RF design

engineering principles and practices

February 1991

Official Show Issue

RF Expo West

Featured Technology

Digital Modulation Techniques

Cover Story

A New RF CAE System

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INFO/CARD 1

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The RF system of the new FM/AM-1600S provides excellent spectral purity along with FM, AM and ΦM capabilities. The instrument's independent RF Generator and Receiver allow frequency offsets to 1 GHz for applications such as cross-band repeater tests and RF injection while signal tracing a receiver front-end and I.F.

The digitized Spectrum Analyzer provides scan widths from 1 kHz/div to full scan, plus store, recall, compare and peak hold trace modes. Log scales of 10 or 2 dB/div and amplitude units scaled in dBm, dBμV, dBmV, dBV or dBμW.

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INFO/CARD 2

Tired of Two-Bit Performance?

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SELECTION CHART

Part No.	Type	Freq/MHz	IL/dB*	Iso/dB*	Sw. Speed/ nsec**	Package	Control
DS0841	SPST	10-200	1.7	70	7.0	14 Pin DIP	TTL
DS0699	SPST	5-1500	0.9	60	23.0	TO-8	TTL
DS0990	SPST	10-2000	1.5	78	25.0	14 Pin DIP	TTL
DS0850	SP2T	DC-2000	0.35	47	5.0	TO-5	0/-7
DS0813	SP2T	DC-2000	0.65	47	200.0	TO-5	TTL
DS0812	SP2T	10-1000	0.5	54	50.0	TO-8	TTL
DS0860	SP2T	10-1500	0.5	47	40.0	.380 sq	TTL
DS0602	SP2T	5-4000	1.5	55	22.0	14 Pin DIP	TTL
DS0842	SP2T	5-1500	0.9	75	50.0	14 Pin DIP	TTL
DS0864	SP4T	5-2000	1.3	55	28.0	16 Pin DIP	TTL
DS0874	SP4T	DC-2000	2.0	60	75.0	16 Pin DIP	TTL
DS0838	SP8T	5-1000	2.2	40	40.0	24 Pin DIP	TTL

*Typical performance measured mid-band.
**50% TTL to 10%/90% RF.

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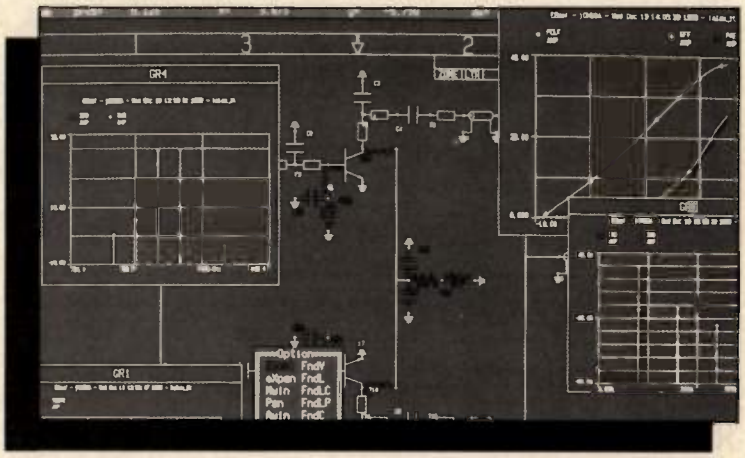

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R.F. DESIGN (ISSN: 0163-321X USPS: 453-490) is published monthly plus one extra issue in September. February 1991. Vol. 14, No. 2. Copyright 1991 by Cardiff Publishing Company, a subsidiary of Argus Press Holdings, Inc., 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111 (303) 220-0600. Contents may not be reproduced in any form without written permission. Second-Class Postage paid at Englewood, CO and at additional mailing offices. Subscription office: *RF Design*, 5615 W. Cermak Rd., Cicero, IL 60650. Domestic subscriptions are sent free to qualified individuals responsible for the design and development of communications equipment. Other subscriptions are: $38 per year in the United States; $48 per year in Canada and Mexico; $52 (surface mail) per year for foreign countries. Additional cost for first class mailing. Payment must be made in U.S. funds and accompany request. If available, single copies and back issues are $5.00 each (in the U.S.). This publication is available on microfilm/fiche from University Microfilms International, 300 Zeeb Road, Ann Arbor, MI 48106 USA (313) 761-4700.

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RF editorial

Obsolete Technologies?

Gary A. Breed
Editor

Note: These comments are "food for thought" only!

Like the steam engine and mechanical adding machines, technologies often are completely replaced by new methods. Others, like woven cloth and wood furniture, don't die away. They keep their basic form while materials and production methods are refined. I'd like to offer for discussion some of electronic technologies that might fit into either of these categories in the not-too-distant future.

Cathode Ray Tubes. The vacuum tube has been relegated to museums, except for high power and special applications. The CRT may follow the same path, as flat-screen technology develops. Although LEDs and LCDs are candidates for replacement of the CRT, one promising flat-screen technology can be considered a refinement, since it uses a vacuum enclosure and a phosphor-coated screen. Instead of using an electron beam to write the image, this method uses a matrix of individual emitters just a short distance behind the screen. Size, weight, and emitted magnetic field advantages will assure that flat screens will eventually take over.

Copper Wire Communications. Mainly, I'm referring to the communications infrastructure of the country: the telephone's "twisted pairs" and cable television's coax. We are already in the early stages of a complete replacement of copper with fiber optics. The advantages of size, weight, cost, and transmission capacity of fiber have already been proven. However, the political and economic decisions that are needed to support such a massive project are monumental.

Fiber-to-home may be the final step in the "wired nation" concept that futurists have been predicting for 40 years. With this technology, telephone, data, audio and video services can reach every home now served by electricity and/or telephone.

Broadcast Television. With fiber to virtually every home, broadcast television would become obsolete. Only a few viewers would require television away from home (in campgrounds, for example), and they could be served by satellite or short-range transmissions. Radio would not be significantly affected, since most listeners will still prefer to listen in automobiles, backyards, walking, biking, etc.

Personal Communications. This will help make inconvenience obsolete. To complement the fiber-optic network, we will have even more wireless communications. The cellular boom is just the start of a process that will eventually lead to a wireless "personal communicator," usable anywhere in the world. Whether this universal system gets implemented with terrestrial radio, satellite systems, or some other technology is still unknown. But, most of us now realize that such capability is no longer science fiction.

It's a lot of fun looking far into the future, but it serves another purpose: We are too easily preoccupied with today's problems, and the next product that is in development. Too often, we don't establish good long-term plans and are caught by surprise. Add your own ideas to my list and remember them now and then, just to remind yourself where your company and your career are headed.

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INFO/CARD 5

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a Cardiff publication — Established 1978

Main Office:
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RF calendar

February

5-7 **RF Technology Expo 91**
Santa Clara Convention Center, Santa Clara, CA
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

24-28 **NEPCON West '91**
Anaheim Convention Center, Anaheim, CA
Information: Janet Schafer, Cahners Exposition Group, 1350 E. Touhy Ave., Des Plaines, IL 60017-5060. Tel: (708) 299-9311. Fax: (708) 635-1571.

March

18-21 **WESTEC '91**
Los Angeles Convention Center, Los Angeles, CA
Information: Event Public Relations Department of SME, One SME Dr., PO Box 930, Dearborn, MI 48121-0930. Tel: (313) 271-0777.

26-28 **International Mobile Communications Expo**
Anaheim Convention Center, Anaheim, CA
Information: April DeBaker, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773-9716.

April

7-10 **1991 US Conference on Gallium Arsenide Manufacturing Technology**
Reno, Nevada
Information: 1991 GaAs MANTECH Conference, Suite 300, 655 15th Street, N.W., Washington, DC 20005.

8-11 **National Electronic Manufacturing and Design Exposition**
McCormick Place, Chicago, IL
Information: Michele Filippi, NEMDE Show Manager, Cahners Exposition Group, 1350 East Touhy Ave., Des Plaines, IL 60018. Tel: (708) 299-9311.

10-17 **Hannover Fair Industry**
Hannover Fairgrounds, Hannover, West Germany
Information: Hannover Fairs USA Inc, 103 Carnegie Center, Princeton, NJ 08540. Tel: (609) 987-1202.

15-18 **36th International SAMPE Symposium and Exhibition**
San Diego Convention Center, San Diego, CA
Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616.

15-18 **HDTV World Conference and Exhibition**
Sands Expo and Convention Center, Las Vegas, NV
Information: NAB, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (202) 429-5350.

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

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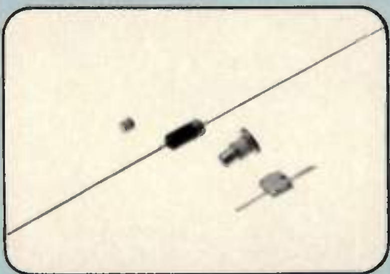

TYPICAL STANDARD MODELS	
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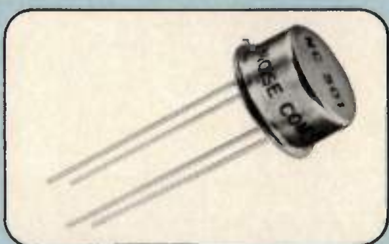

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NC 505	up to 4 GHz
NC 506	up to 5 GHz

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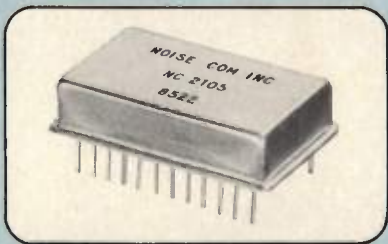

TYPICAL STANDARD MODELS	
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NC 2104	up to 1 MHz
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RF courses

Antennas: Principles, Design, and Measurements
March 13-16, 1991, St. Cloud, FL
Information: Kelly Brown, SCEEE, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146.

Basic Telephony/Digital Switching
March 4-8, 1991, Madison, WI
Cellular Radio
March 25-28, 1991, Madison, WI
ESD - Electrostatic Discharge
April 17-19, 1991, Madison, WI
Information: University of Wisconsin - Madison, Department of Engineering Professional Development. Tel: (608) 262-2061. Fax: (608) 263-3160.

Power Electronic Circuits: Theory and Practice
February 11-15, 1991, Los Angeles, CA
41st Engineering and Management Program
March 24-30, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Advances in Millimeter Wave Applications
February 26-March 1, 1991, Atlanta, GA
Information: Education Extension, Georgia Institute of Technology. Tel: (404) 894-2547.

Introduction to Radar ECM and ECCM Systems
February 20-22, 1991, Washington, DC
Cellular Radio Telephone Systems
February 25-27, 1991, San Diego, CA
Microwave High-Power Tubes and Transmitters
February 25-March 1, 1991, Washington, DC
Broadband Communication Systems
March 4-8, 1991, Washington, DC
Satellite Communications: System Planning, Design, and Operation at Ku and Ka Bands
March 4-8, 1991, Washington, DC
Antennas: Radiation and Scattering
March 11-12, 1991, Washington, DC
Microwave Radio Systems
March 13-15, 1991, Washington, DC
Introduction to Modern Radar Technology
March 13-15, 1991, Washington, DC
Radar Operation and Design: The Fundamentals
March 25-28, 1991, Washington, DC
Lightning Protection
April 11-12, 1991, Orlando, FL
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Thin Films
March 26-28, 1991, Providence, RI
Information: Joseph T. Pittle, URI Conference Office. Tel: (401) 792-2170.

Antenna Analysis, Design, and Measurements
February 19-22, 1991, Tempe, AZ
Fiber Optic Communications
February 25-27, 1991, Tempe, AZ
Current Topics on Fiber Optic Technology
February 28-March 1, 1991, Tempe, AZ

High Resolution Radar, Surveillance and Target Recognition
March 25-27, 1991, Tempe, AZ
Information: Center for Professional Development, Arizona State University. Tel: (602) 965-1740.

Digital Signal Processing Workshop
March 12-14, 1991, Campbell, CA
April 10-12, 1991, Norwood, MA
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Modern Microwave Techniques
February 25-28, 1991, Garmisch-Partenkirchen, Germany
Far-Field, Compact and Near-Field Antenna Measurement Techniques
February 25-28, 1991, Garmisch-Partenkirchen, Germany
Aspects of Modern Radar
February 25-28, 1991, Garmisch-Partenkirchen, Germany
Microwave and Radar Technology
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany
MESFET and Hetrostructure Based MMICs
February 25-March 1, 1991, Garmisch-Partenkirchen, Germany
Modern Digital Modulation Techniques
March 11-15, 1991, United Kingdom
Digital Signal Processing: Filtering and Estimation
March 18-21, 1991, United Kingdom
Broadband Telecommunications
March 18-22, 1991, United Kingdom
Modern Digital Communications for Space, Satellite and Radio
April 15-18, 1991, Italy
RF and Microwave Circuit Design I: Linear Circuits
April 15-19, 1991, Italy
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: +46 (0) 122-17570. Fax: +46 (0) 122-14347.

Introduction to EMI/RFI/EMC
February 20-22, 1991, Orlando, FL
Information: Besser Associates. Tel: (415) 949-3300, Fax: (415) 949-4400.

Electronic Design Techniques and Analysis Required to Meet Electromagnetic Compatibility Requirements
February 19-20, 1991, Novi, MI
Advanced EMC Printed Circuit Board Design
February 21, 1991, Novi, MI
Information: Jastech. Tel: (313) 553-4734.

Introduction to EMI/RFI/EMC
February 19-21, 1991, Orlando, FL
Grounding and Shielding
February 26-March 1, 1991, Anaheim, CA
Practical EMI Fixes
March 11-15, 1991, San Francisco, CA
High Speed Digital Design for EMC
April 9-12, 1991, Seattle, WA
System Integration and Design for EMC
February 19-22, 1991, San Francisco, CA
Information: Interference Control Technologies, Registrar. Tel: (703) 347-0030.

Virginia Tech Starts Mobile and Portable Radio Research Group

Following recent trends in the U.S. wireless communications industry, Virginia Tech recently formed the Mobile and Portable Radio Research Group (MPRG). The program's director, assistant professor Ted Rappaport, built the program around an already strong undergraduate and graduate program in satellite communications, RF/microwave engineering, and electromagnetics. Even though MPRG is less than a year old, it has already attracted several university and corporate sponsors such as Motorola Inc., Northeastern University, Purdue University, Tektronix and others. According to the *Propagator*, the group's newsletter, the group's work focuses on propagation prediction, system design and bit-error simulation for cellular, microcellular and indoor wireless communications; new work is developing analytical tools and computer simulation techniques for spread spectrum wireless networks and robust adaptive equalization techniques for RF and acoustical channels.

The program currently consists of more than a dozen graduate and undergraduate students and three EE faculty and is considered one of the fastest growing programs on campus. MPRG will be hosting a Personal Communications Symposium in June which will provide an open forum for technical discussions and interaction with experts in the personal commmunications industry.

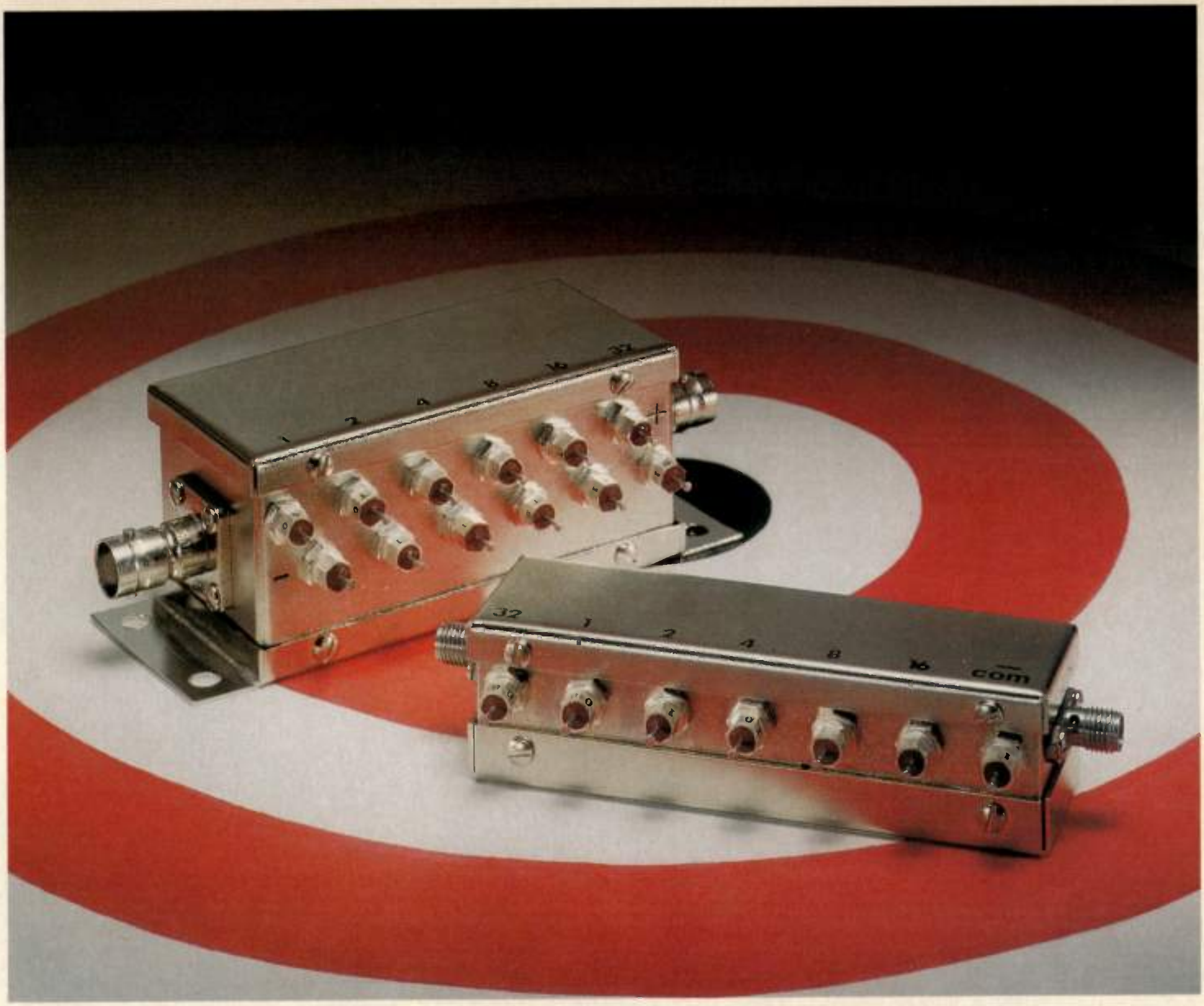
New Class of Amateur Radio License — In response to requests from numerous petitions and public comments, the FCC has created a new Technician class Amateur Radio license which does not require proficiency in Morse code before receiving all amateur privileges above 30 MHz. The FCC has also retained the current Novice Class operator license for those who prefer to pass the 5 word-per-minute Morse code requirement. The codeless Technician license will provide an entry point into Amateur Radio for technically qualified persons who find the Morse code to be a barrier. However, Technicians wishing to gain access to the shortwave bands for worldwide communications must still pass the Morse code test to conform to international requirements.

Time and Frequency User's Manual Updated — NIST has just published a new edition of its *Time and Frequency User's Manual*. Written for readers at all levels of understanding, this edition contains updated information about time and frequency services available from NIST, other federal agencies, and other countries. The indexed publication will be useful to technicians, experimenters, calibration laboratories, and scientists since it covers most aspects of receiving and using time and frequency calibration signals, the history of time services, foreign transmitters, satellite services, and calibration methods. NIST SP 559 is available from the Superintendent of Documents, U.S. Government Printing Office, Washington, DC 20402. Order by stock no. 003-003-03050-3 for $8.50 prepaid.

The Use of International Private Telecommunications Circuits Liberalized — Study Group III of the International Telecommunications Union recently ended its second plenary meeting with agreement on 18 recommendations concerning the international charging and accounting principles of various telecommunications services including message handling services, ISDN services, services provided by digital systems, public data communication services and videotex. One of the decisions concerned the revision of Recommendation D.1 on the general principles applicable to international leased private telecommunication circuits. The new text of Recommendation D.1 advocates a certain degree of liberalization in the conditions of use of international private leased circuits, particularly in respect of the constitution of networks and their use by third parties, including the public, as well as their interconnection to public networks. At the same time, it provides the necessary safety measures to ensure the financial viability of the public service operator. This, along with the other recommendations will be submitted to the next meeting of Study Group III in March 1991 with a view to deciding whether they will be subject to the accelerated approval procedure.

NAB Concerned Over AM Interference Protection in Development of Home Automation Systems — The National Association of Broadcasters has given its support to the development of home automation systems after resolving initial concerns about interference to household AM radio reception. The home automation systems, called "smart houses" use AC power lines to permit household electronic devices to talk to each other, and as a result, help residents perform routine household functions. The Consumer Electronic Bus or CEBus would control the functions of everyday household appliances. The NAB was initially concerned over AM interference but with recent modifications, they are now convinced that devices "can be operated without ... interference." The FCC was also urged to investigate interference problems caused by similar "carrier current" devices that use the frequencies within or near the AM broadcast band, and if necessary, modify the Commission rules to prevent interference to AM radio broadcasts.

U.S. Army Orders Harris Filmless Camera System — Harris Corporation recently received a $5.4 million contract for an electronic filmless camera system (EFCS). The system will allow soldiers to take digital still images without film and rapidly transmit them over HF, UHF and VHF radios or telephone lines. Under the contract, Harris will integrate and supply 126 manportable transmission systems and 36 base stations to receive the still images from remote locations around the world.



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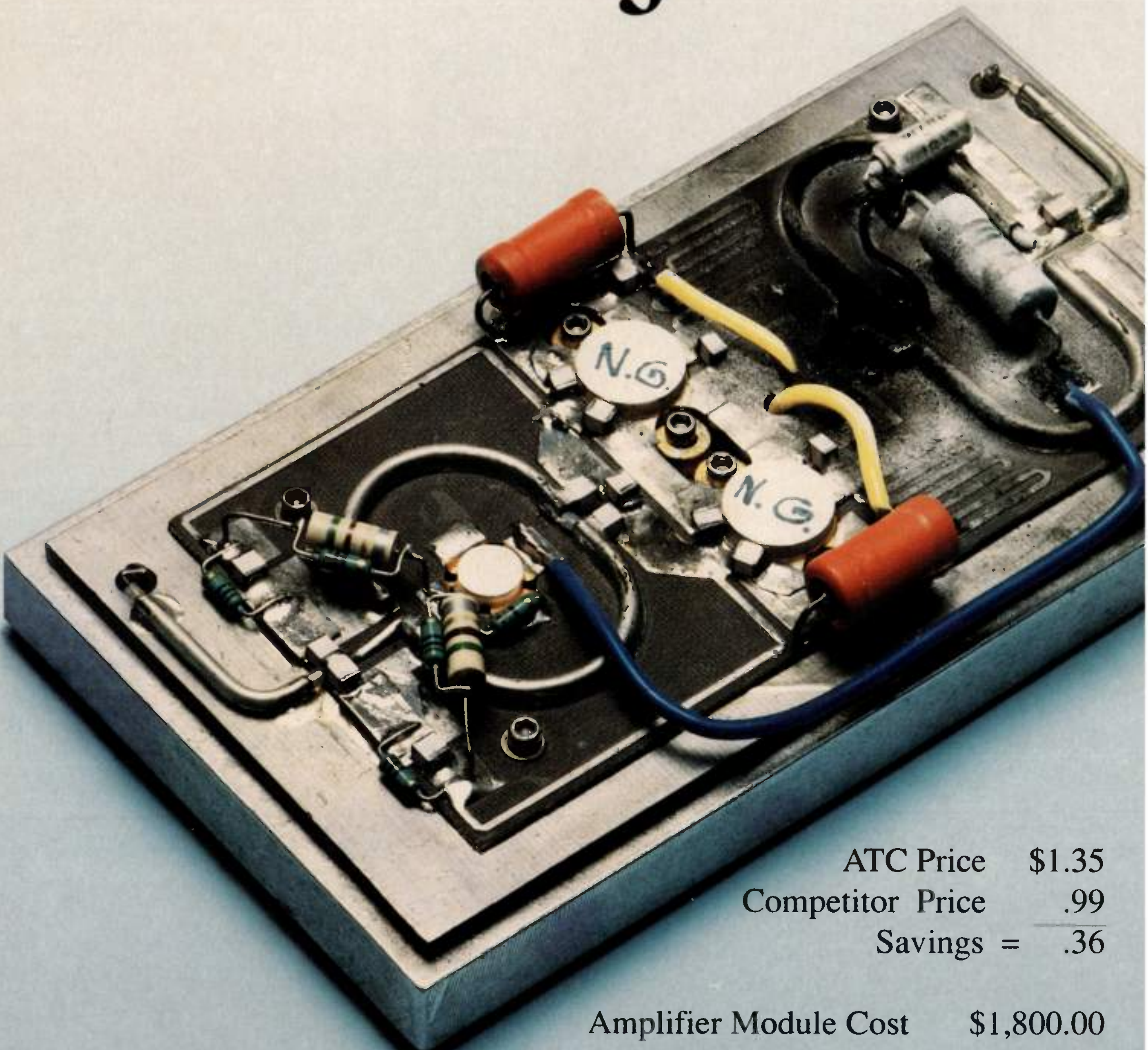

ATC Price	$1.35
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Savings =	.36

Amplifier Module Cost	$1,800.00
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Minus Savings —	.36
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FCC Asked to Begin New FM Technical Proceeding — The FCC has been asked to initiate a new technical proceeding to establish realistic definitions for FM directional antennas. The NAB urged regulators to examine certain technical assumptions about FM signal characteristics to help the FCC accurately determine FM radio station service and interference.

TriQuint Awarded MIMIC Phase 3 Contract — Triquint Semiconductor Inc. has been awarded a $2.65 million MIMIC Phase 3 contract for the volume production of MIMIC chips through a commercial foundry environment. General Electric's Electronics Laboratory is the subcontractor and will design the multi-chip MIMIC function, while TriQuint will fabricate and test the devices using a 0.5 micron depletion mode MESFET process with through wafer vias fabricated on 100mm wafers. More than 300 wafers will be fabricated and tested for this program.

Qualcomm Signs CDMA Agreements — Nokia Mobile Phones and Qualcomm recently announced that they have signed an agreement that will enable Nokia to develop and market mobile phones using Code Division Multiple Access (CDMA) technology. Nokia is also involved in developing equipment for the Pan-European digital network due to be operational next year. Qualcomm's OmniTRACS® was recently honored by the editors of Popular Science as one of 1990's outstanding products and technological achievements in the field of computers and electronics.

In addition, Qualcomm and Motorola, Inc.'s Radio-Telephone Systems Group have signed an agreement to jointly explore and develop products using CDMA technology. Motorola and Qualcomm plan to produce cell site and mobile cellular telephone equipment in sufficient quantity to validate CDMA's capabilities.

CAL Announces Name Change — Canadian Astronautics Limited has announced the change of their name to CAL Corporation as well as the opening of a new subsidiary office in the United States, CAL Systems Corporation.

AEL Awarded $2.7 Million Contract — AEL Defense Corporation was awarded a $2.7 million contract modification by the U.S. Army Aviation Systems Command. The contract is for six RU-21H Guardrail V avionics upgrade kits with an option for 13 additional kits. The upgrade modernizes communication, navigation and flight instrument subsystems, including cockpit lighting for Army secure lighting.

Lucas Aerospace Awarded Naval Contract — Lucas Aul has been awarded a $62 million contract by the Naval Sea Systems Command for the production, integration and support of the AN/SRQ-4 Radio Terminal Set. The AN/SRQ-4 is a shipboard receiver/transmitter used for communication with the SH-60BF Seahawk helicopter as part of the Navy's Light Airborne Multi-purpose System program.

IWL Communications Acquires Controlling Interest in Spacelink Systems — IWL Communications recently acquired controlling interest in Spacelink Systems. IWL designs, assembles, installs and services a variety of radio systems for offshore/inland and emergency communications. Spacelink Systems designs, engineers, install and maintains private satellite-based networks supporting offshore/inland operations for major oil and gas companies worldwide.

Satellite Communications Center Installed at African Mining Site — Halco Mining Inc. has completed installation of a complete multi-channel satellite communications center at their mining site located in the Republic of Guinea, West Africa. The INMARSAT system provides telephone, facsimile, data and telex connections to virtually any location in the world. The system was developed by Magnavox Advanced Products & Systems Company. It consists of two functionally separate operator's stations, connected to a single large dish antenna, which is capable of supporting multiple satellite links simultaneously. Calls to and from the site are routed through an INMARSAT geostationary satellite to a Coast Earth Station, where they are patched into international public switched telephone/telex/data links.

Hewlett-Packard to Supply Portable Spectrum Analyzers to U.S. and Royal Navies — Hewlett-Packard Company announced that it has been awarded contracts to furnish portable

INFO/CARD 17

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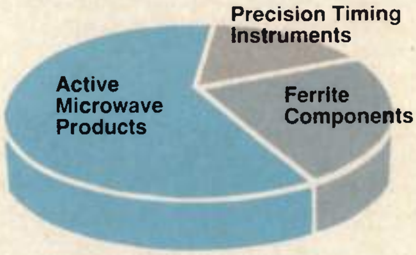

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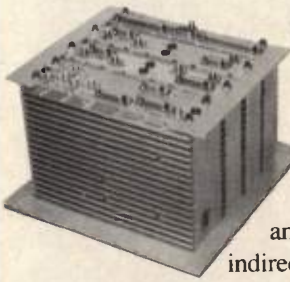

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spectrum analyzers to the U.S. Navy and the Royal Navy of the United Kingdom. The U.S. Navy Award, worth up to $8.1 million over the next three years, will provide HP 8560A-HO1 spectrum analyzers through its General Purpose Electronic Test Equipment program. The Royal Navy award also involves the purchase of a large number of analyzers over the next three years. This award will supply HP 8560A-002 spectrum analyzers with built-in tracking generators through the Common Range Electrical Test Equipment program in the U.K.

Varian To Sell TVT Broadcast Transmitter Business — Varian Associates recently announced that they have signed a letter of intent to sell their TVT broadcast transmitter business to Harris Corporation. Terms of the transaction which is subject to the negotiation of a definitive agreement, were not disclosed.

Plessey Electronic Systems Corporation Changes Name — As a result of last year's acquisition of its parent Plessey, now owned by the General Electric Company plc of the U.K., Plessey Electronic Systems Corporation has changed its name to GEC - Marconi Electronic Systems Corporation.

Cat Wire and Cable Opens New Plant — Cat Wire and Cable Corporation recently opened a new manufacturing plant for production of electronic wire and cable. The 15,000 square foot facility contains equipment for extrusion, cabling and shielding of cables up to 1.5 inches in diameter. Also in the new plant is a complete quality control and test lab.

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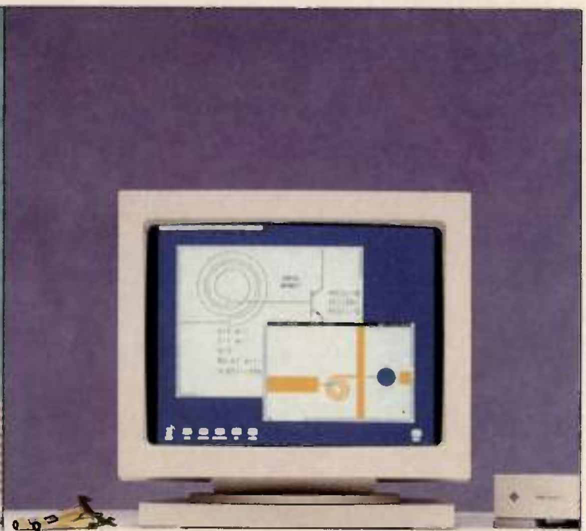

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The Rental Market - Its Success Reflects Hard Times

By Liane Pomfret
Assistant Editor

The rental equipment market has many things working in its favor right now and it shows. Companies report that business is up for all types of rental equipment, including RF equipment. With the current recession there isn't the cash around to buy equipment, so companies are renting instead. The tax incentives of a few years ago no longer exist, adding to the reasons to rent. In addition, the quick evolution of test equipment technology diminishes the cost effectiveness of buying new equipment. Companies are also looking into buying used equipment from rental companies' stock, for example, instead of new.

These days it's logical for companies to look at renting equipment instead of buying because of the recent downturn in the economy. As Ray Van Orden, Vice President and General Manager of Lectronic Research Labs points out, "We're probably running contra-cyclical to the rest of the economy." Business is up for rental companies. "With the slowdown, a lot of people are turning to rentals. They don't have the cash to spend on $15-45,000 equipment, especially when they're doing one shot testing." says Robert Otto, Vice President of Test Equipment Corporation. Companies are having difficulty getting banks to finance new equipment purchases so they are looking into other alternatives. Renting or leasing the necessary equipment is an excellent alternative and can be extremely cost effective.

There have been changes in the types of companies who are renting. It used to be that the small, ten-person companies were the majority of those who rented. Now, the bigger companies are realizing that they can take advantage of renting. "We sell to some of the largest companies in the nation as well as ten-person companies." says Van Orden. Walt Kosmowski, Marketing Manager at Electro Rent Corporation notes that his company follows the same trend, "Our customer base tends to run the gamut from Fortune 100 to Fortune 50,000." He goes on to speculate as to why companies are doing this, "With the present economic and business climate, capital expenditures are being looked at very carefully. Renting and leasing equipment becomes a viable alternative for acquisition of equipment."

For some companies, the decision to rent is based on their method of tax computation (e.g., the alternative minimum tax). Companies who use this method cannot always take accelerated depreciation on new equipment. However, they may be able to deduct their full rental or lease payments under the same plan. Much like the elimination of the investment tax credit a few years ago, this is one more factor in the decision to rent or lease instead of buying outright.

Rental or leasing of equipment has other advantages as well. Rick Worth, Product Manager at Leasametric, offers one suggestion, "Rental fees can be passed on to the customer as line item expenses, whereas buying the equipment is a cost to the company." Companies with this policy would incur virtually no cost for use of the equipment, because the customer would be paying for it. In addition, companies who only need equipment for one particular project would lose money if they were to buy. Jim Keady, Vice President of the Instrument Division at Continental Resources comments, "Basically, if a company is in research and development, they only use the instrument for one particular project so they will rent it for 3, 6, 9 months or however long they need it.

With the rapid pace of technology, state-of-the-art equipment today might be obsolete or outdated by next year. If a company leases on a year-to-year or month-to-month basis, they can always have state-of-the-art equipment. Rick Worth points out, "Ten years ago, a five to seven year depreciation life for an instrument was normal, now it is about three or four years, so if you must have state-of-the-art equipment, it can be less expensive to rent." Many companies come out ahead when they rent or lease. But this depends on factors such as initial purchase price, usage, maintenance, depreciation and eventual replacement costs.

When it comes to market areas doing the most renting, there are two areas which stand out from the crowd. With the upcoming changes in EMC regulations in Europe and changes occurring in the United States, such as Part 15, equipment for testing EMC/RFI has shown some increase in the rental area. "There's a lot of demand for RFI/EMI test equipment, that's where the trend is," says Robert Otto. But, good markets vary from company to company. For other rental companies, cellular test equipment is showing an increase, reflecting the rapid growth in the cellular communication market. Rick Worth notes, "Communication service monitors are very successful for cellular testing and repair."

Used equipment is another alternative for companies who cannot afford new equipment. The used equipment is often from rental or leasing companies and has been kept in relatively good shape. This equipment tends to run from 30 to 70 percent off the new equipment price. Very often, companies who sell used equipment will offer their own warranty in addition to any manufacturer's warranty which may still be in effect.

The rental and used equipment market is doing well despite the problems with the economy. Of the companies interviewed, the ones who indicated an increase in business also noted either a change in marketing strategy or the economy as prime reasons for their success. One company, who experienced a 40 percent increase in business over the past year, acknowledged that this increase was because of a shift to a more customer oriented marketing strategy. In many cases, renting is a sound proposition, both financially and logistically and it makes sense from a convenience standpoint as well. **RF**

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INFO/CARD 22

A Simple Two-Channel Receiver for 8-PSK

By Glen A. Myers, Naval Postgraduate School and
Daniel C. Bukofzer, California State Univ., Fresno

M-PSK is a bandwidth efficient method of transmitting digital data (1). Coherent demodulation is required in order to recover the transmitted symbols. We present here the design, operation, and performance features of a simple two-channel receiver which, for the 8-PSK case, recovers the bits rather than the symbols.

In M-PSK, one of M possible symbols is transmitted by selecting one of M phases of a sinusoidal carrier. Usually, the symbols are a collection of $k = \log_2 M$ bits. Of interest here is the case M = 8 where 3 bits constitute a symbol.

The assignment of symbols to signal phases can be pictorially represented by the signal space diagram of Figure 1. In the demodulation process, the most common error is mistaking a correct symbol with its nearest neighbor. Consequently, in order to reduce the probability of bit error, adjacent symbols are chosen so they differ by only one bit (Gray-Code mapping). One such example is shown in Figure 1.

Two measures of receiver performance are commonly used, namely the symbol error rate (SER) and the bit error rate (BER) as functions of signal-to-noise ratio (SNR). Analyses of the SER for the coherent detection of M-PSK signals have appeared in the literature over the past several years (2). Since the symbols are created for the benefit of the communications engineer, and are usually transparent to the channel user, the focus from a customer point of view is the BER.

Recently, Lee (3) has obtained a closed-form expression for the BER of M-PSK when Gray-Code mapping is used. (Similar results were independently derived in (4).) This analysis assumes coherent signal detection and an additive white Gaussian noise (AWGN) interference model.

Coherent demodulation of 8-PSK implies in principle the use of 8 signal processing channels in the receiver as shown in Figure 2. (While optimum receivers for 8-PSK utilizing fewer signal processing channels exist (5), in all cases the recovered symbols must be mapped to the corresponding set of bits. The receiver to be proposed is not only optimum but also does not require any symbol-to-bit mapping.) This paper presents and explains the operation of a simple coherent receiver which in a sense is an extension of the bit recovery scheme used in 4-PSK. An exact noise analysis shows the BER performance of this simple receiver to be identical to that of the 8-channel coherent system. This is a rather remarkable result in light of the simple implementation being proposed. While it appears that similar receiver methodologies have been proposed and used in the past, the analytical results involving noise performance and the effects of a variable signal phase are new. Furthermore, extensions to 16-PSK signaling schemes are possible using this methodology (6), which is called the "Direct Bit Detection" method.

Other features of this two-channel receiver whose operation is described in the next section are as follows:

(1) Each bit of the 3 bit symbol is recovered separately. Thus, the receiver also functions as a de-interleaver in those situations where the composite bit stream is created by

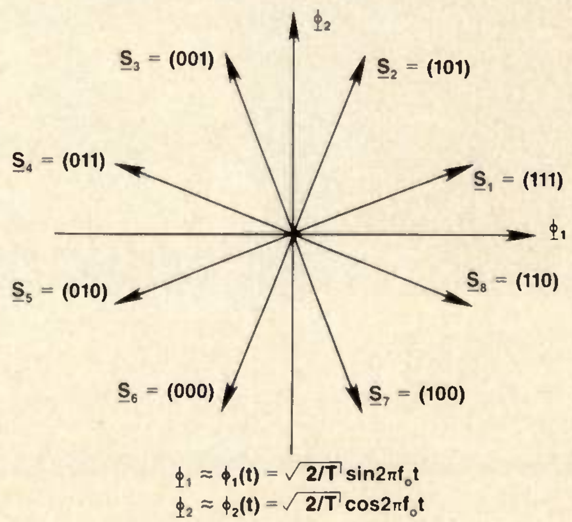

Figure 1. Signal space diagram showing gray level coding of symbols.

interleaving two or three individual bit streams.

(2) Each bit of a symbol is recovered using voltage comparisons with a zero threshold. This means that receiver performance remains robust during fading signal conditions, and the need for the transmission of training sequences and for the adjustment of AGC circuitry (as required, for example, in quadrature amplitude modulation schemes) is eliminated.

(3) The effect of varying the phase angle alpha in Figure 1 on BER for each bit of a symbol is directly obtained when the bit recovery is accomplished with the two-channel receiver described next.

The Two-Channel Receiver

Design of the two-channel receiver whose structure is shown in Figure 3 proceeds from observations concerning the signal space diagram of Figure 1. First, the bits in any symbol are identified as the left bit (LB), middle bit (MB) and right bit (RB). For example, symbol S_5, (010), of Figure 1 has LB = 0, MB = 1 and RB = 0.

Now, consider the RB. All symbols for which the RB = 1 have a projection on the positive Y-axis (labeled the cosine component in Figure 1). All symbols for which the RB = 0 have a projection on the negative Y axis. This means that in the absence of any noise or interference, if the received signal having one of 8 possible phase angles is correlated with a coherent cosine reference, the resulting constant value of this operation will be greater than 0 when the transmitted symbol has RB = 1 and will be less than 0 when RB = 0. A diagram of the receiver which recovers the RB (as well as the MB and LB)

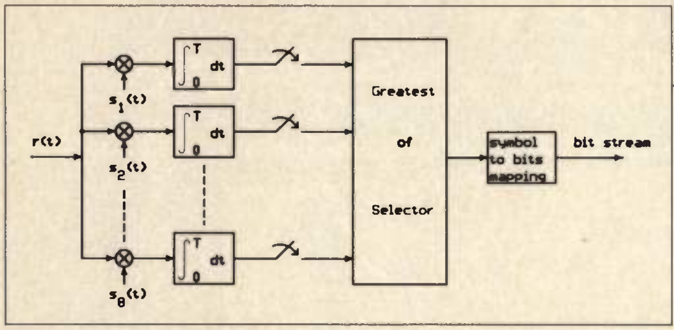

Figure 2. Block diagram of one form of an optimum receiver.

is shown in Figure 3.

Similar considerations apply to recovery of the LB where the projections of interest are now on the X-axis. Correlating the received signal with a coherent sine reference results in a constant positive voltage value whenever the transmitted symbol has LB = 1 and a negative value for LB = 0 in the absence of any noise or interference. Considering the MB, observation of Figure 1 reveals that the symbols for which MB = 1 have projections on the X-axis that are large in an absolute value sense while the corresponding projections on the Y-axis are small in an absolute value sense. The exact opposite result holds true for those symbols for which MB = 0.

The above observations are useful in obtaining a scheme for the recovery of the middle bit by using squaring operations and the axes projections available from the recovery of the RB and LB. Squaring has the advantage of further separating large values from small values. Thus, by squaring the outputs of the two correlators and taking the difference as shown in

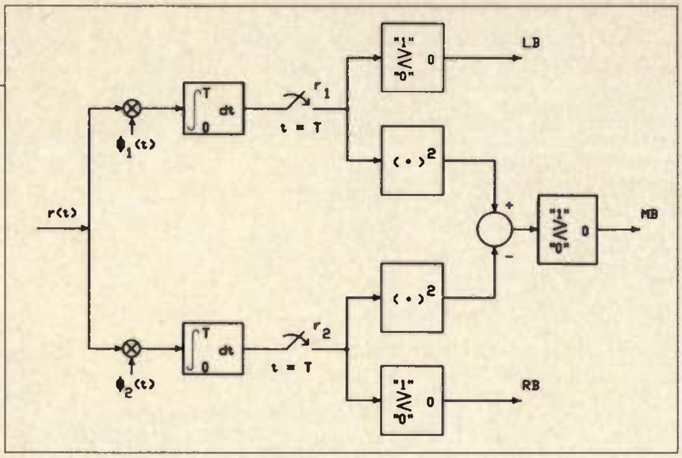

Figure 3. Block diagram of the two-channel receiver for direct bit detection.

Figure 3, a bipolar output is obtained which identifies the MB. In the absence of noise or interference, for symbols with a middle bit which is 1, a positive output results while the converse is true for symbols with a middle bit which is 0. It is apparent that the bit recovery mechanisms create appropriate decision boundaries for demodulation of an 8-PSK signal. As a result, receiver performance can be expected to be optimal; however, simply establishing the existence of these regions does not develop insight gained from further analysis.

Having described the noise-free performance of the proposed receiver, the corresponding noise performance is now derived by developing a mathematical expression for the BER assuming an AWGN interference model.

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Noise Analysis

The possible transmitted signals, which are represented as vectors in Figure 1, are mathematically expressed as

$$\begin{aligned} s_i(t) &= \sqrt{\frac{2E_s}{T}}\sin[2\pi f_0 t + \theta_i(t)] \\ &= \phi_1(t)\sqrt{E_s}\cos\theta_i(t) + \phi_2(t)\sqrt{E_s}\sin\theta_i(t) \end{aligned} \qquad (1)$$

for $0 \leqslant t \leqslant T$, $i = 1,2,\ldots,8$

where (2a)

$$\phi_1(t) = \sqrt{\frac{2}{T}}\sin 2\pi f_0 t \quad ; \quad \phi_2(t) = \sqrt{\frac{2}{T}}\cos 2\pi f_0 t \quad \text{for } 0 \leqslant t \leqslant T$$

T = symbol duration

f_0 = carrier frequency

$$E_s = \int_0^T s_i^2(t)dt = \text{energy per symbol} \qquad (2b)$$

Because in practice the product $f_0T >> 1$ so that the integral of the second harmonic term is negligible, $\phi_1(t)$ and $\phi_2(t)$ form a complete orthonormal set of functions on the interval $(0,T)$. Furthermore,

$$\theta_i(t) = \begin{cases} \alpha + (i-1)\pi/4 & i = 1,3,5,7 \\ (i\pi/4) - \alpha & i = 2,4,6,8 \end{cases} \qquad 0 \leqslant t \leqslant T \qquad (3)$$

where the angle α is a variable parameter shown in Figure 1.

Depending on which symbol is transmitted, because of the AWGN interference model, the received signal $r(t)$ is given by

$$r(t) = s_i(t) + n(t), \qquad 0 \leqslant t \leqslant T \qquad (4)$$

where $n(t)$ is a sample function of the AWGN having a constant power spectral density function equal to $N_0/2$.

Assume first that the symbol $s_1(t)$ (corresponding to LB = MB = RB = 1) is transmitted. As can be seen from Figure 3,

$$(5)$$
$$\Pr\{\text{LB correct}/s_1(t) \text{ transmitted}\} = \Pr\{r_1 > 0/s_1(t) \text{ transmitted}\}$$

$$(6)$$
$$\Pr\{\text{MB correct}/s_1(t) \text{ transmitted}\} = \Pr\{r_1^2 - r_2^2 > 0/s_1(t) \text{ transmitted}\}$$

$$(7)$$
$$\Pr\{\text{RB correct}/s_1(t) \text{ transmitted}\} = \Pr\{r_2 > 0/s_1(t) \text{ transmitted}\}$$

where r_1 is the component of $r(t)$ along $\phi_1(t)$ and r_2 is the component of $r(t)$ along $\phi_2(t)$. From equations 1 - 4 above, when $s_1(t)$ is transmitted

$$\begin{aligned} r_1 &= \int_0^T \left[\phi_1(t)\sqrt{E_s}\cos\theta_1(t) + \phi_2(t)\sqrt{E_s}\sin\theta_1(t) + n(t)\right]\phi_1(t)dt \\ &= \sqrt{E_s}\cos\alpha + n_1 \end{aligned} \qquad (8)$$

where

$$n_1 \triangleq \int_0^T n(t)\phi_1(t)dt \qquad (9)$$

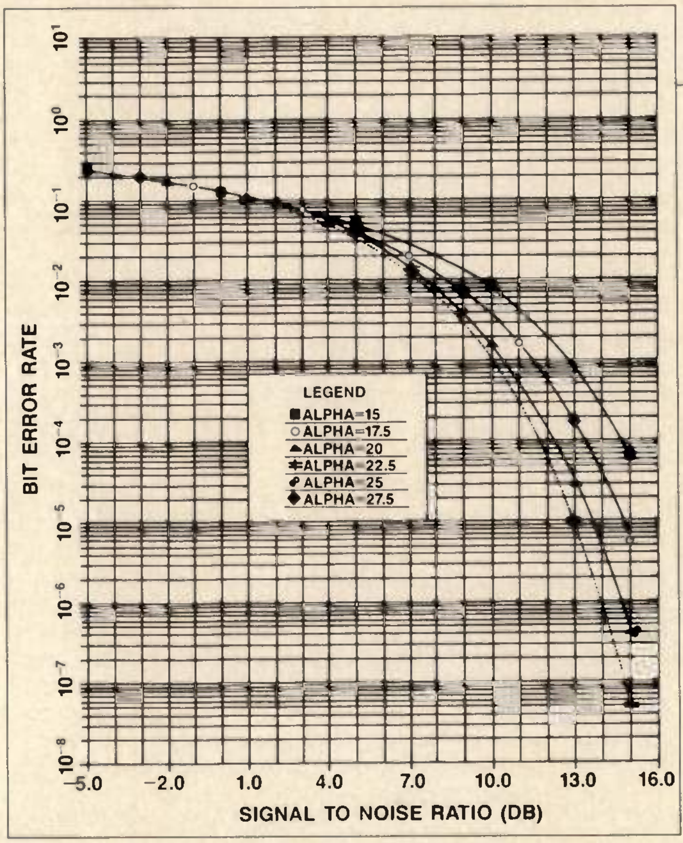

Figure 4. BER versus SNR for various values of phase angle α.

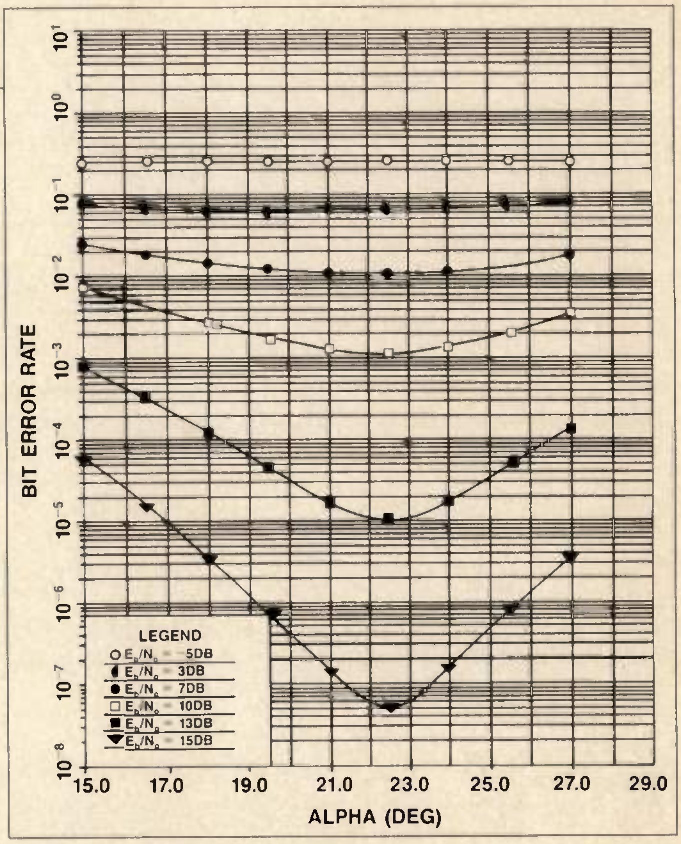

Figure 5. BER plotted as a function of α for fixed values of E_b/N_0.

and

$$r_2 = \int_0^T \left[\phi_1(t) \sqrt{E_s}\cos\theta_1(t) + \phi_2(t) \sqrt{E_s}\sin\theta_1(t) + n(t) \right] \phi_2(t)dt \tag{10}$$

$$= \sqrt{E_s}\sin\alpha + n_2$$

where

$$n_2 \triangleq \int_0^T n(t)\phi_2(t)dt \tag{11}$$

Observe that n_1 and n_2 are zero mean Gaussian random variables having identical variance σ_n^2, where (see (7))

$$\sigma_n^2 = E\{n_i^2\} = E\left\{\left[\int_0^T n(t)\phi_i(t)dt\right]^2\right\} = \int_0^T\int_0^T E\{n(t)n(\tau)\}\phi_i(t)\phi_i(\tau)dtd\tau \tag{12}$$

$$= \frac{N_0}{2} \qquad i = 1,2$$

and furthermore

$$E\{n_1n_2\} = E\left\{\int_0^T n(t)\phi_1(t)dt \int_0^T n(\tau)\phi_2(\tau)d\tau\right\} \tag{13}$$

$$= \int_0^T\int_0^T E\{n(t)n(\tau)\}\phi_1(t)\phi_2(\tau)dtd\tau = 0$$

which means that n_1 and n_2 are uncorrelated. Because n_1 and n_2 are Gaussian, they are also statistically independent random variables. Thus, from equations 5 and 8, when $s_1(t)$ is transmitted

$$Pr\{LB\ correct\} = Pr\{n_1 > -\sqrt{E_s}\cos\alpha\} \tag{14a}$$

$$= \int_{-\sqrt{E_s}\cos\alpha}^{\infty} \frac{1}{\sqrt{2\pi\sigma_n^2}} e^{-N_1^2/2\sigma_n^2} dN_1 = Q\left(-\sqrt{\frac{E_s}{N_0/2}}\cos\alpha\right)$$

where

$$Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}} e^{-u^2/2}du \tag{14b}$$

Similarly, from equations 7 and 10, when $s_1(t)$ is transmitted

$$Pr\{RB\ correct\} = Pr\{n_2 > -\sqrt{E_s}\sin\alpha\} = Q\left(-\sqrt{\frac{E_s}{N_0/2}}\sin\alpha\right) \tag{15}$$

In order to evaluate the conditional probability of the MB being correct, from equation 6

$$Pr\{r_1^2 - r_2^2 > 0/s_1(t)\ transmitted\} \tag{16}$$

$$= Pr\{(r_1 + r_2)(r_1 - r_2) > 0/s_1(t)\ transmitted\}$$

Using equations 8 and 10, define random variables y_1 and y_2 as

$$y_1 \triangleq r_1 + r_2 = \sqrt{E_s}\cos\alpha + n_1 + \sqrt{E_s}\sin\alpha + n_2 \tag{17}$$

$$y_2 \triangleq r_1 - r_2 = \sqrt{E_s}\cos\alpha + n_1 - \sqrt{E_s}\sin\alpha - n_2 \tag{18}$$

Observe that y_1 and y_2 are Gaussian random variable with means

$$E\{y_1\} = \sqrt{E_s}\cos\alpha + \sqrt{E_s}\sin\alpha \triangleq \bar{y}_1 \tag{19}$$

$$E\{y_2\} = \sqrt{E_s}\cos\alpha - \sqrt{E_s}\sin\alpha \triangleq \bar{y}_2 \tag{20}$$

and variances

$$var\{y_1\} = E\{(n_1 + n_2)^2\} = E\{n_1^2\} + E\{n_2^2\} + 2E\{n_1n_2\} = 2\sigma_n^2 = N_0 \tag{21}$$

$$var\{y_2\} = E\{(n_1 - n_2)^2\} = E\{n_1^2\} + E\{n_2^2\} - 2E\{n_1n_2\} = 2\sigma_n^2 = N_0 \tag{22}$$

where equations 12 and 13 have been used to obtain the above variances. More importantly, however, observe that

$$E\{(y_1 - \bar{y}_1)(y_2 - \bar{y}_2)\} = E\{(n_1 + n_2)(n_1 - n_2)\}$$

$$= E\{n_1^2\} - E\{n_2^2\} = 0 \qquad (23)$$

so that indeed y_1 and y_2 are again statistically independent random variables. Thus, when $s_1(t)$ is transmitted, using equations 16, 17, and 18,

$$\Pr\{\text{MB correct}\} = \Pr\{y_1y_2 > 0\} = \Pr\{y_1 > 0 \text{ and } y_2 > 0\} + \Pr\{y_1 < 0 \text{ and } y_2 < 0\} = \Pr\{y_1 > 0\}\Pr\{y_2 > 0\} + \Pr\{y_1 < 0\}\Pr\{y_2 < 0\} \qquad (24)$$

where the last equality in equation 24 is due to the independence of y_1 and y_2.

From equations 19 and 21

$$\Pr\{y_1 > 0\} = \int_0^\infty \frac{1}{\sqrt{2\pi N_0}} e^{-(Y_1 - \bar{y}_1)^2/2N_0}\, dY_1 = Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right) \qquad (25)$$

and from equations 20 and 22

$$\Pr\{y_2 > 0\} = \int_0^\infty \frac{1}{\sqrt{2\pi N_0}} e^{-(Y_2 - \bar{y}_2)^2/2N_0}\, dY_2 = Q\left(-\frac{\bar{y}_2}{\sqrt{N_0}}\right) \qquad (26)$$

Since

$$\Pr\{y_i < 0\} = 1 - \Pr\{y_i > 0\} \qquad i = 1,2 \qquad (27)$$

then equations 24, 25, and 26 yield

$$\Pr\{\text{MB correct}/s_1(t)\text{ transmitted}\} = Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right)Q\left(-\frac{\bar{y}_2}{\sqrt{N_0}}\right) + \left[1 - Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right)\right]\left[1 - Q\left(-\frac{\bar{y}_2}{\sqrt{N_0}}\right)\right] \qquad (28)$$

A similar analysis for probability of correct reception involving LB, MB, and RB given that signals $s_4(t)$, $s_5(t)$ or $s_8(t)$ were transmitted must be carried out. It turns out (and careful analysis of Figure 1 further confirms) that the probabilities in question are identical to those given by equations 14, 15, and 28. It is however necessary to evaluate the previously mentioned probabilities for the case in which $s_2(t)$ is transmitted. Again from equations 1-4, when $s_2(t)$ is transmitted,

$$r_1 = \int_0^T [\phi_1(t)\sqrt{E_s}\cos\theta_2(t) + \phi_2(t)\sqrt{E_s}\sin\theta_2(t) + n(t)]\phi_1(t)dt$$

$$= \sqrt{E_s}\sin\alpha + n_1 \qquad (29)$$

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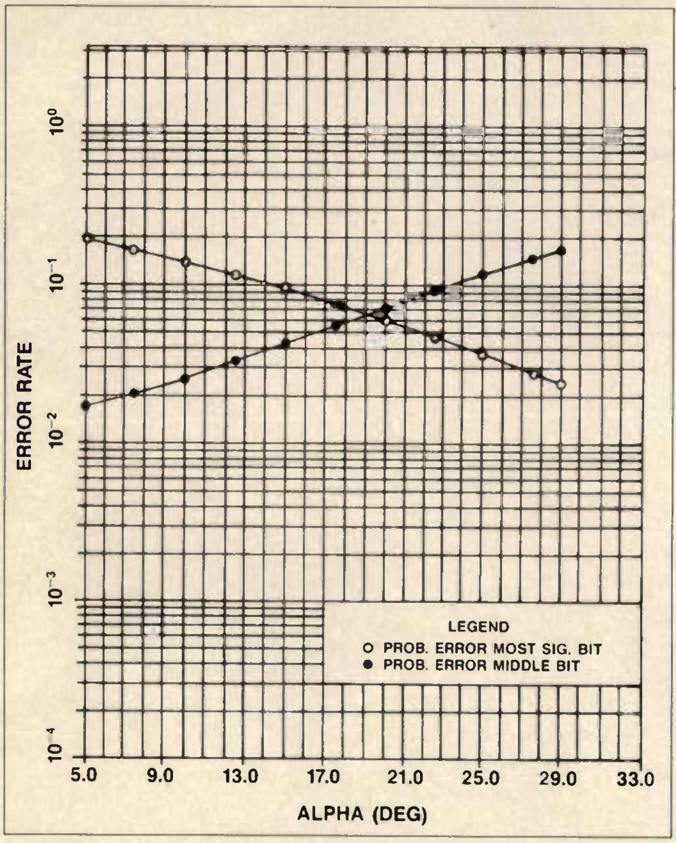

Figure 6. Probability of error for MB and LB as a function of phase angle α with E_b/N_0 = 3 dB.

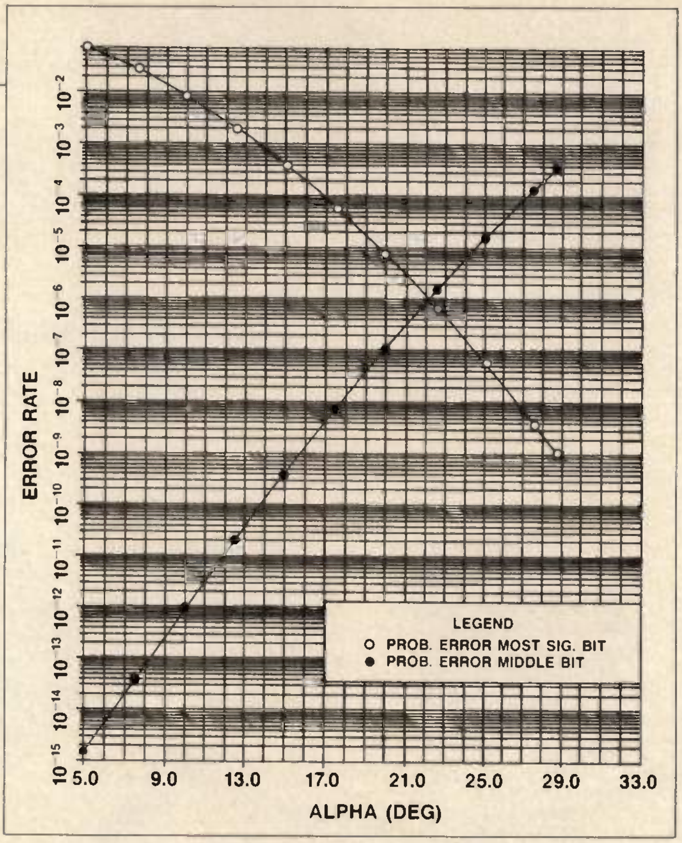

Figure 7. Probability of error for MB and LB as a function of phase angle α with E_b/N_0 = 14 dB.

where n_1 is defined by equation 9. Thus, when $s_2(t)$ is transmitted

$$\Pr\{\text{LB correct}\} = \Pr\{r_1 > 0\} = \Pr\{n_1 > -\sqrt{E_s}\sin\alpha\}$$

$$= Q\left(-\sqrt{\frac{E_s}{N_0/2}}\sin\alpha\right) \quad (30)$$

where equation 12 and the fact that n_1 is a zero mean Gaussian random variable have been used to obtain this expression. Also, assuming $s_2(t)$ is transmitted,

$$r_2 = \int_0^T [\phi_1(t)\sqrt{E_s}\cos\theta_2(t) + \phi_2(t)\sqrt{E_s}\sin\theta_2(t) + n(t)]\phi_2(t)dt \quad (31)$$

$$= \sqrt{E_s}\cos\alpha + n_2$$

where n_2 is defined by equation 11. Similarly, when $s_2(t)$ is transmitted

$$\Pr\{\text{RB correct}\} = \Pr\{r_2 > 0\} = Q\left(-\sqrt{\frac{E_s}{N_0/2}}\cos\alpha\right) \quad (32)$$

Finally, for the MB, when $s_2(t)$ is transmitted,

$$\Pr\{\text{MB correct}\} = \Pr\{r_1^2 - r_2^2 < 0\} = \Pr\{(r_1 + r_2)(r_2 - r_1) > 0\} \quad (33)$$

From equations 29 and 31, the sum $r_1 + r_2$ is given by y_1 of equation 17 while

$$r_2 - r_1 = \sqrt{E_s}\cos\alpha + n_2 - \sqrt{E_s}\sin\alpha - n_1 \triangleq y_3 \quad (34)$$

where

$$E\{y_3\} = \sqrt{E_s}\cos\alpha - \sqrt{E_s}\sin\alpha \triangleq \bar{y}_3 = \bar{y}_2 \quad (35)$$

and

$$\mathrm{var}\{y_3\} = E\{(n_1 - n_2)^2\} = \mathrm{var}\{y_2\} = N_0 \quad (36)$$

Observe that

$$E\{(y_1 - \bar{y}_1)(y_3 - \bar{y}_3)\} = E\{(n_1 + n_2)(n_2 - n_1)\} \quad (37)$$

$$= E\{n_2^2\} - E\{n_1^2\} = 0$$

so that y_1 and y_3 are also statistically independent random variables. Thus, if $s_2(t)$ is transmitted,

$$\Pr\{\text{MB correct}\} = \Pr\{y_1 y_3 > 0\} \quad (38)$$

$$= \Pr\{y_1 > 0\}\Pr\{y_3 > 0\} + \Pr\{y_1 < 0\}\Pr\{y_3 < 0\}$$

where the independence of y_1 and y_3 has been used to obtain equation 38. Since equation 25 specifies $\Pr\{y_1>0\}$, $\Pr\{y_3>0\}$ need only be evaluated, where from equations 35 and 36

$$\Pr\{y_3 > 0\} = \int_0^\infty \frac{1}{\sqrt{2\pi N_0}} e^{-(Y_3 - \bar{y}_3)^2/2N_0}\, dY_3 = Q\left(-\frac{\bar{y}_3}{\sqrt{N_0}}\right) \quad (39)$$

so that

$$\Pr\{\text{MB correct}/s_2(t)\text{ transmitted}\} =$$

$$Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right)Q\left(-\frac{\bar{y}_3}{\sqrt{N_0}}\right) \quad (40)$$

$$+\left[1 - Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right)\right]\left[1 - Q\left(-\frac{\bar{y}_3}{\sqrt{N_0}}\right)\right]$$

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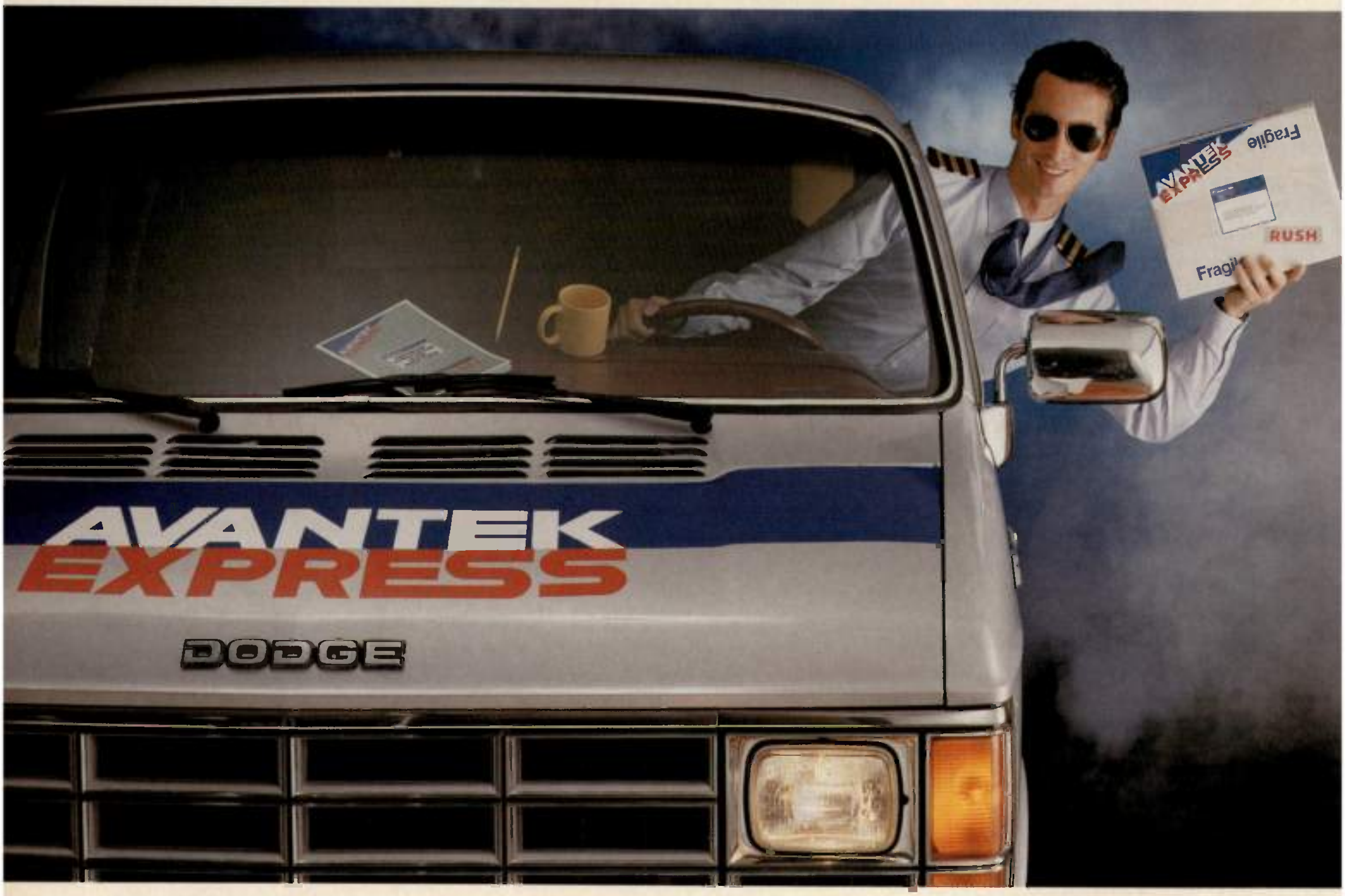

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E_b/N_0 (DB)	ALPHA (DEG)
−5.	3.2188
−4.	6.9688
−3.	10.0000
−2.	12.4688
−1.	14.4375
0.	16.0000
1.	17.2800
2.	18.2184
3.	19.0000
4.	19.6250
5.	20.1563
6.	20.5625
7.	20.9375
8.	21.2188
9.	21.4688
10.	21.6563
11.	21.8125
12.	21.9688
13.	22.0625
14.	22.1563
15.	22.2188
16.	22.2813
17.	22.3438

Table 1. Phase angles that equalize bit error ratios for different signal-to-noise ratios.

A similar analysis for probability of correct reception involving LB, MB, and RB given that signals $s_3(t)$, $s_6(t)$, or $s_7(t)$ were transmitted must be carried out. It turns out here also that the probabilities in question are identical to those given by equations 30, 32, 40.

The results thus obtained can now be used to obtain the receiver BER. Assuming all signals are equally likely to be transmitted,

$$\Pr\{\text{bit correct}\} = \frac{1}{8}\left[4\Pr\{\text{bit correct}/s_1(t)\ \text{transmitted}\} + 4\Pr\{\text{bit correct}/s_2(t)\ \text{transmitted}\}\right] \tag{41}$$

where the word "bit" in equation 41 stands for LB, MB, or RB. Thus using equations 14, 15, 28, 30, 32, and 40, yields

$$\Pr\{\text{LB correct}\} = \frac{1}{2} Q\left(-\sqrt{\frac{2E_s}{N_0}}\cos\alpha\right) + \frac{1}{2} Q\left(-\sqrt{\frac{2E_s}{N_0}}\sin\alpha\right) = \Pr\{\text{RB correct}\} \tag{42}$$

$$\Pr\{\text{MB correct}\} = 1 - Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right) - Q\left(-\frac{\bar{y}_2}{\sqrt{N_0}}\right) + 2Q\left(-\frac{\bar{y}_1}{\sqrt{N_0}}\right) Q\left(-\frac{\bar{y}_2}{\sqrt{N_0}}\right) \tag{43}$$

where $\bar{y}_1$ and $\bar{y}_2$ are given by equations 17 and 18 respectively and the fact that $\bar{y}_2$ and $\bar{y}_3$ are identical has been used to simplify equation 43.

In order to arrive at a final BER expression as a function of signal-to-noise ratio, assume a front-end filter (IF amplifier) is used, having a transfer function given by

$$H(f) = \begin{cases} 1 & f_0 - B < |f| < f_0 + B \\ 0 & \text{otherwise} \end{cases} \tag{44}$$

The noise power N at the output of this bandpass filter (BPF) is given by

$$N = \frac{N_0}{2}[4B] = 2N_0B \tag{45}$$

Assuming $B = T^{-1}$ Hz, then it can be argued that with this minimum bandwidth, the filter causes no distortion of the signal component of r(t) so the signal power S at the output of this filter is

$$S = \frac{E_s}{T} = E_sB \tag{46}$$

Thus the signal-to-noise ratio (SNR) becomes

$$SNR = \frac{S}{N} = \frac{E_sB}{2N_0B} = \frac{E_s}{2N_0} \tag{47}$$

The BER is usually expressed in terms of the energy, E_b, of each bit. Since each symbol contains three bits, then

$$E_s = 3E_b \rightarrow SNR = \frac{3E_b}{2N_0} \rightarrow \frac{E_b}{N_0} = \frac{2}{3}SNR \tag{48}$$

Since for any real x

$$Q(-x) = 1 - Q(x) \tag{49}$$

then from equation 42,

$$\Pr\{LB \text{ in error}\} = 1 - \frac{1}{2}\Big[1 - Q\left(\sqrt{4SNR}\cos\alpha\right) + 1 - Q\left(\sqrt{4SNR}\sin\alpha\right)\Big] = \frac{1}{2}Q\left(2\sqrt{SNR}\cos\alpha\right) + \frac{1}{2}Q\left(2\sqrt{SNR}\sin\alpha\right) = \Pr\{RB \text{ in error}\} \tag{50}$$

Similarly, from equations 19, 20, and 43

$$\begin{aligned}\Pr\{MB \text{ in error}\} &= 1 - [Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha)) \\ &- Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha)) - 1 + 2(1 - Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha))) \\ &(1 - Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha)))] = Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha)) \\ &+ Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha)) - 2Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha)) \\ &Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha))\end{aligned} \tag{51}$$

It is interesting to note that the individual bit error probabilities associated with the receiver of Figure 3 can be expressed solely in terms of complementary error functions.

The results of equations 50 and 51 are useful in various ways. First, they allow for computation of the BER directly as a function of SNR and α. The resulting probability, P_b, of bit error is simply

$$\begin{aligned}P_b &= \frac{1}{3}[\Pr\{LB \text{ in error}\} + \Pr\{MB \text{ in error}\} + \Pr\{LB \text{ in error}\}] \\ &= \frac{1}{3}[Q(2\sqrt{SNR}\cos\alpha) + Q(2\sqrt{SNR}\sin\alpha) \\ &+ Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha)) + Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha)) \\ &- 2Q(\sqrt{2SNR}(\cos\alpha + \sin\alpha))Q(\sqrt{2SNR}(\cos\alpha - \sin\alpha))]\end{aligned} \tag{52}$$

Clearly, equation 52 can be used to compute the BER as a function of SNR (or E_b/N_0) for different values of the phase angle. In fact, it is possible to numerically search for the phase angle α that minimizes P_b for a given value of SNR. Furthermore, equations 50 and 51 can be used to obtain the value of phase angle α that results in

$$\Pr\{LB \text{ in error}\} = \Pr\{MB \text{ in error}\} = \Pr\{RB \text{ in error}\}$$

for a given value of SNR. Obtaining such a solution may be important in cases in which the individual transmitted bits belong to three independent interleaved data streams transmitted via a single carrier. In such cases it may be important to equalize the error rates amongst all three data streams. It must be pointed out that equalization of the individual bit stream error rates can be accomplished without changes in the angle α (away from its optimum value), but with permutations in the bit streams so that a given bit stream is transmitted one-third of the time in the LB position, one-third of the time in the MB position and one-third of the time in the RB position. This, however, is done at the expense of additional hardware both at the transmitter and at the receiver.

Other decision mechanisms for the MB that set up the correct decision boundaries and therefore appear optimal, do not yield as compact and easily computable results. Consider, for example, the recovery of the MB by evaluating $|r_1| - |r_2|$ and comparing its measured value with zero. Using the decision rule

$$|r_1| - |r_2| > 0 \text{ decide } MB = 1 \tag{53a}$$

$$|r_1| - |r_2| < 0 \text{ decide } MB = 0 \tag{53b}$$

it can be seen that in the absence of noise, this decision logic is just as capable of properly recovering the MB. Moreover, any decision logic of the form

$$|r_1|^n - |r_2|^n > 0 \text{ decide } MB = 1 \tag{54a}$$

$$|r_1|^n - |r_2|^n < 0 \text{ decide } MB = 0 \tag{54b}$$

can also be used, with n any integer. However, not only are the corresponding hardware implementations more complex (for n = 3,4,5...), but also the error rate associated with the MB is now quite difficult to obtain analytically. The noise performance for the decision logic given by equation 53 (utilizing n = 1) has been obtained analytically. However, due to the many steps required to carry out the derivation, only the result is given here. That is

$$\begin{aligned}\Pr\{MB \text{ in error}\} &= \frac{1}{2} \\ &+ \frac{1}{2}\Big[\int_{-2\sqrt{SNR}\cos\alpha}^{\infty} Q\left(z + 2\sqrt{2SNR}\sin(\alpha + \pi/4)\right) g(z)dz \\ &+ \int_{2\sqrt{SNR}\cos\alpha}^{\infty} Q\left(z - 2\sqrt{2SNR}\cos(\alpha + \pi/4)\right) g(z)dz \\ &+ \int_{-2\sqrt{SNR}\cos\alpha}^{\infty} Q\left(z + 2\sqrt{2SNR}\cos(\alpha + \pi/4)\right) g(z)dz \\ &+ \int_{2\sqrt{SNR}\cos\alpha}^{\infty} Q\left(z - 2\sqrt{2SNR}\sin(\alpha + \pi/4)\right) g(z)dz\Big] \\ &- \frac{1}{2}\Big[\int_{-2\sqrt{SNR}\sin\alpha}^{\infty} Q\left(z + 2\sqrt{2SNR}\sin(\alpha + \pi/4)\right) g(z)dz\end{aligned} \tag{55a}$$

$$+\int_{2\sqrt{SNR}\sin\alpha}^{\infty} Q\left(z + 2\sqrt{2SNR}\cos(\alpha + \pi/4)\right) g(z)dz$$

$$+\int_{-2\sqrt{SNR}\sin\alpha}^{\infty} Q\left(z - 2\sqrt{2SNR}\cos(\alpha + \pi/4)\right) g(z)dz$$

$$+\int_{2\sqrt{SNR}\sin\alpha}^{\infty} Q\left(z - 2\sqrt{2SNR}\sin(\alpha + \pi/4)\right) g(z)dz\Bigg]$$

where

$$g(z) = \frac{1}{\sqrt{2\pi}} e^{-z^2/2} \tag{55b}$$

Since no known closed form for the integrals in equation 55 is available, the complete Pr{MB in error} expression was evaluated as a function of SNR with α = 22.5 degrees using double precision arithmetic on an IBM 3033, with no significant differences in the resulting values when compared to those obtained via equation 51. The focus of the remaining analysis and the conclusions to be drawn involve recovery of the MB via squaring operations as presented in the section describing the two-channel receiver. This is partly because integrated circuits that can accurately multiply two voltages are readily available, whereas absolute value circuits with comparable performance are not as readily available.

Numerical Results

Computer programs were written in order to evaluate and graphically display the various equations involving receiver performance. In Figure 4, P_b given by equation 52 has been plotted as a function of the ratio E_b/N_0 (see equation 48) for various values of the angle α. The figure demonstrates that a phase angle α of 22.5 degrees, is optimum in the sense of producing a minimum BER. The actual values of P_b were compared with those tabulated in (3) and complete agreement was found. This demonstrates that the receiver is optimum for the interference model used and that an equiphase distribution amongst the symbols is best when analyzing the overall BER. It must be noted however that changes in α have a greater effect on P_b at high values of E_b/N_0 than at low values of E_b/N_0. This is demonstrated by the graphs of Figure 5. Here, P_b has been plotted as a function of α for fixed values of E_b/N_0. At high values of E_b/N_0, there is a distinct minimum in P_b at α = 22.5 degrees, whereas at very low values of E_b/N_0 (where P_b can be expected to be high), P_b is almost insensitive to relative large variations in α. Even though α = 22.5 degrees is still optimum, a different value of α could be used for low values of E_b/N_0 without significantly affecting P_b.

Finally, it must be noted that as α changes, a trade-off between Pr{LB in error}, Pr{RB in error} and Pr{MB in error} exists. For small values of α, the first two probabilities are high while the third probability is low. The converse is true when α is large. It must further be understood that these three probabilities are not identical when α = 22.5 degrees. In fact, the value of α at which these three probabilities are identical is dependent on the E_b/N_0 value. Recalling that Pr{LB in error} is always identical to Pr{RB in error}, in Figures 6 and 7, Pr{LB in error} has been plotted on the same set of axes along with Pr{MB in error} as a function of E_b/N_0. In Figure 6 a representatively low value of E_b/N_0 was used. These figures demonstrate the "cross-over" effect on these probabilities and the fact that as E_b/N_0 increases, a larger value of α is necessary to equalize the individual LB, MB, and RB error rates. Table I displays the α values (obtained via a numerical search) that result in equal LB, MB, and RB error rates for a range of E_b/N_0 values. This is important in light of the fact that if the input data stream is the result of interleaving three unrelated (lower rate) data streams, the receiver should deliver data in which each individual data stream has the same error rate. Table I clearly demonstrates that α = 22.5 degrees (the most commonly used value in practice) would not result in delivered data to individual users with equalized error rates.

Conclusions

The proposed two-channel receiver for 8-PSK modulated signals achieves optimum performance in a minimum BER sense when the channel interference can be modeled as additive white Gaussian noise. The receiver structure is attractive due to its simplicity, its lack of need for any variable threshold settings in the decision circuitry, and the ease with which interleaved data streams from separate sources are delivered directly to the intended users. The selection of the phase angle α depends on the intended application. For minimum overall BER, α should be set at 22.5 degrees. However for applications involving interleaving data streams that must be delivered to users with equal error rates, the necessary phase angle α depends on E_b/N_0. However, the value of α that equalizes the individual data error rates is not a strong function of E_b/N_0.

Acknowledgements

The authors would like to thank Gregory Thompson, Major, United States Marine Corps, who built and measured the noise performance of a modified version of the two-channel receiver presented here. *RF*

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AU-2A-0150	1-500	30	0.5	1.25	1.4	1.5	2:1	+8	55
AU-3A-0150	1-500	45	0.5	1.25	1.4	1.5	2:1	+10	67
AU-1149	10-180	15	0.5	4.5	5.5	6.5	2:1 *	+18	75
AU-1291	.015-500	60	0.75	1.25	1.4	1.5	2:1	+8	90
AM-1300	.01-1000	25	0.75	1.4	1.6	1.8	2:1	+6	50
AM-1052	1-1000	26	0.75	1.4	1.6	1.8	2:1	+6	50
AM-2A-000110	1-1000	27	0.75	1.4	1.6	1.8	2:1	+8	50
AM-1299	1-1000	38	0.75	1.4	1.6	1.8	2:1	+9	75
AM-3A-000110	1-1000	37	0.75	1.4	1.6	1.8	2:1	+9	75

* 75 ohm impedance level

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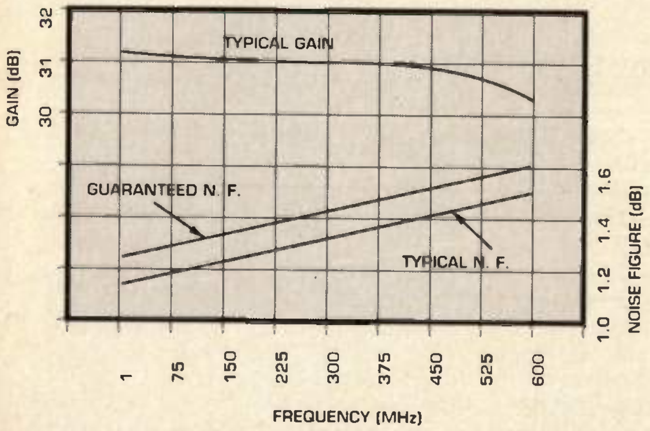

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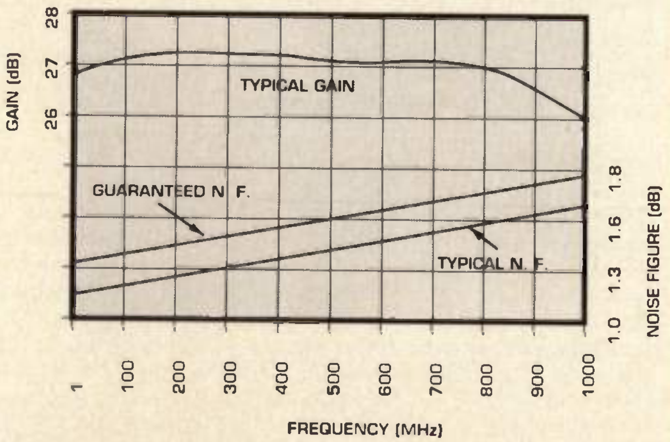

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Susceptibility of π/4 DQPSK TDMA Channel to Receiver Impairments

By K. Anvari and D. Woo
NovAtel Communications Ltd.

The ultimate objective of communications is to enable anyone to communicate instantly with anyone else from anywhere. This can be achieved most readily by cellular mobile radio communication.

The advent of cellular mobile radio (CMR) in recent years has brought access to the public switched telephone network (PSTN) into both commercial and private vehicles. Unlike conventional mobile radio, its main advantage is that it vastly increases the capacity of the system. This is effected by dividing a subscribed area spatially into cells, and by frequency and space division multiplexing those cells. The demand for such systems has already outstripped the existing capacity in some areas of the world and is predicted to grow very rapidly in the near future. To overcome this problem, a number of parameters such as increasing the bandwidth efficiency, reducing cell size, and using smaller cluster size have been investigated. This has spurred the development of a new digital cellular standard that will increase the spectrum efficiency of cellular radios to allow more users having the same frequency spectrum. The Telecommunications Industry Association (TIA) has commissioned a group to define this next generation cellular radio system. Known as the TR 45.3 committee, this group of industry representatives has issued specifications defining the new digital cellular system. The new system requires radios to conform to the old analog specification, as well as the new digital system.

Digital radio signals are susceptible to impairments from both equipment and the channel. The former includes linear distortion, non-linear distortion (caused by amplifiers, mixers and limiters), synchronization errors (both carrier and timing errors), subsystem imperfections (filters, oscillators) and echo (caused by mismatches), while the latter includes interference (adjacent channel and co-channels) and multipath fading.

A time and frequency domain simulation package has been developed which augments analytical modeling and hardware simulation activities and provides a flexible and useful tool for isolating sources of channel impairments while evaluating tradeoffs, and making overall performance predictions. Typical facilities allow the channel to be modeled from a block diagram. Simulated signals are passed sequentially through the elements of the channel model, and performance is then evaluated. The advantage of this simulation is its flexibility and the insight into channel performance, provided both by the modeling process itself and by the experience gained from the simulation experiments.

This paper investigates the susceptibility of the digital channel to impairments from different components of the receiver for the North American π/4 DQPSK modulation scheme. In the course of this investigation different demodulation schemes have been considered. These schemes are limiter discriminator demodulation, IF differential detection, and baseband differential detection.

Receiver Architecture

Current receiver development in mobile radio is characterized by an ever increasing complexity in frequency range and modulation characteristics. At the same time, there is a need to reduce the size and weight (bearing in mind battery requirements). The solution to this latter problem is large scale integration, but to date the radio receiver has been resistant to this approach, except for some easily integrated subunits. High frequency, large dynamic range and frequency selectivity are

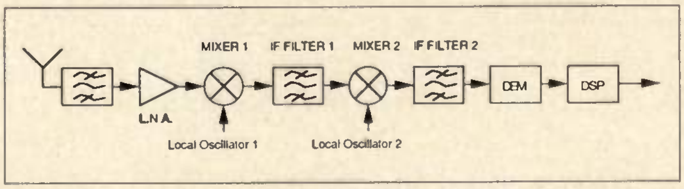


Figure 1. Block diagram of double conversion receiver.

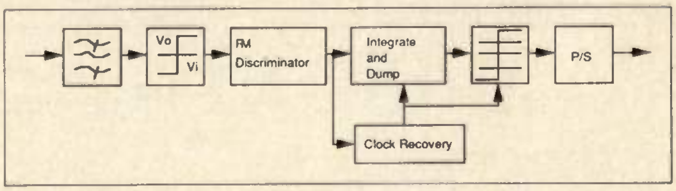


Figure 2. Block diagram of limiter discriminator detector.

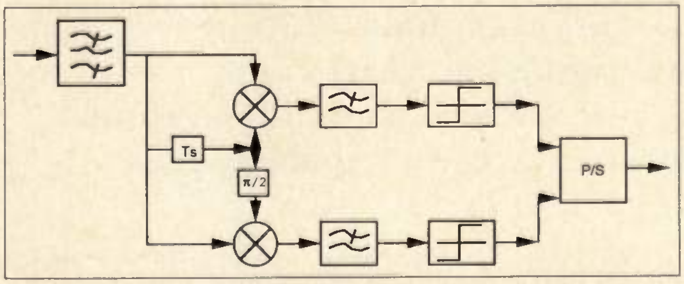


Figure 3. Block diagram of IF differential detector.

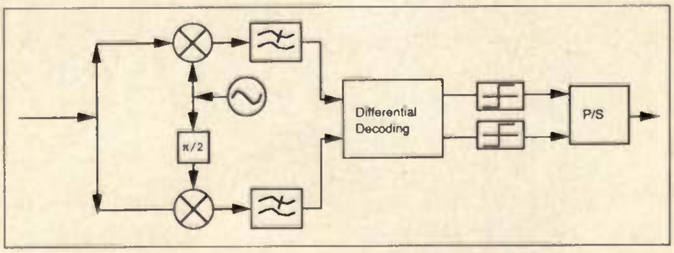


Figure 4. Block diagram of baseband differential detector.

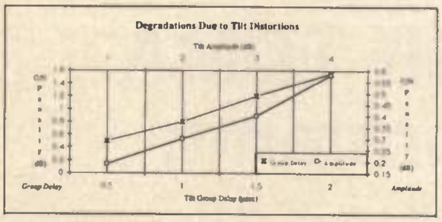


Figure 5. Carrier to noise penalties versus linear distortion.

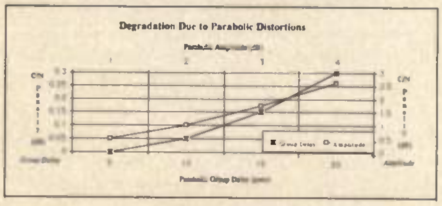


Figure 6. Carrier to noise penalties versus quadratic distortions.

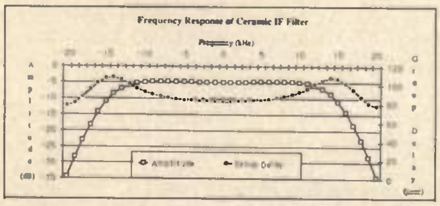


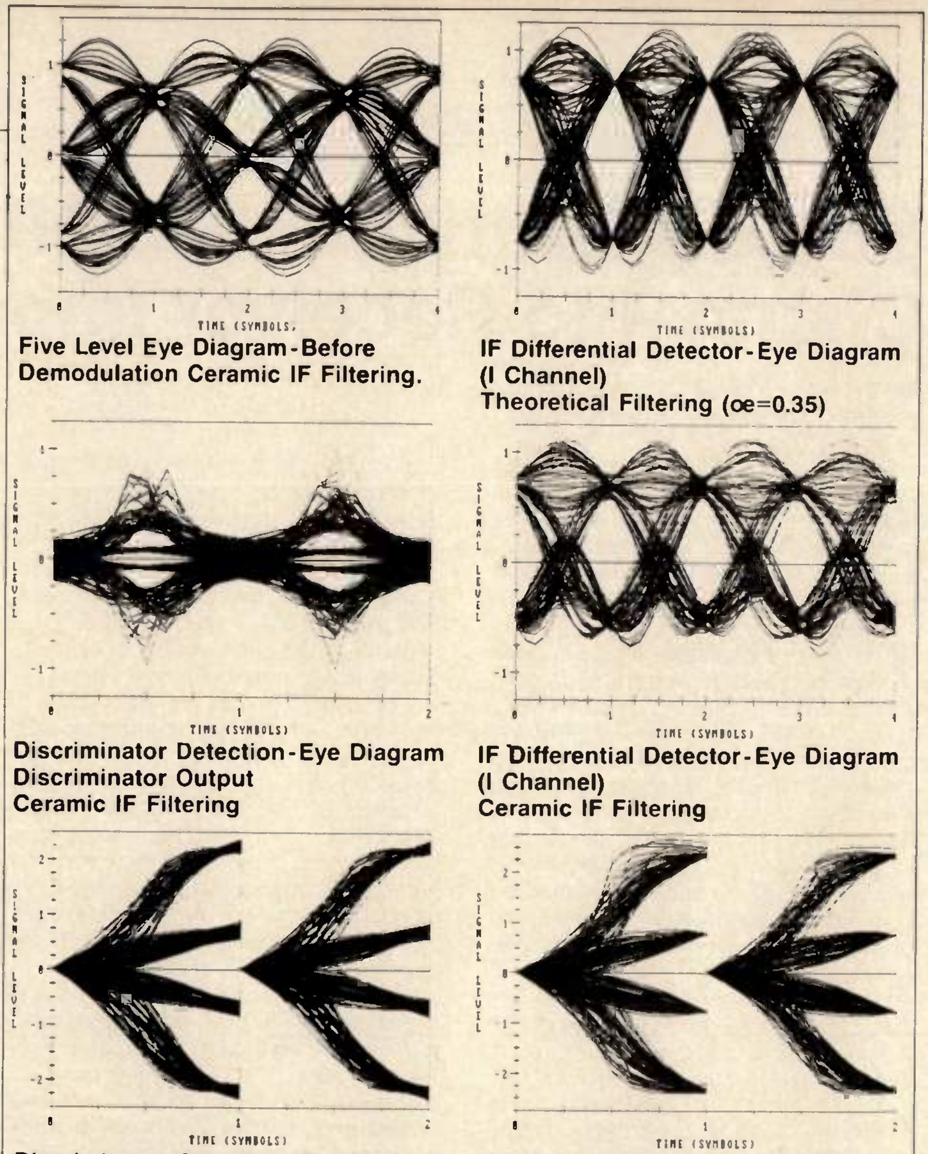
Figure 7. Frequency response of ceramic IF filter.

fundamental difficulties.

There are four receiver architectures which can be used for mobile receivers. They are:

a) double conversion receiver
b) single conversion receiver
c) direct conversion receiver
d) direct digital radio frequency receiver

A double conversion receiver is the



Five Level Eye Diagram - Before Demodulation Ceramic IF Filtering.

IF Differential Detector - Eye Diagram (I Channel) Theoretical Filtering (αe=0.35)

Discriminator Detection - Eye Diagram Discriminator Output Ceramic IF Filtering

IF Differential Detector - Eye Diagram (I Channel) Ceramic IF Filtering

Discriminator Detection - Eye Diagram Integrate and Dump Filter Output Ceramic IF Filtering

Discriminator Detection - Eye Diagram Integrate and Dump Filter Output Theoretical Filtering (αe=0.35)

Figure 8. Eye diagrams - limiter discriminator and IF differential detection.

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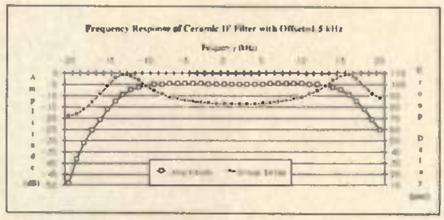

Figure 9. Ceramic IF filter amplitude and group delay response for offset 1.5 kHz.

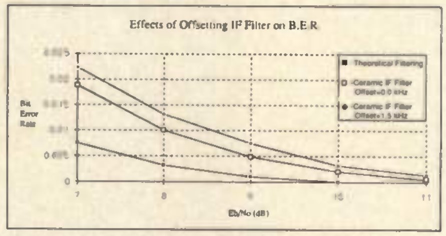

Figure 10. Bit error rate degradation for various IF filter offsets.

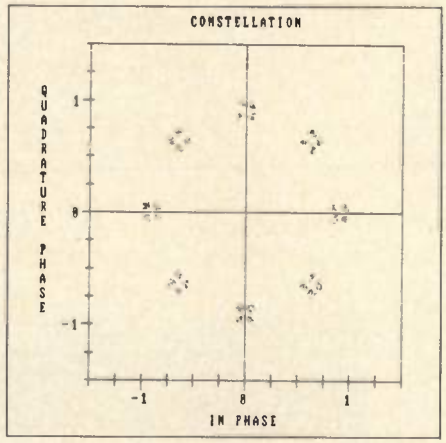

Figure 11. Constellation of channel with parabolic group delay.

architecture which is being used in the existing cellular mobile radios. Two of the most important characteristics of the North American cellular system are receiver selectivity and dynamic range requirements. With present filter technology, the double conversion receiver appears to be a suitable architecture. The block diagram of this receiver is shown in Figure 1.

The demodulator in this diagram can take any of the three configurations mentioned earlier. In the following sections these demodulation schemes are studied and the effect of imperfections which are fundamentally due to filters, oscillators and synchronization errors are investigated. For those results which are bit rate dependent 46.8 kbit/s has been assumed.

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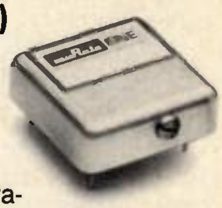

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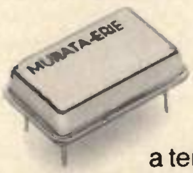

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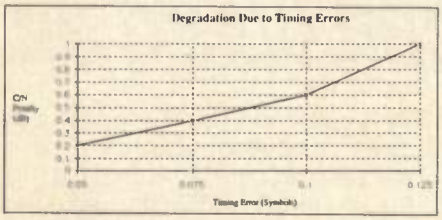

Figure 12. C/N penalties versus sampling timing errors.

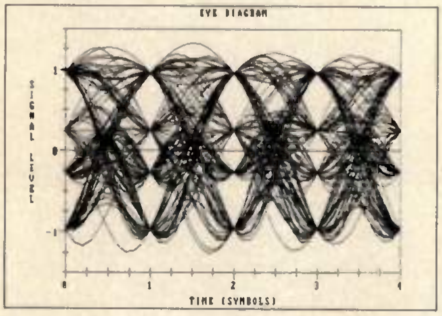

Figure 13. Two level eye diagram baseband differential detection 2.0 kHz carrier frequency error.

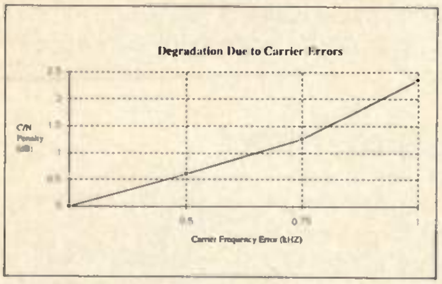

Figure 14. C/N penalties versus carrier errors.

IF Differential/Limiter Discriminator

Figures 2 and 3 show the block diagram of an IF differential and a limiter discriminator detector. These two schemes have two main features. First, they do not require any carrier recovery circuit, and second, the total IF filter frequency response when combined with the transmitter base band filter should provide the Nyquist raised cosine spectral shaping. Among the impairment factors mentioned in the introduction, the linear distortion is the deviation of the overall 6 dB bandwidth from the Nyquist bandwidth. There are a wide variety of deviations from designed channel characteristics that are observed in practice (eg: ripples, slopes etc.) in both amplitude and group delay. To a first order filter, the departure from the ideal of both the amplitude response and the group delay response can be described entirely in terms of linear and quadratic variations. Linear amplitude distortion is defined as the decibel difference of the amplitudes at the two edges of the Nyquist bandwidth when the amplitude response of the channel is linear with frequency. When the amplitude response (in dB) of the channel (in the Nyquist bandwidth) is a parabolic function of frequency the distortion is quadratic. Within the Nyquist bandwidth when the delay response of the channel is linear, the resultant distortion is also linear; however, when the delay response is parabolic the distortion is quadratic.

Figures 5 and 6 show carrier to noise penalties caused by linear and quadratic amplitude and delay distortions at an error rate of 10^{-2}. It can be seen that quadratic amplitude distortion causes more significant penalties than linear amplitude distortion. The opposite tendency is observed for delay distortion.

For an overall Nyquist channel, the total amplitude response of the IF filters has to be a square root of a raised cosine with x/sinx pulse shaping. The group delay response of the IF filters has to be flat at least across the 6 dB bandwidth of the channel. Analog implementation of such a filter is extremely difficult and the alternative is either digital implementation or baseband implementation by down and up-conversion. The choice of implementation technique is a debatable subject and is not in the scope of this paper. In our investigation, the commercially available and easily implementable ceramic filters have been considered. Figure 7 illustrates the frequency response of a realizable ceramic filter which provides optimum inband performance and adjacent channel rejection. The first IF filter is assumed to be wide enough not to contribute to channel shaping and narrow enough to reject unwanted signals and image frequencies.

Figure 8 shows the eye diagram for an ideal and a realizable channel using these two detectors. It can be seen that due to imperfection some eye closure will be experienced.

Although these two detection schemes are almost insensitive to small carrier frequency changes, the offset of IF filters due to these frequency errors causes performance degradation as well as reduction of the adjacent channel protection factor. The filter center frequency offset can also result from temperature changes. Figure 9 shows the IF amplitude and group delay response for 1.5 kHz of carrier error. The bit error rate degradation due to the carrier error is shown in Figure 10.

In calculations to this point, the sampling time has been optimized for minimum eye closure. Errors in the timing phase cause regenerator decisions to be made at instants other than the optimum ones, at which intersymbol interference is minimum. Figure 11 shows the constellation of a channel with parabolic group delay distortion. It can be seen that at each state of the constellation there are four points which are the result of intersymbol interference. An optimum sampling time is required to minimize the performance degradation. For an ideal channel with a timing error, a similar constellation will result. Figure 12 shows the carrier-to-noise penalties versus sampling time error.

Baseband Differential

The fundamental difference between this demodulation scheme and the ones which were discussed in the previous section is that the detection is performed at baseband (Figure 4). This feature allows all the channel shaping to be done at baseband. In order to maintain the linearity throughout the channel, all amplifiers and mixers have to operate in their linear region. The transmitter portion of the channel can be kept linear by adaptive pre-distortion. Although the receiver deals with small signals, overloading of the mixers, amplifiers etc. can occur when received adjacent and alternate channel signal levels are significantly higher than the designed level. Therefore the IF filters have to provide sufficient rejection for adjacent and alternate channel interferences. Although these filters do not contribute to channel shaping they are narrow enough to provide the distortions we have already discussed.

The local oscillator for this demodulator has to be locked to the incoming carrier in order to avoid any consequent phase error after the differential detector. This requires a carrier recover circuit which can be complex and sensitive to multi-path fading. Figure 13 shows the two level eye diagrams of this detector for 2 kHz of frequency error. In Figure 14 the carrier to noise penalties against the carrier error are shown. This figure only considers the resultant phase error and does not take into account any distortion resulting from filter offset due to carrier error. Up to 1 kHz carrier error can be tolerated without significant

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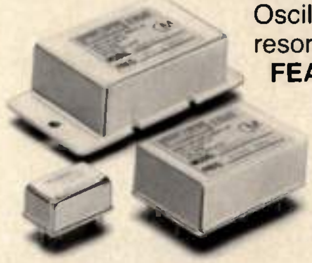

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FEATURES

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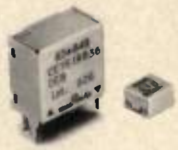

These dielectric resonator devices are used as interstage matching devices to absorb antenna mismatching and maintain stable transmission.

FEATURES

■ Power Ratings from 2 to 10 Watts ■ 15 dB min. Isolation ■ Less than 1 dB Insertion Loss ■ Sizes Ranging From 6.9 mm × 6.8 mm × 4 mm to 15 mm × 15 mm × 9 mm ■ Surface Mount Packages Available

Applications include Cellular Mobile Telephone (CMT), Microwave Test Equipment, etc.

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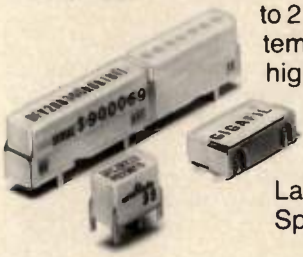

These microwave filters are based on Dielectric Resonator technology and cover the frequency range from 400 MHz to 2 GHz. They offer extremely good temperature stability (± 5 ppm/°C), high selectivity and low insertion loss. Applications include CMT, GPS, Data Transmission, Navigation Systems, Cordless Telephones, Land Mobile Radio (LMR) and Spread Spectrum Systems.

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performance degradation.

Conclusion

This paper has described three types of demodulation schemes - IF differential, baseband differential and a limiter of discriminator detection. We have shown that simulation of the digital channel for prediction of performance and effect of subsystems impairments on channel performance is essential.

The susceptibility of a $\pi/4$-DQPSK digital channel to various receiver impairments was discussed. It was shown that receivers with IF differential and discriminator detection schemes require an IF match filter before demodulator. We also discussed the baseband differential detection technique and showed the effect of carrier offset on its performance. ***RF***

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Practical Performance Prediction Techniques for Spectrally Efficient Digital Systems

By Dr. Andy Kucar and Dr. Kamilio Feher
Bell Northern Research, Ltd., and University of California, Davis

The performance of any coherent digital modulation system is degraded by an excessive amount of phase noise. In this contribution, the term phase noise is a synonym for random interferences such as phase noise of the frequency sources (local oscillators, up/down converters, voltage controlled oscillators), etc. In the past, a considerable effort has been devoted by various authors to this subject (1-11). However, neither an exact mathematical solution nor practical approximate solutions for higher than 4-ary modulation schemes have been published. In this correspondence we adopt a practical approach, which yields useful performance curves in which the degradation due to phase noise can readily be seen.

In the next section, the phase noise characterization is given. After which the results for probability of error performance of MPSK, MQAM and MQPR modulation systems in the presence of thermal and phase noise are presented.

Phase Noise Characterization

The expressions for the performance of coherent digital modulation systems in a Gaussian channel are well known (1). However, in practice we rarely have perfect knowledge of the local phase reference. The output signals from carrier and symbol timing recovery circuits are really random processes and produce errors in the estimation of phase and time, respectively. These random processes consist of two components: first contributed by thermal white Gaussian noise, and second contributed by all other random sources which is termed phase noise. The effects on the error probability performance due to this phase noise can be obtained in principle. However, this is a formidable task with rather complex solutions existing (2-4) for two- and four-state modulation systems only.

We compute the degradation of a system in the presence of phase noise in the following manner. Considering the sources of phase noise as independent random variables, the central-limit theorem says that, under certain general conditions, the resultant equivalent phase noise probability density function approaches a normal Gaussian curve as the number of sources increase (12). Since the number of individual sources tends to be large in practice, and their magnitudes are of a comparable order, the assumptions of the central limit theorem should apply. This is the case of a well designed system. However, in a particular example, an individual source might dictate the overall system performance and an overall system might not be Gaussian, which might result in an increased degradation of the performance. The phase noise is combined with assumed white Gaussian noise channel to produce a total carrier-to-noise ratio $(C/N)_T$ given by

$$(C/N)_T = [(N/C) + (N/C)_p]^{-1} \qquad (1a)$$

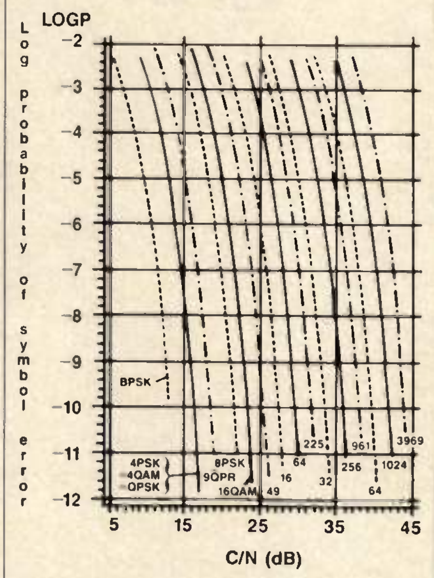


Figure 1. The average probability of error performance.

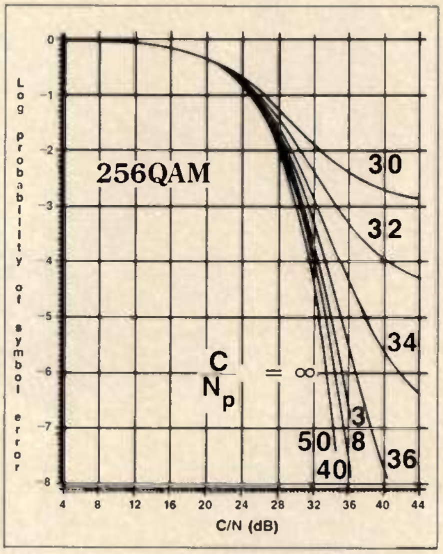


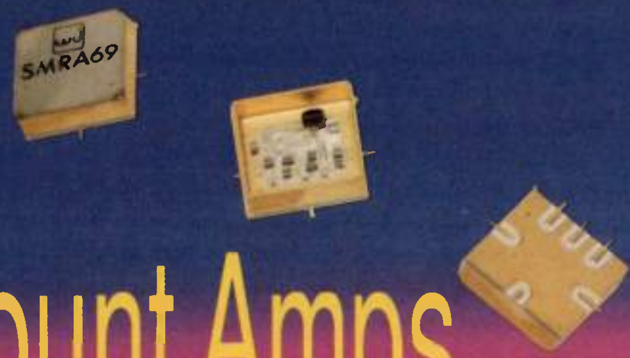
Figure 2. The 256QAM average probability of error performance .

$$(N/C)_p = (N/C)_1 + (N/C)_2 + \dots + (N/C)_n \qquad (1b)$$

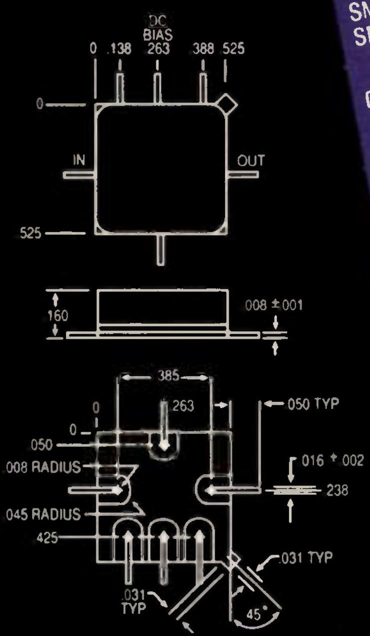
where (N/C) is the thermal noise-to-carrier ratio, $(N/C)_i$ (i=1,...,n) is the noise-to-carrier ratio of the i-th random source and $(N/C)_p$ is the equivalent phase noise-to-carrier ratio. We use known expressions (7) to calculate performance of coherent digital modulation systems in the presence of phase noise by replacing (C/N) by $(C/N)_T$ and having $(C/N)_p$ as a variable parameter.

Because of a simplistic phase noise model, the accuracy of results might depend on how close the assumed Gaussian probability density function fits the probability density function of the real phase noise. However, in practice we are dealing with the small amount of phase noise which causes degradations of less than 1 dB at $P_e = 10^{-6}$. In that case, and for our purpose, the Gaussian probability density function seems to be a good approximation for a real but unknown distribution of a phase noise.

High Gain Surface Mount Amps



Model	Frequency Range	Small Signal Gain dB Min.	Gain Flatness ±dB Max.	Noise Figure dB Max.	Power Output @ 1 dB Compression dBm Min.	Intercept Point dBm Typ.	VSWR I/O Max.	DC Volts Nom.	DC mA Typ.
5 to 2000 MHz									
SMRA89	5-500	25.5	0.7	4.5	20.5	35	1.8	15	130
SMRA89-1	10-500	29.0	0.7	4.5	20.0	36	1.8	15	130
SMRA76	10-500	38.5	0.9	3.8	13.3	26	1.8	15	81
SMRA69	10-1000	24.0	0.7	5.0	20.0	34	2.0	15	130
SMRA66	10-1000	35.0	1.0	4.5	14.0	30	1.8	15	81
SMRA26	10-1500	26.0	0.8	6.0	13.5	27	2.0	15	82
SMRA36	100-2000	23.0	0.9	6.5	12.0	22	1.8	15	76
SMRA38	200-1800	14.5	1.0	6.7	17.5	25	2.3	15	127
	200-2000	14.5	1.0	7.5	16.0	25	2.3	15	127
0.5 to 5 GHz									
SMRA43	1-4	19.5	0.9	5.3	10.5	22	2.0	5	115
SMRA46	1-4	24.0	0.8	5.2	17.0	30	2.0	12	175
SMRA53	1-5	19.5	0.9	5.3	10.5	22	2.1	5	115
2 to 6 GHz									
SMRA62	2-6	14.0	0.7	5.0	12.0	28	2.0	5	65
SMRA63	2-6	16.5	1.0	7.0	10.0	25	2.2	5	120
SMRA63-1	2-6	16.5	1.0	7.0	13.0	32	2.2	5	120



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Our results of the measurement presented in the next section are in close agreement with previous assumptions.

Performance Evaluation

The probability of error performance of Binary PSK (BPSK = 2PSK), MPSK (M > 2), MQAM and MQPR modulation schemes is given by the following expressions (7)

$$P_B \approx \frac{1}{2}\operatorname{erfc}(\sqrt{\gamma_b}) \qquad \text{(BPSK) (2)}$$

$$P_M \approx \operatorname{erfc}\left(\sqrt{k\gamma_b}\sin\frac{\pi}{M}\right) \qquad \text{(3) (MPSK, M > 2)}$$

$$P_L \approx \left(1-\frac{1}{L}\right)\operatorname{erfc}\left(\sqrt{\frac{3}{M-1}\,\frac{1}{2}\,\gamma_{av}}\right) \qquad \text{(MQAM) (4a)}$$

$$P_L \approx \left(1-\frac{1}{M}\right)\operatorname{erfc}\left(\sqrt{\frac{3}{M-1}\,\frac{1}{2}\left(\frac{\pi}{4}\right)^2\gamma_{av}}\right) \qquad \text{(MQPR) (5a)}$$

$$P_M = 2P_L\left(1-\frac{1}{2}P_L\right) \qquad \text{(4b, 5b) MQAM and MQPR}$$

where (8)

$$\gamma_b = \frac{E_b}{N_o} = \frac{C}{N}\,\frac{B}{f_b} \qquad (6)$$

where P_B is the probability of error performance of BPSK, P_M is the corresponding symbol error rate for the MPSK (M > 2), MQAM and MQPR systems, while P_L corresponds to the probability of error of the baseband signal in each of the two quadrature components of QAM or QPR. γ_{av} is the average signal-to-noise ratio per k-bit symbol, i.e. the average energy per symbol-to-noise ratio, where $k = \log_2 L$, L is the number of levels and $M = L \times L$. γ_b is the energy per bit-to-noise ratio (C/N) is the carrier-to-thermal noise ratio, f_b is the bit rate bit/s, and B is the double-sided noise bandwidth in Hz.

We assume B is equal to the double-sided Nyquist bandwidth. Equations 2-5 give the probability of error performance curves illustrated in Figure 1. Inclusion of $(C/N)_p$ as a variable parameter produces a series of curves, one of which

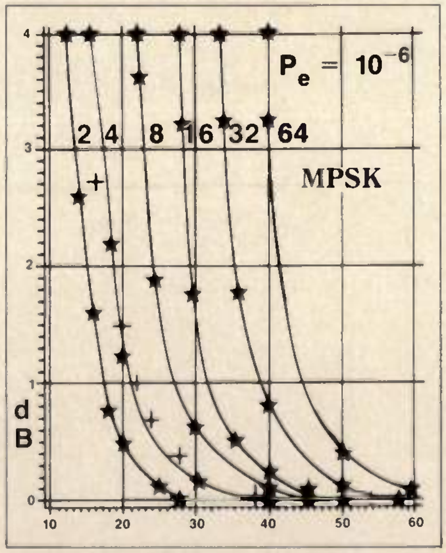

Figure 3. The degradation of the MPSK systems in dB.

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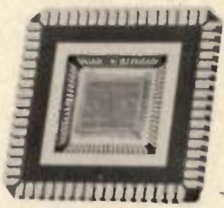

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(256 QAM) is shown for illustration in Figure 2. The degradations $(C/N)_T-(C/N)$ for $P_e = 10^{-6}$, dBr, versus the $(C/N)_p$ are summarized in Figures 3-5. Figure 3 applies for the MPSK systems, the MQAM systems are shown in Figure 4 and the MQPR systems in Figure 5.

In the BPSK system, Figure 3, a $(C/N)_p > 20$ dB causes the degradation to be less than 0.5 dB. If $(C/N)_p > 30$ dB the degradation is negligible. This is an easily achievable goal except in very low speed systems, where the spectral purity of signal source(s) could be critical (4). However, the 256QAM system - Figure 2 and 4 - requires $(C/N)_p > 45$ dB to limit the phase noise caused by degradation to the amount below 1 dB. Additionally, the high-ary modulation schemes are more fragile in a multiple fading and interference environment. Further, the interference equalizer, carrier and symbol timing recovery and data generation and decision circuits are more complex for the case of high-ary modulation schemes, see (9). Consequently, the performance required of each component in the high-ary system should be higher. In all practical cases we find that if we take $(C/N)_p$ to be 10 dB higher than the $(C/N)_6$, the degradation is less than 1 dB (Figures 2-5). If $(C/N)_p > (C/N)_6 + 20$ dB the degradation is negligible. Results of measurements, Figures 3-4, which will be briefly assessed, are in close agreement with our engineering rule-of-thumb approximation.

In a low speed data system (e.g. single-channel-per-carrier SCPC satellite link) a phase noise of frequency sources might limit overall performance of the system. To verify this, we performed sets of measurements on SCPC satellite links consisting of two different 70 MHz 32 ksymbols/s 4-state modems (staggered QPSK and IJF), University of Ottawa earth station with 70 MHz/14 GHz up-converter and 12 GHz/70 MHz down-converter and two different "space segments" (Telesat Canada ANIK 14/12 GHZ satellite transponder and a laboratory transponder with a similar performance). Each link consisted of five local oscillators of a comparable quality, whose phase noises were measured by an automated spectrum analyzer. Phase noise power of each oscillator was integrated over the error transfer function (1-H(f)) of the modem's carrier recovery loop (Costas loop type with a

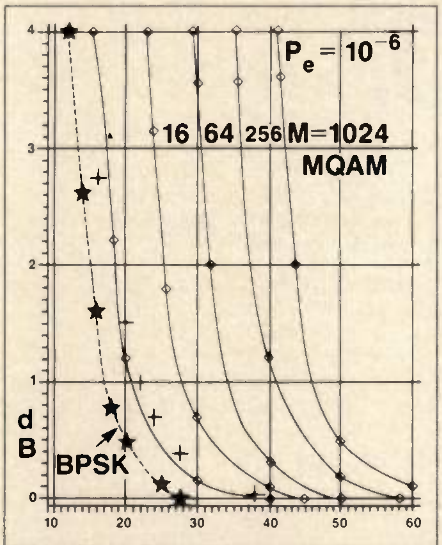

Figure 4. The degradation of the MQAM systems in dB.

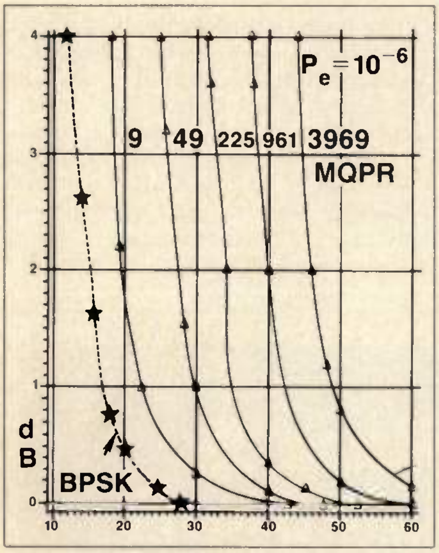

Figure 5. The degradation of the MQPR systems in dB.

second order passive filter). The average of multiple measurements gave a noise power within the 16 kHz Nyquist bandwidth, which after multiplying by 2 and dividing by a total carrier power gave an $(N/C)_i$ of a particular phase noise contributor. (More details on the phase noise measurements can be found in Reference 10). To get different values of $(N/C)_i$, the frequency stability of one of five local oscillators was intentionally degraded. Even with these unbalanced phase noise sources the measured results were in close agreement with previously assumed curves, Figures 3 and 4. Although the phase noise of a frequency source itself is nonwhite (it might be represented by $\Sigma c_i f^{-v_i}$, where c_i and v_i are appropriate constants), after it passes through a "whitening filter" (i.e. error transfer function of the loop, which might be approximated by the cf^v near the carrier where the noise contribution is the most important) it becomes near white, but not necessarily Gaussian, see Figure 6.

For example, the carrier phase noise introduced by a long haul (1000 km) nonregenerative FDM analog system, illustrated in the upper trace of Figure 6, may have a significant impact on the performance of the 256QAM and 1024QAM data-and-voice modems (8,9).

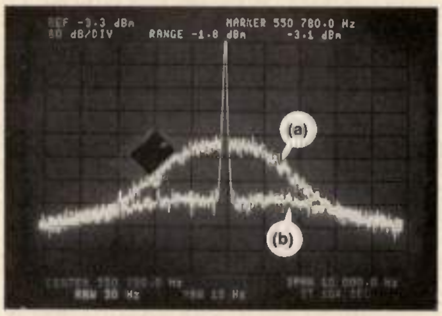

Figure 6. (a) The measured phase noise introduced by a long haul (1000 km) nonregenerative FDM analog system, (b) Canceled phase noise. This picture is a courtesy of Karkar Electronics Inc.

Careful design of frequency sources and every single component of these high-ary QAM schemes is necessary (11).

Conclusion

In this contribution, the performance of MPSK, MQAM and MQPR modulation systems in the presence of phase noise

and additive white Gaussian noise is presented in graphic forms. If a carrier-to-phase noise ratio is at least 10 dB higher than a carrier-to thermal noise ratio required for the $P_e = 10^{-6}$, the degradation due to phase noise will be less than 1 dB. If $(C/N)_p > (C/N)_6 + 20$ dB the degradation is negligible. The analysis is supported with results of measurements on 4-state SCPC satellite modems and terrestrial radio links. Based on previous results, a first order estimation of components requirements might be made. To minimize the degradation in the higher-ary modulation schemes, the rigorous selection of components and use of advanced technologies is necessary. ***RF***

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Antenna Products Follow The Action in Communications

By Gary A. Breed
Editor

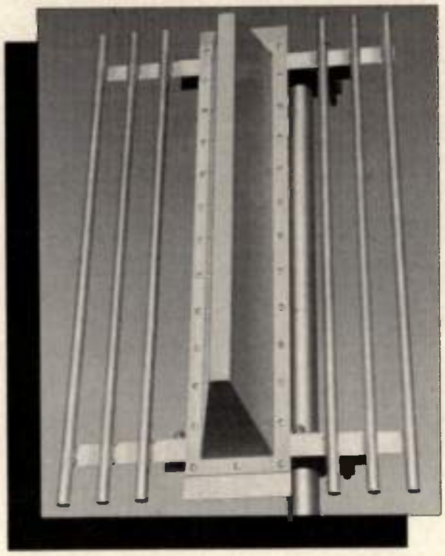
Antennas are essential for RF communications. Whatever the application, if it's not contained in a cable it relies on radiation from a transmit antenna to a receive antenna. Not surprisingly, the market for antennas is following the activity in commercial and consumer electronics for more wireless convenience and greater information transfer via radio communications.

Cellular radio is the application most often noted for current RF activity, with many antenna companies specializing in antennas for cell sites and mobile radios. Antenna Specialists has invested heavily in mobile cellular installations using its "on-glass" technology. Sinclair Radio Labs emphasizes cellular base station antennas, and Sales Administrative Supervisor Meg Driggers notes that business is, "...great, it has been busy...We have had a lot of cellular business lately." This view is echoed by Al Crego, Vice President of Marketing for Radiation Systems, Inc., Mark Antenna Division, "Our big load is in common-carrier cellular antennas. The international market for cellular antennas is also very big."

How good is cellular for antenna makers? Crego adds to his observations by noting that cellular should peak out this year, but remain solid for at least three more years. As the market matures, replacement antenna makers like Centurion are ready. According to Steven C. Bowles, Centurion's Director of Sales and Marketing, "Products in portability are becoming more popular. We have new cellular replacement antennas for improved range and signal quality."

Government business is slowing down, but not for everyone. Mark Antenna's Crego proudly announced that his company just closed a deal on the world's largest moveable radio telescope, a 100 meter diameter antenna. Most companies have experienced the expected reduction in government spending for conventional radio systems, but ongoing upgrades and replacements will remain a significant market for some manufacturers.

For broadcasting, the best markets have been outside the United States. China, South American countries, and other nations are still developing television systems. Their approach is often different than the U.S. market. For example, the use of broadband antennas that allow a single antenna to serve transmitters for several television channels saves construction costs. Designing the antennas and the combining/isolating hardware is a significant business for companies specializing in broadcast applications.



Cellular antennas like this directional unit from Mark Products are a big part of the antenna market.

Computer technology has made a tremendous impact on the design of antennas for all applications. "We save about 25 to 30 percent of our design time," Al Caplan of Telex/HyGain says about the MININEC-based design tools his company uses. Jennie Allen, Sales and Advertising Manager at Micro Communications is just as enthusiastic, "We use CAD tremendously; we can design a whole base system in software."

The development of the Numerical Electromagnetics Code to implement the method-of-moments analysis technique began a true revolution in antenna design. The development of the PC-based MININEC by engineers at the Naval Ocean Systems Center made this analysis method available to anyone.

As a result, notes Centurion's Bowles, "We can research areas we couldn't research before." The power of these programs is such that nearly any imaginable physical arrangement of conductors and dielectric materials can be modeled. The PC-based versions of these software tools are generally limited to wire or tubing construction, modeled either in free space or over a dielectric ground plane (real earth). Experimental configurations can be explored simply by defining their physical structure and feedpoint(s).

Coming Applications

The future of antenna markets after the cellular market has stabilized remains promising. Crego sees several new things coming, "Pipelines and utilities will really boom, but that won't come for a few years." He adds PCN (Personal Communications Network) and other commercial communications to the list of potential applications. Other areas include satellite systems for navigation and communications, and other short-range wireless systems such as remote utility meter reading.

Conventional two-way communications is changing, as well. Systems are in various stages of development and implementation for many service industries such as package delivery, trucking, railroads, taxis and buses. It is no longer sufficient to rely on simply on voice communications, so more advanced systems are needed. These industries share the common desire to know where a particular vehicle or shipment is located in a highly mobile system — at all times. Both terrestrial and satellite communications are being used in these applications.

Antennas remain an essential part of RF communications, and as new applications are developed, antennas with the required performance naturally follow. With much new development work underway, antenna manufacturers are enthusiastic about the future. **RF**

New CAE Software Brings Design Automation to the RF Engineer

By Dr. Thomas M. Reeder
EEsof, Inc.

The design of RF circuits in the 1990s has put greater demands on the design engineer. New applications in telephone and radio communications, remote data acquisition, and electronic navigation have fueled heavy competition in the marketplace, and with the competition, a strong demand for faster time-to-market. With these volume oriented commercial applications has come a greater need for both lower cost manufacturing and increased demand for high reliability. RF circuits continue to grow in function and integration complexity, often requiring unusual packaging which integrates RF with digital circuits, displays, and sometimes even antennas.

In the past, CAE design tools for the RF engineer have been limited to specific point tools for circuit simulation and board layout. Integration of simulation and layout for RF circuit boards is more difficult than most digital computer boards, because many RF components exhibit coupling effects that require handcrafted layout and insightful simulation to achieve a reliable, manufacturable finished product. Moreover, until now, CAE tools for the RF engineer have not included the kind of built-in circuit and data entry, documentation, and project management capability that has recently become popular in major CAE software for digital circuit design.

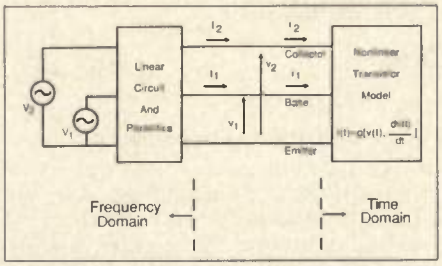


Figure 1. A block diagram view illustrating for a simple one transistor circuit how the harmonic-balance simulation method partitions analysis into linear and nonlinear circuit analysis blocks.

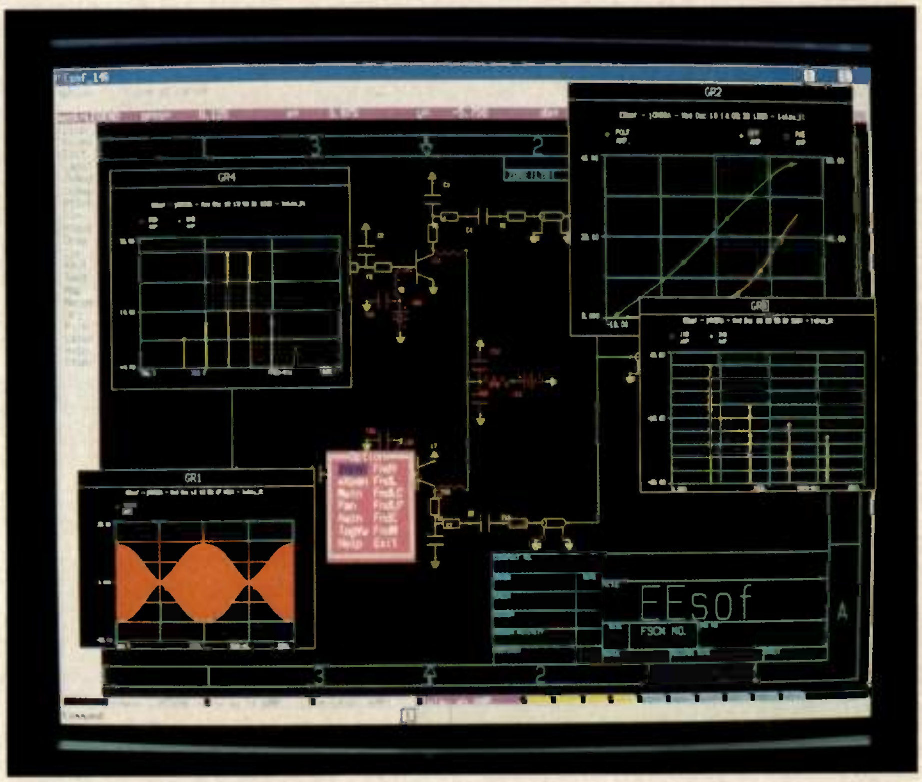


Figure 2. Screen photo showing the application of jOMEGA in the design of a push-pull RF power amplifier.

EEsof, Inc. has just announced a new CAE design tool developed to solve these types of problems specifically for RF circuit design. Called jOMEGA™, the new CAE product combines a powerful harmonic-balance linear/nonlinear circuit simulator with an advanced graphics package that provides schematic entry, multi-window simulation control, engineering documentation, and RF board design layout and floor planning. jOMEGA includes a wide variety of passive and active models useful to RF circuit design, and a new large-signal silicon transistor library, including models for the transistor packages.

jOMEGA's capabilities are shaped by two important CAE software modules built into the new product: 1) a graphical environment that provides multi-window display and control of simulations, and 2) a harmonic-balance simulator engine that provides both frequency and time analysis for circuits driven by sinusoidal inputs.

The Harmonic-Balance Technique

In particular, the use of a harmonic-balance (HB) simulator is interesting because it has significant new benefits in the design of nonlinear circuits, and it is the first new simulator technology to be made commercially available to the RF engineer since the introduction of Touchstone® and other linear simulators in the early 1980s. Harmonic-balance was a university research topic in the early 1980s, using large mainframe computers to carry out the required analysis. But with the introduction of powerful desktop computers using the IBM OS/2 and AT&T UNIX

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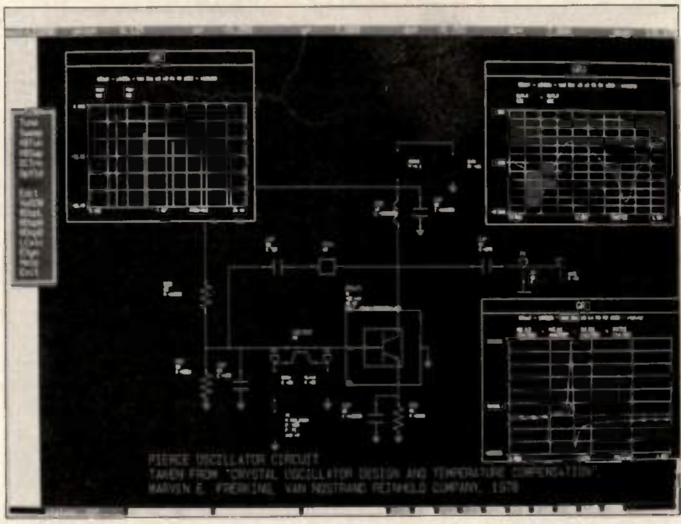


Figure 3. Application of jOMEGA to the design of a high-Q VHF crystal oscillator.

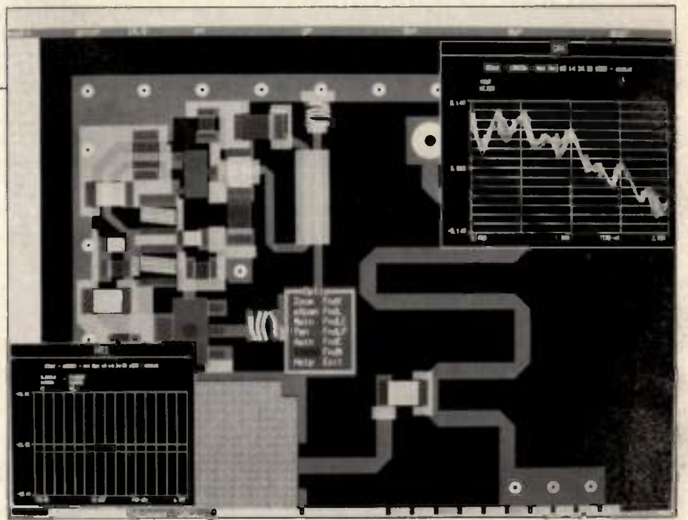


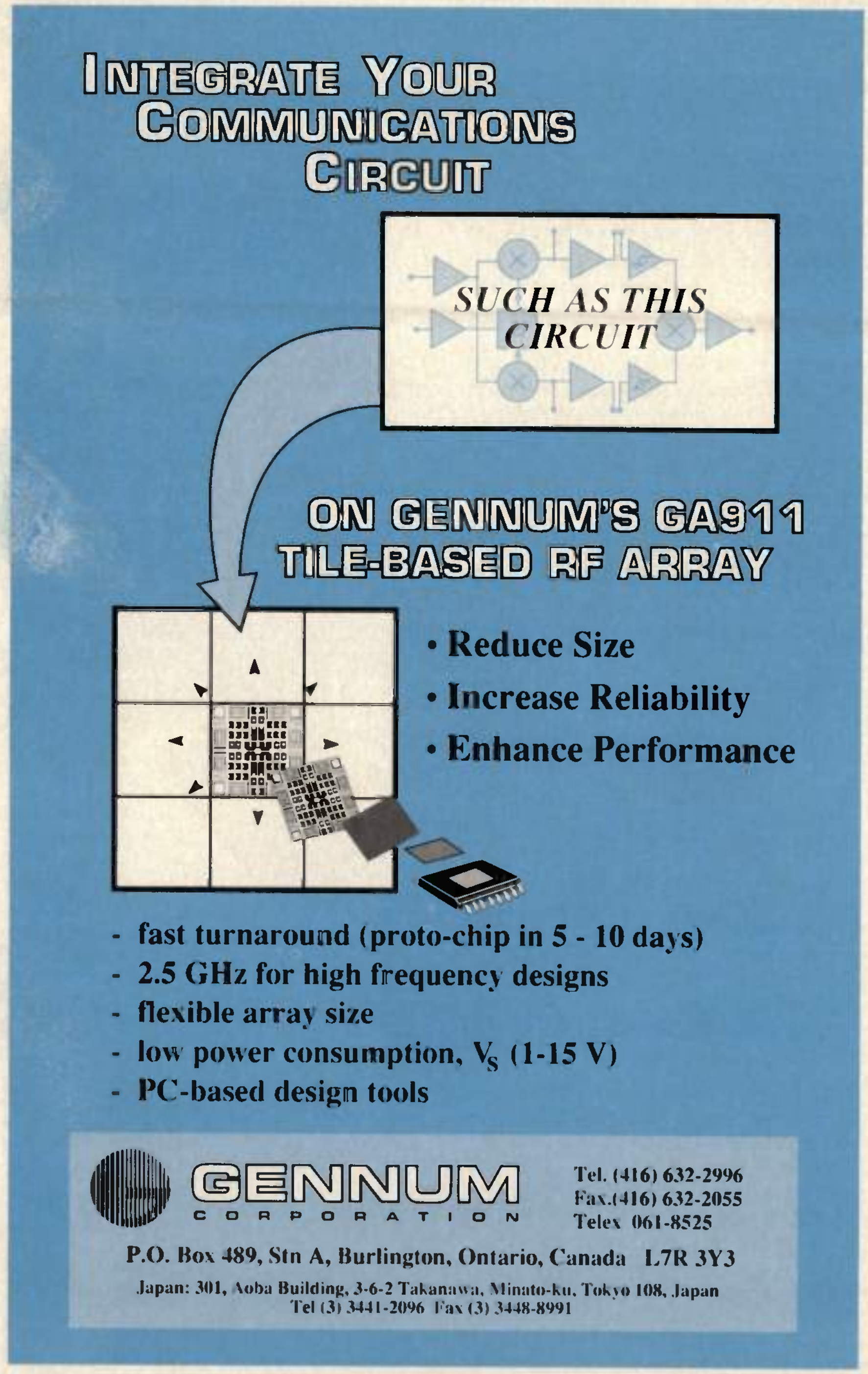
Figure 4. jOMEGA screen showing an example of VHF board layout using jOMEGA's layout option.

operating systems, it has become practical to implement the harmonic-balance technology in a commercial simulation software. Harmonic-balance works by partitioning the circuit into linear and nonlinear portions, followed by an iterative analysis to match boundary conditions at the partitions (1).

Figure 1 illustrates the process of circuit partitioning for a simple one-transistor circuit. The nonlinear part of the circuit is analyzed by assuming that the nonlinear currents and voltages can be represented by a time-domain harmonic series with unknown amplitude coefficients. By using the Fast Fourier Transform, the nonlinear circuit time-domain boundary variables are converted to frequency-domain, where ordinary nodal equations can be written and solved. An error in current-voltage match is noted at the linear-nonlinear circuit boundaries, which leads to a steadily decreasing error with analysis iteration. The process stops when the error is sufficiently small.

Harmonic-balance provides very fast and efficient linear and nonlinear simulation of sinusoidally driven circuits, including amplifiers, mixers, oscillators and other circuits of central interest to the RF engineer. The HB simulator makes it relatively easy to analyze and optimize amplifier nonlinear response, mixer conversion loss, and oscillator at frequency characteristics. HB is complimentary to, but, much more efficient than SPICE for steady-state sinusoidal circuit analysis. HB also has the advantage that it is a natural extension of frequency-domain simulators RF engineers already know.

For analysis of the linear circuit portions, harmonic-balance utilizes the same linear circuit model set and simulation engine used in the past, so the HB approach has linear circuit analysis "built-in." But it also has the ability to analyze steady-state nonlinear circuit operation with the most complete linear and nonlinear model set possible. Transmission-line dispersion, wave propagation loss, and transistor large-signal dynamic operation are all included in a package that looks and feels like the well-known linear simulators of the early 1980s. And since HB uses both time-and frequency-domain operations in its analysis, the designer has time-domain waveforms and frequency-spectra plots in addition to the usual frequency-dependent s-parameter circuit simulations.



A Power Amplifier Design

Figure 2 shows a typical example of a design session in jOMEGA. A power amplifier designed for use at cellular radio frequencies (820-900 MHz) has been entered through the schematic entry screen. Achieving maximum power-added-efficiency (PAE) was the issue in this design, so attention is directed toward optimizing transistor bias and circuit match conditions to achieve the best PAE possible while retaining good amplifier linearity. Starting from the upper right-hand corner and moving clockwise, windows have been opened to display:

- P_{out} and power-added efficient versus P_{in} for CW signal
- The harmonic spectrum with two-tone drive
- The two-tone single sideband modulation envelope
- The corresponding two-tone odd-order intermodulation spectrum

The PAE displayed is a jOMEGA measurement defined by $100(P_{out} - P_{in})/P_{dc}$. Other definitions for PAE or most other output measurements can easily be defined for display versus power or frequency using jOMEGA's output equation capability. The tuning and nonlinear optimization of the displayed circuit can now proceed with clear update to all important circuit quantities immediately visible to the designer.

Oscillator Design

Another difficult RF design problem where jOMEGA provides new capability is in oscillator design. Many engineers already use EEsof's Touchstone linear circuit simulator to estimate oscillator frequency based on linear, start oscillation conditions. jOMEGA's harmonic-balance algorithms and a new software oscillator design element called OSCTEST™ provide a straightforward way to analyze and optimize oscillator operation under loaded-circuit, fully nonlinear operation. Basically, OSCTEST provides a noninvasive coupler that detects forward and reflected waves excited at a suitable point of measurement in the oscillator circuit. jOMEGA includes os-

cillator design algorithms that automate the task of finding the steady-state operating conditions and voltage waveforms for the circuit. Figure 3 shows a simulation of a high-Q oscillator circuit used in a VHF remote signalling application. Note that for oscillators with Q above 10,000 it is not practical to simulate circuit performance in SPICE because simulation time becomes unreasonably long. Since SPICE is time-domain based, it must simulate over a time period greater than the product of steady-state time period multiplied by the oscillator Q. On the other hand, harmonic-balance simulation time is limited only by the number of frequency points desired.

Physical Layout Option in jOMEGA

Another feature of jOMEGA that will be appreciated by RF engineers is its physical layout option, which is very useful for physical design and floor planning of RF circuit boards. This option lets you switch back and forth from the schematic entry view to a physical layout view where actual package outlines and circuit interconnections can be placed. Parasitics that often greatly influence RF circuit operation can be seen during the initial design phase when costs are reasonably low. In contrast, other RF-oriented CAE packages available today are netlist or schematic entry only, which means that layout must be done separately, often by graphics draftsmen in a separately located facility or group. This leads to late detection of physical parasitic problems and little opportunity to remove and re-optimize around them.

Figure 4 shows a jOMEGA board floor plan design for a portion of a VHF radio communication circuit. Standard circuit elements can be immediately entered in this layout, and simulation can be done as shown from this view as well as from schematic. The layouts for new, unusually shaped elements and components can be user-defined through a built-in macro system. And when the board design is finished, the jOMEGA layout can be transferred to the board finishing system through either Gerber or IGES standard interfaces.

EEsof has developed jOMEGA as an extension of its well-known ACADEMY® graphical design and Libra® harmonic-balance software packages, which have become well known for MIC and MMIC design. Knowing that RF designs must often be integrated with other designs that range from lower frequency analog and digital to microwave, jOMEGA has been designed to work closely with other EEsof products, giving the RF engineer access to a full range of integrated CAE/CAD solutions. E-Syn™ circuit synthesis program, Microwave SPICE® time-domain simulator and their OmniSys® hardware system simulator can be "plugged into" jOMEGA so that the simulator graphics input/output and simulator control are provided through jOMEGA, making it a truly "hierarchical" CAE software tool for the RF engineer.

Further development of jOMEGA will emphasize the unique needs of the RF engineer for passive and active models, for layout design and integrated electromagnetic simulation, and in easy interface to other design automation systems, such as those from Mentor Graphics, Valid, and Racal-Redac, that are often the systems used in final RF board manufacturing layout. Thus, EEsof intends to further develop jOMEGA as a premium design tool for the RF engineer but at the same time will insure that this product maintains interfaces through industry standards to both other EEsof products and to other major CAE/CAD products utilized by the RF engineer.

jOMEGA will ship first on the Sun SPARCstation, beginning in April. Shipments on other platforms, including popular HP/Apollo platforms and the new IBM RS/6000 computer will be announced later in the year. For further information and to receive a detailed product brochure on jOMEGA call EEsof at (818) 991-7530, FAX at (818) 889-4159 or circle info/card # 230. *RF*

References

1) S. El-Rabaie, V.F. Fusco, and C. Stewart, "Harmonic-Balance Evaluation of Nonlinear Microwave Circuits - A Tutorial Approach," *IEEE Microwave Theory and Tech.*, Vol. MTT-36, pp. 181-192, August 1988.

About the Author

Dr. Thomas Reeder is Vice President of Marketing at EEsof. He may be reached at 5601 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (818) 991-7530.

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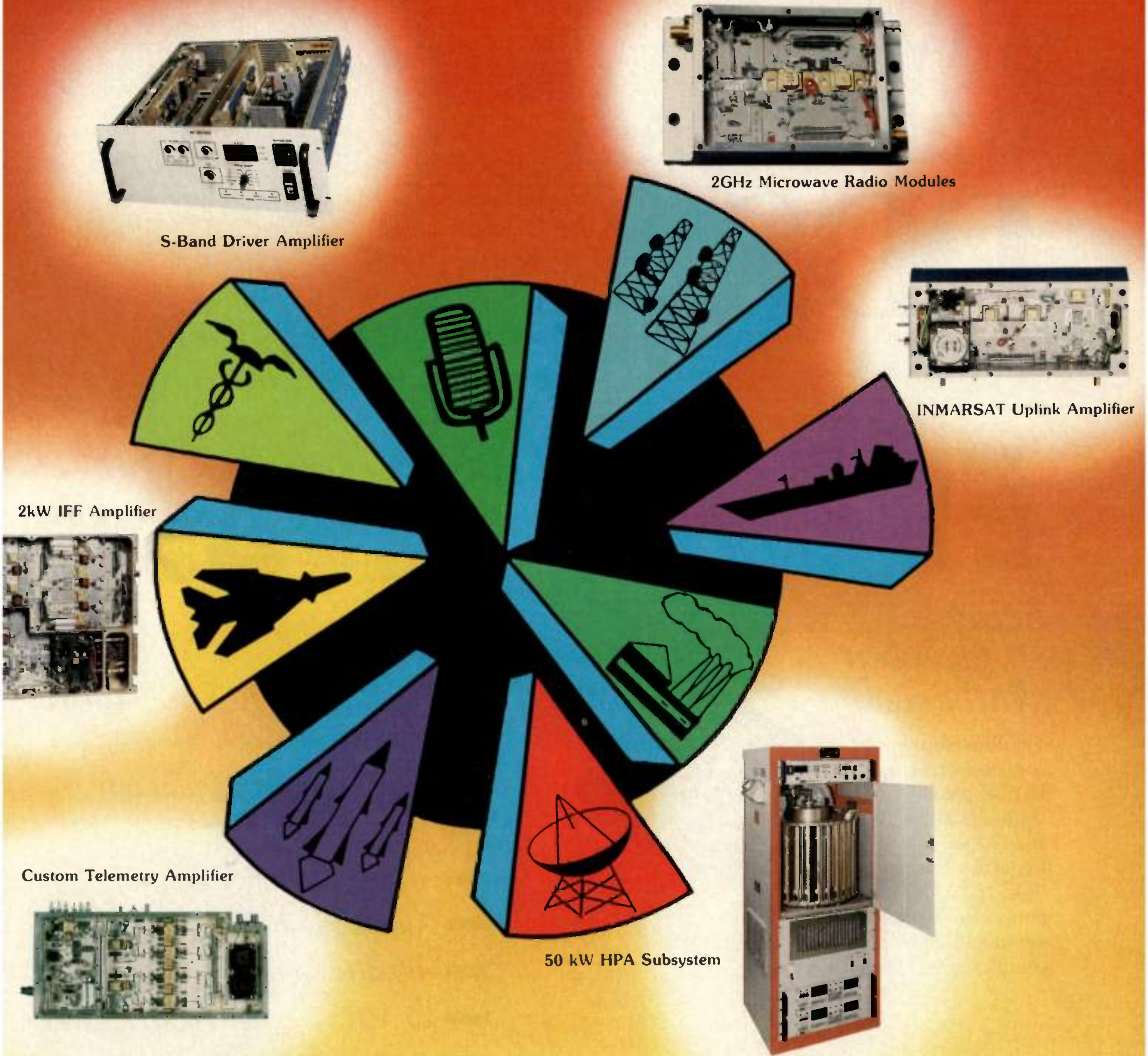

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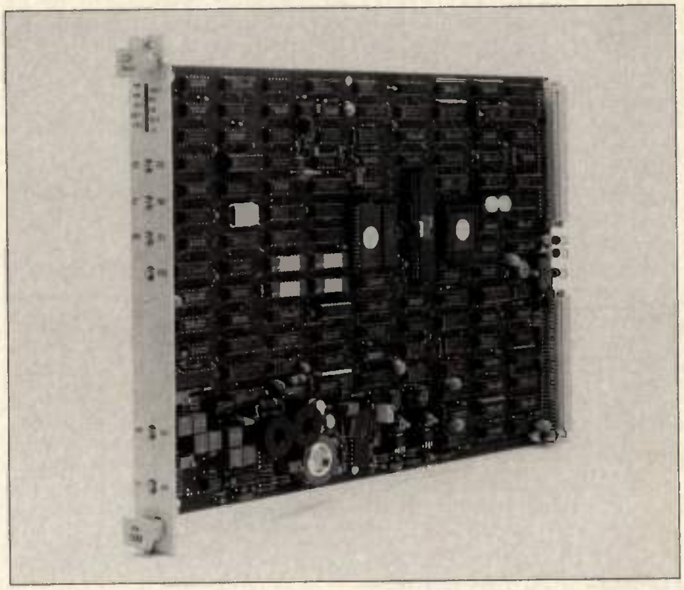
Arbitrary Waveform Generator



Tektronix and Colorado Data Systems (CDS) have announced the 73A-243 VXIbus Arbitrary Waveform Generator Module (ARB). The ARB can generate waveforms of 16,384 voltage samples at a rate of 25 MHz. The 73A-243 has the ability to vary the frequency and amplitude of a stored waveform without accessing the system controller. The ARB has sine, square, sawtooth, and triangular waveforms available. The module output is capable of driving a 50 ohm load in three ± voltage ranges: 10.22 V with 5 mV resolution, 5.11 V with 2.5 mV resolution, and 0.1 V with 50 uV resolution (Option 01). A programmable attenuator and low pass filter (5 MHz, 500 kHz, or 50 kHz) is provided at the output. Breakpoints can be set in any non-consecutive voltage sample, and an output voltage during a breakpoint is available. Individual waveform points can be edited without a memory reload. The ARB has a 16 KByte input buffer. Programming can be done in ASCII, binary, or CIIL. The stored waveform can repeat continuously or from 1 to 255 times. Pricing for the 73A-243 is $4,100.
Colorado Data Systems
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Crystal Clock Oscillator

Champion Technologies' MSO Series surface mountable crystal clock oscillators are now available over a frequency range of 1.25 to 40 MHz. They measure 0.560 by 0.360 in. (14.22 × 9.14 mm) with 0.160" (4.06 mm) height, and have a frequency stability of ±0.01 percent from 0 to +70 C (50 ppm optional). The oscillators are in a hermetically sealed ceramic package, are available in a "C" lead configuration, and are track-compatible with standard equipment. Features include a 3,000g shock rating, CMOS or TTL compatible output with tight symmetry (45/55 percent), and enable/disable optional. The MSO oscillators are assembled automatically in a Class 100 Clean Room.
Champion Technologies
INFO/CARD #138

UHF/Cellular GaAs MMICs

Oki Semiconductor announces a broad line of analog and digital GaAs MMICs. The KGF-1145 dual-gate buffer amplifier has 3 dB noise figure, and draws 4 mA at 850 MHz. The KGF-1146 two stage buffer has +4.2 dBm output

at a low 2.5 mA supply current. Front-end components include the KGF-1175 low-noise amplifier and the KGF-1155 dual-gate mixer amplifier. The KGF-1165 feedback wideband amplifier has +7 dBm output, 4 dB NF, and draws less than 25 mA in the 800-900 MHz range. Four power amplifiers offer +17 to +31.5 dBm outputs, gains to 13 dB, and low large-quantity pricing. Also available is the KGL-2115 two-modulus prescaler with 0.7 to 1.0 GHz toggle frequency, and divide-by-128/129 output.
Oki Semiconductor
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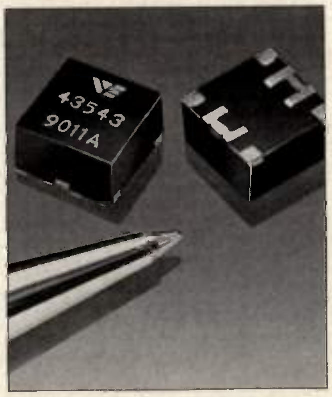
Surface Mount Leadless Transformers

Surface mount leadless transformers are available from Vanguard Electronics, designed and built to customer specification to meet MIL-C-15305, MIL-T-55631, or MIL-T-27. Features include a compact low profile and a surface mountable package having tin plated phosphor bronze terminations. These devices are ideal for automatic insertion and can be supplied as RF transformers with up to 6 connections, wideband transformers with a frequency range from 100 kHz to 300 MHz, pulse transformers with 100 ns rise/fall time and 10 percent droop, and power transformers up to 1 watt. 100 piece pricing is $7.00 to $9.00 each.
Vanguard Electronics Company, Inc.
INFO/CARD #136



Surface Mount Hyper-Abrupt Tuning Diodes

The MSI 1M1401 series tuning diode, in a surface mount package with Style 1206 solder pad geometry footprint, satisfies many RF variable capacitance requirements consistent with dense component packaging. Design efforts leading to surface mount, mass

production circuits are facilitated with glass packaged counterparts, the MV1401 - MV1412 series diodes, that have capacitance ranges from 400 pF for the 1M1401 to 10 pF for the 1M1412 with capacitance ratios exceeding 10:1 from 2 to 10 volts bias. Solder contacts to the diode are made to 0.020 in. (3 mm) long mounting side. The environmental resistant 1M1401 series with their 12 volt reverse voltage rating are especially suitable for RF applications that range from mobile radios to cellular telephones. The 1M1401 is $10.35, and the 1M1403-1412 are $7.50 (100-500).
MSI Electronics, Inc.
INFO/CARD #135

RF Modules

Pan Datacomm has announced the availability of OEM and custom made RF Modules. The VHF, UHF, and 800-900 MHz transmitter, receiver, and frequency synthesizer modules are very low power and ultra compact. They are made for applications such as RF links, walkie-talkies, hand held telephones, portable 2-way radios, FSK receivers, scanners, etc.
Pan Datacomm Inc.
INFO/CARD #134

Axial Leaded Inductors

The Delevan Division of American Precision Industries announces the availability of an

expanded line of Established Reliability axial leaded inductors designed for military and commercial applications. The inductance values range from 0.10 uH to 1000 uH and meet the requirements of MIL-C-39010 standards in slash number configurations including -01, -02, and -06 through -10. The inductors are available in tolerance ranges from ± 5 to ± 10 percent, depending on the individual inductive value. Unit cost ranges from $1.92 to $3.00.
Delevan Division,
American Precision Industries
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Directional Detector

Veritech Microwave has introduced the Model VMDD 0518B-01 which incorporates the functions of a directional coupler and low-impedance diode detector in a size of 1.0 × 1.0 × 0.22 inches. Detected voltage varies ±0.2 dB over −54/85 C and ±1.4 dB over 0.5 to 18 GHz. Insertion loss is 3.0 dB, with a ripple of ±0.8 dB. VSWR is 1.7:1, video impedance is 800 ohms. Nominal output is −22 mV per mW, and directivity is 13 dB. The price is $575 for 10-19 pieces.
Veritech Microwave, Inc.
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Surface Mounted Crystals

TELE QUARTZ is offering surface mounted crystals in a plastic housing which is resistant to deformation and suitable for infra-red-reflow and wave soldering. These crystals are available over the total frequency range from 4 or 8 MHz up to 360 MHz. The housings XSO-4 and XSO-4L have been submitted as a proposal for DIN standard. The appropriate blister-pack belt has a width of 16 or 24 mm.
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ECM Devices has introduced Model 2F002 which is an uncased single pole double throw PIN Diode Switch Driver with gold wire bonds. The 2F002 has a size of 0.24 × 0.28 × 0.05 inches, an output current of 20 mA (2 inverting, 2 noninverting), an output voltage of ±4 VDC, an Operating Voltages of +5 and −12 VDC at 30 mA, and is TTL compatible. Other features are 10 MHz Repetition Rate, and a 40 ns Switching Speed. Also available is a 3/8 × 3/8 Flatpack. For 10-24 units, the price is $52.00 each.
ECM Devices, Inc.
INFO/CARD #249

BNC Connectorization Kit

A BNC Coax Connectorization Kit for RG-58/59/62 coax is available from ALPHA Wire Corp. The AT-KIT-1 is the latest addition to Alpha's "Wire Management" program. The kit contains a "3-cut, one-step" AT-140/3 Stripping Tool, an AT-330 BNC Coax Ratchet Crimping Tool, and 10 compatible BNC crimp connectors with all the required parts. The AT-KIT-1 comes with a comprehensive selection guide.

Alpha Wire Corporation
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ACMOS Compatible Oscillator

The Conner-Winfield Corporation is now offering a reduced size, ACMOS compatible oscillator. The A50 series is available in a 50 MHz through 150 MHz frequency range. Accuracies to 10ppm, and operating temperature ranges from −40 C to +85 C are available. Pricing for prototype quantities, 10 pieces, 100 MHz, is $42.50 each.
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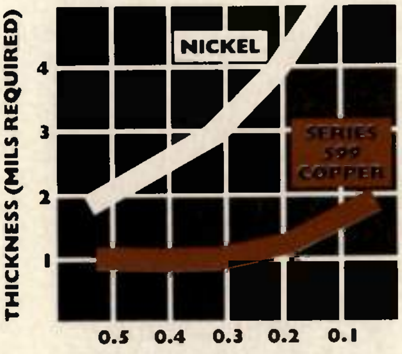

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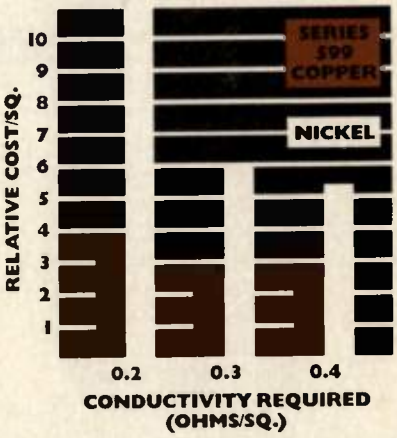

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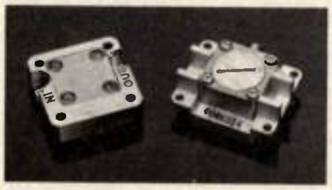
Polaris Electronics Corp.
INFO/CARD #246

Space Qualified Isolators

1.5 to 30 GHz isolators are available from Sierra Microwave Technology, offering up to 100 percent bandwidths, less than 0.4 dB insertion loss, and isolation and return loss better than 26 dB. Drop-in models incorporate a shielded microstrip alumina substrate launch. Connectorized versions are available.



Sierra Microwave Technology
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E-Field Antenna

The ADA-120/A is a low-noise active receiving dipole antenna designed for accurate measurement of electric filed intensity up to about 3 V/m, over a 1 kHz to 200 MHz frequency range. A broadband balun allows use without a ground plane. Dynamic range is 110 dB, and each antenna is individually calibrated to insure accurate measurements.
Antenna Research Associates
INFO/CARD #244

Surface-Mount Filters

RLC Electronics introduces its line of surface mount filters, with typical package heights of .110 and .225 inches. Leads may be true SMT or formed to customer specifications. Lowpass, highpass, and bandpass responses are available over 10-2000 MHz. Devices may be supplied and tested in accordance with MIL-F-18327. Pricing starts at $275.
RLC Electronics, Inc.
INFO/CARD #243

Spectrum Analyzer

The new 26.5 GHz Spectrum Analyzer FSM is introduced by Rohde & Schwarz. Features include low noise of −142 dBm (typical) with 6 Hz maximum resolution, intermodulation-free range

of 100 dB, 1 dB frequency response flatness to 5 GHz, and a comprehensive self-calibration routine. A low noise synthesizer, together with the narrow resolution bandwidth, allows close-to-carrier measurements over the entire frequency range.
Rohde & Schwarz
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VCOs for 5 Volt Supplies

Vari-L announces the VCO-305, the first in a series of +5 volt supply VCOs with octave tuning. This model covers 200-400 MHz with +10 dBm output, phase noise −75 dBc at 1 kHz offset and −125 dBc at 100 kHz offset, with 11 mA supply current. The full octave tuning is available with a +1 to 20 volt tuning voltage. The family will cover 25-2000 MHz.
Vari-L Company
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Open Circuit Terminations

For improved calibration of vector automatic network analyzers (VANAs), Maury Microwave has developed the 8048A1 and 8048B1 female and male 3.5 mm open circuit terminations. Calibration using these terminations will typically result in an effective source match of better that 35 dB (typically better than 40 dB).

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The Model 3022 Sweep/Function Generator from B&K Precision covers 0.02 to 2 MHz in seven ranges, and includes a 5-digit frequency counter. Operating modes include sweep, AM, FM and voltage-controlled. Outputs are available for TTL, CMOS

and other waveforms, with a variable DC offset available at the output. Low sine wave distortion of less than 1 percent, and 2 percent square wave symmetry are other features. The 3022 is priced at $450.
B&K Precision
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High Dynamic Range Amplifier

Phoenix Microwave introduces the PA923 TO-8 hybrid amplifier, featuring 22 dB reverse isolation over the 5-1100 MHz range, with 10 dB small-signal gain and +10 dBm output power at 1 dB compression. The amplifier is guaranteed stable for any source or terminating impedance.
Phoenix Microwave Corp.
INFO/CARD #238

MOS-Type Capacitors

Thin film MOS-type capacitors, designed for hybrid circuits, are available from Mini-Systems. The available capacitance range is 4.7-1000 pF in 15-180 working voltage. Features include low dissipation factor (0.1 percent) and high insulation resistance (1011 ohms).
Mini-Systems, Inc.
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Amplifonix announces the TM6119 amplifier, offering 6.5 dB gain over its 30-250 MHz frequency range. Output power is +20 dBm at 1 dB compression, noise figure is 3.5 dB maximum, and the current draw is 50 mA from a +15 VDC supply. Available packaging includes TO-8, flatpack, surface mount and connectorized versions.
Amplifonix
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Reduced Bandwidth FM Communications

By Mark A. Kolber
Honeywell Business and Commuter Aviation Systems

Increased usage of communications frequencies brings with it an increased potential for interference among users occupying adjacent channels. This article presents a technique for reducing the occupied bandwidth of FM voice modulated systems by using modified modulation processing (1). The method has the potential for materially reducing interference to communications on nearby frequencies, while maintaining the same audio bandwidth on the transmitted signal.

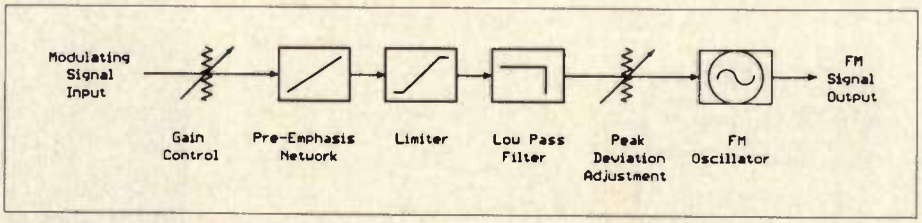

Figure 1. Conventional modulation processing sequence of gain, preemphasis, limiting, filtering, and deviation control.

The technique is based upon the fact that the peak bandwidth of an FM signal is primarily determined by the maximum deviation of the highest modulating frequencies, and to a lesser degree, the maximum deviation of the lower modulating frequencies. With conventional modulation processing, the peak deviation of the higher modulating frequencies determines the peak bandwidth of the signal. The modulation processing presented here employs a frequency-dependent modulation limiter in which the higher modulation frequencies are limited to a lower value of peak deviation and therefore occupy more nearly the same bandwidth as the lower modulating frequencies, reducing the peak bandwidth.

To begin the discussion, we need to examine what determines the peak deviation of an FM transmitter. Peak deviation occurs at the peak amplitude of the modulating signal. If a transmitter were to be directly modulated by a complex waveform such as a voice signal, the peak deviation would occur at the peak amplitude of the voice signal. Since the peak amplitude of voice signals are not well controlled, the peak deviation and, therefore, the bandwidth of the transmitter would not be well controlled. In order to guarantee that the transmitted signal will occupy no more than the allocated bandwidth, it is necessary to control the characteristics of these voice signals with a modulation processing system. This usually consists of a preemphasis network, amplitude limiter, and a lowpass filter.

The preemphasis network compensates for the skewed power spectral density characteristics of voice signals, which tend to have power concentrated in the lower frequency region. The preemphasis results in a modulating signal that has a more constant amplitude versus frequency characteristic.

The amplitude limiter then limits the

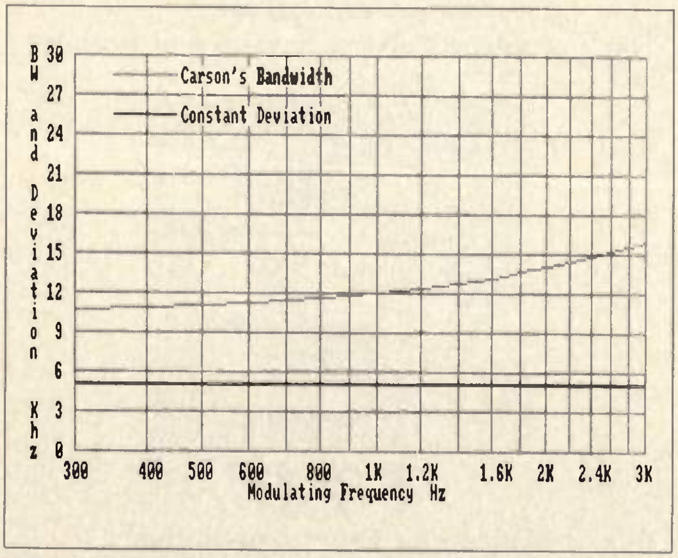

Figure 2. Bandwidth according to Carson's rule versus modulating frequency for constant 5 kHz deviation.

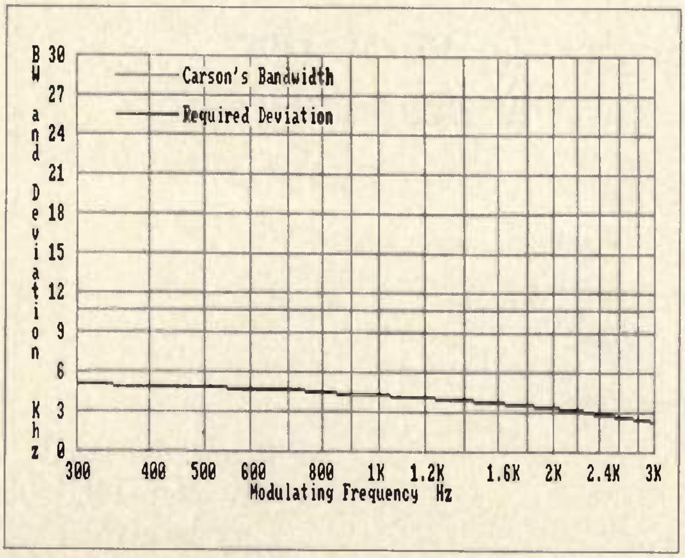

Figure 3. Deviation versus modulating frequency for a constant 10.6 kHz Carson's bandwidth.

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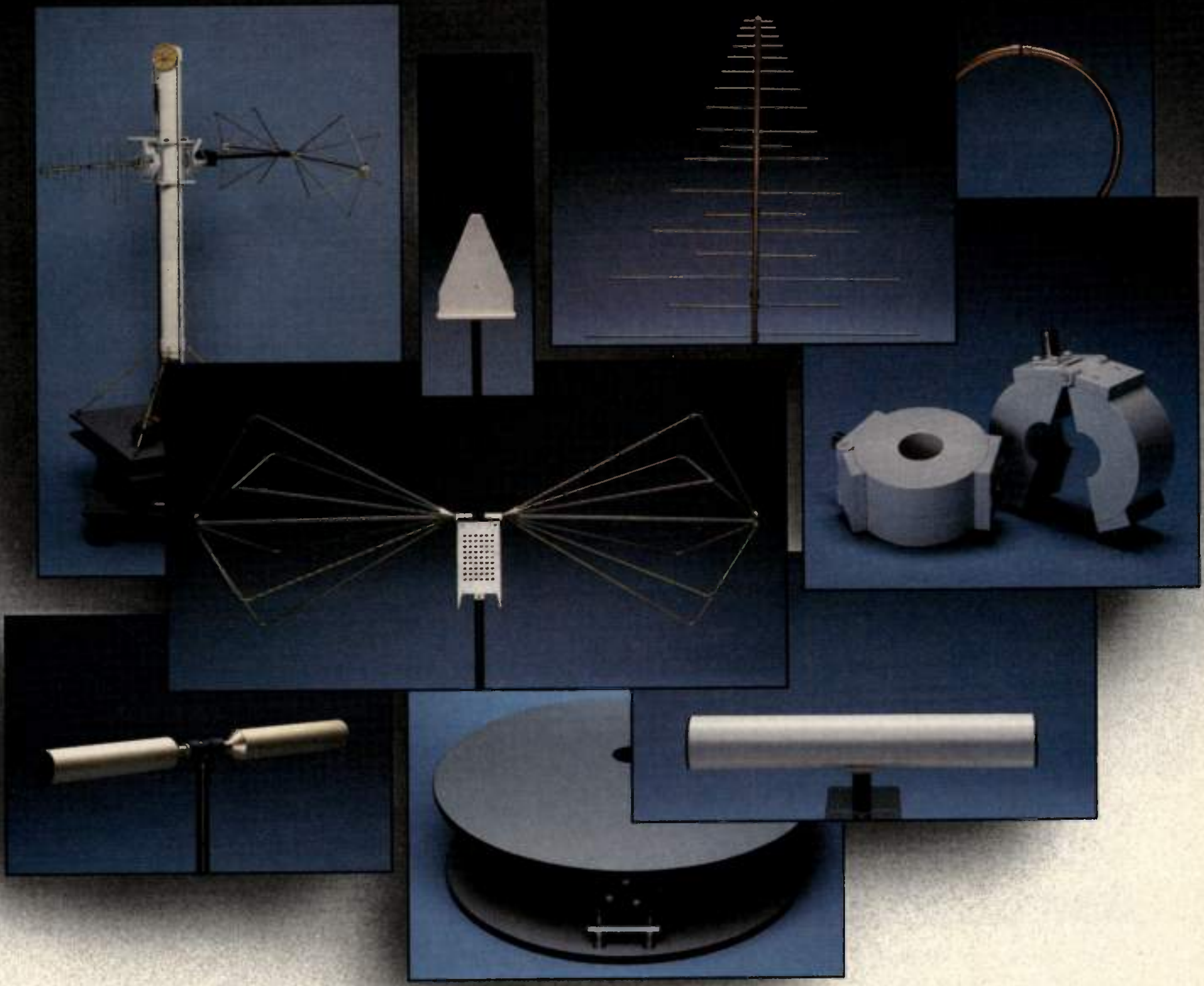

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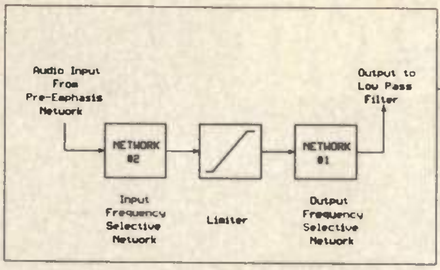

Figure 4. Block diagram of a frequency-dependent limiter using complementary frequency selective networks and conventional limiter.

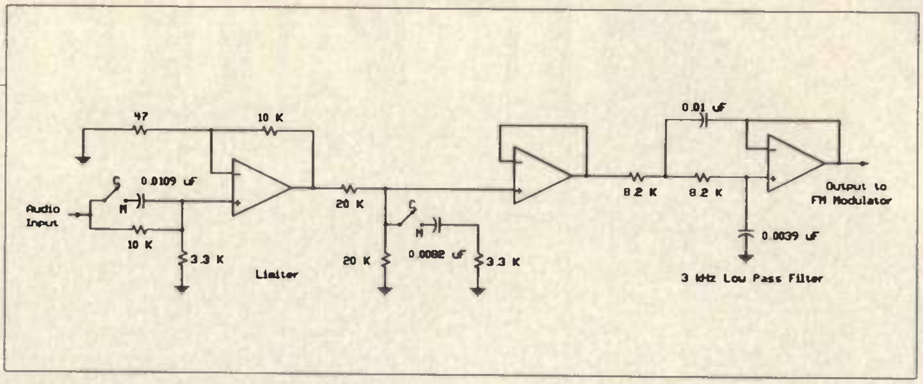

Figure 5. Diagram of the modified modulation processing circuit.

signal peak positive and peak negative excursions to a defined limit, therefore establishing a peak deviation (see Figure 1). Regardless of the setting of the gain control or the exact nature or amplitude of the input modulating signal, the peak amplitude at the output of the limiter cannot be exceeded. Since the preemphasis network precedes the limiter and has no effect on the system frequency response following the limiter, the limited peak deviation is not a function of the preemphasis. The preemphasis does cause the gain to be a function of frequency, but not the peak deviation.

The lowpass filter following the limiter establishes the highest allowed modulating frequency and removes some of the harmonics generated in the limiter. For example, land mobile systems typically use a modulating frequency range of 300 Hz to 3 kHz. Therefore, these systems have a lowpass filter with a cutoff of 3 kHz. The result is that, in spite of the preemphasis which provides higher gain at higher audio frequencies, the peak deviation for all audio frequencies below 3 kHz is nominally constant, normally ±5 kHz. For fundamental components and harmonics above 3 kHz, the maximum deviation is reduced by the filter.

To compare the bandwidth of conventional and modified modulation processing, we need a way to analyze the occupied bandwidth of the FM signal. The simplest analysis makes use of "Carson's Rule" (2) which results in a approximation for bandwidth, B:

$$B = 2(F_d + F_m) \quad (1)$$

where,

B = Bandwidth of FM signal

F_d = Peak frequency deviation

F_m = Modulating frequency

Figure 2 shows the results of calculating the Carson's Rule bandwidths for the

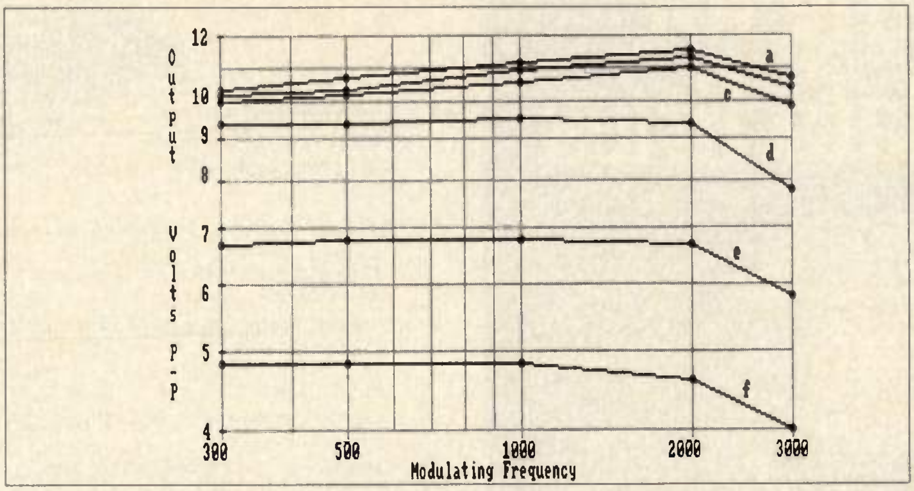

Figure 6. Frequency response of conventional modulation processing for six input levels.

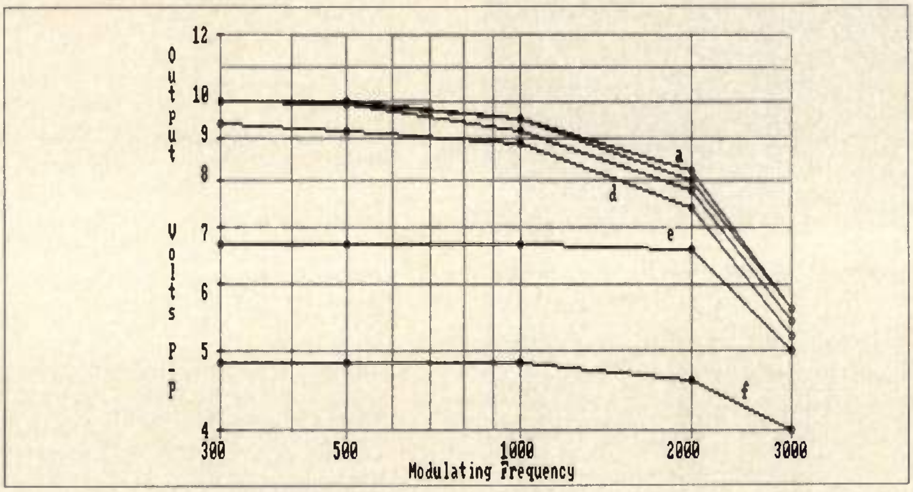

Figure 7. Frequency response of the modified modulation processing circuit for the same six levels as Figure 6.

range of audio frequencies, assuming that conventional modulation processing is employed and considering only the fundamental component of the modulating signal. The constant level of ± 5 kHz deviation is shown on the lower trace of the graph and the resulting Carson's bandwidth is on the upper trace. On these graphs, the deviation traces are in units of ± kHz while the bandwidth traces are in total kHz bandwidth. As can be seen, the resulting bandwidths range from about 10.6 kHz up to 16 kHz, a variation of about 1.5 to 1.

If we modify the modulation processing such that the deviation is not held to a constant level, but has a decreasing level at the higher audio frequencies, we can obtain a constant bandwidth that is less than that obtained with conventional processing. To determine the required deviation for a constant bandwidth at each audio frequency, equation (1) is rearranged to solve for F_d:

$$F_d = B/2 - F_m \qquad (2)$$

B is set to 10.6 kHz, which is the bandwidth required for 5 kHz deviation at 300 Hz audio. Figure 3, in the lower trace, shows the values of deviation obtained at the various audio frequencies to achieve a constant 10.6 kHz bandwidth which is in the upper trace. As shown by the graph, the required deviation decreases as the modulating frequency increases. At 300 Hz the deviation is ± 5 kHz, while at 3000 Hz the deviation must be limited to ± 2.3 kHz.

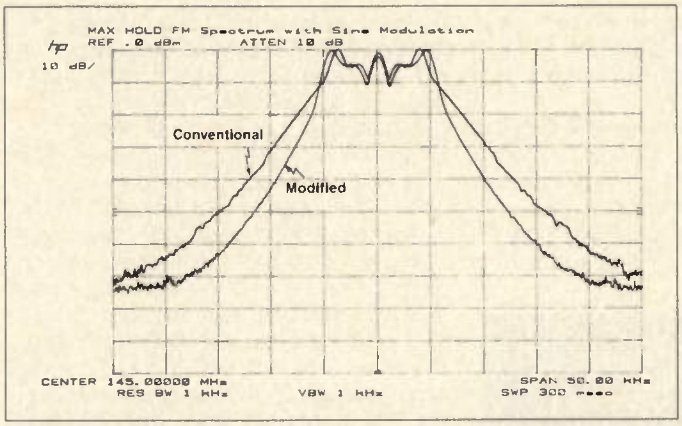

Figure 8. Occupied bandwidth plot of conventional and modified processing, using a swept sine wave signal source.

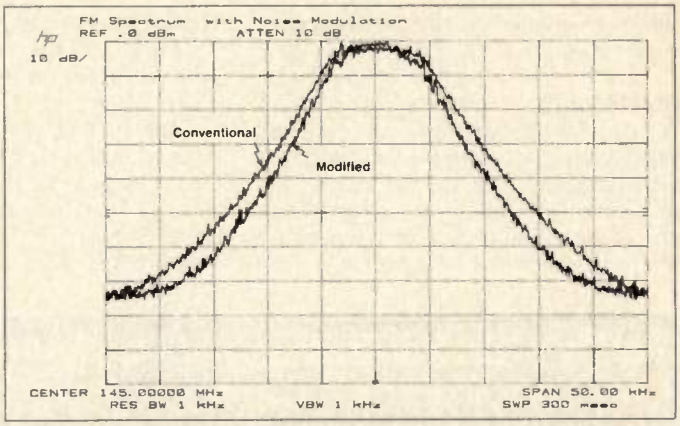

Figure 9. Occupied bandwidth of conventional and modified processing, with a band-limited white noise modulation source.

Circuit Description

The modified modulation processing described here requires a modulation limiter that is designed to limit the higher frequency audio signals to a reduced peak deviation so that they can occupy more nearly the same bandwidth as the lower frequency audio signals. We don't want to simply attenuate the higher frequency modulating signals. We want to amplify them the same as the lower frequency components, but we want to *limit* them to a lower value of deviation. We need a frequency-dependent limiter.

A limiter circuit with a frequency-dependent limit value can be implemented using a conventional non-frequency-dependent limiter with frequency-dependent networks preceding it and following it. Such an implementation is shown in Figure 4. The frequency selective networks are designed to have a complementary frequency response. Since the two frequency selective networks are complementary, for signal amplitudes below the limiting point of the limiter the circuit passes all audio frequencies equally. That is, below the limiting threshold, the circuit has a flat frequency response. For signal amplitudes above the limiting threshold, the output amplitude versus frequency characteristic of the combination is simply that of frequency selective network #1.

The operation of the conventional and frequency-dependent limiters were characterized by measuring the frequency responses at various input amplitudes. The circuit diagram is shown in Figure 5 and the results are shown in Figures 6 and 7. The graphs show a family of six curves, each curve representing the frequency response at a particular input level. The six input levels ranged from .212 $V_{p\text{-}p}$ to 1.2 $V_{p\text{-}p}$ and were increased in steps of 3 dB. The output levels ranged from 4 $V_{p\text{-}p}$ to almost 12 $V_{p\text{-}p}$. The vertical scales of the graphs are logarithmic.

The results indicated that at low input levels where no limiting occurs, both circuits provide a flat response except for the roll-off above 2500 Hz due to the characteristics of the lowpass filter. As the input level was increased, the output level begins to limit. At these levels, the conventional circuit shows a peak near the higher frequencies due to the Gibb's truncation effect (3). The modified circuit shows a lower output level for the higher modulating frequencies. This is the desired characteristic required to control the bandwidth.

Test Results

The effectiveness of the modified modulation processing was evaluated by connecting the output to an FM generator. The output of the generator was in turn connected to a spectrum analyzer and peak deviation meter. The resulting FM spectrum was then recorded. Two input signals were used for the evaluation, a swept sine wave and band-limited white noise.

The swept sine signal was 2 $V_{p\text{-}p}$ and was slowly varied from 300 Hz to 3000 Hz. The gain of the FM generator was set so that the peak deviation was 5 kHz. The spectrum analyzer was operated in the "MAX HOLD" mode so that the highest level that occurred any time during the sweep was recorded. The results are shown in Figure 8 for modified and conventional processing. The unmodulated carrier level is at the reference at the top of the scale. The observed −50 dB bandwidth was reduced from over 30 kHz to about 23 kHz.

The evaluation was also performed using band-limited white noise instead of the swept sine wave. The white noise source amplitude was approximately 2 $V_{p\text{-}p}$ and was band-limited to 5 kHz. Because it is difficult to measure the peak deviation with noise modulation, the FM generator gain was adjusted using the sine wave as above. After the gain had been set, the white noise was applied and the resulting spectrum was recorded again using the MAX HOLD mode. The observed −50 dB bandwidth was reduced from about 28 kHz to 24 kHz, as shown in Figure 9.

Conclusions

A method of reducing the required bandwidth for the transmission of voice modulated FM signals was explored. The method makes use of modified modulation processing that utilizes a frequency-dependent limiter. This results in a lower limit of peak deviation for the higher modulating frequencies thereby equalizing and reducing the required bandwidth. A simple method of realizing the frequency-dependent limiter was demonstrated, using a standard limiter with complementary frequency shaping networks at the input and output. ***RF***

References

1. M.A. Kolber, "Bandwidth Reduction in Voice Modulated FM Communications Systems," MSEE Thesis, Arizona State University, May 1989.
2. H.S. Black, *Modulation Theory*, D. Van Nostrand Co., Inc., Princeton, NJ.
3. P.M. Chirlian, *Basic Network Theory*, McGraw-Hill, New York, NY, 1969, pp. 578-592.

About the Author

Mark A. Kolber is Principal Engineer at Honeywell Business and Commuter Aviation Systems, 5260 W. Phelps Road, Glendale, AZ 85308. He can be reached at 602-863-4727. Mark received his BSEE from Newark College of Engineering (now New Jersey Institute of Technology). The work presented here was done as his thesis research for a MSEE at Arizona State University.

A Novel, Wide Band, Crystal Controlled FM Transmitter

By Thomas G. Xydis

Recently, several manufacturers have produced RF ICs which implement a complete crystal controlled FM receiver. These devices are typically applicable to both wide band and narrow band applications. The availability of similar ICs for the implementation of crystal controlled transmitters however, is currently limited to narrow band designs. Thus, wide band FM crystal controlled transmitter designs are generally limited to discrete implementations, especially if low cost and battery operation are desired. This paper describes a novel wide band FM transmitter topology which utilizes a commonly available FM receiver to implement a Frequency Locked Loop (FLL) control circuit (1).

A wide band ($\Delta f = \pm 50$ kHz), 90 MHz low power crystal controlled FM transmitter built using this technique is described.

To help motivate the new design, several well-known techniques for generating FM are briefly reviewed. The reader is encouraged to refer to References 1 and 2 for further information.

Direct Modulation of a Crystal Oscillator

The simplest way to generate a crystal controlled FM signal is to directly modulate the crystal oscillator. Since, typically, crystals can only be pulled a small amount while retaining linear deviation, the oscillator must be multiplied in frequency to increase the deviation requiring both power and space for filtering. This method was deemed undesirable for my application due to these requirements.

Indirect FM

As long as the modulation can be AC coupled, a reactance modulator is a nice solution, since in theory, it can work directly at the carrier frequency, thus eliminating much of the output filtering requirements. Reactance modulators are variable delay circuits which are used to phase modulate the carrier. By integrating (or in most cases low pass filtering) the applied modulation, FM is produced. However, active implementations are usually limited in deviation, thus in most applications the deviation is increased by frequency multiplication which again requires additional space, power and cost for filtering. Higher cost implementations usually utilizing mutual inductances are generally required to generate both high deviation and good linearity.

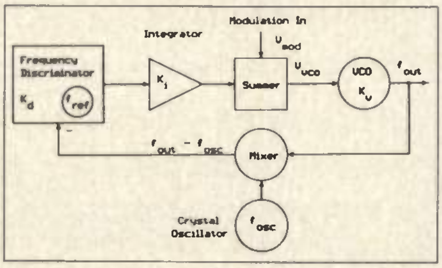

Figure 1. Operation of the FLL transmitter.

PLL Modulator

The PLL modulator (3) is a very good solution. Here, a narrow bandwidth PLL is used to control the frequency of a VCO which is then directly modulated. Since the VCO can run at the carrier

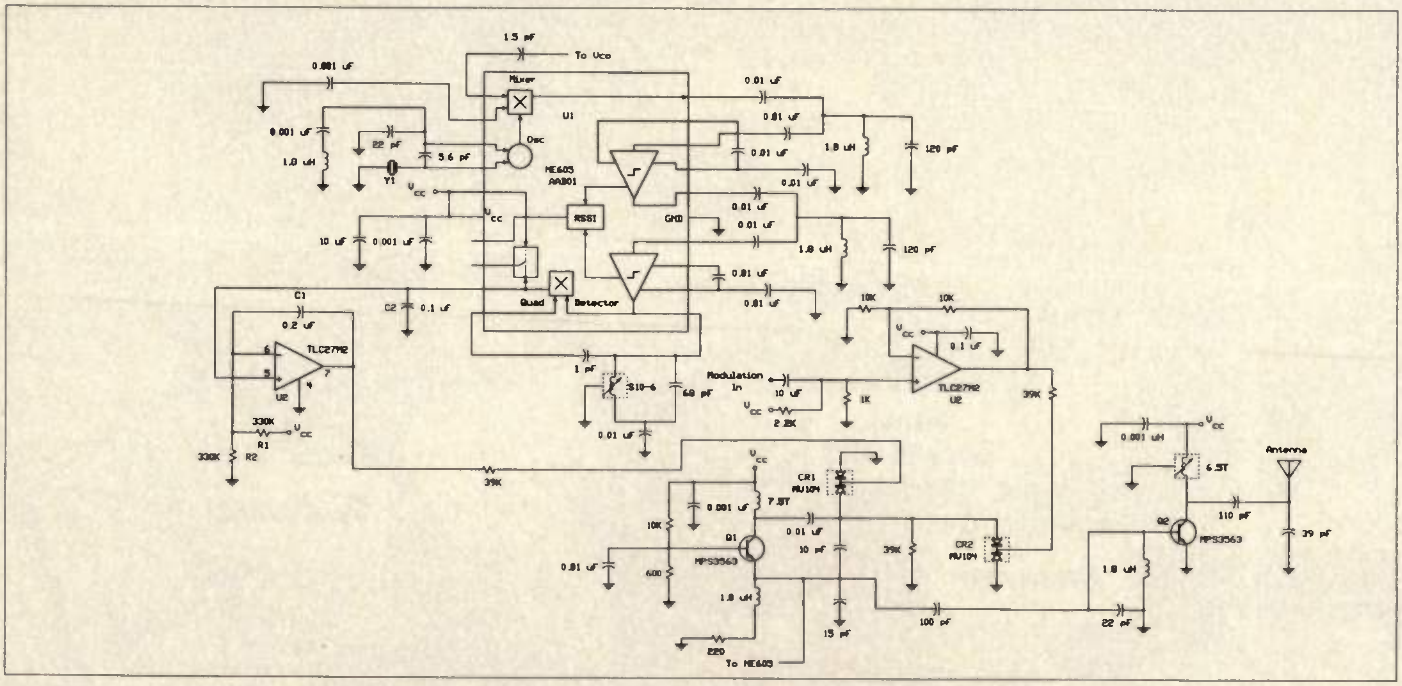

Figure 2. Wide band FM transmitter.

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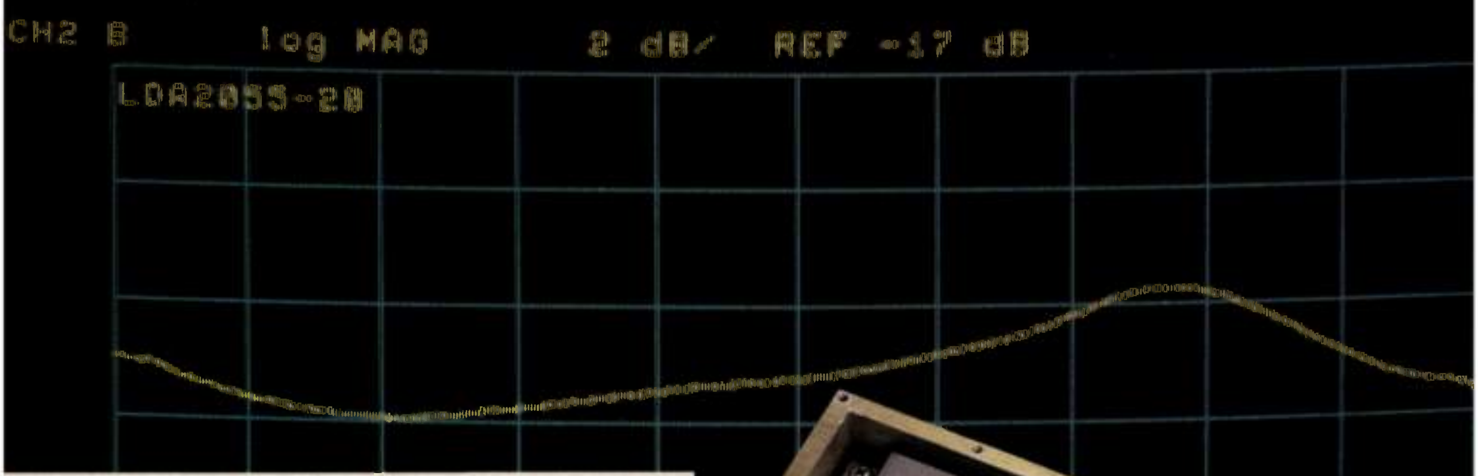

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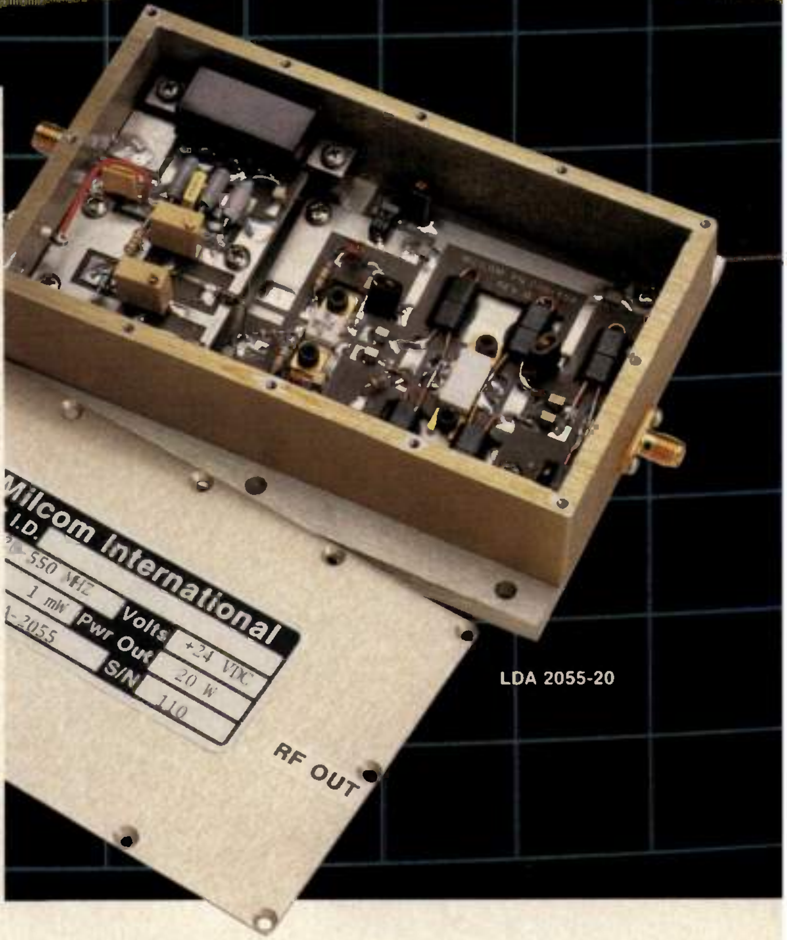
A0150-40	100-500 MHz	40 W	46DB	$2,295	In Stk
A0150-150	100-500 MHz	150 W	9DB	$2,995	Stk-6wks
A0240-10	225-400 MHz	10 W	43DB	$ 750	In Stk
A0240-20	225-400 MHz	20 W	43DB	$1,195	In Stk
A0240-30	225-400 MHz	30 W	45DB	$1,395	In Stk
A0240-50	225-400 MHz	50 W	47DB	$1,995	In Stk
A0240-150	225-400 MHz	150 W	7DB	$1,795	In Stk
A0250-20	200-500 MHz	20 W	43DB	$1,295	In Stk
A0250-30	200-500 MHz	30 W	45DB	$1,495	In Stk
A0250-40	200-500 MHz	40 W	46DB	$1,995	In Stk
A0350-30	300-500 MHz	30 W	45DB	$1,395	In Stk
A2055-10	20-550 MHz	10 W	46DB	$ 895	In Stk
A2055-20	20-550 MHz	20 W	46DB	$1,395	In Stk
A2050-40	20-500 MHz	40 W	46DB	$2,495	In Stk
A0510-10	500-1000 MHz	10 W	40DB	$1,595	Stk-4wks
A0510-20	500-1000 MHz	20 W	43DB	$1,795	Stk-4wks
A0510-30	500-1000 MHz	30 W	45DB	$1,995	Stk-4wks
A0510-75	500-1000 MHz	75 W	6DB	$1,795	Stk-6wks
A0809-20	800-900 MHz	20 W	43DB	$1,695	Stk-4wks

Actual Frequency Response for LDA 2055-20 Amplifier



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frequency, multiplication is not required. By using a relatively high gain VCO, good deviation and linearity can be obtained. To allow a wide modulation bandwidth, a narrow band loop is required. The use of a narrow loop generally necessitates some form of aided acquisition. (The use of a phase-frequency detector common in integrated PLLs is adequate.) The requirement for a phase frequency detector or other means for aided acquisition means that the implementation of this approach is limited by cost, especially for the single channel application considered here. Also, VHF designs usually require a prescaler which adds both to cost and power consumption.

The New Solution

The new solution is to use a low cost FM IC as a Frequency Locked Loop control circuit (1) for a VCO which can be modulated separately. These ICs are now available in very low power technologies and can operate well into the UHF region.

This approach affords a reduction in complexity and spurious emissions over direct crystal modulation or indirect FM for use in VHF/UHF transmitters since the modulated oscillator can be operated directly at the final frequency, eliminating the need for frequency multiplication stages. Also, due to the use of a low cost FM IC, this implementation has both cost and power consumption advantages over designs using PLL ICs, especially in single-channel applications.

Although the specific application described here is a crystal controlled wide band transmitter, the use of this design in a transceiver would allow a significant cost savings since both the transmitter and receiver could share the same crystal and FM IC. This will be discussed in more detail later.

Referring to Figure 1, the operation of the FLL transmitter can be described as follows: The output of a VCO is sampled and mixed down to an intermediate frequency where it is applied to a frequency discriminator (A DC coupled quadrature detector). The output of the discriminator is a voltage proportional to the applied frequency and serves as an error signal. This error signal is integrated and used to control the center frequency of the VCO. If the "sense" of the VCO control signal is correct, (i.e. when the frequency of the VCO deviates, the control signal changes in a manner to counteract the deviation) the VCO frequency will be stabilized. If the gain of the loop is infinite at DC, the VCO will lock to the frequency, $f_{osc} \pm f_{disc}$. Note

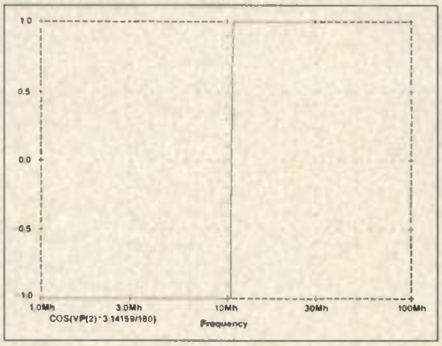

Figure 3. Relative discriminator response vs. frequency.

that two possible output frequencies are available from the topology shown in Figure 1, $f_{osc} + f_{disc}$ and $f_{osc} - f_{disc}$. The particular output frequency is selected by the sense of the loop, i.e. if the loop locks to $f_{osc} + f_{disc}$, it will drive the VCO away from lock at $f_{osc} - f_{disc}$. In this design the loop sense yielding $f_{osc} + f_{disc}$ was used.

Also note that the crystal oscillator is not the frequency reference for the loop, it only serves to translate the frequency of the VCO to an acceptable range for the frequency discriminator. In a sense the crystal oscillator/mixer serves a similar function to a prescaler in a PLL. The "reference" for the FLL is the center frequency of the phase shifting network used in the quadrature detector itself.

System analysis shows that the closed loop response is:

$$f_{out} = (f_{disc} + f_{osc}) \left[\frac{K_d K_i K_{VCO}}{s + K_d K_i K_{VCO}} \right] \quad (1)$$

a standard first order response. Here the finite frequency response of the quadrature detector and VCO have been neglected since these additional poles lie approximately 4 and 6 decades above the dominant pole respectively. A reader familiar with Phase Locked Loops may wonder why the inclusion of an integrator did not result in a second order response. The reason is that in a PLL, the second pole arises from the use of a phase detector in place of the frequency discriminator used in the FLL. A first order response has the advantage of being unconditionally stable (at least in theory).

The response of the loop to modulation, specifically the relationship between V_{mod} and V_{VCO} can be shown to be 1 - H(s) where H(s) is the loop response given in equation 1. Analysis of Figure 1, assuming steady state and letting $s \rightarrow j\omega$ yields a single pole high pass response:

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$$V_{VCO} = \left[\frac{1}{1 - j\left(\frac{K_d K_I K_{VCO}}{\omega}\right)} \right] V_{mod} \qquad (2)$$

By suitably designing the integrator based on the gain of the VCO and discriminator, the low frequency corner of the modulation response can be made arbitrary low. However, one does not have complete freedom in this regard. By examining equation 1 it is apparent that the settling time of the loop is inversely proportional to the low frequency corner in equation 2. (The settling time of the loop can be determined by computing the step response of equation 1.) I have found that if the VCO is well designed, a low frequency corner of 50 Hz (resulting in a 1 percent settling time of 14 ms) provides adequate loop response to compensate for the effects of mechanical shocks and bumps on the VCO even if the equipment is handheld. If desired, a modulation response extending to DC could be obtained by modulating the frequency discriminator. This yields a low pass modulation response. Modulation at both points can be combined to yield an all pass response limited only by the modulation bandwidth of the VCO itself.

Frequency Stability

Aside from the availability of oscillators and mixers in most FM receiver ICs, frequency translation has stability advantages over the use of a frequency divider, specifically it minimizes instabilities arising from the inaccuracies in the frequency discriminator (1). Assuming that discriminator frequency deviates by δ_d and the crystal oscillator deviates by δ_o the VCO frequency will be:

$$f_{out} = f_{disc} + f_{osc} + \delta_d + \delta_o \qquad (3)$$

Note that the frequency errors directly add. If the crystal oscillator/mixer were replaced by a frequency divider then the VCO error would be:

$$f_{out} = N(f_{disc} + \delta_d) \qquad (4)$$

In this case the error of the output frequency is equal in percent to the error of the discriminator. Thus if the crystal oscillator is more stable than the discriminator, it provides a significant stability advantage. Since my application was the generation of wide band FM, this FLL property allowed me to use an LC tank for the discriminator yet retain a stability of better than ± 10 kHz at 100 MHz. To obtain greater accuracy, a ceramic or crystal discriminator could have been used.

Circuit Description and Design

Referring to Figure 2, one can see that most of the functions depicted in Figure 1 are incorporated in the FM receiver IC U1, a Signetics NE605 (4). Many other devices are also appropriate. Note that although the NE605 was designed to be used in a receiver, none of the functions used by the FLL circuit are outside the design limits of the IC.

U1 has an internal crystal oscillator/mixer circuit which has been successfully operated to over 150 MHz using the circuit shown. Higher frequencies, up to approximately 500 MHz have been accommodated using a separate oscilla-

tor/multiplier chain. This approach is preferable to multiplying the VCO directly since it simplifies the suppression of spurious products.

A sample of the VCO (Q1) output is lightly coupled into U1 with a 1.5 pF capacitor. The interstage LC tanks, coupled between pins 14 and 16 and 18 and 20 respectively, are used in place of the ceramic filters usually employed when using the IC for receiver operation.

A single tuned 10.7 MHz quadrature tank coupled between pins 10 and 11, is used as the "frequency reference." Thus the output frequency is $f_{out} = f_{Y1} +$ 10.7 MHz. The resulting frequency discriminator has the characteristic depicted in Figure 3. Note that the discriminator response is monotonic with frequency. This is desirable since it insures that the loop will have gain and the proper sense over a very wide frequency range. Discriminators with a familiar "S curve" characteristic can be used but the VCO range should be limited to insure that operation does not occur outside the center of the characteristic. Here the VCO range is limited to approximately ± 10 MHz.

The integrator is implemented with 1/2 of the CMOS opamp U2. This yields a K_i of 1/(R1||R2)C1. Also note that the response of the quadrature detector has been limited by C2. As stated earlier, this pole is well removed for the dominant loop response pole. It was added to prevent nonlinear distortion in the integrator. When used in this circuit, the integrator constant yields a loop bandwidth of approximately 50 Hz.

The VCO is a standard grounded base Colpitts design. Note that the addition of the modulation signal is implemented by the addition of the capacitances of CR1 and CR2 in the oscillator tank. The use of two varactors allows the bias point of the modulation varactor to be controlled independently from the operating point of the loop. Both CR1 and CR2 are dual devices which were used to limit spurious emissions caused by RF modulation of the varactors. The 1/2 of U2 is used to buffer the modulation input and to bias CR2.

Q2 forms a standard class C amplifier yielding approximately 50 mW output at 90 MHz.

Performance Results

The circuit described above provided greater than ± 200 kHz deviation (deviation was limited to ± 50 kHz due to regulatory considerations) with excellent linearity while retaining a frequency stability of better than ± 10 kHz at 90 MHz which was deemed adequate for the application. Higher frequency stability (at additional cost) could be obtained, without loss of modulation capability, by replacing the LC discriminator tank with a ceramic or crystal discriminator. The use of the FLL topology allowed spurious response to be suppressed by greater than 50 dB (harmonics were less than −30 dBc) using only a single tuned filter at the transmitter output.

A method for the inexpensive implementation of a Frequency Locked Loop wide band FM transmitter has been described. It was shown to provide excellent frequency stability and wide, linear deviation in a low cost/low power design.

Although the application described here was a crystal controlled transmitter, there are several other exciting applications for this technique. For example, if used in a transceiver which required only simplex operation, both the transmitter and receiver could share the same FM IC and crystal. This can be explained as follows, since the output frequency of the transmitter is actually $f_{crystal} + f_{disc}$ and the IF frequency of the receiver is by definition, f_{disc}, the same crystal can be used to make a transceiver which transmits and receives at the same frequency. All that is needed in addition to the circuitry shown in Figure 2 is some switching, (or a coupler) at Pin 1 of U1, and the replacement of the tank circuits, coupled between pins 14 and 16 and 18 and 20 respectively, with ceramic filters. Since the NE605 has two discriminator outputs there would be no need to use a switch there.

For another example, consider that the use of an LC tank as the discriminator in Figure 2 allowed the output frequency to be tuned over several hundred kHz. Also recall that the absolute stability of the discriminator was directly reflected in the output frequency. These facts suggest the application of this design as a reasonable stable tunable LO for receiver applications. Specifically, LC tanks at 455 kHz with stabilities of approximately ± 1 kHz are frequently built for use as discriminators in narrow band FM receivers. Such a circuit could be used in a FLL controlled LO at 200 MHz while retaining ± 1 kHz stability (assuming a stable crystal). This stability figure is entirely reasonable for a continuously tunable receiver in many applications. ***RF***

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VSWR Protection of Solid State RF Power Amplifiers

By H.O. Granberg
Motorola, Inc., SPS

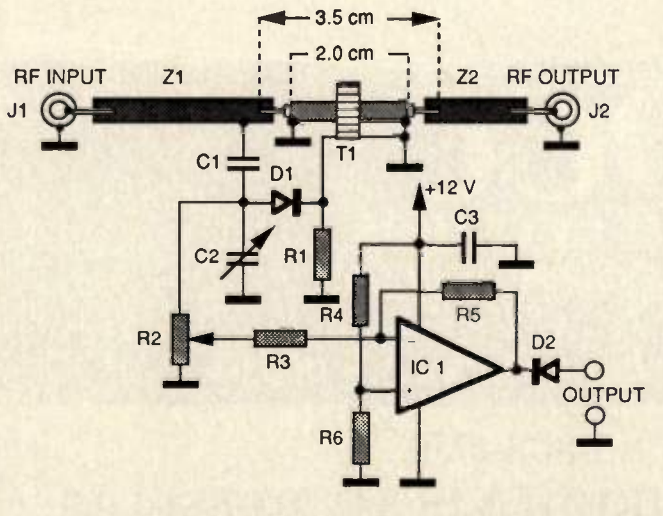
Most transistor failures in solid state amplifiers occur at load mismatch phase angles presenting a high current mode of operation to the transistor(s), which results in an increase in the power dissipated. Since the temperature time constant of a typical RF power transistor die is 0.5 - 1.0 ms, any protection system employed (Including all delays in the AGC/ALC loop) must react faster than this.

Although there are a number of methods to accomplish solid state RF power amplifier protection against load mismatches, the reflectometer VSWR sensing is the most commonly used method. The reflectometer is usually located in series between the amplifier output and the load. A voltage proportional to the amount of output mismatch is obtained from the reflectometer, which is processed accordingly, and fed back to either the power amplifier input or one of the preceding stages in a manner to reduce the power gain gradually or to a complete shut-down.

Principle of Operation

A standard reflectometer principle, presented in numerous publications (1,2,3,4,5) over the years, is used in this design to detect the RF power amplifier output mismatch. It is also commonly known as a VSWR bridge, and its use can be extended up to the microwave frequencies with proper mechanical design. The UHF and microwave designs commonly employ stripline techniques, whereas lower frequency circuits favor lumped constant implementations. In fact up to UHF, the lumped constant concept is probably the most practical means to approach the coupling coefficient required between the current line (amplifier output) and the sample line in order to produce an output voltage of a practical level. A tight coupling in a lumped constant system is achieved by passing the current line through a multiturn pick-up coil, thus forming a transformer, where the current line is the primary and the multiturn coil the secondary. The multiturn winding is usually in the form of a toroid, which allows magnetic material to be used as the core to increase the low frequency response. The inductive reactance of the multiturn winding must be greater than the load impedance of the current line (usually 50 ohms) at the lowest frequency of operation. The high frequency end is limited by leakage inductances and the physical length of the multiturn winding. There will be resonances at each multiple of a fraction of the wavelength, getting stronger the closer one full wavelength is reached. The same can be noted with the design of transmission line transformers (6).

The principle of operation is as follows: the voltage across the multiturn secondary of the transformer is proportional to the current passing through the current line as well as to the frequency of operation. Without duplicating the formulas for determining the coupling coefficient, the secondary voltage available, etc., presented in reference 2, it can be noted that most of these parameters are not critical in an application such as this. In addition to the voltage derived from the secondary of the toroidal transformer, a voltage sample



C1 — 0.25 pF (2 each 0.5 pF chip capacitors in series)
C2 — 5-30 pF ceramic, or equivalent film variable
C3 — 0.1 μF ceramic multilayer
D1 — 1N5711 Schottky Barrier diode or equivalent
D2 — 1N4148 or equivalent
IC 1 — MC34071 or equivalent
J1, J2 — Female chassis-mount BNC or Type N connectors
R1 — 51 Ohms, 1/2 W
R2 — 1 k Ohm one turn trimpot
R3, R6 — 10 k Ohms, 1/4 W
R4 — 1.2 Megohms, 1/4 W
R5 — 1.0 Megohm, 1/4 W
T1 — Primary indicated length of 50 Ohm, 0.014″ O.D. semirigid co-ax, secondary 35 turns of AWG #26 enameled wire wound on Micrometals toroid T 37-6
Z1, Z2 — 50 Ohm etched lines on circuit board

Figure 1. Schematic diagram of VSWR protection circuit.

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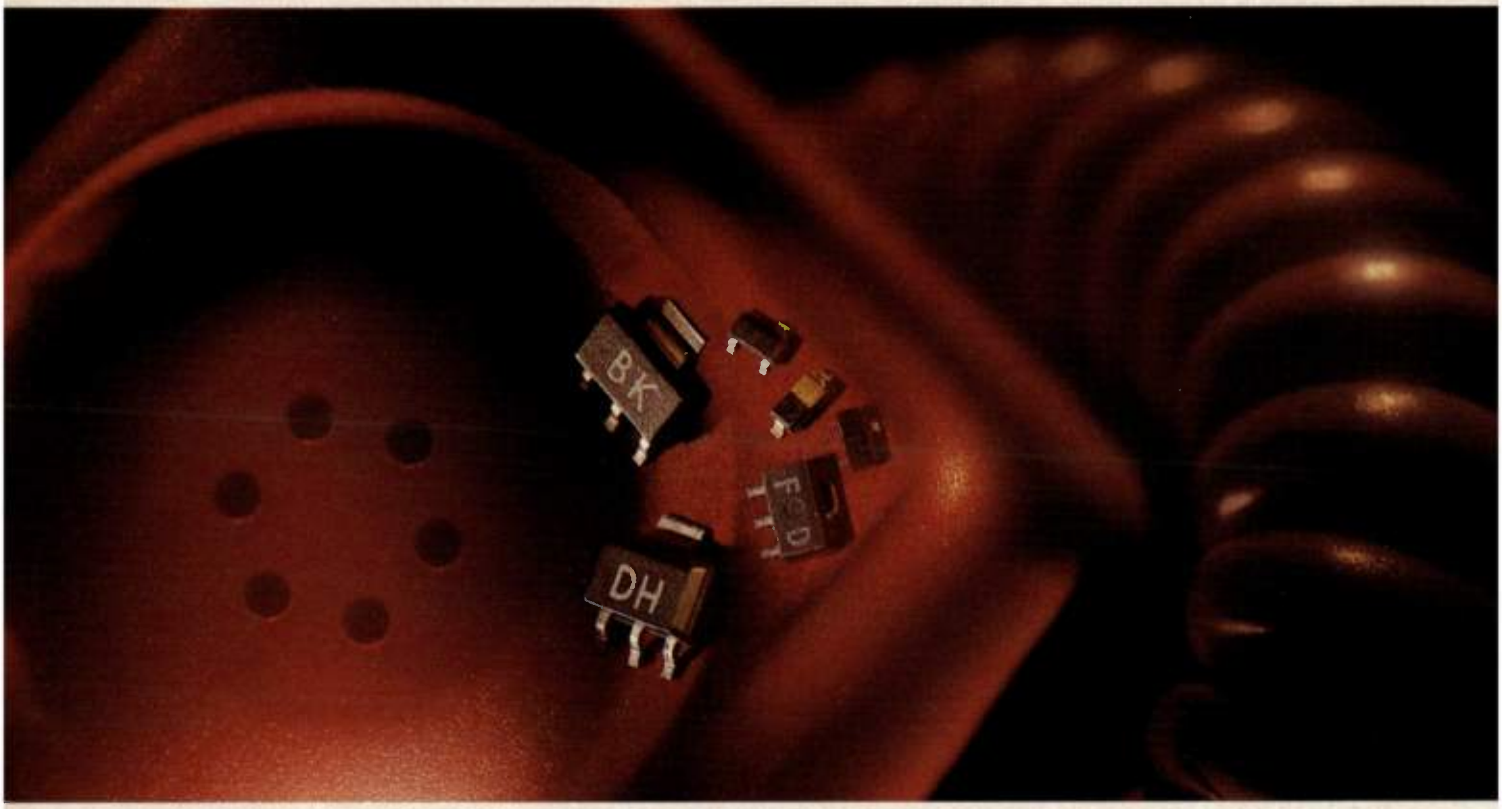

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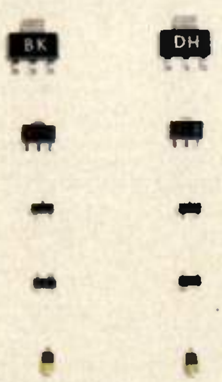

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BFP 180	8	3	1.9	15	1 to 3/0.2 to 2.5 paging system
BFP 280	8	10	1.4	18	1 to 5/0.2 to 8 low-noise pager
BFP 181	12	20	1.3	19	1 to 8/0.5 to 10 low-noise pre-amplifier
BFP 182	12	35	1.2	19	1 to 8/1 to 20 low-noise amplifier
BFP 183	12	65	1.2	19	1 to 8/2 to 28 low-noise amplifier
BFP 193	12	80	1.2	19	3 to 8/5 to 40 low-distortion output stage
BFP 196	12	120	1.4	18	3 to 8/30 to 100 low-distortion output stage

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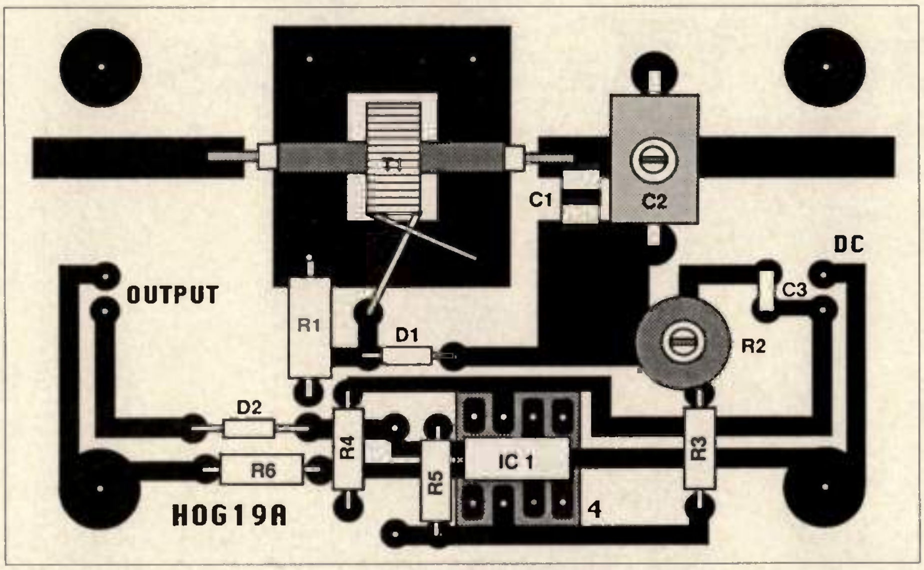

Figure 2. Component layout diagram of VSWR protection circuit viewed from the top side of the circuit board. Note C1 being located at the bottom.

is taken from the current line by means of a capacitive divider C1-C2 (Figure 1). A half wave rectified voltage is brought to the junction of C1 and C2 from the transformer secondary. These voltages are 180 degrees out of phase in case of a non-mismatched load. The amplitudes are adjusted equal with C2 which reduces the voltage to near zero at the junction of C1 and C2 until a mismatch in the load is present which results in the phase shift deviating from 180 degrees.

Electrical Aspects

Because of the principles of operation described in the preceding paragraph, it is difficult to design extremely wideband and high power systems, since for high frequencies the toroidal pick-up coil should be as small as possible, and for low frequencies it should be large enough for the minimum reactance required. High permeability (μ=100 and higher) ferrites in the toroid are usually too lossy at high frequencies and will heat up even at moderate power levels. For example at 150 MHz, materials with μ of only up to 15 have been found acceptable.

A Faraday shield is usually employed between the current line and toroidal winding to prevent the capacitive coupling between the two. It can be best accomplished with a length of coaxial cable of proper characteristic impedance, where the inner conductor forms the current line and the outer conductor the Faraday shield. The circuit details of the unit described here are shown in Figure 1. The forward port has been omitted since the system is not intended for forward power measurements.

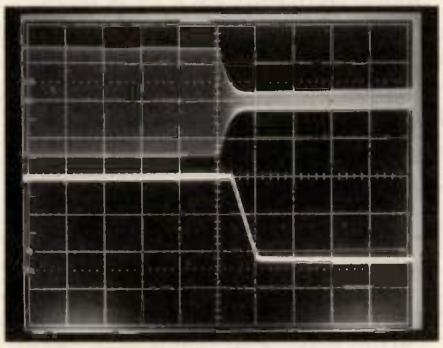

Figure 3. RF envelope and amplifier output. Horizontal scale 2 μs/div. Vertical scale 5.0 V/div.

Mechanical Design

It was an objective to design a unit physically compact enough to extend the frequency response at least 200 MHz and still be able to handle relatively high power levels - at least 500 W at the high frequency end and 1 kW at lower frequencies. Operation down to 2 MHz was expected, but as seen in Figure 4, the sensitivity degrades rapidly. Changing to the -8 material from -6 material for the toroidal core would correct this problem. In addition, a fast operational amplifier was included to provide a shut down of the RF amplifier or gradual power reduction. The slope can be controlled by adjusting the gain of the op-amp with R5 and the sensitivity adjustment R2. This will be discussed later in more detail.

The component layout is shown in Figure 2. A continuous ground plane is located on the lower side covering the top half of the board. It is important that this ground plane be connected to the metal housing at each end. Also the four corner mounting spacers must be grounded to the bottom of the housing. Otherwise the resulting ground loops will affect the operation of the circuit at higher frequencies. This means that the unit must be completely mounted in its housing before it can be reliably tested.

Testing the Unit

The test results shown in Figures 3 and 4 were taken simulating a load mismatch of 5:1 using an open and shorted load through a 2 dB power attenuator and a pin diode switch driven by a pulse generator under the maximum gain and sensitivity conditions. A complete shut-off is accomplished in 2.0 μs (Figure 3), which is fast enough to protect most RF power amplifiers. The operational amplifier (MC34071) used has a 13 V/μs slew rate, and is capable of sinking currents up to 20 mA. This can be used to turn off the bias voltage of a MOSFET or an emitter follower can be added for larger current requirements for controlling a PIN diode attenuator for example. A power handling capability test was performed using three separate amplifiers, specifically 1 kW at 30 MHz, 300 W at 150 MHz and 200 W at 220 MHz. This does not mean that the circuit cannot handle more than the power levels indicated at 150-220 MHz, but these were the highest power signal sources available at the time. There is no reason however, that the power handling capability would considerably degrade at increasing frequencies.

The output of the operational ampli-

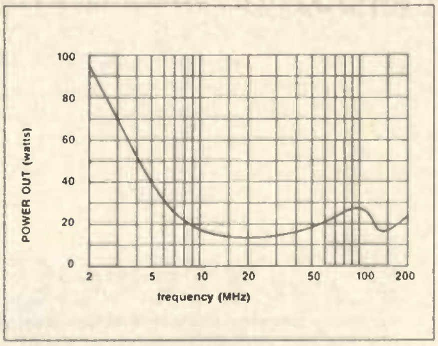

Figure 4. Sensitivity versus frequency response of the VSWR sensor at 5:1 load mismatch.

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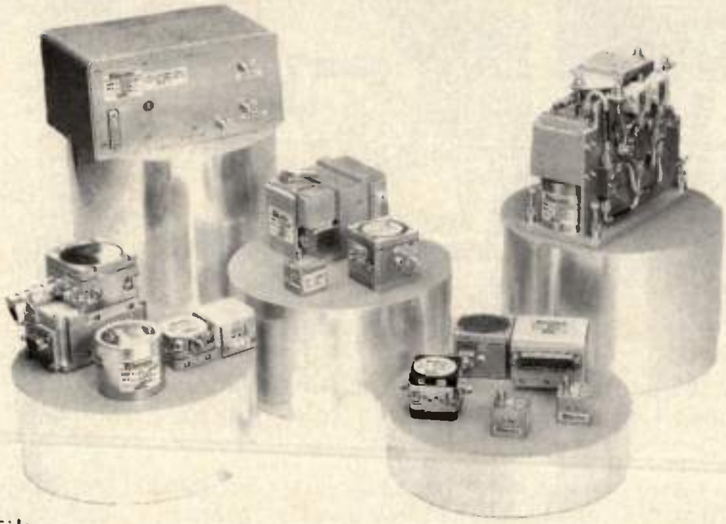

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fier can in fact be made the main bias source to provide the MOSFET gate bias voltage (Figure 7b). Controlling the gate voltage of a MOSFET for a gradual gain reduction, linear operation would not be possible since a steady idle current is required.

In such cases some type of voltage of current controlled RF attenuator must be used, preferably in the low level pre-stages, which are usually operating in class A, and insensitive to variations in the output load. This would be the only way to control the power gain of bipolar transistor amplifiers since the AGC function of MOSFETs is not available. One possibility is a PIN diode attenuator shown in Figure 7c. Depending on the attenuator characteristics and the power level, the power output can be adjusted as desired for a given output mismatch with the combination of R2 and R5. For this, as well as the circuit in Figure 7b, D2 (Figure 1) must be shorted in order to employ the output of IC1 (Figure 1) for a voltage pull-up function. If only a fast shut down of the amplifier is desired without linearity requirements, the circuits in 7a and 7b are adequate and simple. It is recommended that an early stage in the amplifier chain be controlled since low power MOSFETs have low gate input capacitances, which speeds up the shut off. In Figure 7a the FET bias is supplied by an external source, whereas in Figure 7b the bias source is the op-amp output of the VSWR sensor.

Final Notes

The only vulnerable component in the design is C1, which under certain load mismatch conditions may be subjected to high RF voltages. Since C2 is much higher in value than C1, most of the RF voltage will be present across C1. Failures of this voltage sampling capacitor have been known to occur, and for this reason two capacitors were connected in series to increase the voltage rating. In some designs it is comprised of a length of coaxial cable being a part of the current line, where a portion of the outer conductor forms one of the capacitor electrodes.

The operational amplifier (IC1) can be operated at voltages up to 44 V, making it possible to power it directly from the RF power amplifier DC supply. When used with amplifiers operating from supply voltages higher than 44 V, the operational amplifier voltage must be adjusted by a zener diode or a voltage regulator if a single

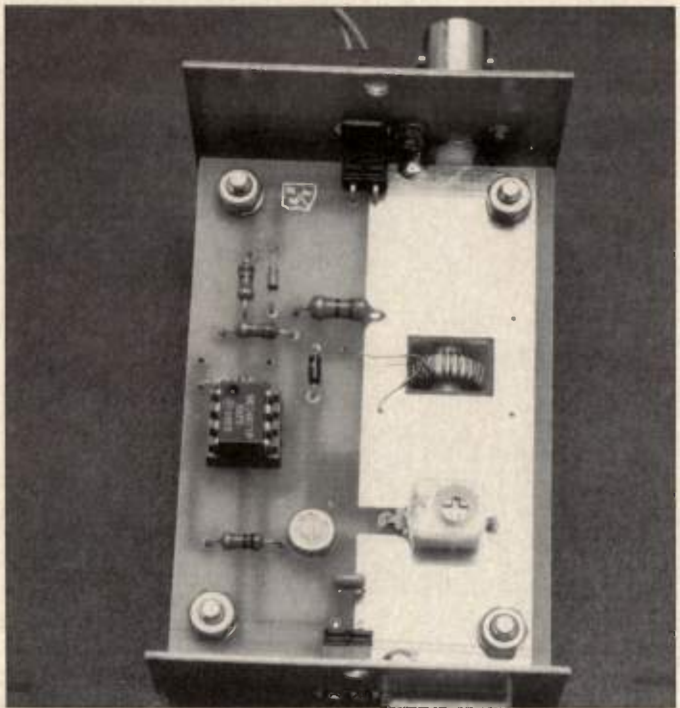

Figure 5. Photo of top of the unit. Note the location of the ground plane in the area of the current line and the pick-up coil.

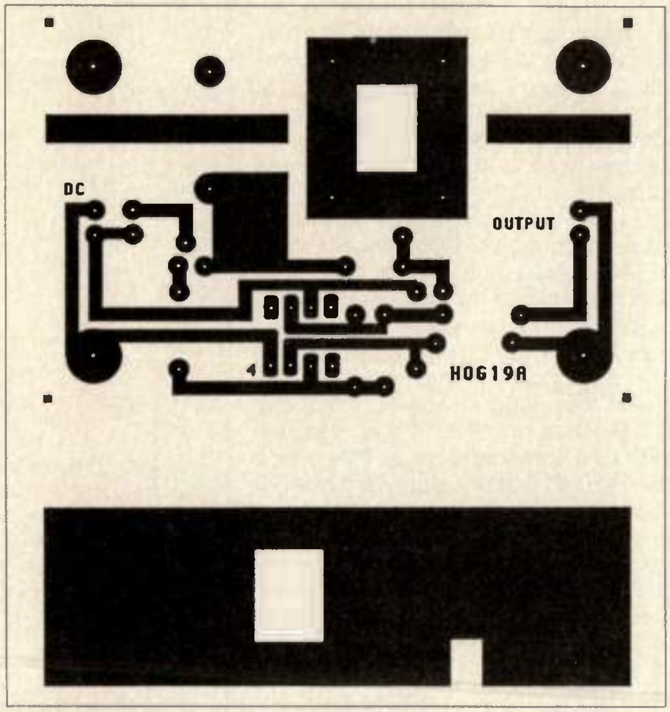

Figure 6. The photomaster of the VSWR sensor. The lower segment is the ground plane in the area of the current line and the pick-up coil (shown actual size).

supply operation is desired. The tests as shown in Figures 3 and 4 were performed using a 12 volt separate power supply. *RF*

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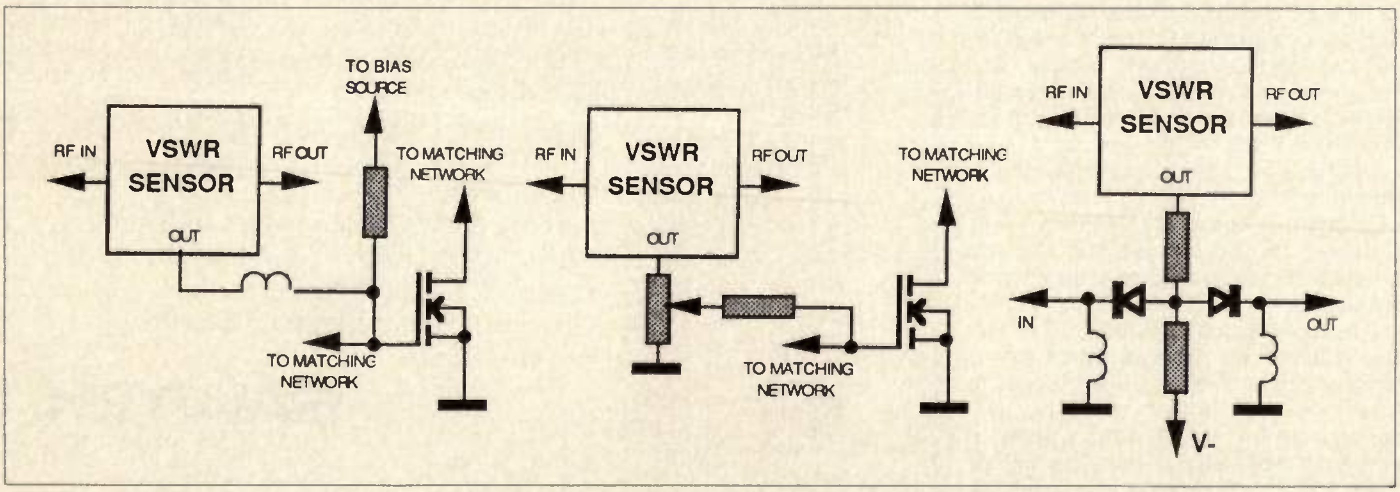

Figure 7. Possible control methods for an AGC feedback loop. 7a and 7b are not suitable if linear operation of the amplifier is required.

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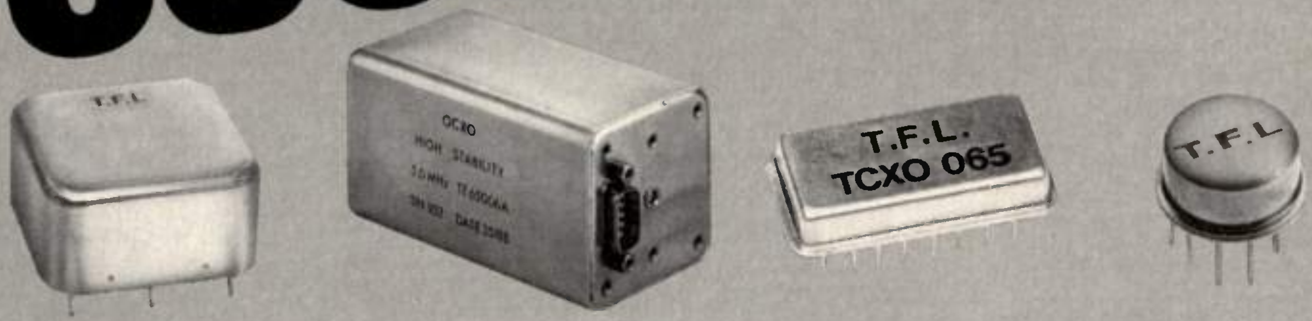

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(0° to +50°C)
Low Aging Rate:
$< 5 \times 10^{-11}$/Day
Low Noise:
< −157 dBc@
10 kHz Offset
Low Vibration Sensitivity:
3×10^{-10}/g
Temperature Range:
−55° to +120°C

TCXO
Temperature Controlled Crystal Oscillator

Frequency Range:
0.02 Hz to 20 MHz
Frequency Stability:
±0.8 PPM
(−40° to +85°C)
Aging: ±1.0 PPM/yr typ.
Supply Voltage:
2 to 15 Vdc
Supply Current:
As low as 1.0 mA
Size:
Standard:
1.5″ × 1.5″ × 0.5″
As small as:
0.960″ × 0.5″ × 0.2″

Complete line of quartz crytals are also available. All oscillators can be manufactured to meet MIL-O-55310, MIL-STD-202, and MIL-STD-810.

CXO
Crystal Clock Oscillator

Frequency Range:
TTL: 10 Hz to 100 MHz
C-MOS: 1 Hz to 5 MHz
ECL: 5 MHz to 500 MHz
Sinewave: 1 Hz-1 GHz
Frequency Stability:
Typ. ±50 PPM (−20° to +70°C, Industrial)
Typ. ±50 PPM (−55° to +125°C, Military)
Up to ±10 PPM available (−20° to +70°C)
Aging:
±10 PPM/yr (Industrial)
±5 PPM/yr (Military)
Outputs: TTL, C-MOS, ECL, Sinewave
Packages: TO-5, TO-8, DIP, Hermetically Sealed Metal Case

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Sampling Delay — Is It Real?

By William Egan
ESL, Inc.

The pages of RF Design *have, over the years, featured various articles concerning the delay associated with the sampling process in phase-locked loops (PLLs) and the resulting phase shift that must be accounted for in loop design, especially in synthesizers. That so much of the technical discussion on this important subject should be associated with this one publication is remarkable. But perhaps as remarkable is the fact that much of what has been said is in apparent disagreement. My intention here is to review the history of this discussion and then to explain the contradictions, both apparent and real. I think that many readers, especially those who already have an understanding of phase-locked synthesizer design, will find this both instructive and interesting. Unfortunately, due to space limitations, there will be little discussion of a tutorial nature but I will try to make up for that by providing references.*

Several articles (1,2) have stated that there is an inherent delay of one reference period, $T=1/F_{REF}$, that must be included in the open-loop transfer function. I wrote a letter (3) stating that the only inherent delay is from an approximation of the hold function (Figure 1) such as is used in a sample-and-hold phase detector (S&H PD) and that its value is T/2. James Crawford then wrote an article (4) agreeing with the T/2 value and stating that such a delay exists even with the more common phase-frequency detector (PFD). Crawford provided convincing evidence of that for two types of loops. Dan Baker heard about this at RF Expo 89 in Santa Clara, probably from a paper by Dan Gavin (5) which referenced Crawford's article. Baker doubted the existence of such a delay and he built a special loop to permit him to test his theory. He presented his evidence, again convincing, in the July 1989 issue of *RF Design* (6). The interesting thing is that both Crawford and Baker were right.

The Sample-and-Hold Phase Detector

For comparison we will start with a discussion of the effect with a S&H PD (7). A S&H PD contains a real sampler and a real hold circuit. When phase modulation at frequency f_m passes through the sampler, many new modulation components are produced (4,8) at $nF_{REF} \pm f_m$ and these circulate around

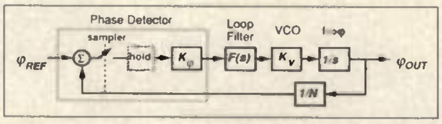

Figure 1. Generic synthesizer loop.

the loop, are sampled again, etc. The process can be represented by z transforms, which do a nice job of taking all of this into account. Two approximations are involved. First, the sampler, which actually is a switch that closes for a very small part of the cycle, is approximated as an impulse sampler, in which sampling occurs instantly. Second, the synthesized frequency should not change too much (on a relative or percent basis) because that would change the sampling frequency and z transforms represent sampling at a fixed rate.

Usually the higher frequencies are greatly attenuated by the loop filter (and even by the hold circuit). Then the products produced by sampling can be ignored and the open-loop transfer function $G(f_m)$ can be represented by the response $G_O(f_m)$ that considers the fundamental alone, as in a continuous system. When the loop is narrow compared to F_{REF}, the effect of the hold

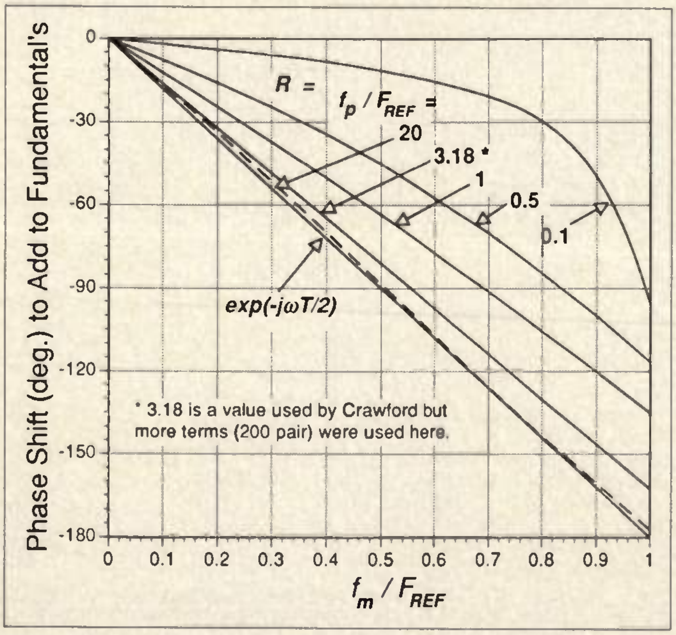

Figure 2. Added phase shift due to sampling, 2nd order type-1 loop with low-pass filter.

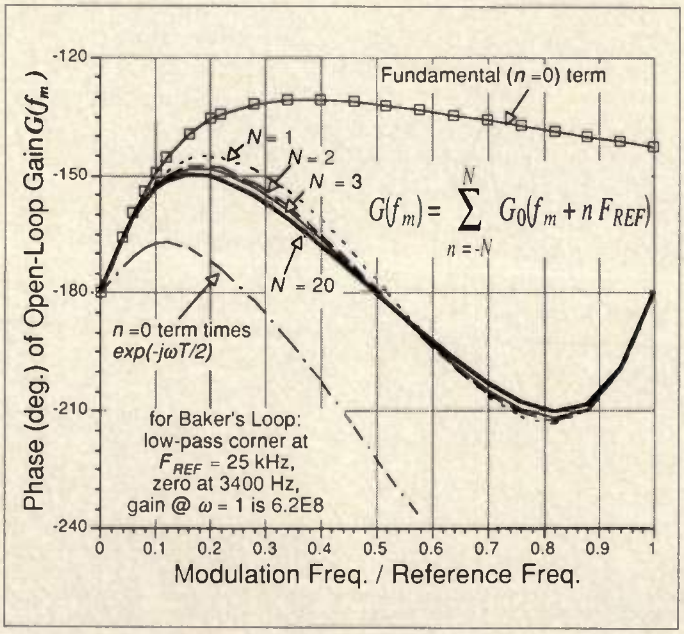

Figure 3. Added phase shift due to sampling, 3rd-order type-2 loop with low-pass filter.

circuit on $G_O(f_m)$ can be ignored. As the bandwidth widens, though, the phase shift produced within the loop bandwidth by the hold circuit becomes more important and it is beneficial to include a phase shift term $\exp(-j\omega T/2)$ to represent this effect. There is also an effect on amplitude but we will concentrate on phase in this article.

The Phase-Frequency Detector

A PFD (9) contains no real sampler and no hold. Phase at the PFD output is represented by the width of a pulse. If the phase change is not too great, we can represent it approximately as an impulse having a value (area) that is proportional to the change in pulse width (10). Then z transforms can be used to represent this loop also. Once again, if higher frequencies are attenuated sufficiently we can concentrate on the fundamental component of the modulation and use LaPlace transforms. As Baker has stated, the hold "model should thus not be included for most PLL designs, including ones using the 4044, 4046 type sequential phase/frequency detector." Whereas with the S&H PD there was a hold circuit to be represented, there is none with the PFD. We just ignore the sampling.

However, if the loop filter does not cut off at a low enough frequency, some of the other frequencies produced by the sampler must be taken into account, as was done by Crawford (11). This is illustrated, for a PLL with a low-pass filter, in Figure 2 where the phase that must be added to the fundamental term is shown for various ratios R of low-pass pole frequency f_p to F_{REF}. As f_p goes higher and higher, ever more products of the sampling process become significant and, eventually, the sum of these components produces a net phase shift that approaches $-\omega T/2$. Since R was 3.18 in Crawford's example (12), $-\omega T/2$ turned out to be a pretty good approximation of added phase. But at more usual values, R = 0.1 for example, it is not.

Figure 3 relates to Baker's loop which contains an integrator-and-lead filter. The phase obtained by adding 20 pairs of sampler-generated products corresponds closely to what Baker shows

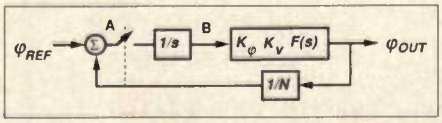

Figure 4. Loop with linear elements reordered.

from his z transform analysis. Because of his loop's unusual bandwidth, frequencies out to $f_m = 0.3\ F_{REF}$ are important. By that bandwidth it is apparent that the fundamental (n=0) term $G_O(f_m)$ is very inaccurate as is the approximation, $G_O(f_m)\exp(-j\omega T/2)$. R is too high for the former and too low for the latter. However, the addition of even one or two pairs of sampling products drastically improves the approximation that uses only the fundamental.

Why is exp(-jωT/2) Sometimes a Good Approximation of Added Phase?

Figure 1, without the hold, could represent a loop with a PFD. In Figure 4 the order of the linear elements has been changed so that we can concentrate on how the sampler interacts with the 1/s term that appears in every PLL. The reordering does not affect ϕ_{OUT}. Let A(t) be a sinusoid that represents the phase error at **A** in Figure 4. During each sampling interval T, an impulse, whose area is equal to A(t)T, is input to the 1/s block, which integrates it (Figure 5). The result is B(t), a stepped output. We will show that D(t), the fundamental component of B(t), is the integral of A(t). It does appear that this may be so in Figure 5; note that D(t) peaks when A(t) goes through zero. If so, then D(t) is the same as the signal that would exist at point **B** if there were no sampling. If, also, the filter function F(s) completely suppresses the other frequencies, the sampling process can be ignored. At the other extreme though, where there is no filter and the wave shape at **B** actually reaches the sampler unchanged, the level at the next sample is the same as that of a sinusoid E(t) which has been delayed by T/2 relative to the fundamental D(t). Thus, in this extreme case, a phase shift of $-\omega T/2$ produces an accurate representation of the delay around the loop.

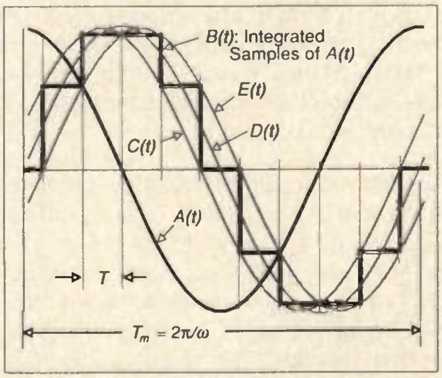

Figure 5. Waveforms at points A and B of Figure4.

Mathematical Basis

Refer to Figure 5. Let $A(t) = \cos(\omega t)$. Then at each sample an impulse of area A(t)T is integrated by 1/s to produce a change in magnitude equal to that area. By using a trigonometric identity (13), this can be written:

$$TA(t) = C(t) - E(t) \tag{1}$$

where

$$C(t) = \frac{\sin\left[\omega\left(t + \frac{T}{2}\right)\right]}{\left[\omega\ \mathrm{sinc}\left(\frac{T}{T_m}\right)\right]} \tag{2}$$

$$E(t) = \frac{\sin\left[\omega\left(t - \frac{T}{2}\right)\right]}{\left[\omega\ \mathrm{sinc}\left(\frac{T}{T_m}\right)\right]} \tag{3}$$

and

$$\mathrm{sinc}(x) \equiv \frac{\sin(\pi x)}{(\pi x)} \tag{4}$$

However, B(t) is the same as a sampled-and-held version of C(t) and that process is known to produce a fundamental component equal to the input sinusoid multiplied by $\exp(-j\omega T/2)\ \mathrm{sinc}(T/T_m)$ (14). Multiplying C(t) by this we obtain the fundamental

$$D(t) = C(t)\exp\left(\frac{-j\omega T}{2}\right)\ \mathrm{sinc}\left(\frac{T}{T_m}\right) \tag{5}$$

$$= \sin\left(\frac{\omega t}{\omega}\right) = \int A(t)dt$$

Q.E.D. Also, since (6)

$$E(t) = C(t)\exp(-j\omega T) = \frac{D(t)\exp\left(\frac{-j\omega T}{2}\right)}{\mathrm{sinc}\left(\frac{T}{T_m}\right)}$$

the phase of E(t) is delayed by T/2 relative to the fundamental.

It would be very nice to have a rule to tell us when we should take the trouble to include the additional terms in the open-loop transfer function. It is difficult to give an all-inclusive rule but I have looked at some loops of the two general types that we have discussed and it appears that, as long as $f_p \leqslant 0.1F_{REF}$, the errors at $f_m \leqslant 0.1\ F_{REF}$, due to not including the extra terms, are less than 2 degrees and 0.4 dB. On the other hand, the effects can be significant with higher values of f_p or f_m. For example, if, using only $G_O(f_m)$, we should design a Type 1 loop to have 32 degrees of phase margin at f_L, if $f_p = 0.1\ F_{REF}$ (apparent $f_L = 0.14\ F_{REF}$), our simplified analysis would lead us to set our gain about 2 dB high. This high gain and the extra phase shift would reduce the true phase margin to only 28 degrees. Far worse, though, would be the effect if f_p should be increased to $0.2\ F_{REF}$ (apparent $f_L = 0.23\ F_{REF}$). We would then be led to set the gain high by more than 6 dB and the loop would be unstable.

Summary

A phase shift of $-\omega T/2$ should be included in the continuous-system representation of a loop with a S&H PD but practically never with a PFD. Accuracy can always be increased by including additional terms in the loop gain to account for the effects of sampling. This is particularly important when the loop bandwidth is wide and the low-pass corner frequency is high. When that corner is very high and the loop has a PFD, the phase shift from all of the additional gain terms approaches $-\omega T/2$. ***RF***

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1. A.B. Przedpelski, "PLL Primer Part III," *RF Design*, July/August 1983, pp. 48-58.
2. S. Goldman, "Divider Delay, The Missing PLL Analysis Ingredient," *RF Design*, March/April 1984, pp. 58A-66A.
3. W. Egan, letter to the editor, *RF Design*, March/April 1984, pp. 9A-12A.
4. J. Crawford, "The Phase/Frequency Detector," *RF Design*, February 1985, pp. 46-57.
5. D. Gavin, "Phase Locked Loop Synthesizer Switching Speed Optimization and Measurement Techniques," *Proceedings RF Technology Expo 89*, pp. 355-372.
6. D. Baker, "A Reference-Canceling Phase/Frequency Detector," *RF Design*, July 1989, pp. 35-40.
7. W. Egan, *Frequency Synthesis by Phase Lock*, Wiley, 1981 and Krieger, 1990, pp. 108-115.
8. J. Ragazzini and G. Franklin, *Sample-Data Control Systems*, McGraw-Hill, 1958.
9. *Frequency Synthesis*, pp. 115-123.
10. We also assume that there is no appreciable feedback while the sampling switch is closed, which should be true in most cases. Otherwise we are forced to some insightful analysis of the representation of the PFD that appears to be related to Crawford's contention (op. cit.) that some low-pass filter, no matter how high in frequency, is required in the model. But that is another subject.
11. Data for Figures 2 and 3 were generated by a slightly modified version of programs provided in Crawford's article (referenced above).
12. $f_p = 2\pi 2\times10^3$ Hz, $F_{REF} = 1\times10^4$ Hz.
13. $\sin a - \sin b = 2\cos[(a+b)/2]\sin[(a-b)/2]$ where $a \to \omega(t+T/2)$, $b \to \omega(t - T/2)$.
14. *Frequency Synthesis*, pp. 99-102, 123-125.

About the Author

Bill Egan is a Principle Engineer at ESL, Inc. He also teaches courses in phase-locked loops and frequency synthesis in Santa Clara University's graduate engineering program. He can be reached at ESL, Inc., Box 3510, Sunnyvale, CA 94088. Tel: (408) 752-2439.

New Software Features EE Utilities and Information

EEPal is a memory resident (pop-up) program designed for an engineer's convenience. It can be accessed while using a wordprocessor or CAD program to view data, use the calculator, and evaluate a formula. The results can then be put into the application program. It handles computations using complex numbers, matrices, and different bases. There are over 200 built-in screens in EEPal covering vendors, parts, units, math, networks, fields, and systems. Up to 520 data, formula, and minisheet screens can be created by the user and added to the program, and data can be moved to and from other programs. Help screens are included and can be made for any user-developed screens. It requires an IBM PC or compatible, 384K of RAM, a hard disk, DOS version 3.0 or later, and is priced at $479.
Eagleware
INFO/CARD #214

Digitally Tuned Closed Loop Filters

Ferretec has introduced a line of digitally tuned closed loop Ferretrac[R] filters covering the 0.5-40 GHz frequency range. This is an extension of the analog tuned Ferretrac filters for use in microwave receiver and test equipment applications.
Ferretec, Inc.
INFO/CARD #213

Toggle Switch Attenuator

Alan Industries has introduced a 75 ohm version of their TX series miniature toggle switch attenuator. Models are available with up to 82.5 dB of attenuation in 0.5 or 1.0 dB increments. The units operate up to 1 GHz and are rated for 1 watt average power and 500 watts peak power. Prices start at $215.
Alan Industries, Inc.
INFO/CARD #212

Medium Power SMT Amplifiers

AML has released a family of low cost medium power PC surface mount amplifiers covering the 10 to 1200 MHz frequency range with output powers from 100 mW to 1 W. Prices start at $119 in quantities of 1-9.
AML, Inc.
INFO/CARD #211

RF Calorimeter

Model 6091 high accuracy calorimeter is used for RF power measurements up to 200 watts and 1.25 percent accuracy. The unit is fully automatic and displays power by digital display. An IEEE 488 bus interface is an integral part of the unit.
Bird Electronic Corporation
INFO/CARD #210

Spectrum Analyzer

The R3271 Microwave Spectrum Analyzer from Advantest America covers the 100 Hz to 26.5 GHz frequency range. It provides more than 100 dB of true dynamic range and a displayable dynamic range of 100 dB. Its uses range from mobile radio equipment to communications/broadcast satellite applications.
Advantest America
INFO/CARD #209

Susceptibility Testing Antenna

Model AT5000 is a combination E-field and H-field generator simulating free-space conditions in a shielded room for RF susceptibility testing of relatively large assemblies. It generates more than 500 volts per meter over a 10 kHz to 100 MHz bandwidth.
Amplifier Research
INFO/CARD #208

Clock Oscillator

CTS' model CXO-63GAU is a high frequency clock oscillator available up to 110.0 MHz. The oscillator features CMOS/TTL compatible outputs, tristate capability, and 47/53 percent typical symmetry.
CTS Corp., Frequency Control Div.
INFO/CARD #207

Low Cost 12-Bit DAC

Burr-Brown has released a low cost 12-bit DAC for direct digital synthesizers and high-speed automatic test equipment. The DAC65P settles to 0.012 percent in 35 ns and has exceptional linearity. It comes in a plastic DIP package.
Burr-Brown Corporation
INFO/CARD #206

Phase-Locked Sources

Magnum Microwave has developed automatic phase-locked sources, modulated phase-locked sources, and crystal oscillators.
Magnum Microwave Corporation
INFO/CARD #205

Miniature PC Mount Filters

Trilithic has developed a line of miniature PC mount filters with a frequency range from 800 kHz to 3 GHz and bandwidths of 1 to 150 percent.
Trilithic, Inc.
INFO/CARD #204

Broadband Quadrature Network

Maxtech introduces the QH100-6A quadrature hybrid covering the entire 2-32 MHz HF band with maximum error of ±0.75 degree phase difference and 0.25 dB amplitude. Isolation between outputs is 20 dB minimum.
Maxtech, Inc.
INFO/CARD #203

Feed-Through Filters

Ceramic Devices has released 0.086 inch diameter, solder-in, feed-through filters from 0.6 to 10,000 pF available in C or L sections. Also available are 0.105 inch diameter screw-in filters available in C, L, T, or Pi configurations.
Ceramic Devices, Inc.
INFO/CARD #202

High Reliability HEMTs

The NE21283A and NE21200 are designed for low noise applications in the 0.5 to 18 GHz frequency range and feature mushroom shaped gates for decreased resistance and improved handling capacity. The noise figure is 1.0 dB typically at 12 GHz and gain is typically 10.5 dB at 12 GHz.
California Eastern Laboratories, Inc.
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Miniature Chip Inductors

Stetco has a new line of miniature chip inductors in a standard 1008 chip package. The chips are available in an inductance range of 4 nH to 12,000 nH, and they come in bulk packaging, waffle packs, or 8 mm tape and reel.
Stetco, Inc.
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SMA Programmable Attenuator

The 50P-766 SMA programmable attenuator covers the DC to 5 GHz frequency range, has a switching speed of 10 milliseconds.
JFW Industries, Inc.
INFO/CARD #199

Cable Bending Tool

The Betta model 04 is a manually operated cable bending tool that easily converts many Mil-D-1000 semi-rigid cable drawings into accurately formed cables. Shipping weight is 70 pounds.
Applied Specialties, Inc.
INFO/CARD #194

RFI/EMI Shielded Enclosure

Compac Development has developed a shielded enclosure which provides 80 dB attenuation to 20 GHz off the shelf. It can be modified to fit various specifications including, hardware, plating, and connectors.
Compac Development Corporation
INFO/CARD #197

SMT Connectors

AEP has released surface mountable coaxial connectors in SMA, SMB, SMC styles. They are available for end- or top-mounting and pull-off forces exceed line to board bonding strength. These parts are suitable for automatic placement machines.
Applied Engineering Products
INFO/CARD #196

Microwave Harmonica™ RF

Microwave Harmonica RF is a nonlinear simulator for RF designs up to 3 GHz. It operates on PCs with OS/2 or DOS. Microwave Success™ is a PC and UNIX workstation-based system simulator with a state-of-the-art user interface.
Compact Software, Inc.
INFO/CARD #195

Laminated Absorbers

AEMI has produced a lightweight, multi-layer laminated radar absorber composed of open cell polyurethane foam layers laminated with flexible adhesives and latex. The AEL series of absorbers is flexible and wrappable, and may be cut to fit most standard shapes.
Advanced ElectroMagnetics, Inc.
INFO/CARD #198

Quadrature Synthesizer Board

The DRFS-DX4660 is based on the DRFS-3250 NCMO™ and operates at clock frequencies from DC to 40 MHz. Quadrature accuracy is within 2 degrees at output frequencies less than 1/4 clock, with typical accuracy better than 0.1 degrees.
Digital RF Solutions Corporation
INFO/CARD #193

Phase Adjustable Connectors

Model 5999 incorporates an SMA male connector and mates with 0.141, or 0.085 semi-rigid cable. Model 5018 is an in-series adapter with male and female SMA terminations, and model 5998 is a similar configuration. All three devices are adjustable coaxial phase shifters operating from DC to 26 GHz.
Connecting Devices, Inc.
INFO/CARD #191

RF Circuit CAE Tool

jOmega combines a harmonic-balance linear/nonlinear circuit simulator with a graphics package that provides schematic entry, multi-tasking simulation control, engineering documentation, and RF board design layout.
EEsof, Inc.
INFO/CARD #192

Zero-Biased Schottky Diode Detectors

The D265B and D265BR feature a frequency range of 0.1 to 26.5 GHz, 550 mV/mW sensitivity, and a VSWR of less than 2.0:1 to 26.5 GHz. These detectors can be used with common oscilloscopes since they do not require DC bias current.
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Broadband Linear Amplifier

ETO's model SL-1DP is a broadband (5-30 MHz) solid state RF linear amplifier capable of providing up to 60 dB of gain and 1000 watts peak RF output for MRI/MRS and similar applications.
Ehrhorn Technological Operations, Inc.
INFO/CARD #189

Tunable SMT Air Wound Coils

Goguen Industries has released new tunable surface mount air wound coils. Goguen has also started a new delivery service that facilitates delivering requested coils as quickly as overnight.
Goguen Industries, Inc.
INFO/CARD #188

Divide-By-8 Module

K&L Microwave has designed a frequency divider module that has an input frequency of 600-1000 MHz and an output frequency of 75-125 MHz. It has an input power range of 0 dBm to −10 dBm and an output level of +2 dBm nominal. VSWR is maintained at 1.7:1 and phase noise is better than 130 dBc per Hz at 100 Hz offset.
K&L Microwave, Inc.
INFO/CARD #187

QPSK Modulator

The HPMX-2001 is a silicon IC which can replace numerous discrete components in the design of a digital radio transmitter. The device can be used by manufacturers of mobile/portable digital cellular telephones, cordless telephones, and RF LANs. It has a typical LO operating frequency range of DC to 2000 MHz and a typical I/Q bandwidth of DC-700 MHz.
Hewlett-Packard Company
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EMP Protectors

Huber+Suhner's new EMP-protectors are used to protect RF and microwave communication equipment against the destructive effects of lighting and nuclear EMP. Quarter wave, waterproof, and EMP small-size protectors are also available.
Huber+Suhner, Inc.
INFO/CARD #184

30 Watt Power Amplifier

Milcom International introduces a 500 to 1000 MHz, 30 watt linear power amplifier.
Milcom International, Inc.
INFO/CARD #185

Test Fixture Mainframe

Inter-Continental Microwave has produced a microwave test fixture that provides accurate measurement data up to 50 GHz in association with a 50 GHz TRL/LRM Microstrip Calibration Standards Kit. The TF-3001-G is priced at $5,000, and the TRL-2007 or 2008 is $4,400.
Inter-Continental Microwave
INFO/CARD #183

TO-8 SMT Package

This open-tooled TO-8 SMT package from Kyocera functions in the DC to 6 GHz frequency range and offers high reliability and compatibility with automated assembly techniques. The SMT TO-8 package meets industry standards for mounting compatibility, electrical performance, solderability, and MIL-STD-883 requirements.
Kyocera America, Inc.
INFO/CARD #182

SMT PIN Diode

The MA4P1250, a square surface mountable PIN diode in a non-rollable MELFT package, has been developed by M-A/COM Semiconductor Products. The unit is a low inductance ceramic package with no ribbons or whisker wires and is designed for use as a low loss switching

element from HF through UHF.
M/A-COM, Inc.
INFO/CARD #181

Bandpass Filters

Anthony RF Products has released miniature and lumped component bandpass filters that cover the 5 MHz to 5 GHz frequency range. Also newly released are tunable bandpass filters available in 3 and 5 resonant sections covering the 40 MHz to 1.5 GHz frequency range.
Anthony RF Products
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Wide Band IF Filter

The MC13155 is a wide band FM IF filter designed for analog video and high speed data applications. It operates up to 130 MHz input frequency and has a demodulation bandwidth greater than 20 MHz. It is available in an SO-16 package.
Motorola Semiconductor Products
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Radar Absorber

Cuming Corporation has released C-RAM SFC, a series of radar absorbing materials designed for applications requiring the simulation of a free space environment. These materials can be used to construct anechoic chambers with −40 dB or better quiet zones at frequencies as low as 500 MHz.
Cuming Corporation
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Cellular Radio Cell Set

Alpha's GaAs MMIC cellular radio cell set consists of control devices for the cellular radio market from 1 MHz to 1 GHz. These MMIC FET devices are designed by a well established, reliable production process.
Alpha Industries
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Coax Connector

The Series BFA miniature 50 ohm coaxial connectors feature a VSWR of less than 1.2:1 through 4 GHz and are offered in both straight and right angle PCB mountable jacks. RF leakage is typically better than −60 dB through 2 GHz.
Murata Erie North America
INFO/CARD #175

Ultra Low Noise Oscillators

The 9000 Series crystal oscillator is available between 40 and 125 MHz and delivers an output power level of +20 dBm and has SSB phase noise of −98 dBc/Hz at 10 Hz up to −173 dBc/Hz at 100 kHz from the carrier.
Techtrol Cyclonetics, Inc.
INFO/CARD #176

Low Power Compandors

The NE577 and NE578 are low power compandors intended for portable communication systems. They typically draw 1.4 mA at 3.6 V. Both chips have one expander channel and one compressor channel, and the unity gain level is programmable between 10 mV and 1 V.
Philips Components-Signetics
INFO/CARD #174

Quadrature Mixer/ Modulator

Synergy Microwave has developed a 34:1 Bandwidth ratio quadrature IF mixer/QPSK modulator.
Synergy Microwave Corporation
INFO/CARD #173

Low Profile OCXO

Piezo Crystal has produced a low profile, low phase noise ovenized crystal oscillator with a frequency range of 5 to 16 MHz. Typical SSB phase noise at 10 MHz is −120 dBc at 10 Hz from the carrier and −150 dBc at 100 Hz from the carrier.
Piezo Crystal Company
INFO/CARD #172

Unit Amplifiers

The UNA family of unit amplifiers cover the 1 MHz to 1 GHz frequency range in values between one-half watt to five watts. They are 50 ohm units and 0.25" x 0.50" in size. Also on display is Transclamp-1, a hold-down assembly for use with flange mounted transistors.
Richardson Electronics
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Digitally Tuned Hopping Filter

The Mini-Pole™ digitally tuned hopping filter operates in the 10 MHz to 1 GHz frequency range with a tuning speed of 10 usec. The filters are two pole Butterworth designs with 3 dB bandwidths ranging from 3 percent to 30 percent.
Pole/Zero Corporation
INFO/CARD #170

SMT Crystals

Raltron Electronics has produced a family of surface mount crystals covering frequencies up to 300 MHz and families of thermally stable SMT networks.
Raltron Electronics Corp.
INFO/CARD #168

Single Chip 1.6 GHz PLL Frequency Synthesizer

QUALCOMM has developed a single chip (44 pin PLCC package) 1.6 GHz PLL frequency synthesizer with an on-chip divide by 10/11 prescaler and digital phase/frequency comparator. Phase noise contribution is as low as −150 dBc/Hz at 20 kHz.
QUALCOMM, Inc.
INFO/CARD #169

Shielded SMT Chip Inductors

The GLD series of inductors range from 10.0 to 330.0 uH in inductance and include ferrite filler in their encapsulation to reduce magnetic coupling. They are 3.2 x 2.5 x 2.2 mm in size and start at $0.41 each in quantities of 1,000.
Sprague-Goodman Electronics, Inc.
INFO/CARD #167

SMA Connectors

The SMA series includes connectors for semi-rigid and flexible cables, bulkhead connectors, stripline and microstrip, hermetic sealed, and PC boards in frequencies ranging from DC to 25 GHz with 50 ohms impedance.
Rosenberger/Micro-Coax
INFO/CARD #166

Coaxial Switches/ Connectors

Automatic Connector introduces lines of coaxial switches and blind mate connectors.
Automatic Connector, Inc.
INFO/CARD #165

Broadband Isolators

HYPER-MODE™ broadband isolators span 2-7 GHz, 4-18 GHz, and 3-20 GHz frequency ranges. They feature up to 16 dB and excellent insertion loss levels. All HYPERMODE isolators are rated for 10 watts average power.
Teledyne Microwave
INFO/CARD #164

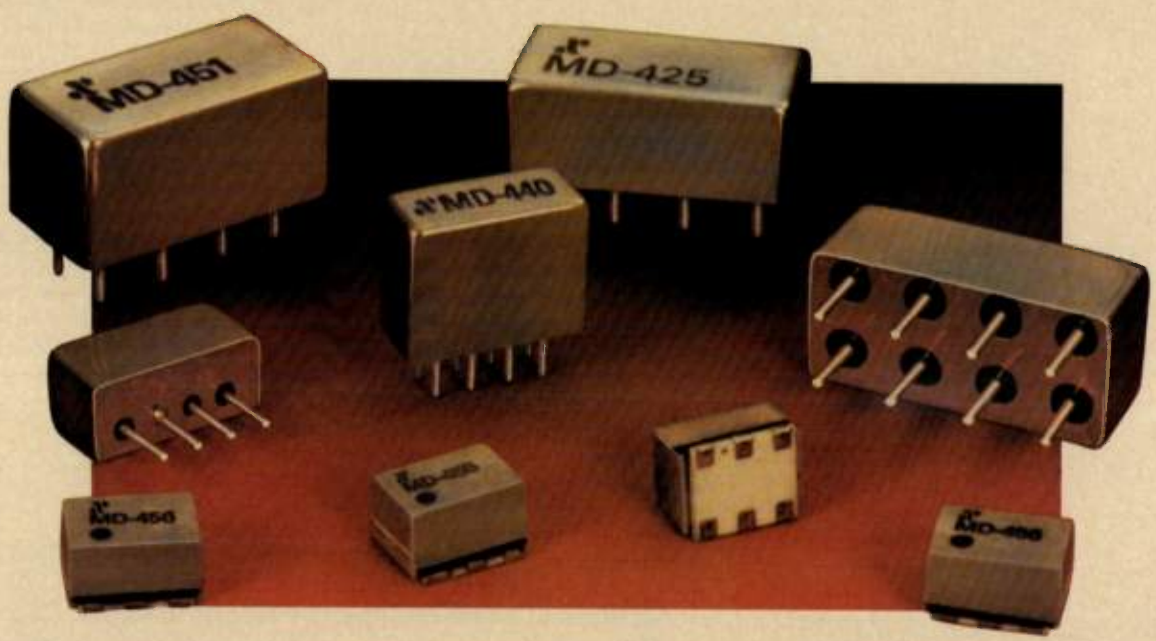
Wirepac™ Hybrid

Model FWP604 Wirepac hybrid operates from 1.7-2.5 GHz with an insertion loss less than 0.2 dB, VSWR of 1.20:1 maximum, and isolation greater than 20 dB. Power handling capability is 500 watts average and 2500 watts peak.
Sage Laboratories, Inc.
INFO/CARD #163

12 Bit, 100 MSP DAC

The TKDA30 is a 12-bit, 100 MSPS minimum DAC used for applications in digital frequency synthesis, arbitrary waveform generation, professional video reconstruction, instrumentation, and CAD/CAM workstations.
Tektronix, Inc.
INFO/CARD #162

ANZAC DELIVERS *The Alternative*



ANZAC's MD-400 Series of Mixers — The Quality Alternative.

Until now, buying a low cost mixer may have meant receiving a low quality device. With ANZAC's new MD-400 Series of mixers you get everything you'd expect from an ANZAC product at prices you might expect from the other guy.

Performance

Model No.	Frequency Range RF/LO (MHz)	IF (MHz)	Conv. Loss (dB) Typ	Isolation (dB) Typ LO-RF	LO-IF	Comp. Point (dBm) Typ	LO Drive (dBm)
MD-405	0.1-250	DC-250	5.0	48	45	+13	+17
MD-401	0.2-400	DC-400	6.0	50	45	0	+7
MD-428	1-400	DC-400	5.5	55	45	-1	+7
MD-410	0.5-500	DC-500	5.0	50	40	-2	+7
MD-413	0.5-500	DC-500	6.0	50	45	+9	+17
MD-411	0.6-500	DC-500	5.0	50	35	-2	+7
MD-426	1-500	DC-500	5.5	50	40	0	+7
MD-455	1-500	DC-500	7.0	50	45	+1	+7
MD-403	2-500	DC-500	6.5	50	45	+13	+17
MD-441	2.5-500	DC-500	6.0	60	45	+1	+7
MD-450	5-500	DC-500	6.5	45	45	+17	+23
MD-435	0.5-600	DC-600	5.5	55	45	+2	+7
MD-440	1-1000	DC-1000	5.5	45	40	-1	+7
MD-400	1-1000	DC-1000	6.5	50	45	+2	+7
MD-412	1-1000	0.5-500	5.0	40	30	-1	+7
MD-414	2-1000	DC-1000	6.5	45	35	+9	+17
MD-404	5-1000	DC-1000	6.5	48	45	+14	+17
MD-456	5-1000	DC-1000	7.0	40	35	+2	+7
MD-427	10-1000	5-500	5.5	50	40	+1	+7
MD-406	5-1200	DC-1200	7.5	45	40	+13	+17
MD-402	5-1250	DC-1250	6.5	45	40	+2	+7
MD-415	2-2000	0.5-500	7.0	45	35	+5	+10
MD-425	5-2000	10-600	8.0	45	35	+2	+7
MD-407	10-3000	10-800	7.5	40	25	+5	+10
MD-416	10-3000	10-1000	7.5	35	25	+10	+17

Specs per Adams-Russell Components Group 1989 RF & Microwave Signal Processing Components catalog. For more information contact the factory.
*MD-426 1-24 units (higher quantity pricing available).

Quality

MD-400 Series mixers meet stringent military standard and come with ANZAC's 1 year warranty.

Competitive

Series 400 mixers start as low as **$5.90*** and are available **FROM STOCK**, in a variety of relay header and surface mount package styles.

Call today for more information about our complete line of quality, low cost mixer alternatives.

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RF expo products continued

Hybrid Log Amplifiers

Teledyne Microelectronics has developed DLVA, SDLA, and true log amplifiers with frequency ranges of 0.5 to 26 GHz, 50 to 800 MHz, and 5 kHz to 100 MHz respectively. Each has a log accuracy of ±0.75 dB, ±1.0 dB, and ±1.0 dB respectively and a dynamic range greater than 70 dB.
Teledyne Microelectronics
INFO/CARD #161

Frequency Synthesizer

The VDS-1306 frequency synthesizer has a bandwidth between 55 MHz and 85 MHz with a step size of 100 Hz standard. Spurious signals are less than −55 dBc and phase noise is less than −95 dBc/Hz at 10 Hz to than −115 dBc/Hz at 100 kHz.
Sciteq Electronics, Inc.
INFO/CARD #160

Unity Gain Delay Module

The MRH801 is a unity gain, wideband microwave delay module based upon BAW technology. It provides a 1 usec delay from 2.5 to 5.3 GHz and uses high linearity MMIC amplifiers to achieve ±3 dB loss variation over a military temperature range.
Thomson-ICS
INFO/CARD #159

Non-Polarized Capacitor

With a frequency range from 100 kHz to 100 MHz, the 900 C capacitor is rated at 200 VDC, from −55 to +125 degrees Celsius, and is a high current component designed to replace tantalum and aluminum electrolytic capacitors. Pricing is $3.00 in quantities of 10,000.
American Technical Ceramics
INFO/CARD #158

SMT Helical Filters

Series 5CH is available in center frequencies from 400-600 MHz, and Series 5CHL is available in center frequencies from 800-1100 MHz. The 5CH series comes in 2 and 3 pole designs with 10 MHz minimum bandwidths at 1 dB. The 5CHL series comes in 2 pole designs with 25 MHz minimum bandwidths at 1 dB.
Toko America
INFO/CARD #157

Microwave Probe Manipulators

The PR/PV210 series manipulators are designed for use in conjunction with the Tektronix TMP-9600 series of microwave probe feet. Features include precise positioning control, magnetic or vacuum base, and radial arm movement for flexible positioning.
Wentworth Laboratories
INFO/CARD #156

RF Signal Generators

The PM 5327 and PM 5328 RF signal generators offer FM stereo modulation capabilities and are used in the test and repair of consumer audio electronics. The 180 MHz signal generators can be used either as a stand-alone or, using the GPIB interface, in test systems.
John Fluke Manufacturing Co.
INFO/CARD #155

RF Power Transistors

The F-1500 series of RF power MOSFET transistors from Polyfet RF Devices is designed for applications in the 500 MHz region. Power outputs range from 10 to 35 watts with gain of 12 to 13 dB.
Polyfet RF Devices
INFO/CARD #154

Non-Reflective SPST

This non-reflective SPST switch with integral TTL driver covers the 0.5 to 6.0 GHz frequency with isolation greater than 80 dB up to 2 GHz. It has field-replaceable connectors and is priced at $550.
Custom Microwave Components, Inc.
INFO/CARD #153

Limiter Schottky Detectors

Models ACLS-3562N through -3575 series of coaxial and module limiter-schottky detectors cover the frequency ranges from 100 MHz to 20 GHz and include narrow and broadband devices which are rated at 100 watts peak power and 2 watts CW.
Advanced Control Components
INFO/CARD #152

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★ ★ ★ ★ ★ ★ ★ ★ ★ ★ ★ **RF Design Awards** ★ ★ ★ ★ ★ ★ ★ ★ ★ ★ ★

The <u>DESIGN</u> Contest

Grand Prize
HP 8591A Spectrum Analyzer!
provided by Hewlett-Packard Company

The Grand Prize for the Design Contest is a HP 8591A portable spectrum analyzer for design, production, or field service. A few key specs include 9 kHz to 1.8 GHz coverage, stable synthesized LO with 2 kHz accuracy and −95 dBc/Hz phase noise at 30 kHz offset, −115 dBm noise floor of +30 dBm maximum signal, 70 dB calibrated display range, and a 1 kHz to 5 MHz resolution bandwidth. IEEE-488 interface is provided, with internal firmware for limit lines, trace math and high level functions.

Second Prize
RF System Simulation Software!
provided by TESOFT

The TESS block diagram simulator is shown here operating with OrCAD/SDT III schematic capture under Windows 3.0. Second prize includes this software plus the MODGEN model generator and Microsoft FORTRAN, with which users can add new models to TESS.

Third Prize (Both Contests)
RF Component Package!
provided by Adams-Russell, ANZAC Division

The Third Prize in each contest is a package of RF components from ANZAC. Stock your lab with these mixers, couplers, transformers, and other devices.

For Contest Rules, See the January Issue of *RF Design*, Pages 30-31

Entries Must Be Postmarked By March 30, 1991 and Received No Later Than April 10, 1991!

Send Entries to:
RF Design Awards Contest
***RF Design* magazine**
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111

A Special Prize That Any Entrant Might Win!

This Consolation Prize will be awarded to an engineer drawn at random from all of the non-winning entries. Even if you don't capture First, Second or Third Place in either contest, you could win this valuable prize!

★ ★

★ ★ ★ ★ ★ ★ ★ ★ ★ ★ ★ RF Design Awards ★ ★ ★ ★ ★ ★ ★ ★ ★ ★ ★

The PC SOFTWARE Contest

Grand Prize
Macintosh computer system with RFDesigner software!

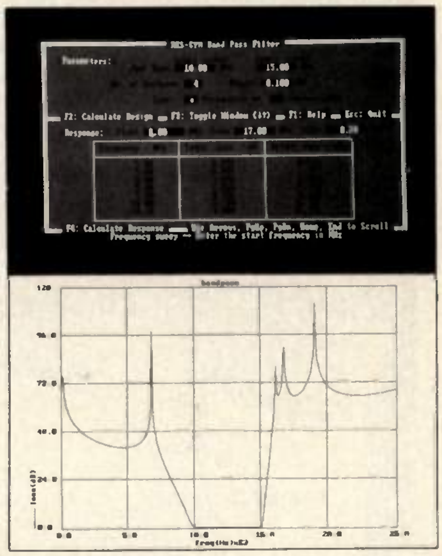
provided by ingSOFT Limited

The Software Contest Grand Prize includes a complete RF design computer system. Hardware includes a Macintosh model SE/30 with monochrome monitor, 4MB RAM, 80MB hard disk, and dot matrix printer. The RF/microwave design software package includes ingSOFT's RFDesigner™ for analysis and optimization, RFSynthesist™ for circuit synthesis, and Engineering Utilities, plus DesignWorks™, MiniCad+™, TASTE™, Excel™, Stepping Out™, DiskTop™, Suitcase™, and Prograph 2.0.

Second Prize
S/FILSYN Student Edition!

provided by DGS Associates

S/FILSYN provides designers with highly professional filter design and synthesis software, handling all filter formats—passive, active, microwave, digital, and all passband/stopband configurations. The student edition, distributed exclusively by Besser Associates, contains nearly all features of S/FILSYN.



Everyone Wins — An RF Design Awards T-Shirt!

Entrants in both contests will receive a T-shirt for their hard work. The block diagram identifies you as a true RF engineer. Provided by TESOFT.

Consolation Prize
RSG-10 Signal Generator

provided by Ramsey Electronics, Inc.

The RSG-10 is a synthesized signal generator designed for engineering, general testing, or production. 1 kHz to 1 GHz frequency coverage is provided, with calibrated AM/FM modulation, and output of −130 to +10 dBm. A programmable memory can store and execute up to 20 test setups.

★ ★

Power Amplifiers

Model A650 series of HP amplifiers delivers up to 500 watts of CW output power the 1.0-18.0 GHz frequency range. They are suited for use as communication amplifiers in susceptibility test systems and antenna test systems and are available in two chassis or ruggedized versions.
LogiMetrics, Inc.
INFO/CARD #151

IF SMT Filters

FSY has released a line of IF surface mount filters covering the DC to 1 GHz frequency range in standard sizes of 0.45 × 0.45 × 0.17 and 0.40 × 0.80 × 0.17. Other frequencies and configurations are available.
FSY Microwave, Inc.
INFO/CARD #150

Attenuator Drivers

Impellimax has developed a family of attenuator drivers that provides tailored currents for PIN attenuators or tailored voltages for GaAs attenuator MMICs. Analog and digital inputs are available, and only set resistors are needed for a complete attenuator driving circuit.
Impellimax, Inc.
INFO/CARD #149

INFO/CARD 107

Phase Shifter

The PWM-84C, an 8-bit digitally driven, analog phase shifter, offers 360 degrees of shift with 1.4 degrees resolution. The phase shifter is available in various bandwidths up to an octave with center frequencies from 20 to 3,000 MHz.
Merrimac Industries, Inc.
INFO/CARD #148

TV Linear Amplifiers

Lucas introduces solid-state RF amplifier products with power levels from 5 watts to several kilowatts.
Lucas Aerospace Communications and Electronics
INFO/CARD #147

Monolithic Schottky Bridges

M-Pulse Microwave has released monolithic Schottky bridge mixers/detectors designed for high volume, high performance, applications where a simple solder or epoxy installation is desired. They are available in multiple barriers with a capacitance range of 0.10 pF to 0.70 pF.
M-Pulse Microwave
INFO/CARD #146

System Analyzer

Motorola announces the R-2600A RF System Analyzer. Includes a 1 GHz generator and receiver, spectrum analyzer, counter and storage scope. Prices start at $10,700.
Motorola Test Equipment Products
INFO/CARD #145

Microwave Laser Plotter

The N524000-MLPP system plots a microwave mask from an AutoCAD drawing file in less than one-half hour, with no post processing. The system uses a high power laser diode and unique optics that have a long depth of field, which eliminates the need for auto focus.
Newport Electro-Optics Systems
INFO/CARD #144

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ANNOUNCING...

RF DESIGNS' FILTER HANDBOOKS

Each handbook contains 80 compiled pages of our best (and most requested) articles on Filter Design, and Filter Applications. These Filter articles and tutorials come directly from the pages of *RF Design* magazine, and are written by experts from across the RF industry. Each Handbook is filled with articles that target every level of design experience and application...articles which will bring new engineers up to speed quickly on Filter basics...as well as articles that will teach experienced engineers how to maximize Filter performance at the most complex level.

Volume I
Filter Applications Handbook

Volume II
Filter Design Handbook

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Cable Equalizer Design

By Frederick J. Radler
Trilithic, Inc.

A problem that has plagued RF engineers in the CATV industry and others when designing systems that transfer RF energy over long coaxial cables, is the loss phenomena that attenuates the higher frequencies more than the lower frequencies. This can be a severe problem when the bandwidth is several octaves as it is in some broadband distribution systems.

The cable loss is related to the square root of the frequency which is similar to an LC resonance curve. Using the characteristic of an LC network is the simplest solution by inserting this network before any gain stage. The ideal case is a circuit that has exactly the opposite transfer loss characteristics of the cable being used, so that the net system amplitude response would be flat.

Two common types of networks will be presented and a computer program will be made available to design, analyze, and edit several configurations of the Bridged T equalizer networks.

Although many network configurations exist, the one chosen for presentation is the basic Bridged T device because of its ability to maintain relatively constant impedance match over all frequencies in the design range.

The Cable

First an investigation of the cable losses is necessary to understand and establish the parameter that must be compensated.

The standard matched line loss equation is:

$$\frac{\text{Att.}}{100\text{ ft.}} = \frac{4.34R_t}{Z_0 + 2.78FF_p\sqrt{e}} \text{ in dB} \qquad (1)$$

Where: F_p is the dielectric power factor at frequency

e is the dielectric constant relative to air

F is the frequency in MHz

Z_0 is the characteristic impedance of cable

R_t is given by

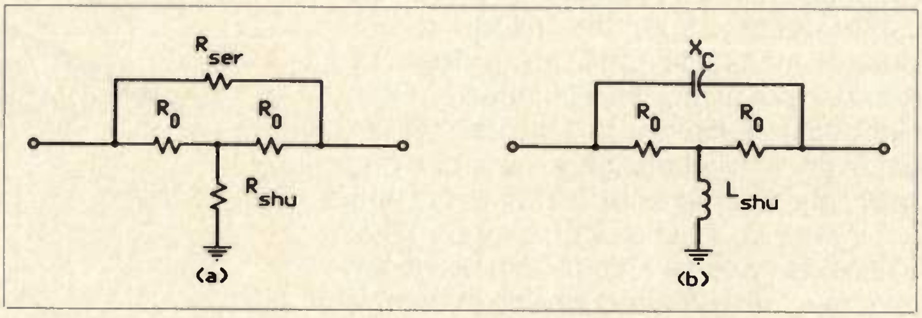

Figure 1. (a) A bridged T attenuator. (b) A bridged T equalizer.

$$R_t = .1\left(\frac{1}{d} + \frac{1}{D}\right)\sqrt{F} \qquad (2)$$

This equation holds for copper lines, see below for other material

Where: d is the diameter of the center conductor

D is the inner diameter of the outer conductor

For other conductor materials the terms (1/d) and/or (1/D) must be multiplied by

$$\sqrt{\frac{f\mu_r p}{p_{cu}}} \qquad (3)$$

Where: μ_r is the relative permeability of conductor material.

μ_r = 1 for copper and other non-magnetic materials

p is the resistivity of material

p_{cu} is the resistivity of copper, 172.4 x 10^{-6} ohms/meter

f is frequency in cycles per second

F_p, if not given, can be estimated by

$$F_p = \frac{G}{2\pi fC} \qquad (4)$$

Where: G is the conductance of line in mhos/unit length - Use G_0 for low loss line, and C is the line capacitance in pF.

These are generic equations (2) and are used in the computer program to calculate the amplitude response parameter of the line being compensated.

The Equalizer

To understand the approach for the circuit design, the Bridged T attenuator as it relates to the basic Bridged T equalizer must be understood first. A comparison is shown in Figures 1a and 1b. For Figure 1a the relationships are:

$$R_0 = R_0' = R_L = Z_0 \qquad (5)$$

$$(R_{shu})(R_{ser}) = R_L^2 \qquad (6)$$

$$R_{ser} = R_L(K - 1) \qquad (7)$$

Where: K = antilog(dB/20)

The transfer function:

$$\frac{E_{in}}{E_{out}} = K = \frac{R_{ser} + R_L}{R_L} \qquad (8)$$

Ignoring distributed effects, Figure 1a would work from DC to light but our goal of compensating the cable loss would not be satisfied, so we have to relate Figure 1a to Figure 1b in order to solve for the reactive element values. Since the following should already be known: F_{Lo} is the lowest frequency of operation, F_{Hi} is the highest frequency of operation, and attenuation at F_{Hi} is the loss at the highest frequency (Cable span/dB) and from the relationships of Figure 1a where:

$$\frac{E_{in}}{E_{out}} = K = \frac{R_{ser} + R_L}{R_L}, \text{ then } K = \frac{X_c + R_L}{R_L} \qquad (9)$$

Four reasons to buy Kay Attenuators.

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✓ **Price**
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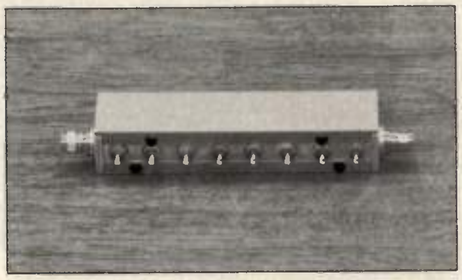

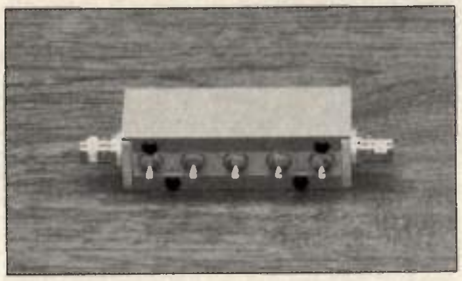

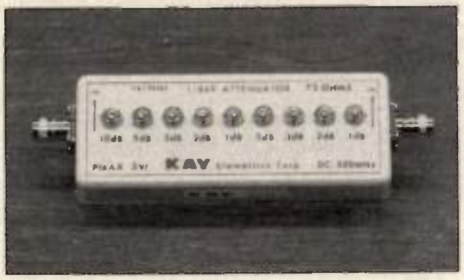

Kay Elemetrics Corp offers a complete line of Programmable, In-Line, Toggle Switch and Continuously Variable Attenuators. We can supply you with the right model to fit your specific application.

If you don't see an Attenuator that fits your requirements call us. Kay can design an Attenuator to fit your specific OEM needs. We have designed over 250 Attenuators in the past. Listed below are some typical Attenuator models.

Model#	Imped.	Freq. Range	Atten. Range	Steps
Programmable				
4440	50Ω	DC-1500MHz	0-130dB	10dB
4450	50Ω	DC-1500MHz	0-127dB	1dB
1/4450	50Ω	DC-1000MHz	0-16.5dB	.1dB
4457	75Ω	DC-1000MHz	0-127dB	1dB
1/4457	75Ω	DC-750MHz	0-16.5dB	.1dB
4460	50Ω	DC-1500MHz	0-31dB	1dB
4467	75Ω	DC-1000MHz	0-31dB	1dB
4480	50Ω	DC-1500MHz	0-63dB	1dB
4487	75Ω	DC-1000MHz	0-63dB	1dB
In-Line				
837	50Ω	DC-1500MHz	0-102.5dB	.5dB
838	50Ω	DC-1000MHz	0-41dB	1dB
839	50Ω	DC-2000MHz	0-101dB	1dB
1/839	50Ω	DC-1000MHz	0-22.1dB	.1dB
847	75Ω	DC-1000MHz	0-102.5dB	.5dB
849	75Ω	DC-1500MHz	0-101dB	1dB
1/849	75Ω	DC-500MHz	0-22.1dB	.1dB
860	50Ω	DC-1500MHz	0-132dB	1dB
870	75Ω	DC-1000MHz	0-132dB	1dB
Continuously Variable				
0/400	50Ω	DC-500MHz	1-13dB	—
0/410	75Ω	DC-400MHz	2-14dB	—

For more information ***call Kay's Product Specialist today at (201) 227-2000, ext. 104.***

KAY

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✓	Model No.	Frequency Response	SS Gain (dB) Min.	Flat-ness (dB) Max.	Noise Figure (dB)	PO@ 1dB C (dBm) Min.	VSWR In/Out	VDC	Cur-rent (mA)	Case/ Connector
	Wide Band / Low Noise Amplifiers									
	W40C	1MHz—40MHz	42	±.5	1.0 Typ 1.2 Max	+5	2:1	+15	20	C/SMA
	W50ETC	10KHz—50MHz	24	±.5	5.3 Typ 6.0 Max	+23	2:1	+15	125	E-75/BNC
	W50ATC	10KHz—50MHz	50	±.5	1.3 Typ 1.5 Max	+5	2:1	+15	25	C-75/BNC
	W110F	5MHz—110Mhz	55	±.5	1.1 Typ 1.2 Max	+15	2:1	+15	80	C/SMA
	W110H	5MHz—110MHz	30	±.5	1.2 Typ 1.4 Max	+5	2:1	+15	30	C/SMA
	W500K	1KHz—500MHz	30	±1	1.7 Typ 2.2 Max	+3	2:1	+15	25	C-75/BNC
	W500C	5MHz—500MHz	40	±.5	1.4 Typ 1.6 Max	+10	2:1	+15	50	C/SMA
	W500EF	5MHz—500MHz	60	±.5	1.3 Typ 1.4 Max	+20	2:1	+15	190	A/SMA
	W500H	5MHz—500MHz	33	±.5	1.2 Typ 1.4 Max	+5	2:1	+15	25	C/SMA
	W1G2M	10KHz—1000MHz	30	±1	2.0 Typ 3.0 Max	+5	2:1	+15	35	C-75/SMA
	W1G2H	5MHz—1000MHz	30	±.5	1.3 Typ 1.5 Max	+5	2:1	+15	40	C/SMA
	W2GH	500MHz—2000MHz	22	±1	4.0 Typ 4.5 Max	+5	2:1	+15	30	C/SMA
	WFR1-4GA-14	100MHz—4000MHz	28	±1	3.5 Typ 4.0 Max	+14	2:1	+15	100	A-75/SMA
	Medium Power Amplifiers									
	P150D	35KHz—150MHz	27	±.5	5.0 Typ	+30	2:1	+24	400	H/SMA
	P150M	500KHz—150MHz	26	±.5	5.0 Typ	+30	2:1	+24	600	H/BNC
	P150ML	400KHz—150MHz	24	±1	11 Typ	+29.5	2:1	±24	600	H/BNC
	P500A	2MHz—500MHz	37	±.5	4.5 Typ	+30	2:1	+24	500	H/SMA
	P500L	5MHz—500MHz	17	±.7	10 Typ	+30	2:1	+24	420	H/BNC
	P500ML	2MHz—500MHz	16	±1	11 Typ	+28	2:1	+24	600	H/BNC
	P1GB	50MHz—1000MHz	30	±1	5.5 Typ	+30	2:1	+20	800	A-S/SMA
	P1000M	5MHz—1000MHz	20	±.5	6 Typ	+21	2:1	+20	200	H/SMA
	P2GF-2	10MHz—2000MHz	32	±1	7.5 Typ	+30	2:1	+15	1000	FW1/SMA
	P42GA-29	.5GHz—4.2GHz	30	±1.5	6.5 Typ	+29	2:1	+20	1200	FW75/SMA

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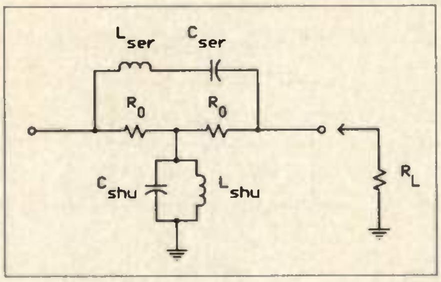

Figure 2. Resonant arm equalizer — an alternate form of a bridged T equalizer.

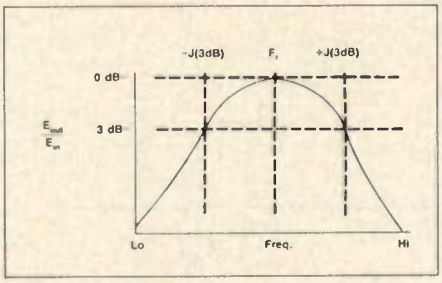

Figure 3. Response curve of a resonant armequalizer.

Then a convenient point for the design procedure would be where:

$$X_c = R_L,\ X_L = R_L \text{ or} \tag{10}$$

$$K = 1.414 \text{ or the 3dB point}$$

Then,

$$C = \frac{1}{2\pi F_x X_c} \tag{11}$$

and

$$L = \frac{X_L}{2\pi F_x} \tag{12}$$

All that remains is to solve for the frequency point, F_x, which is the point where the cable loss is 3 dB less than F_{Hi}.

From equations 1, 2, 3, and 4, after some manipulation, F_x can be solved for any desired attenuation point

$$F_x = \left[\left(\frac{\text{Atten}}{X_2 + \text{Const.}^2}\right)^{1/2} - \text{Const.}\right]^{1/2} \tag{13}$$

Where

$$\text{Const.} = \frac{(X_1/X_2)}{2} \tag{14}$$

and

$$X_1 = 4.34\left[.1\left(\frac{1}{d}\right)\left(\left(\frac{p}{p_{cu}}\right)^{1/2}\right) + .1\left(\frac{1}{D}\right)\left(\left(\frac{p}{p_{cu}}\right)^{1/2}\right)\right] \Big/ Z_0 \tag{15}$$

$$X_2 = 2.78\sqrt{e}\left(\frac{2\pi FC.01}{G_0}\right) \tag{16}$$

Where

$$C = 1.016\,\frac{\sqrt{e}}{Z_0}\left(10^{-9}\right) \text{ in } \frac{\text{pF}}{\text{ft.}} \tag{17}$$

Another form the equalizer can take is shown in Figure 2.

This is known as a resonant arm equalizer. The design procedure is similar to the previous discussion with one exception. Because it is a resonant device, it exhibits the response curve shown in Figure 3.

To operate as a cable equalizer, the characteristic of the −J slope would be used (capacitive). The X_c 3 dB point has the same relevance it had in the previous discussion. The resonant frequency point (F_r) should be located approximately 10 percent above the desired F_{Hi} frequency to avoid having the equalizer response roll off at the high frequency end. This can be seen in Figure 3 if F_{Hi} and F_r are the same frequency.

The transfer function for the resonant arm equalizer is:

$$\frac{E_{in}}{E_{out}} = \frac{R_L + J(X_L - X_c)}{R_L} \tag{18}$$

The equations for the element values are:

$$L_{ser} = \left(\frac{R_L}{2\pi F_x}\right)\left(\frac{1}{m^2 - 1}\right) \tag{19}$$

$$C_{ser} = \left(\frac{1}{2\pi F_x R_L}\right)\left(\frac{m^2 - 1}{m^2}\right) \tag{20}$$

Where:

$$m^2 = \frac{F_r^2}{F_x^2} \tag{21}$$

F_x is from equation 13

The shunt elements can be obtained from duality:

$$L_{shu} = (R_L^2)(C_{ser}) \tag{22}$$

$$C_{shu} = \frac{L_{ser}}{R_L^2} \tag{23}$$

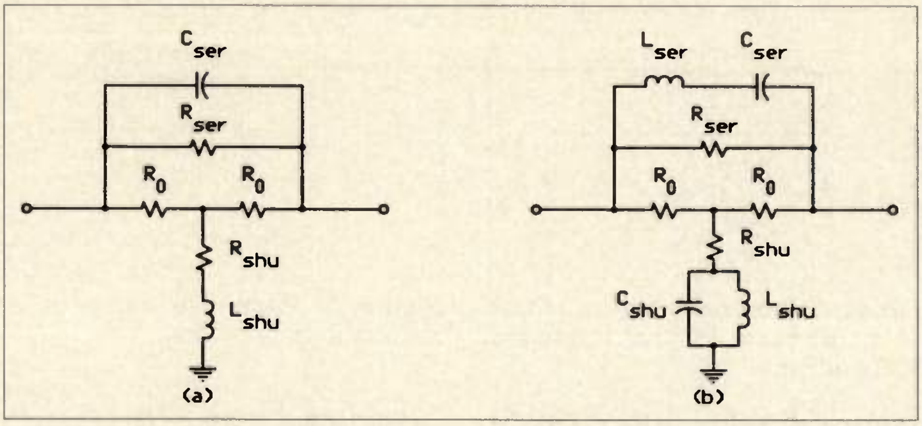

Figure 4. (a) An equalizer with an attenuator. (b) A resonant arm equalizer with an attenuator.

Circuit Application

Figure 4 shows the two circuits under discussion which are:

a) Equalizer + attenuator
b) Resonant arm equalizer + attenuator

As can be seen from the respective transfer functions ignoring R_{ser} (series) and R_{shu} (shunt):

$$\text{a)}\quad \frac{E_{in}}{E_{out}} = \frac{R_{ser} + R_L}{R_L} \tag{24}$$

$$\text{b)}\quad \frac{E_{in}}{E_{out}} = \frac{R_L + J(X_L - X_c)}{R_L} \tag{25}$$

The slope rate versus the frequency scale of the resonant arm equalizer is twice that of the single reactive element equalizer, as would be expected. In practice this establishes the following practical limits on the application of the two circuits relative to achieving the optimum compensation of the cable attenuation slope.

a) Equalizer + attenuator Maximum slope, 10 dB/decade

b) Resonant arm equalizer + attenuator Maximum slope, 20 dB/decade

Although the two circuits will perform several dB beyond these limits, an increase in the compensation ripple amplitude will result. This may be acceptable depending on the particular system specifications.

Variable Slope Circuits

Figure 4 shows the two equalizer versions that have been discussed with an attenuator inserted as shown. The insertion of the resistive elements R_{ser} and R_{shu} form the basis for a variable Bridge T equalizer. When the attenuator has a high value (20 dB or more), the circuit provides the maximum available slope from F_{Lo} to F_{Hi} as if the R_{ser} and R_{shu} elements are not present. As the value of attenuation is decreased, the slope compensation also decreases as more energy passes through R_{ser}, thus decreasing the effect of reactive elements in the series and shunt arms.

If R_{ser} and R_{shu} are replaced with suitable pin diodes which are operated from a bias network that controlled the R_s of the diodes so that they traced required values of a Bridged T attenuator, the circuits will perform as variable slope compensation networks.

The Program

The program supplied uses the above principals in designing five circuit configurations.

In addition to the design section, the program will also analyze your design (1). An editor is supplied to optimize the desired response or replace calculated values with available standard values and analyze their effect on the circuit's performance. The analyzed data can be presented as a graph within the program. It will present the circuit amplitude response, the return loss, and the calculated cable amplitude response. Also available graphically is the net difference in amplitude of the cable and the equalizer circuit. This is called "ripple" in the program.

The program was compiled to use a math coprocessor if it is available in the user system. If the user system does not have a coprocessor, the program will emulate it. If screen copies of the graphs are desired, the DOS routine GRAPHICS must be resident in memory.

A "README.EQU" file is also supplied in ASCII. It provides detailed instructions for using the program.

This program is available on disk from the RF Design Software Service. See page 8 for ordering information. ***RF***

References

1. Ken Wyatt, "Ladder Program Analysis," *RF Design,* November 1986.
2. Reference Data for Radio Engineers, ITT Forth *Edition,* 1956.
3. Frank Eganstafer, *Bridge T Equalizer Analysis,* 1970.
4. Brainard, Koehler, Reich and Woodruff, *Ultra High Frequency Techniques,* D. Van Norstand, July 1943.
5. Thomas Cuthbert, Jr., *Circuit Design Using Personal Computers,* J. Wiley & Sons, 1983.

About the Author

Frederick Radler is a staff engineer with Trilithic Inc. He can be reached at 9202 East 33rd Street, Indianapolis, IN 46236. Tel: (317) 895-3600.

RF software

Schematic Editor

CapFast™ CF3000M is a schematic design tool used to do a majority of circuit design on personal computers and then to port the circuit to a Mentor Graphics workstation for final integration. It features open databases and ASCII file formats, EDIF 200 translators, general purpose symbol editor, and a programmable netlist library. CapFast requires an IBM PC or compatible, with 2 Mbytes of extended memory, a two- or three-button mouse, a hard disk drive, and EGA/VGA card with 256 Kbytes of on-board memory.
Phase Three Logic, Inc.
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Filter Design Software

A free software program called FILTER PERFECT™, from Burr-Brown, helps users design low-pass filters. It includes an application note which describes the operation of the program and measured results of actual filters designed with the program. FILTER PERFECT supports Butterworth, Chebyshev, and Bessel filters, computing capacitor and resistor values.
Burr-Brown Corporation
INFO/CARD #142

Power Amplifier Design Tool

Design Automation has released a demonstration disk of the HEPA-PLUS™ RF power amplifier design package. The program designs, simulates, and optimizes high-efficiency switching-mode (Class E) RF power amplifiers. The demonstration disk is available for $30, and runs on IBM PC compatible computers.
Design Automation, Inc.
INFO/CARD #217

DSP Code Generation Software

The Code Generation System™ (CGS) for Comdisco Systems' Signal Processing WorkSystem™ (SPW) generates C code to implement DSP algorithms on Motorola, AT&T, Texas Instruments, and Ariel processing chips. After the CGS produces the C code, it is downloaded from SPW, compiled, and executed in the PC system containing the board that is being designed for.
Comdisco Systems, Inc.
INFO/CARD #216

Verilog Translator

A new translator converts IC designs simulated in Verilog™ hardware description language for high-speed logic simulation and verification on hardware-assisted simulators from Ikos Systems. VLIT™ reads Verilog netlist files and translates them into Ikos-compatible files. It works on Sun Microsystems and HP/Apollo workstations and is available for a license fee of $5,000 per node.
Ikos Systems, Inc.
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Scientific Plotting Software

MicroMath announces the release of GRAPH version 3, for plotting of scientific and engineering data. New features of this version include extended memory support, multiple y-axis plotting, import/export of Lotus, DIF, dBase and ASCII files, enhanced statistics, digital data smoothing, and PostScript output, GRAPH is priced at $149 for a single-user copy. The update alone is specially priced at $49 until February 10, after which it will be $69.
MicroMath Scientific Software
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Shared-Resource Manager

HP E2085A is a new shared-resource manager (SRM) on the HP-UX server software (SRM/UX). The SRM/UX provides users of HP BASIC with access to a standard local area network to share files and programs among network users. Peripheral sharing and remote files are supported in the HP-UX system, which is compatible with USL's UNIX operating system. Up to 63 clients are allowed on SMR-style interfaces, or 265 client on LAN interfaces. The server software is priced at $4,495, including Series 200/300 HP BASIC/WS client software.
Hewlett-Packard Company
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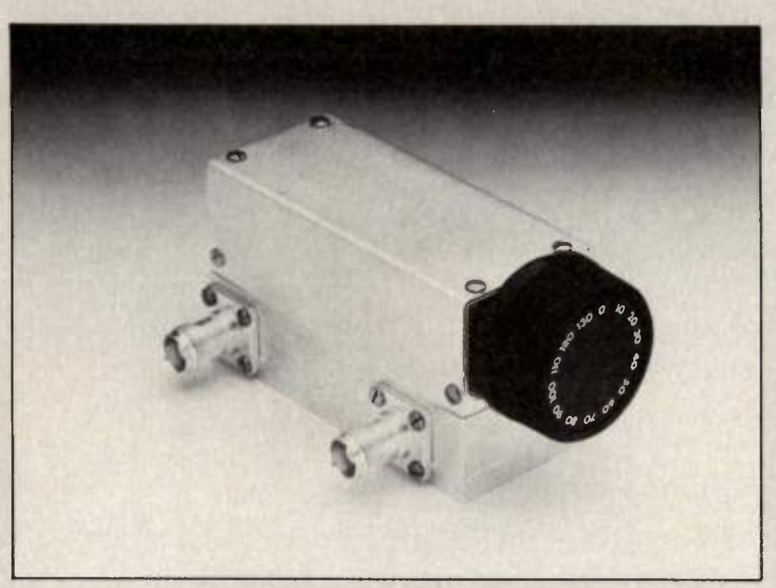
JFW Rotary Attenuators



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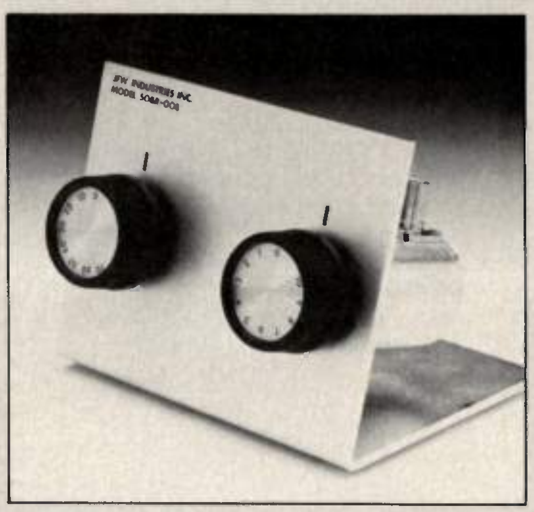

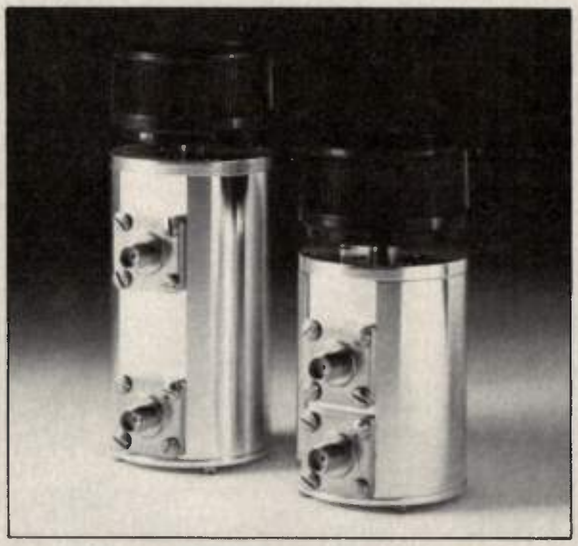

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50R-019 (right)
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Attenuation Range: 0-10 dB in 1 dB steps
Connectors: SMA female (standard) BNC available upon request

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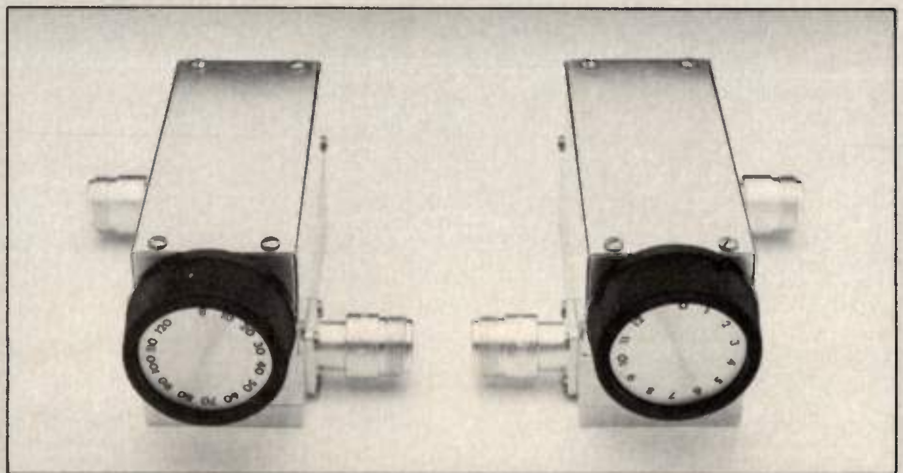

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RF literature

Analog Dialogue

Analog Devices has produced their latest issue of *Analog Dialogue*, a quarterly journal on circuits, systems, and software for real-world signal processing. Discussed in this edition are specifications and applications of a mixed-signal processor which combines the ADSP-2101 DSP and both A-D and D-A converters on a single chip.
Analog Devices
INFO/CARD #229

Low Noise Amplifiers

Miteq has released a catalog detailing data on their low-noise, wideband, bipolar amplifier product line, covering the frequency range from a few kHz to 2 GHz. Special purpose and custom designed amplifiers are also available through the catalog.
Miteq
INFO/CARD #228

Solid State Switches

A catalog of solid state switches covering the 10 MHz to 18.5 GHz frequency range has been provided by New England Micronetics. Single pole single throw switches through single pole five throw switches are detailed and performance characteristics and custom options are included.
New England Micronetics
INFO/CARD #227

Monolithic Capacitors

A new catalog describing Murata Erie's line of axial leaded monolithic conformal coated ceramic capacitors is now available. The catalog contains specifications on temperature, capacitance tolerance, and mechanical and environmental details. Capacitor values range from 10 picofarads to 1 microfarad.
Murata Erie North America
INFO/CARD #226

Solid State Amplifiers

ENI has prepared a pamphlet describing their line of solid state power systems covering the 10 kHz to 1000 MHz frequency range with power outputs ranging from milliwatts to kilowatts. Included are models designed for magnetic resonance imaging, nuclear magnetic resonance spectroscopy, and broadband applications.
ENI
INFO/CARD #225

Measurement Antennas

Anritsu has released a brochure detailing its full line of measurement antennas covering the 9 kHz to 2000 MHz frequency range. Comprehensive descriptions of dipole, log-periodic, biconical, loop and rod antennas are included. A chart detailing the specific components of each antenna, and diagrams providing measurements of each model are also contained int he brochure.
Anritsu America, Inc.
INFO/CARD #224

Crystals and Oscillators

CTS has released a catalog describing their lines of crystals, hybrid oscillators, and precision crystal oscillators. Diagrams, specifications, and performance features are included for each product, and ordering forms are attached.
CTS Corporation, Knights Division
INFO/CARD #223

Miniature Coax Connectors

The 50 ohm MMCX-Connector Series from are described in this catalog released by Huber + Suhner. The connectors cover the DC to 6 GHz frequency range and come in 25 different types, which include connectors for flexible and semi-rigid cables in straight, right-angle, and bulkhead versions.
Huber + Suhner AG
INFO/CARD #222

Inventory Update

Tucker Electronics has released an update on their lines of used instruments. Listed are universal counters, network analyzers, step attenuators and more from companies such as Hewlett-Packard, Fluke, Tektronix, and others. Also included are various power supplies and cable assemblies for 75 ohm terminations.
Tucker Electronics Company
INFO/CARD #221

Electronic Materials Catalog

Electrical Insulation Suppliers has released their Electronic Materials Catalog which describes materials for electronic manufacturing and assembly. It features EMI/RFI shielding, adhesives, coatings and other products.
Electrical Insulation Suppliers, Inc.
INFO/CARD #220

Electronic Components Catalog

Digi-Key has published Catalog No. 911, featuring new additions to their product offerings. A wide range of components is listed, including inductors, capacitors, resistors, crystals, filters, connectors, digital and linear ICs, and many other products. Also described is their rapid shipment policy, pricing, and discount information.
Digi-Key Corp.
INFO/CARD #219

Test System Products

The 1991 IEEE-488 and VXIBus Control, Data Acquistion and Analysis catalog is now available from National Instruments. The catalog features software products for thw control of programmable instruments, plug-in data acquisition boards with sampling rates up to 1 MHz, analog signal conditioning modules, VXIBus data acquisition and control systems, plus software tools for control of RS-232 based instruments.
National Instruments
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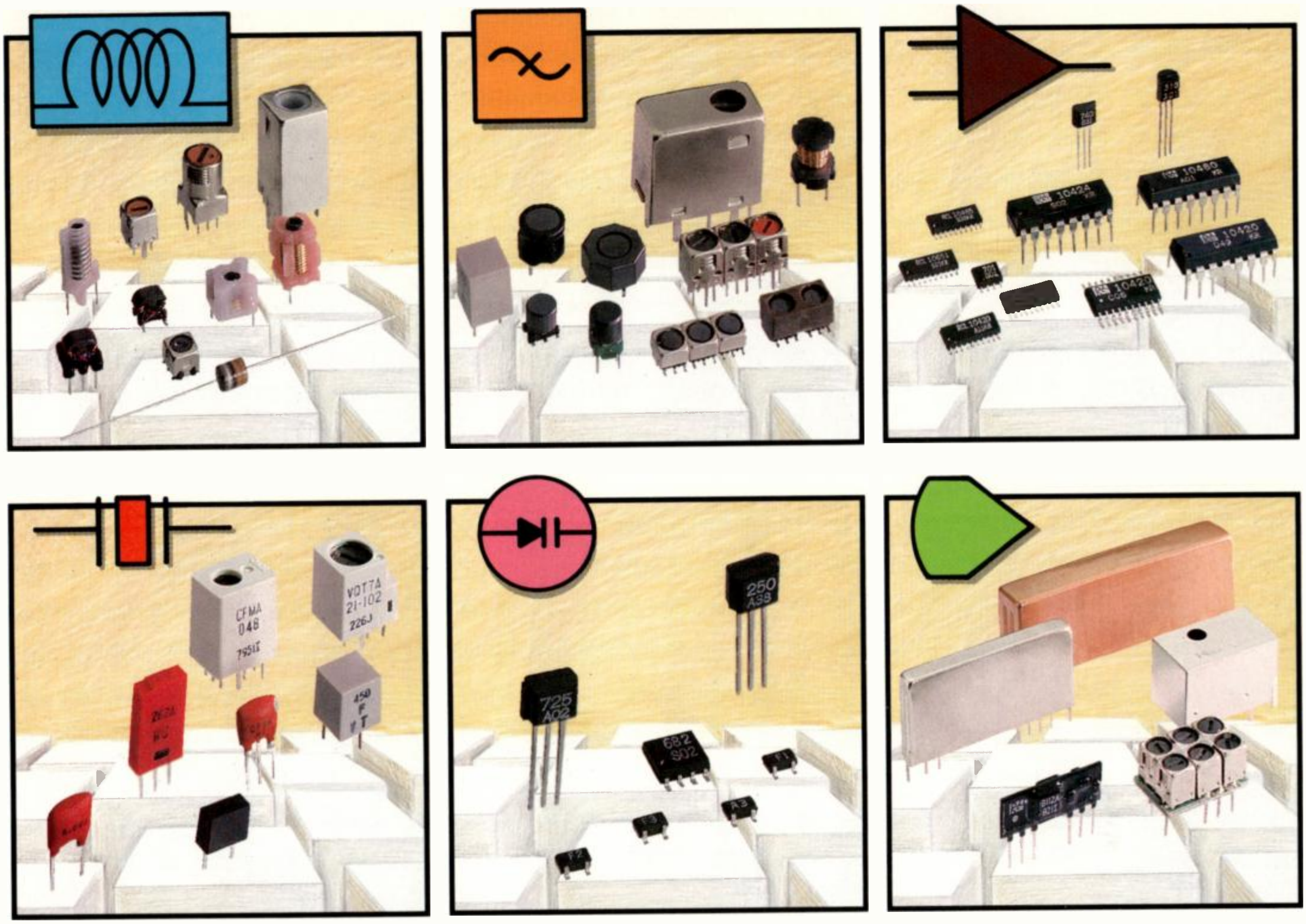

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