300 Hz. If the jump is much bigger than 300 Hz. The locking process will be slow.

Select $T_s = 10^{-3}$, $f_0 = 5$ kHz, this is the reference frequency. In the case of jumping 5 kHz instantly, we have

Tl = 1.327×10^{-5} sec. $T2 = 2.22 \times 10^{-4}$ sec. $T3 = 4.26 \times 10^{-5}$ sec.

The locking speed is much faster than before, which is directly proportional to T_s. Because the time constant is much smaller, we must consider the frequency response performance of opamp. If this device has sufficiently large gain bandwidth product, it is a good choice. Or, we increase T, to make a compromise.

Type 3 PLL

6

I use the conventional symbol and definition. The Laplace Transform of Type 3, represented by Figure 8, is the following equation (15):

$$\frac{\frac{\theta_{e}(s)}{\theta_{i}(s)}}{s^{3} + s^{2}K\left(\frac{T_{2}}{T_{1}}\right)^{2} + s2K\left(\frac{T_{2}}{T_{1}^{2}}\right) + K\left(\frac{1}{T_{1}^{2}}\right)}$$

98



Figure 6. T_2/T_s (dashed line) and $T_1/K^{0.5} T_s^{1.5}$ (solid line) are functions of F_s (x-axis) only.

The above equation is correct, only the all initial conditions are zero. K is still defined by (1b).

As with Type 2, assume the denominator of equation (15) in the form of equation (2). Also, there are three cases of input reference signal:

- Case (1): The input reference is constant phase ϕ , $\Phi_i(s) = \phi / s$. Never deal with this case.
- Case (2): The input reference is constant frequency f_0 , $\Phi_i(s) = 2\pi f_0/s^2$. The initial condition of $\theta_e(t)$: $\theta_e(0) = 0$, $\theta_e(0) \neq 0$. In equation (a1),

$$b = K (T_2/T_1)^2$$
(16)

• Case (3): The input reference is constant frequency ramp f_0 , $\Phi_i(s) = 2\pi f_0/s^3$. The initial condition of $\theta_e(t)$: $\theta_e(0) = 0$, $\theta_e(0) = 0$. In equation (a2),

$$b = 0$$
 (17)

Because the solution of $\theta_e(t)$ in Case 2 or Case 3 still uses (a1) or (a2), we define F_S and T_s as before, equations (3) and (4) are still effective. Due to variables in the denominator of (15) being only T_1 and T_2 , parameter F_S and T_s are sufficient to determine this problem. Using the internal relationship of (2), we get:

$$\alpha = \frac{\omega n^2}{2\left(\omega n + \frac{1}{\tau}\right)}$$
(18)

$$T_2 = \frac{2\frac{\alpha}{\tau} + \omega n^2}{2\alpha\omega n^2}$$
(19)

$$T_1 = \frac{1}{\omega n} \sqrt{\frac{K}{\alpha}}$$
(20)

We have all equations now. Analysis shows T_2/T_s and $T_1/(K^{0.5} T_s^{1.5})$ are

function of F_S only. In Figure 6, the dashed line represents T_2/T_s , and the solid line represents $T^{1/}(K^{0.5} T_s^{1.5})$. The X-axis is also F_s .

The X-axis is also F_S . Phase margin is shown in Figure 7. X-axis is also F_S , and the Y-axis is phase angle in degrees. Further analysis proves the same behavior as Type 2 on the solution of two cases. Let the 3rd order Type 3 Case 2 solution (a1), with $b = K(T_s/T_1)^2)$ be $\theta_3(t)$. $\theta_3(t)$ is proportional to T_s . If we define $F_0 = f_0^*T_s$, and keep F_0 constant. $\theta_3(t)$ is only a function of F_s and F_0 on the T_s scaled time axis. The Case 3 solution (a2, with b=0) is denoted as $\theta'_3(t)$. $\theta'_3(t)/T_s$ is also independent of T_s .

Example

A 960 MHz transmitter is based on a



3rd order Type 3 PLL, the reference frequency is 15 MHz, where:

 $\begin{aligned} &K_{d} = 0.25 \text{ V/rad} \\ &K_{a} = 1 \\ &Ko = 3 \times 10^{9} \text{ rad/s/V} \end{aligned}$

Calculate N = .960/15 = 64. Select F_S = 2.7. With Figure 10, we have phase margin 40.8°. Assume any T_s value, calculate ω n and τ with (3) and (4). Calculate a with (18). Calculate b for Case 2 with (19),(20) and (16). $F_0 = f_0$ $T_s = 7$. Substitute all parameters in (al) and (a2). We then get the curves of Figure 8.

We can see maximum value of solid line is about 140°. And both curves tend to zero. These curves are independent of T_s . We have the freedom to select T_s . Select $T_s = 0.001$. Using Figure 9 or (19) and (20), get the following results:

 $\begin{array}{l} T_1 = 2.31 \times 10^{-3} \; {\rm sec.} \\ T_2 = 9.37 \times 10^{-5} \; {\rm sec.} \end{array}$

With this choice, $f_0 = 7$ kHz. If the

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Figure 7. Phase margin versus F_S for the Type 3 PLL



Figure 8. $\theta_e(t)$ for Case 2 (solid line) and $\theta_e(t)$ of Case 3 (dashed line) in degrees.

reference frequency jumping more than 7 kHz instantly, the maximum swing of $\theta_e(t)$ may be bigger than 180°. That may cause slow locking. Selecting small T_s can reduce capture time, but that is limited by the performance of device.

Comparing Type 2 and Type 3, one more op-amp is needed in Type 3. However, continuous frequency ramp, Case 3, makes steady phase error $\theta_{e}(t)$ not zero in Type 2. *RF*

References

1. Fred Salvalti, "Technique eases design of high-order PLLs," *EDN*, June 9, 1994, pg 172

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3. High Performance Frequency Conlrol Products, Motorola, Inc. 1993

4. U.L. Rohde, Digital PLL Frequency Synthesizer Theory and Design, Prentice Hall, 1983 ibid 3, page 142.

About the Author

Fu Nian Ku can be reached at 35 Violet Avenue, Mineola, NY 11501. He is currently working as a consultant at Mitsubishi's Advanced TV Lab. Previously, he worked at ADEMCO. He received BSEE and PhD degrees from Tsinghua and Peking Universities, respectively.

Appendix:

Equation (a1) — The function:

$$\frac{2\pi f_0(s+b)}{(s+\alpha)\left(s^2+\frac{2}{\tau}s+\omega n^2\right)}$$

has the inverse transform:

$$\frac{2\pi f_0 \tau}{-2\alpha + \alpha^2 \tau + \omega n^2 \tau} \bullet$$

$$\begin{vmatrix} e^{-\alpha t} (b - \alpha) \\ + \left(e^{-1/\tau} \frac{\tau \omega n^2 + \tau a b - a - b}{\sqrt{\omega n^2 \tau^2 - 1}} \right) \sin\left(\frac{1}{\tau} \sqrt{\omega n^2 \tau^2 - 1}\right) \\ + e^{-1/\tau} (a - b) \cos\left(\frac{1}{\tau} \sqrt{\omega n^2 \tau^2 - 1}\right) \end{vmatrix}$$

Equation (a2) — The function:

$$\frac{2\pi f_0 \left(1 + \frac{b}{s}\right)}{\left(s + \alpha\right) \left(s^2 + \frac{2}{\tau}s + \omega n^2\right)}$$
has inverse transform:

$$\frac{2\pi f_0}{-2\alpha + \alpha^2 \tau + \omega n^2 \tau} \bullet$$

$$\left| \begin{array}{c} e^{-\alpha t} \left(\tau - b\frac{\tau}{a}\right) + b \left(\frac{\alpha \tau}{\omega n^2} + \frac{\tau}{a} - \frac{2}{\omega n^2}\right) \\ + \left[\frac{e^{-1/\tau}}{\sqrt{\omega n^2 \tau^2 - 1}}\right] \left[\tau^2 (\alpha - b) - \tau \left(1 + \frac{\alpha b}{\omega n^2}\right) + \frac{2b}{\omega n^2}\right] \\ \cdot \sin \left(\frac{1}{\tau} \sqrt{\omega n^2 \tau^2 - 1}\right) \\ + e^{-1/\tau} \left[-\tau + \frac{b}{\omega n^2} (2 - \tau \alpha)\right] \cos \left(\frac{1}{\tau} \sqrt{\omega n^2 \tau^2 - 1}\right)$$

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ENGINEER'S NOTEBOOK

Simple Direct-Conversion Receiver Checks Frequency Counters

By Michael A. Covington The University of Georgia

Many frequency counters use a 10 MHz precision crystal oscillator and include an "oscillator out" jack where this frequency can be sampled. The circuit presented here is a direct-conversion receiver that tunes in WWV on 10 MHz, using the frequency counter timebase as its local oscillator, thus providing immediate confirmation that the oscillator is on frequency.

The circuit is an adaptation of the American Radio Relay League's "Neophyte Receiver" [1] and can be built on a printed circuit board made for the receiver [2]. It has been modified to have a 9-volt supply, a 10 MHz front end, and no local oscillator. The audio section has also been simplified.

The two ICs are an NE602 mixer-oscillator (with the oscillator disabled) and an LM386 audio amplifier. Note that the LM386 is fed in differential mode by the differential outputs of the NE602. This greatly improves rejection of nearby AM signals. If the output of the NE602 were taken single-ended, it would include some audio from shortwave broadcasters in the 9.5 MHz band and possibly even some local AM stations.

Antenna requirements depend on location. In Georgia, a 7 MHz dipole mounted in an attic was found to be sufficient; a 10 MHz dipole would, of course, be better. Near the WWV and WWVH transmitters in Colorado and Hawaii, an indoor whip antenna might suffice. Radiowave propagation will affect the reliability of reception. With the low sunspot activity that is occuring in the present solar cycle, it is possible that WWV and WWVH will not be heard during nighttime hours. In periods of high solar activity, daytime reception of the stations at distant locations may be hindered by absorption.

In normal use, this receiver serves to confirm that the frequency counter has not drifted drastically, and it renders the timebase NIST-traceable. For critical adjustment of the TCXO, an oscilloscope can be connected across the speaker to observe the subaudible beat frequencies. Note, however, that the signal from WWV is reflected off the ionosphere. If the reflective layer is rising or falling at the time, Doppler shifts can change the frequency by 1 Hz or more. RF

References

1. John Dillon, "The Neophyte Receiver," QST, February 1988, pp. 14-18.

2. One source for the printed circuit boards is: FAR Circuits, 18N640 Field Ct., Dundee, IL 60118-9269; tel/fax: (708) 836-9148.

About the Author

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Figure 1. Circuit diagram of the WWV direct-conversion receiver.



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3dbm, Inc.	72	F. W. Bell, Inc.	22	Oscillatek	6
3M Electrical Specialties Div.	91	Fluke Corporation	109	Pacific Monolithics	60
ASM Products, Inc.	92	Frequency Electronics	69	Philips Components	68
Advanced Control Components	65	Fujipoly America Corp.	91	ProTek, Devices	9:
Advanced Messaging Systems Division (AMSD)	26	Fujitsu Towa Electron Ltd.	22	Pulse	6
Aerovox Group	109	Gage Applied Sciences	67	Ray Proof, EMC Test Systems	9
Allied Bsiness Intelligence, Inc.	24	Group Technologies Corporation	22	Reeves-Hoffman	70
Alliance Semiconductor	24	Harris Semiconductor	22	Richardson Electronics Ltd.	7:
Anadigics Inc.	66	Hewlett-Packard Co.	65, 108	Rodgers Corporation	2:
Analog Devices, Inc.	56	IFI, Inc.	92	Rohde Schwarz GmbH & Co.	9
ARLON, Inc.	22	IEEE	109	SGS-Thomson Microelectronics	2:
Arnold Engineering Co.	84	Interad, Ltd.	22	SiS	2.
Audra System	26	JEFA International	22	STB Systems	24
Avista Design Systems	108	Jameco Electronics	109	Scientific-Atlanta	20
Baltelle R&D	22	K&L Microwave, Inc.	68, 109	Semtech Corporation	9:
Berndt Associates, Inc.	84	LCF Enterprises	65	SERSYNTH, Inc.	10
Bird Component Products	68	Level One Communications	66	Sierra Semiconductor	2.
Booton Electronics Corp.	67	Lindgren RF Enclosures, Inc.	84, 91	Sony	2
Cadence Design Systems, Inc.	22	Logistics Technology News	109	Sprague-Goodman Electronics	6
Canadian Satellite Communicatons (Cancom)	26	Loral Communication Systems	62	Systems Science, Inc.	2:
Champion Technologies, Inc.	70	James Walker & Co.	91	Tait Electronics-USA, Inc.	69
Cirrus Logic	22	MF Electronics	109	Tech-Etch, Inc.	9:
COMSAT RSI, Inc.	22	Metrum, Inc.	22	Technology International Inc.	84
Comlinear	74	Metric Systems Corp.	72	TEMIC Telefunken	60
Comtech Microwave Products Corp.	22	Micro Networks	69	Toko America, Inc.	68
Comtech Telecommunications Corp.	22	MicroWave Technology	66	Transport Technology Publishing	109
Connec t-Tech Products	72	Mitsubishi Electronics	68	Trident	24
Cubic Communicatons, Inc.	22	Molex, Inc.	109	Trompeter Electronics	109
Current Technology, Inc.	84	Motorola	26, 66, 108	Tru-Connector Corp.	75
Dale Electronics	91	NIST	24, 84	Tusonix	95
Densitron Corporation of America	22	Noble Publishing	8	ULSI Systems	24
Dynamic Sciences International	91	Noise Com, Inc.	65	Ultraview Corporation	6'
EMC Consulting, Inc.	86	Oak Frequency Control Group	70	Vectron Labs	69
Ecliptek Corporation	69	Oak Technology	24	Vectron Technologies Inc.	70
Elanix Inc	108	Ocean Applied Research Corp	22	Waisshal Corporation	00

RFadvertising index

			ADVERTISER	PAGE #	
ADVERTISER	PAGE #	READER SVC #	Milliren Technologies, Inc		
A.H. Systems			Mini Circuits	4-5,634-35, 75.6	2.3,26.27,64,93,4,5.61,62,87,88
Advantage Instruments			Miteq		
Amplifier Research			National Semiconductor	CVR 4	
California Eastern Labs			National Instruments		
Centurion			Noise Com. Inc		
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Fujitsu Microelectronics			Sprague Goodman		
Giga-tronics Inc.			Surcom Associates		
Henry Radio			Synergy		
Hewlett Packard			Telcom Industries		
Hexawave			Tele Ouarz		58
International Crystal Mfg.			Temex		
Interad Ltd.			Toko America, Inc		
ITT GTC			TPL Communications		
Jan Crystals			Tusonix. Inc.		
Johnson Components			Univeristy of Oxford		
Kalmus Engineering	9		Valpey-Fisher		
KVG North America			Vari-L Company, Inc		
Lap-Tech			Vectron Technologies		
Lindgren RF Enclosures			Vectron Laboratories		
LPKF CAD/CAM Systems. Inc			Voltronics		
M/A Com (3 Divisions)			Wayne Kerr, Inc		
Metuchen Capacitors, Inc.			Werlatone, Inc.		
MicroNetworks Corp			XTAL Technologies, Ltd		32
Milcom International			Z Communications		

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