RF design

engineering principles and practices

May 1991

Featured Technology

Test and Measurement

Crystal Oscillators

Quadrature Detectors

Wavecom SP2T coaxial switches

The most choices. The largest line.

Wavecom offers the largest line of SP2T switches in the business. What's more you can get exactly the specs you want – at commercial prices and with products that are manufactured to Wavecom standards of space age reliability.

That's because we designed our basic switches to accommodate an almost infinite variety of spec combinations.

Frequency range: DC through 26.5 GHz.

RF connectors can be SMA, TNC, N, BNC, SC or PC board mount.

Actuation can be failsafe, latching, or normally open. Make-before-break available for hot switching applications.

Options can include 50 ohm terminations, indicator circuitry, TTL logic, BCD TTL decoder, self de-energizing circuitry, and manual operation.

We also offer a complete line of multi-position (SP3T to SP12T) and transfer switches – all available with the same variety of spec combinations as SP2T switches.

All switches can be manufactured to meet any spec requirements, whether commercial, mil-spec or deep space probes.

For more information or a free catalog, call or write Loral Microwave-Wavecom, 9036 Winnetka Avenue, Northridge, CA 91324. Tel: (818) 882-3010. Fax: (818) 709-8204.

Specifications – Wavecom SP2T coaxial switches

Frequency Range (GHz)	DC - 3	3 - 8	8 - 12.4	12.4 - 18	18 - 26.5
VSWR (Max)	1.2:1	1.3:1	1.4:1	1.5:1	1.6:1
Insertion Loss (Max dB)	0.2	0.3	0.4	0.5	0.6
Isolation (Min dB)	80	70	60	60	50

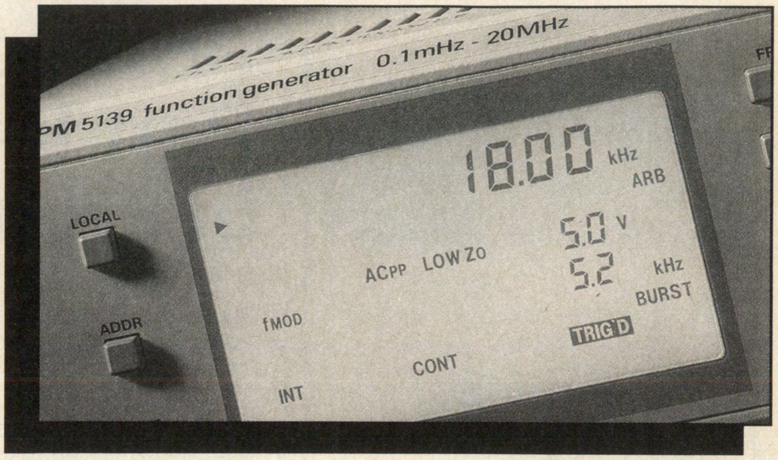
LORAL
Microwave-Wavecom

INFO/CARD 1 Please see us at MTT-S Booth #520

RF design

contents

May 1991



featured technology

emc corner

design awards

departments

Our cover, provided by John Fluke Mfg. Co., represents our emphasis on engineering test and measurement. More information on this product can be found on page 48.

R.F. DESIGN (ISSN: 0163-321X USPS: 453-490) is published monthly plus one extra issue in September. May 1991. Vol. 14, No. 5. 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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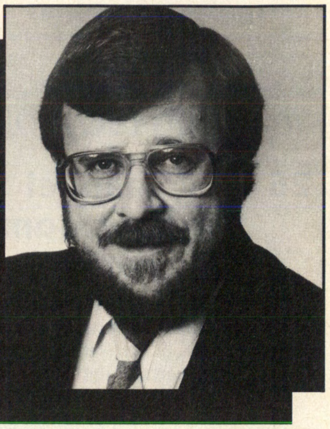
Engineering Productivity Part II — Test Instruments



By Gary A. Breed
Editor

Last month, I noted some of the organizational changes that are occurring in the U.S. electronics industry. To effectively improve efficiency, these changes require that companies get the maximum results from their employees' efforts. For RF companies, this is an especially difficult assignment, since experienced RF designers are in relatively short supply and many are already extremely busy.

One method of improving an RF engineer's productivity is to provide support that allows him more time for creative engineering. Recent developments in the world of test and measurement certainly fit this description, with instruments that not only offer improved performance, but have operating features that minimize setup time.

The principal enhancement in newer test instruments is the incorporation of computing power. This allows the user to store standard setups, eliminating all the knob-twisting that used to be needed. Errors in those dial settings are eliminated, too. Computations based on the measured data can be performed as part of the pre-programmed test procedure. For example, a spectrum analyzer can list the harmonic levels relative to the fundamental, or identify spurious signal levels and frequencies.

All this power can be utilized in a multi-unit test setup. A modulated signal generator, spectrum analyzer, power supply and DC multimeter could be connected into a power amplifier test system, controlled by a host computer. A little extra time to program the system eliminates lengthy and tedious manually-operated measurements. You might not need a computer, either. The controller might be inside one of these instruments, since a few already have a keyboard option that is all they need to be stand-alone PCs!

Performance features improve the accuracy and reliability of measurements, especially in high performance testing. Improved measurement confidence reduces the need to repeat a test "just to make sure." Lower noise synthesizers in signal generators, spectrum analyzers and network analyzers have increased the dynamic range of these instruments' measurements. Direct digital synthesis and digital modulation allows generation of complex waveforms that previously required combinations of several sources. Baseband FFT analysis in spectrum analyzers dramatically increases the precision of their measurements. Internal calibration routines make sure that these powerful functions are operating as intended when you need them.

These advantages are not limited to the designer's bench — production line testing may receive an even bigger benefit. Faster, more accurate measurements are exactly what the production test manager wants! Pre-programmed setup takes most of the human error out of the instrument operation and data collection process. To meet the needs of manufacturing, a new generation of instruments with "practical" rather than "ultimate" performance specifications is beginning to show up.

So, the next time you find yourself wishing for some new test equipment, take the time to evaluate how much time you spend preparing for tests, running them by hand, and manually tabulating the results. If you are under pressure to deliver your designs on increasingly shorter timetables, this information might be all that is needed to justify the investment. Your company will get more out of your talents, and you'll get greater satisfaction doing it.

RF design

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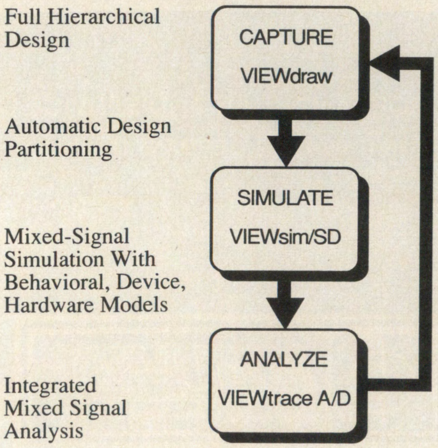
For Most People, It Was Just Another Warm September Day.

For design engineers, it was the day mixed analog/digital design came of age.
The event was the mixed-signal design demonstration at the IEEE Bipolar Circuits and Technology Meeting (BCTM). The goal was to give credence to mixed-signal simulation and to benchmark companies in the marketplace. The results were conclusive.

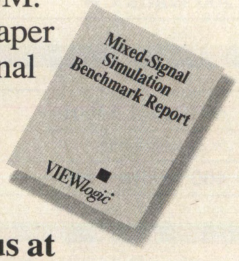
Viewlogic came up with the right answer first.
But more to the point, what we did at BCTM in September, we can do for you now. We're the only company with a proven technology and a three year track record of success. The only one that integrates design capture, simulation and analysis.

But that's just the beginning.

Performance and Flexibility available nowhere else.
With VIEWsim/SD, you'll get the choices you need. You'll be able to mix behavioral models with gates and SPICE primitives. Choose from leading analog simulators like PSPICE and HSPICE. Include physical hardware models for devices when software models are not available. Use the most popular platforms from SUN, DEC and IBM.



Our white paper "Mixed-Signal Simulation Benchmark Report" proves the point. **Call us at 1-800-422-4660, Ext. 103.** You'll like the climate we're creating for mixed-signal design.



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RF letters

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

Obsolete or Absolute?

I read your editorial on obsolete technologies with much interest and a bit of puzzlement.

I think you are way off base, 180 degrees out of phase with reality, about personal communications making "inconvenience obsolete." The plain truth is that such a system would make inconvenience *complete*. If you doubt this point of view, I challenge you to make a simple test. Just give all the secretaries around you a week off and for one week, answer your phone yourself every time it rings. And, of course, you will do this with no lessening of your regular work output.

Hanging one of those infernal machines around your neck so that people could call you up and talk about a subject of their choosing at a time of their choosing, with absolutely no thought for your convenience would be — well, the inner circle of hell! The phone is getting to be an invasion of privacy already, but what would it be like if there were nowhere you could go to get away from it?

Add this to your list of pet peeves, and contemplate that we now have the technology to bring all of them to pass.

C.A. Chrestien
Sunnyvale, CA

QUAD COMPARATORS 8ns, 18mW–ONLY $1.50*

- **+5V Operation Cuts Power Drain and Heat Dissipation**
- **Input Range Includes Ground and Eliminates Need For Negative Supply**
- **Four Comparators in One Package Saves Board Space and Reduces Cost**
- **Separate AGND and DGND Minimizes Noise, Improves System Performance**
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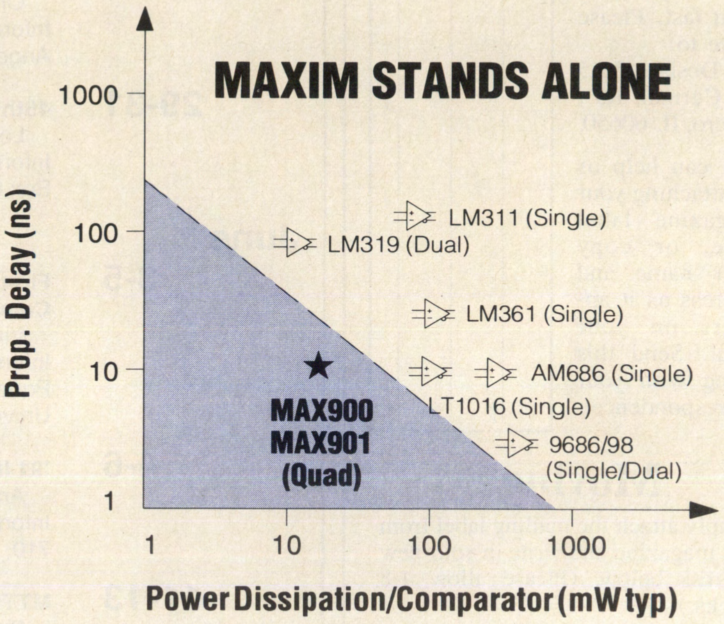

One Stop Shopping For High Speed/Low Power

Maxim's new MAX900 and MAX901 quad comparators can deliver both high speed *and* low power at a price that compares favorably to high speed single or dual equivalents. MAX900/MAX901 offer you propagation delays of only 8ns with a 5mV overdrive, power dissipation of 18mW per comparator (when powered from a +5V supply) and space saving 0.3" DIP or small outline (SO) packages - for only $1.50 per comparator.

Maxim's High Speed Comparator Family

Part Number	# Comps	Logic	Delay (typ)	Latch	Package	Price†
MAX900	**4**	**TTL**	**8.0ns**	**Yes**	**DIP, SO**	**$7.01**
MAX901	**4**	**TTL**	**8.0ns**	**No**	**DIP, SO**	**$5.98**
MAX9685	1	ECL	1.3ns	Yes	DIP, SO, Can	$3.38
MAX9686	1	TTL	6.0ns	Yes	DIP, SO, Can	$2.31
MAX9687	2	ECL	1.4ns	Yes	DIP, SO	$5.12
MAX9690	1	ECL	1.3ns	No	DIP, SO	$3.29
MAX9698	2	TTL	6.0ns	Yes	DIP, SO	$3.92

* MAX901, 1000-up F.O.B. USA price per comparator † 1000-up F.O.B. USA

Call your Maxim representative or distributor today for applications information, datasheets and samples. Or, write Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, (408) 737-7600, FAX (408) 737-7194.

MAXIM

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RF calendar

May

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

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

June

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

4-6 **'91 International Communications Association Expo**
Anaheim Convention Center, Anaheim, CA
Information: Angela Gallagher, ICA, 12750 Merit Drive, Suite 710, LB-89, Dallas, Texas, 75251. Tel: (214) 716-4124.

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

13-14 **ARFTG**
Boston, MA
Information: ARFTG, c/o Henry Burger, 1061 E. Frost Drive, Tempe, AZ 85282. Tel: (602) 839-6933.

24-27 **Test Engineering Conference**
Georgia World Congress Center, Atlanta, GA
Information: Miller Freeman Expositions. Tel: (800) 223-7126 or (617) 232-3976.

25-27 **EMC Expo '91**
Walt Disney World Resort, Orlando, FL
Information: Ellen Lunsford, Registration. Tel: (703) 347-0030.

July

15-18 **Communication Networks West '91**
Moscone Center, San Francisco, CA
Information: Conference Sales, Tel: (800) 225-4698.

22-24 **Fifth International Conference on HF Radio: Systems and Techniques**
Edinburgh Conference Centre, Edinburgh UK
Information: Secretariat, Conference Services, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 240 1871 ext. 222.

August

13-15 **IEEE 1991 International Symposium on Electromagnetic Compatibility**
Cherry Hill, NJ
Information: IEEE International Symposium on EMC, PO Box 609, Lincroft, NJ 07738. Tel: (201) 386-2378.

Fundamentals of Radar Cross Section
May 20-24, 1991, San Diego, CA
Information: Kelly Brown, SCEEE, Tel: (407) 892-6146.

Microwave/Millimeter Wave Monolithic Integrated Circuits
June 4-7, 1991, Los Angeles, CA
Radiation Hardening of Electronic Systems
June 10-14, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Frequency Hopping Signals and Systems
May 20-22, 1991, Washington, DC
Principles of High Frequency Radio Communications: Applications for Operators and Managers
May 20-23, 1991, Washington, DC
New HF Communications Technology: Advanced Techniques
June 3-7, 1991, Washington, DC
Spread Spectrum Communications Systems
June 10-14, 1991, Washington, DC
Introduction to Modern Radar Technology
June 26-28, 1991, Washington, DC
Ionospheric Radio Propagation for System Planners
July 8-11, 1991, Washington, DC
Radio Frequency Spectrum Management
July 8-12, 1991, Washington, DC
Synchronization in Spread-Spectrum Systems
July 15-19, 1991, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

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

Digital Signal Processing Workshop
June 12-14, 1991, Norwood, MA
July 16-18, 1991, Campbell, CA
Mixed Signal Design Seminar
May 15, 1991, Pleasanton, CA
May 16, 1991, Bellevue, WA
May 17, 1991, Beaverton, OR
May 20, 1991, Woburn, MA
May 21, 1991, Montreal, Canada
May 22, 1991, Bloomington, MN
May 23, 1991, Houston, TX
May 24, 1991, Dallas, TX
May 28, 1991, Phoenix, AZ
May 29, 1991, Denver, CO
May 30, 1991, Arlington, Heights, IL
May 31, 1991, Rochester, NY
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Touchstone/Academy
June 3-7, 1991, Gilching, Germany
Libra/Academy
June 17-21, 1991, Baltimore, MD
June 17-21, 1991, Gilching, Germany
Information: EEsof. Tel: (818) 991-7530. Fax: (818) 991-7109.

Frequency-Time and Spatial-Time Signal Processing
June 10-14, 1991, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

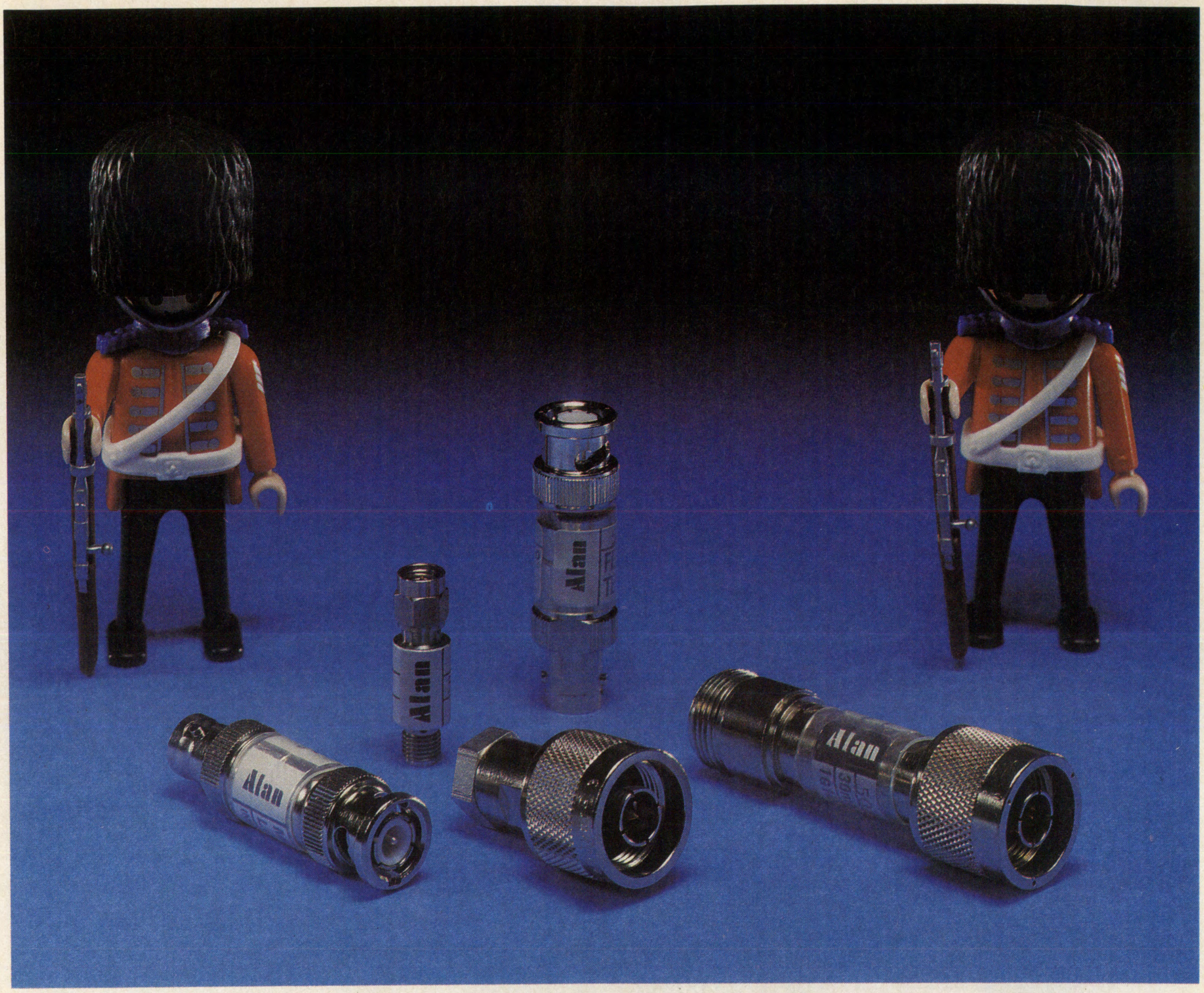
Modern Power Conversion Design Techniques
May 20-21, 1991, San Rafael, CA
July 15-19, 1991, Chicago, IL
Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

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

Basic Network Measurements Using the 8510B Network Analyzer
May 29-31, 1991, Los Angeles, CA
June 11-13, 1991, San Francisco, CA
June 18-20, 1991, Atlanta, GA
June 25-27, 1991, Los Angeles, CA
Designing for EMC
July 11-12, 1991, Atlanta, GA
Information: Hewlett-Packard Company. Tel: (714) 999-6700.

Introduction to Fiber Optic Communications
June 4-7, 1991, San Francisco, CA
Digital Signal Processing: Techniques & Applications
June 4-7, 1991, San Francisco, CA
June 18-21, 1991, Washington, DC
Hands-On Datacomm Troubleshooting
June 4-7, 1991, Washington, DC
June 25-28, 1991, San Francisco, CA
Introduction to Telecommunications
June 11-14, 1991, Los Angeles, CA
June 25-28, 1991, Washington, DC
Information: Learning Tree International. Tel: (800) 421-8166, (703) 893-3555, (203) 417-8888.

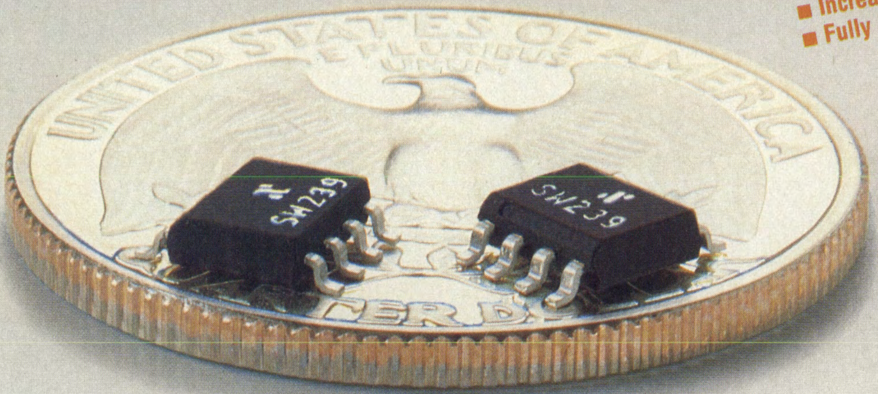
Introduction to EMI/RFI/EMC
May 21-23, 1991, Washington, DC
June 24-25, 28, 1991, Orlando, FL
EMI Control Methodology and Procedures
May 21-24, 1991, Philadelphia, PA
Practical EMI Fixes
June 3-7, 1991, Las Vegas, NV
Grounding and Shielding
June 24, 27, 28, 1991, Orlando, FL
System Integration and Design for EMC
June 24, 25, 28, 1991, Orlando, FL
Information: Interference Control Technologies, Registrar. Tel: (703) 347-0030.



Two SPDT Bits.

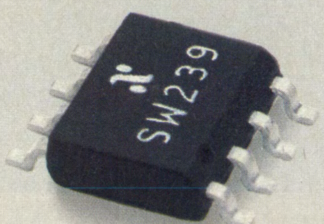
FIELD TESTED AND IMPROVED!

- Improved Frequency Performance
- Increased Operating Temperature Ran
- Fully Passivated Die



ANZAC Does It Again!

ANZAC offers yet another GaAs Monolithic alternative to pin diode switch technology. The SW-239 SPDT RF switch features nanosecond speed with only microwatts of DC power. Utilizing an 8-Lead SOIC package these units come ready for full automatic insertion.



MODEL SW-239 Performance Characteristics	
Frequency Range	DC-2000 MHz
Operating Temperature	-40°C to +85°C
Insertion Loss	0.5 dB
VSWR	1.1 : 1
Isolation	50 dB
1 dB Compression	+27 dB
Switching Speed	4 nsec TYP
Package	8 Lead SOIC (3.8mm x 4.8mm)

Tomorrow's GaAs MMIC technology can be yours today for just a handful of quarters. (40 to be exact.)

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INFO/CARD 16

Bell Atlantic to Expand Microcell Technology — Bell Atlantic Mobile has announced that it will expand its use of microcellular technology to enhance cellular reception indoors in public locations. The first location will be Amtrak's Union Station in Washington, DC. Microcell technology utilizes the cellular concept on a smaller scale than the traditional car usage. It will use extremely low-power cell sites in heavily trafficked areas. A cell site, housing computer-controlled transmitters and receivers, is a key piece of the cellular system, serving as a gateway to the traditional telephone network. In Union Station, a suitcase-sized transmitter and antenna will be placed in the Amtrak passenger waiting area. The initial equipment was developed by NovAtel Communications, Inc. Regardless of what network equipment is in use at a given time, customers will be able to make calls on their cellular phones while within the microcell's coverage area. Initially, the microcell will not be connected to the outside "macro-cellular" system. Later this year, the two will be interconnected and customers will be able to make and receive calls, as well as hand-off calls as they leave the station.

Tektronix Announces Equipment Grant — Tektronix Inc. recently presented a grant to the University of Minnesota's Institute of Technology School totaling approximately $130,000 in the form of state-of-the-art engineering equipment. The contribution is comprised of test and measurement equipment manufactured by Tektronix and includes such instruments as digital oscilloscopes, spectrum analyzers, digital voltmeters, function and pulse generators, and color plotters. The equipment is dedicated for use at the newly-constructed senior design lab at the school.

USSR Air Traffic Control Modernization — AT&T, Westinghouse, IBM, C. Itoh and Daimler-Benz have committed themselves to an approximately $10 billion modernization of the Soviet air traffic control system. The plan is slated for completion by the year 2005. The plan calls for merging of Soviet national airspace plans with air traffic control plans of the United States, Canada, Europe and Asia.

Researchers Develop New Class of Antenna — Engineers at the Georgia Tech Research Institute have developed a new class of microstrip antenna that combines the broad-band performance typical of spiral and sinuous antennas with the surface mount capabilities, efficiency and low cost of microstrip antennas. The new antenna design was developed through a research program supported by the U.S. Air Force's Wright Research & Development Center. Though designed for military use, the new antenna could have widespread commercial applications such as cellular telephones, identification, satellite communications, inter and intra-building communications and the tracking of ground shipments. The antenna uses existing technology and therefore could be mass-produced at a low-cost.

The antenna was developed using

two different antenna technologies: the broad-band spiral technology and the microstrip technology. While most microstrip antennas offer a bandwidth of just 10 percent, the new design offers bandwidths as high as 500 to 600 percent.

Direct Conversion Technique Relocates — Direct Conversion Technique (DCT) recently announced the opening of its new headquarters facility in Des Plaines, Illinois. Their new address is 1245 Forest Avenue, #12, Des Plaines, IL 60016. Tel: (708) 827-2282. Fax: (708) 827-2280.

MIC Technology Acquires Thin Film Coating and Patterning Business — MIC Technology has acquired the U.S. portion of the thin film coating and patterning business of Materials Research Corporation's Hybrid Products Division. MIC will operate this business as the Thin Film Division of MIC Technology. Materials Research Corporation will retain its European based Hybrid Products Division and continue to service the European market.

EEsof Users' Group to Meet at 1991 IEEE MTT Symposium — EEsof Inc. will sponsor the MTT-S meeting of its Users' Group, June 11 from 5:00 p.m. to 7:30 p.m. in room 311 at the Hynes Convention Center, Boston, MA. The EEsof Users' Group invites both members and non-members to attend the meeting. Technical papers will be presented investigating the use of CAE/CAD software applications in microwave and RF circuitry design. Proceedings will be available at the meeting.

Magnavox to Provide Pilot Automatic Vehicle Location System —Magnavox/Nav-Com has won a contract to provide the Swedish Road Administration with a pilot Automatic Vehicle Location System (AVLS), which will be used to track the location and movements of vehicles at a central facility. The vehicle navigation system uses a small multi-channel GPS receiver, which determines the vehicle's location from signals broadcast from the satellite based Global Positioning System. The Swedish Road Administration will provide the data link for the vehicles, based on the established Mobitex radio data communications network. The vehicles will be equipped with Volvo VTS MC100 data terminals, which will be interfaced with the Magnavox navigation system, and will transmit AVLS messages to the central facility. There, using Magnavox-developed software, the vehicle positions are displayed on a 19 inch color graphics monitor.

IITS Purchases Dash, Straus & Goodhue — Inchcape Inspection and Testing Services recently announced the purchase of Dash, Straus & Goodhue (DS&G). DS&G is involved with compliance testing for electrical and electronic products, particularly in the area of electromagnetic compatibility, including testing for electromagnetic interference and radio frequency interference. DS&G will become a member of IITS's Manufactured Goods Testing Group.

Microwave Hybrid Circuits Call For Papers — A call for papers has been issued for the 1991 Microwave Hybrid Circuits Conference. The conference will be held October 13-16 in Sedona, Arizona. The conference theme is "Designing for Production — Issues in the Transition from the Drawing Board to the Factory Floor." Papers describing new work will be considered on the following subjects: innovative MIC and MMIC design; MIC processing and quality issues; advancement in MIC materials; MIC materials testing; MMIC packaging techniques; unsolved problems in MIC design or implementation; and Taguchi methods and design centering techniques. Abstracts are due June 1, 1991. Abstracts should be sent to: MHCC, c/o Toshikazu Tsukii, Program Chairman Raytheon ESD, 6380 Hollister Avenue, Goleta, CA 93117. Tel: (805) 967-5511. Fax: (805) 964-0470.

New Report on Future of Broadcast Transmission Systems — A new report issued by Elsevier Advanced Technology discusses the future of broadcast transmission systems to the year 1995 and beyond. The report, "The World Market for Broadcast Transmission Systems to 1995," discusses the effects of slow economic growth on the market, the changing political climate, increased competition and long term trends. Areas of special interest include the continuing expansion of alternative broadcasting technologies such as satellite and cable used for broadcasting transmission. The report predicts a doubling of the number of households who receive satellite or cable within a few years. Direct broadcasting by satellite is also expected to increase rapidly. The falling demand for LF and HF transmissions is also discussed and the subsequent changes in companies who will serve that market. On the brighter side, the growth of FM and TV transmission systems are expected to show significant growth especially in less developed countries. Problems are expected though in areas such as southern Europe where stricter spectrum management policies will be put into place.

Federal Rules on Antennas Should Supersede Local Rules — The NAB has called for broader federal authority over local zoning restrictions on satellite dishes and other types of radio and TV antennas. The NAB made these comments to the Federal Communications Commission because many of the nation's municipal rules on radio and TV antennas are "undefined and unreasonable." These requests were made as a result of a Deerfield, New York zoning ordinance on satellite dishes.

Arlon Announces Plant Expansion — Arlon, Microwave Materials Division has announced a 75,000 square foot building addition and equipment acquisition. The new facility will contain production, quality control, and testing laboratories for the company's PTFE laminated substrate materials. The addition is expected to be completed in February 1992 and will complement the company's existing 60,000 square foot facility.

RF Technologies Announces Opening — A new company has opened who will provide high power and super high power waveguide and coax networks, as well as transmission systems and components. The company, RF Technologies Corporation, will also offer a line of RF instrumentation and control equipment. Their address is 482 Congress Street, Suite 101, Portland, Maine 04101. Tel: (207) 773-7778 or (800) 634-4075.

Motorola and Northern Telecom to Develop Open Cellular System Standards — Motorola and Northern Telecom recently announced an agreement to develop major enhancements to their existing cellular products for global markets, based on new open system standards. The two companies intend to jointly specify, develop and implement, in connection with the appropriate standards bodies, open cellular system interface standards which will permit operators to purchase cellular system components from a number of vendors. The standards will be based on existing international switching and cellular standards, such as CCS7, GSM, ISO and ISDN.

New Cellular Network Antenna — CAL Corporation and Bell Cellular recently introduced a new cellular antenna to improve network performance for cellular users. The low profile sector-beam cellular antenna was developed through a joint venture. The antenna tilt, used to adjust antenna frequency and scope, can be controlled remotely and adjustments no longer need to be done manually.

Antenna Specialists Relocates — Antenna Specialists Company have announced the moving of its Site Management Products Research and Development Group to new facilities in Reno, Nevada. The new facility, with 50 percent more space, will include additional test equipment and environmental test chambers.

Varian Completes Restructuring — Varian Associates recently completed the initial phase of its financial turnaround with the sale of its last remaining non-core business, a broadcast transmitter operation in England, to the Harris Corporation. The sale of this unit and four other non-core businesses and six product lines have generated combined proceeds in excess of $60 million. The funds will be used for various purposes, including the second phase of Varian's turnaround which involves achieving operational excellence in each of its core businesses — electron devices, analytical instruments, medical equipment and semiconductor equipment.

S.T. Research Receives HPI Receiver System Contract — S.T. research has been awarded a contract for the development and product of the High Probability of Intercept (HPI) Receiver System from the Naval Sea Systems Command. The contract, depending upon options exercised and ancillary system and support, will exceed $50 million.

INFO/CARD 21

HDTV: The World is Waiting

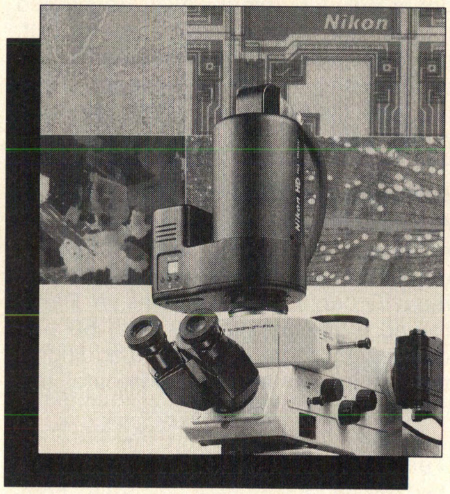
By Gary A. Breed
Editor

High definition television (HDTV) has clearly reached critical mass in its development. Regular developmental broadcasts are taking place in Europe and Japan, studio video equipment is readily available (although expensive), and component manufacturers are actively developing products for HDTV receivers and monitors.

The components industry is nearly ready for high-volume HDTV manufacturing, although production will wait until the market requires it. Thomson-ICS, Sawtek, and other SAW device houses have announced HDTV IF bandpass filters. Video amplifier and switching components from Maxim, National Semiconductor, Comlinear, Elantec, Burr-Brown and Analog Devices are capable of handling the required bandwidths. Direct satellite broadcast of HDTV is attractive to GaAs front-end suppliers such as Pacific Monolithics, Triquint and Anadigics.

Also required are high performance analog and digital signal processing components. A-to-D and D-to-A converters with at least 10-bit resolution must accurately handle greater than 25 MHz video bandwidth. Digital signal processing circuits have the same constraints. To put the required performance in perspective, current NTSC studio video requires 6 to 10 MHz bandwidths, depending on the ultimate quality level desired, while HDTV requires 25 to 50 MHz.

Studio equipment is already in place, and manufacturers are clamoring to supply networks, stations and video production houses with new systems. Test equipment, cameras, recorders, graphics and special effects, signal routing equipment, signal processing systems, and monitor makers are waiting for the doors to open on a huge market. However, the list of current suppliers is a Who's Who of Japanese electronics. NHK, the Japanese broadcasting agency, sponsored an HDTV "open house" at the recent National Association of Broadcasters convention, and the list of participating companies included 20 major firms such as Sony, Canon, NEC, Panasonic, and Toshiba.



Nikon offers HDTV imaging for microscopic observation.

Non-broadcast applications are helping keep enthusiasm high while broadcast standards evolve. Medical imaging, industrial monitoring and inspection, photo archives, and military display systems are but a few of the more promising applications. For example, Nikon is promoting high resolution imaging systems for microscopy in medical, metallurgical and microelectronic applications. Although alternative high resolution display technologies are available now, many potential applications will not become widespread until the large quantity manufacturing for consumer HDTV receivers brings prices down.

Terrestrial Broadcast

The selection of a method to bring HDTV to homes in the U.S. is the single major problem area at this time. FCC testing of the five applicants for consideration as the broadcast standard has been delayed again, this time due to delivery problems with a standards converter being built by Tektronix. Final development of this complex unit has required hand-picked A to D converters because standard components cause visible defects in the picture.

The goal of the FCC is to complete on-the-air testing by Spring 1992, in order to have evaluation ready for Commission consideration by September of that year. Selection of a standard is currently planned for early 1993. Many industry and government observers believe that this is the latest possible date that is acceptable, and any further delay will certainly jeopardize the U.S. market position in the world HDTV market.

One industry participant, choosing to remain anonymous, muses, "Why are terrestrial standards holding things up when DBS and fiber are perfectly capable of delivering HDTV?" The strength of the broadcast lobby is one part of the answer to his question, but the situation is not quite that simple. First, neither fiber nor DBS is in widespread use in North America, and the time required to bring those technologies to "universal" status is unknown, perhaps longer than that required for a changeover to a broadcast system that at least approaches the quality available with HDTV.

With HDTV VCRs already developed, and given the current popularity of rented video programming, it will be extremely interesting to see how the average consumer reacts to broadcast television that has somewhat lesser picture quality than taped programming. Currently, the opposite is true.

Which leads us to the biggest question — will the consumer accept HDTV at all? The quality difference is analogous to that of the compact audio disc versus vinyl records, and CDs are enormously successful. However, the improvement offered by stereo AM broadcasting and quadraphonic FM has been generally ignored. Even stereo television sound has been under-utilized.

Will reruns of Gilligan's Island become more interesting on HDTV, or will program producers react to the quality of the medium and produce higher quality programs? The answer to this question is probably the key to HDTV's future, not the technical standards or hardware capabilities. **RF**

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A Channel Sounder For Indoor Communications Systems

By Fernando Casadevall and Xavier Barba
Department of Signal Theory and Communications
Catalunya Polytechnic University

From the beginning of the 80s the possibility of giving total mobility to the increasing number of communications terminal equipment in large buildings like factories, hospitals, warehouses, commercial centers etc., has emerged strongly due to the important advantages it brings. Even though an infrared optical system could be considered for achieving such an objective, it becomes inoperative when obstructed. For this reason, radio systems must be considered.

In radio communication systems, the multiple reflections of the transmitted signal from the structure of the building and surrounding inventory create multipath interference. Multipath has historically been identified as the most important factor limiting mobile and portable radio communication systems. For narrow-band systems (where the baseband digital symbol duration is several times greater than the extent of multipath-induced propagation delays, that is, flat fading conditions), multipath causes large fluctuations (fading) in the received signal voltage due to the changing phasor sum of the signal components arriving at the receiver antenna via different paths. Temporal variations of the channel as well as changing multipath geometry seen by a mobile user are the mechanisms for the fading. Therefore, an accurate characterization of the operating channel is a mandatory prerequisite for the development of reliable indoor radio systems.

Usually we could distinguish between a large scale model, used in determining coverage areas and co-channel interference levels, and a small scale model that describes the fading due to temporal variations of the received signal and used in determining the bit error rate and outage probability.

The expression that characterizes the large scale model is the Path Loss (L) given by $L=Kd^n$ where K is an appropriate constant, d is the distance between transmitter and receiver in meters and n is the mean path loss exponent. For stationary transmitters and receivers, the small scale model is usually characterized by a Rician distribution. In this case the most relevant parameter is the ratio of the received power of the specular radio signal to the average power of the scattered signal.

Measurements reported in References 1 and 2, indicate that n usually ranges between 2 to 6 and K between 1 to 16 dB depending on the building structure. So, there is a great value dispersion and it could be convenient to characterize the operating channel in each case.

In this paper we present a simple channel sounder appropriate for characterizing narrow-band indoor communications channels. The first part is devoted to analyzing the transmitter structure, the second studies the receiver system and the last presents some practical results and conclusions.

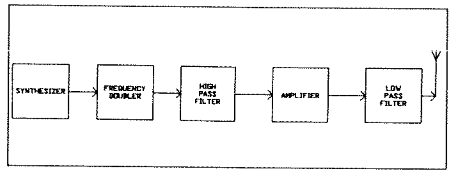

Figure 1. Schematic drawing of the CW transmitter.

Transmitter

The schematic drawing of the CW transmitter is shown in Figure 1. We can distinguish five parts. The synthesizer produces a sine signal whose carrier frequency ranges between 800-850 MHz. The frequency doubler changes the carrier frequency up to the 1.6-1.7 GHz band. This signal is not synthesized directly in the 1.6-1.7 GHz area because the voltage controlled oscillator (VCO) and counters (Prescalers) are cheaper and more readily available in the 800 MHz band than in the 1.6 GHz band. The high pass filter eliminates the

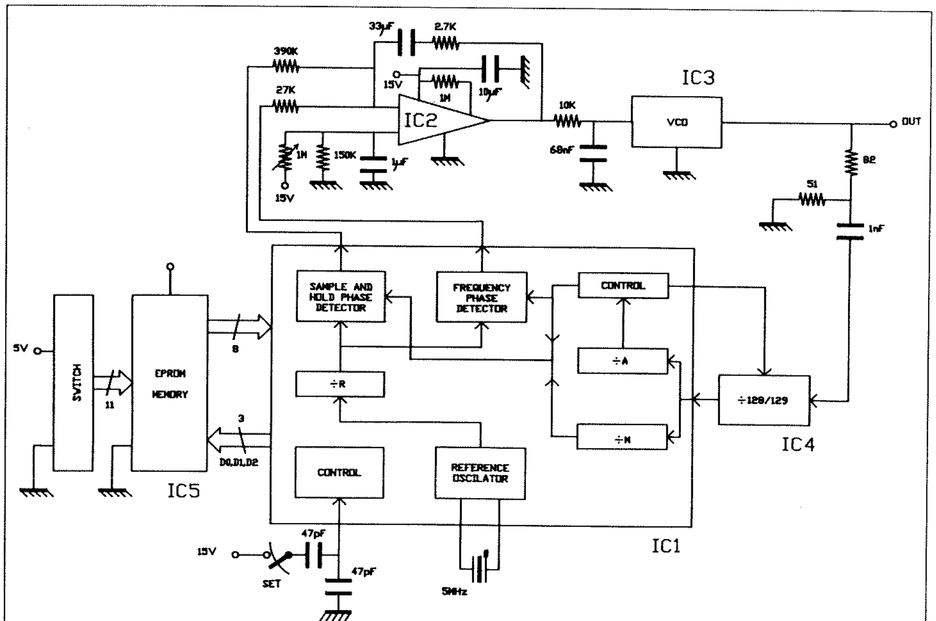

Figure 2. Schematic diagram of the synthesizer.

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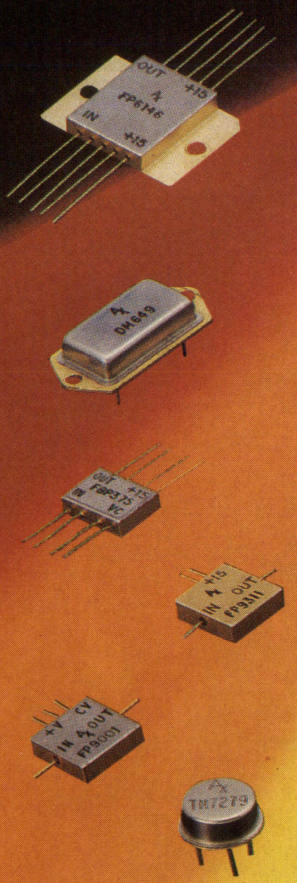

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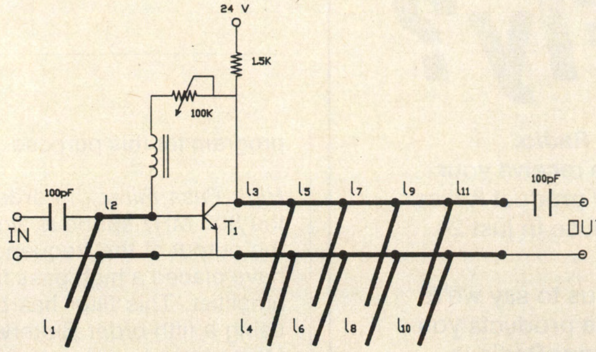


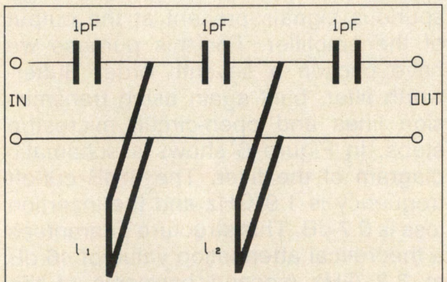
l_1 - 7.4 mm, 50Ω
l_2 - 22.5 mm, 50Ω
l_3 - 5 mm, 50Ω
l_4 - 12 mm, 50Ω
l_5 - 20 mm, 50Ω
l_6 - 8 mm, 50Ω
l_7 - 48 mm, 115Ω
l_8 - 48 mm, 45Ω
l_9 - 48 mm, 115Ω
l_{10} - 48 mm, 45Ω
l_{11} - 48 mm, 115Ω
T_1 - HXTR3615
Epoxy Substrate (E_r = 4.75, thickness 0.75 mm)

Figure 3. Frequency doubler.

spurious signals around 800 MHz that appear at the frequency doubler output. The output amplifier increases the useful signal level up to 20 dBm. Finally, the low pass filter attenuates the level of the harmonics present in the output signal.

In the paragraphs below we describe the most relevant aspects of each part.

Synthesizer - The synthesizer is built around the integrated circuit NJ8820 from Plessey. This circuit contains a reference oscillator, an 11 bit programmable reference divider, two phase comparators: one digital and one sample and hold type, two programmable dividers: one, called A, that counts down to zero from the programmed input, with a 7 bit length, and the other, called M, that counts up from zero to the programmed input and with a 10 bit length, and finally the necessary control and latch circuitry for accepting and latching the input data. By means of an external dual modulus counter (N/N+1), the synthesizer is able to program a total division ratio given by NM+A, where A is between 0 and 127, M is always greater than or equal to A and the minimum division ratio is N^2–N. However, the value of N must also be large



l_1 = 10.2 mm, 50Ω
l_2 = 10.2 mm, 50Ω
Epoxy Substrate (E_r = 4.75, thickness 0.75 mm)

Figure 4. High pass filter.

enough to guarantee that the output frequency from the dual modulus counter not exceed the maximum input frequency of the integrated circuit.

The frequency synthesizer is completed by an active loop filter and a VCO, as shown in Figure 2.

For each synthesized frequency, the NJ8820 circuit reads eight 4 bit words from an external memory. The timing signals needed are provided by the same integrated circuit. With this structure the synthesizer can produce a carrier frequency that ranges from 800 MHz to 850 MHz, with a frequency step of 25 KHz and whose level is around -5 dBm.

Frequency Doubler - The schematic of the frequency doubler is shown in Figure 3. It uses an HXTR3615 transistor as active device. The input matching network is composed of an open circuit microstrip stub and a transmission line, whereas the output matching network uses two open circuit microstrip stubs with two sections of transmission line between them (3). In this case the stub lengths are designed in order to present a 1/4 length at the second and the third harmonic of the output signal, respectively. In this way the network not only matches the output impedance but also introduces a selective effect to the undesired signals. A band stop filter, tuned at 800 MHz, is placed behind the output matching network. This filter is built using two open circuit microstrip stubs and three transmission line sections. Obviously this filter also eliminates the undesired signals placed around the third, fifth and seventh harmonics of the input signal.

It is important to emphasize that this circuit, and everything described below, have been designed using computer aided design (CAD) techniques. In particular, we used the SuperCompact program for this purpose.

High Pass Filter - In order to attenuate the 800 MHz spurious signal present at the output of the frequency doubler, we have placed a high pass filter before the amplifier. This filter has been designed using a fifth order Butterworth structure (4) and made using transmission lines and short-circuit stubs as shown in Figure 4. The 3 dB cut-off frequency is 1.56 GHz and the insertion loss is 1.34 dB. This structure has a theoretical attenuation value of 50 dB at the 800 MHz band and it guarantees sufficient attenuation of the spurious signals.

Amplifier - The signal level at the output of the frequency doubler is −5 dBm. Therefore, in order to obtain an useful signal of 20 dBm at the transmitter output, the amplifier gain must be approximately 25 dB. This amplifier has three stages with power gains of 10, 10 and 5 dB respectively. The first two stages make use of a HXTR3102 transistor as an active device whereas the output stage uses a BFG34 transistor able to produce the required output level. The matching network between stages is built by means of transmission lines and microstrip stubs in open and short circuit.

In Figure 5 we show the schematic diagram the amplifier. The last two microstrip stubs, placed at the output of the amplifier, are designed in order to attenuate the second and third harmonic respectively. It is important to emphasize that we have chosen for the second stage a short-circuit microstrip stub to diminish the low frequency gain of the amplifier and consequently to increase its stability.

Low Pass Filter - The low pass filter is devoted to attenuation, as much as possible, of the harmonics and other spurious signals present at the output of the amplifier. For this purpose we have chosen a seventh order Butterworth filter, built again using transmission lines and open-circuit microstrip stubs. In Figure 6 shows a schematic diagram of the filter. The 3 dB cut-off frequency is 1.9 GHz and the insertion loss is 0.7 dB. This structure guarantees a theoretical attenuation value of 46 dB at 3.2 GHz (second harmonic of the output signal) and more than 70 dB at 4.8 GHz (third harmonic). These values are sufficient to obtain an output rejection to harmonics and spurious signals greater than 60 dB with respect to the

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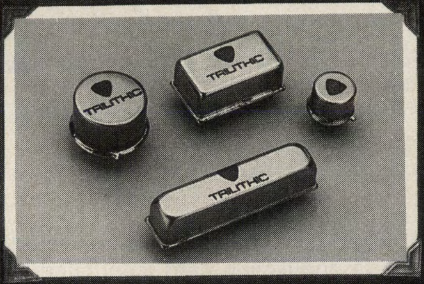
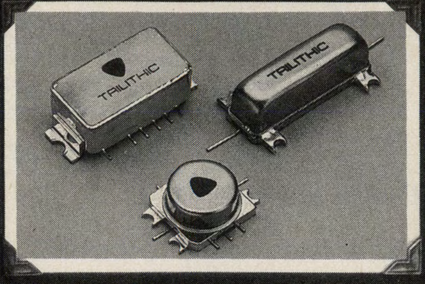
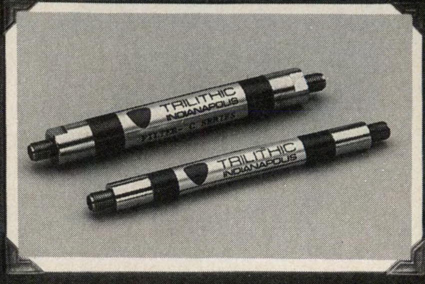


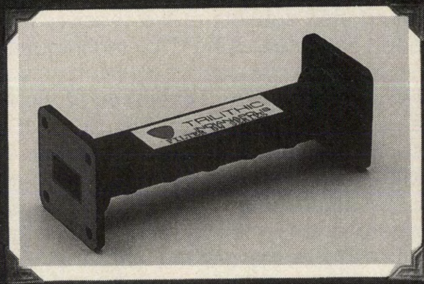
Miniature filters



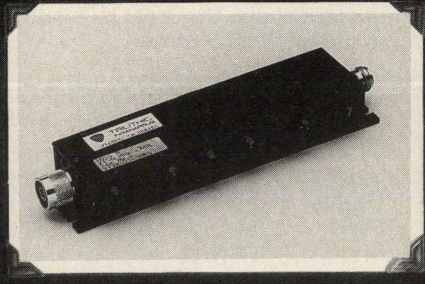
Surface mount filters



Tubular filters

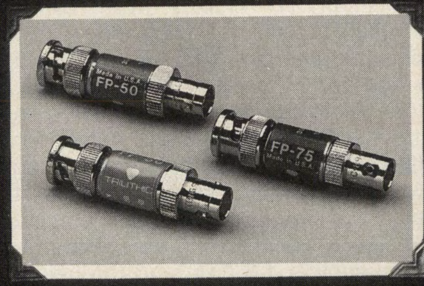


Waveguide filters

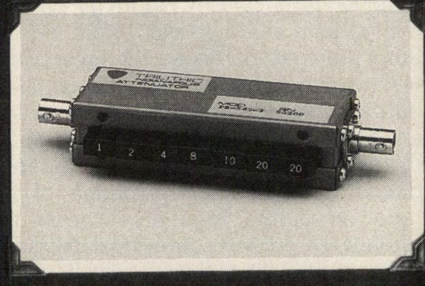


Cavity filters

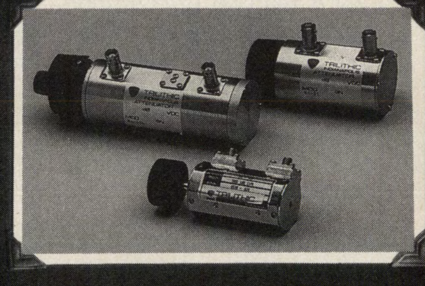
Tunable filters



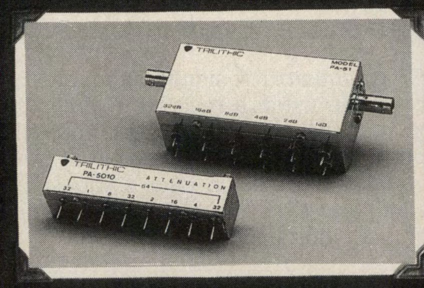
Coaxial attenuators



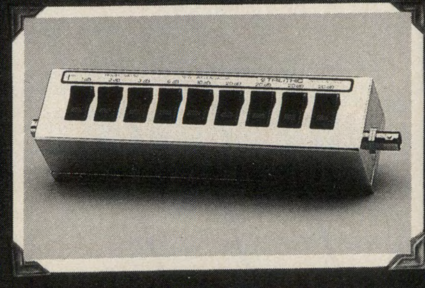
Pushbutton attenuators



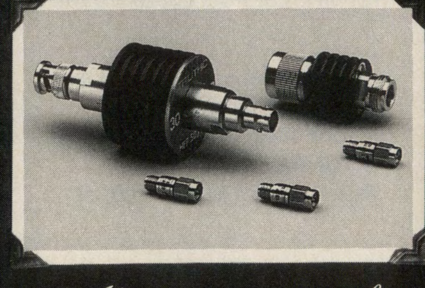
Rotary step attenuators



Programmable step attenuators



Switch attenuators



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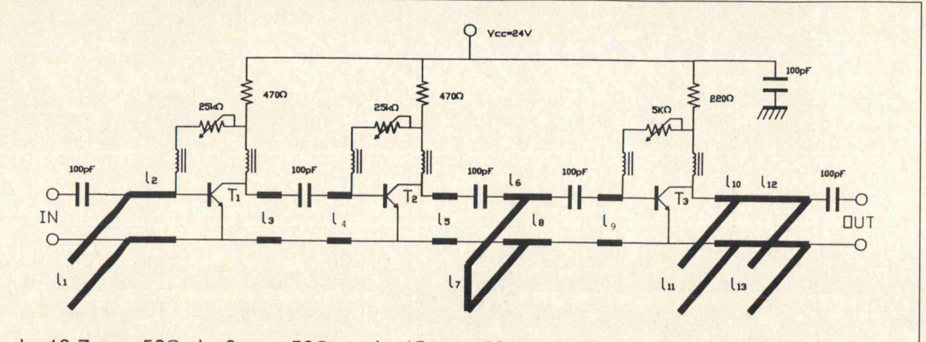


l_1=19.7 mm, 50Ω	l_4=6 mm, 50Ω	l_7=15 mm, 50Ω	l_{10}=1 mm, 50Ω	l_{13}=12.5 mm, 50Ω
l_2=6.8 mm, 50Ω	l_5=3.7 mm, 50Ω	l_8=3.5 mm, 50Ω	l_{11}=8.4 mm, 50Ω	T_1=HXTR3102
l_3=6 mm, 50Ω	l_6=3.7 mm, 50Ω	l_9=3.5 mm, 50Ω	l_{12}=20 mm, 50Ω	T_2=HXTR3102
T_3=HXTR3102				

Epoxy Substrate (E_r = 4.75, thickness 0.75 mm)

Figure 5. RF amplifier.

useful signal.

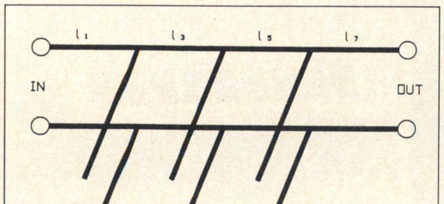
Transmitter Performance - Once the transmitter was made, a complete set of measurements was taken for its characterization. The most relevant features are described below:

- -Output level: 18.5 dBm - 0.5 dB
- -Frequency Range: 1.6 GHz to 1.7 GHz, with 50 kHz steps
- -Spurious and harmonic rejection: Greater than 60 dBC
- -SWR: 1.5 at 1.65 GHz, <1.85 between 1.6-1.7 GHz

Receiver

The schematic drawing of the CW receiver is shown in Figure 7. We can distinguish six different parts: an RF amplifier that increases the level of the weak signal present at the antenna output; an RF attenuator that generates an appropriate level at the input of the mixer; the mentioned mixer that produces a frequency translation of the received signal to a new carrier frequency of 70 MHz; the local oscillator associated with the mixer; an IF amplifier, and finally a logarithmic amplifier that produces an output signal proportional to the received power at the antenna port. This signal is delivered to a data acquisition card, controlled by a personal computer (PC), in order to be processed properly.

As with the transmitter, all the above mentioned subsets of the receiver have been designed using CAD techniques.



l_1= l_5= 9.2 mm, 115Ω
l_2= l_6= 10 mm, 22Ω
l_3= l_7= 8.2 mm, 115Ω
l_4= 10 mm, 11Ω
Epoxy Substrate (E_r = 4.75, thickness 0.75 mm)

Figure 6. Low pass filter.

RF Amplifier - The schematic of the RF amplifier is shown in Figure 8. It is composed of two amplifier stages and a high pass filter. The first amplifier stage is designed to optimize the noise figure whereas the linearity aspects are emphasized in the second amplifier. Both use common emitter structure with transmission lines and open circuit microstrip stubs for the matching circuits. The transistor HXTR3615 from Hewlett-Packard is used as an active device. The power gains are 12.5 dB for the first stage and 13.5 dB for the second, and the 3 dB bands are 1.5-2 GHz and 1.5-1.9 GHz respectively. A high pass filter is placed between both stages which eliminates the noise and all the spurious signals below 1 GHz that could appear in the antenna port. The cut-off frequency is 1.3 GHz and the insertion loss is 0.3 dB. This high pass filter is placed after the first amplifier in order to optimize the noise figure of the receiver.

RF Attenuator - A simple attenuator can be made using a resistive pi structure. If the resistor values are controlled properly this circuit is capable of giving variable attenuation over a wide bandwidth and provide matched impedances in both ports. On the other hand, for

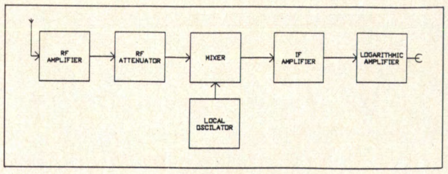
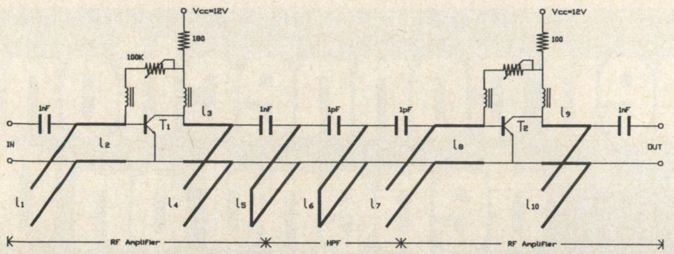


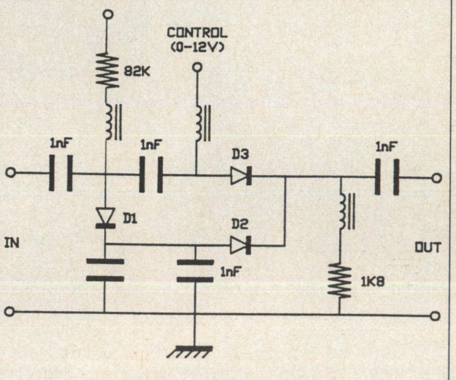
Figure 7. Schematic drawing of the CW receiver.



l_1= 6.4 mm, 22Ω
l_2= 8 mm, 50Ω
l_3= 9 mm, 82Ω
l_4= 37 mm, 50Ω
l_5= 12.5 mm, 50Ω
l_6= 12.2 mm, 50Ω
l_7= 12.7 mm, 22Ω
l_8= 2.5 mm, 45Ω
l_9= 11.6 mm, 74Ω
l_{10}= 33 mm, 45Ω
T_1= HXTR3615
T_2= HXTR3615
Epoxy Substrate (E_r = 4.75, thickness 0.75 mm)

Figure 8. Receiver part: RF amplifier.

small signals at RF frequencies, the PIN diode, when it is conducting, appears essentially as a pure resistor whose value is roughly inversely proportional to the bias current. Moreover, the quality of this resistor, i.e. its linearity and lack of distortion products, can be very good if the diode is used under the right conditions. So, for these reasons the RF attenuator has been made using a PIN diode attenuator, as shown in Figure 9, where the series and parallel bias currents are properly controlled by means of a single control voltage (5). The insertion losses are 1.5 dB, the attenuation reaches up to 18 dB and the SWR ratio is lower than 2 dB in the full range of the attenuation values.



D1 = D2 = D3 = 1N5767 Diode PIN

Figure 9. Receiver part: RF attenuator.

Mixer - This circuit is built around the SRA-2000 double balanced mixer from Mini-Circuits. In order to guarantee that the RF port and the IF port have 50 ohms at any frequency, special matching networks have been used (6), as shown in Figure 10. The input network is composed of an impedance transformer, made by means of a transmission line, and a short circuit microstrip stub in series with a 50 ohm resistor. At the working frequency the stub appears as an open circuit. Then, the impedance transformer changes the output impedance of the RF amplifier (50 ohms) to the nominal mixer impedance of 200 ohms. At lower frequencies the stub appears as a very small inductive impedance. A 50 ohm resistor is placed at the input of the impedance transformer regardless of the source impedance value. Really, at these frequencies, the 10 pF capaci-

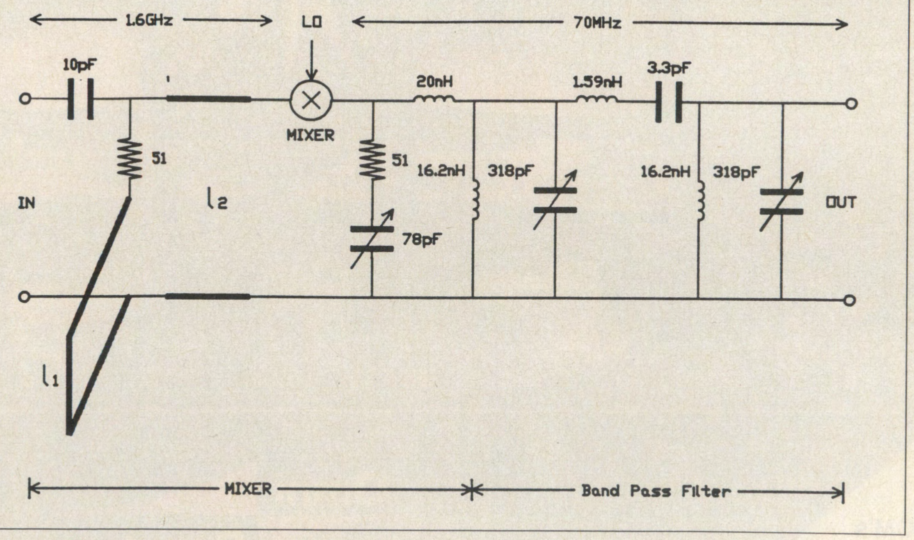


Figure 10. Receiver part: mixer.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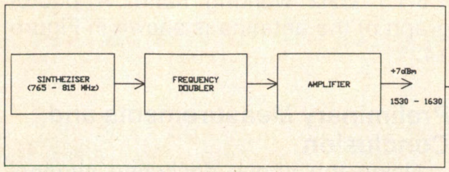


Figure 11. Local oscillator chain of the receiver part.

tor guarantees a joint impedance (source plus capacitor) greater than 50 ohms. The output network acts as a low pass filter with a cut-off frequency around 400 MHz. The insertion losses of the mixer and matching networks are approximately 9 dB.

A bandpass filter is placed at the output of the above mentioned circuit in order to reduce the noise bandwidth of the receiver. This filter is a third order Butterworth structure with 10 MHz bandwidth.

Local Oscillator - The schematic drawing of the local oscillator chain is shown in Figure 11. We can distinguish three parts. The synthesizer that produces a sine signal whose carrier frequency ranges between 765-815 MHz and has the schematic shown in Figure 2, but with different words recorded in the memory; the frequency doubler that changes the carrier frequency up to the 1530-1630 MHz band (its schematic is shown in Figure 3) and the amplifier that increases the useful signal level to 7 dBm and whose schematic corresponds to the first two amplifier stages shown in Figure 5.

IF Amplifier - The IF amplifier provides the additional gain necessary to guarantee that, even with a signal level equal

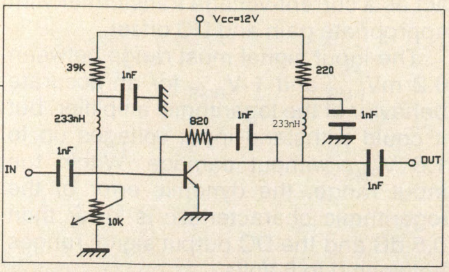
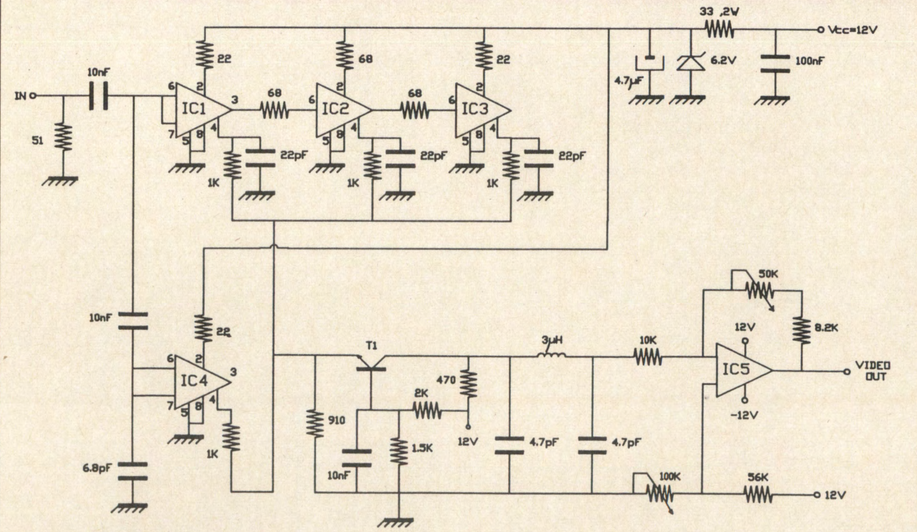


Figure 12. Receiver part: IF amplifier.

to the receiver sensitivity, the IF signal delivered to the input of the logarithmic amplifier has an appropriate level. The schematic is shown in Figure 12, where we can see that the amplifier uses a common emitter structure with an AC feedback between the collector and the base. This feedback permits us to provide the appropriate 50 ohm value to the input and output impedance. The power gain at 70 MHz is 24 dB and the amplifier band ranges between 20 to 140 MHz.

Logarithmic Amplifier - The logarithmic amplifier is a device that produces a video output which is proportional to the logarithmic value of the IF input amplitude (7). It uses a cascade of logarithmic video amplifiers which permits almost 70 dB of the dynamic range. This amplifier whose schematic diagram is shown in Figure 13 is basically built using three SL623B and one SL621B integrated circuits from Plessey. These integrated circuits are intended primarily for use in successive detection logarithmic IF strips, operating at center frequencies between 10 and 140 MHz. For the SL621B the mid-band gain voltage



IC1=IC3=IC4=SL523 from Plessey
IC2 = SL521 from Plessey
IC5 = μA741

Figure 13. Receiver part: logarithmic amplifier.

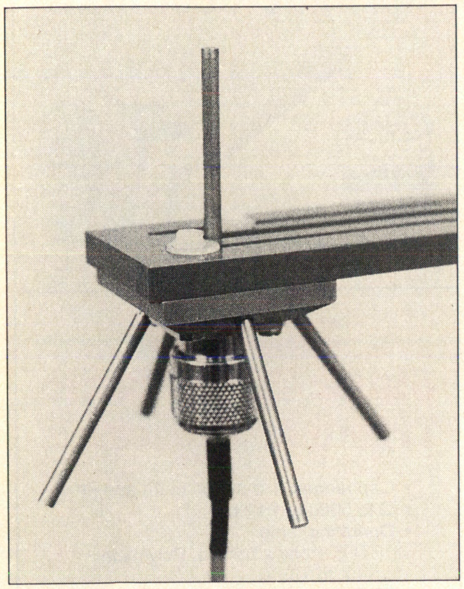


Figure 14. Photograph of the antenna.

is around 12 dB whereas for the SL623B it is 24 dB. The former provides a logarithmic characteristic over a 10 dB range and the later over 20 dB. As the integrated circuits provide a current whose level is proportional to the logarithm of input signal (with maximum level equal to 2.1 mA), the common base transistor and the operational amplifier act as a current-voltage transformer with appropriate gain and DC offset.

The input signal must range between 0.2 mV_{RMS} and 1 V_{RMS} for an accurate behavior of the logarithmic amplifier, but it could withstand input voltages up to 1.8 V_{RMS} without damage. Within the input range, the dynamic error of the logarithmic characteristic is lower than 0.5 dB and the DC output signal ranges between 0 to 5 Volts.

Receiver Performance - Once the receiver was built, we obtained a complete set of measured characteristics, whose most relevant aspects are described below:

- Sensitivity: −97 dBm
- Maximum input level: 0 dBm
- IF Bandwidth: 10 MHz
- Logarithmic amplifier output: 0-5 Volts
- CAG: 18 dB manually controlled
- SWR: 1.9 at 1.65 GHz, <2.2 between 1.6-1.7 GHz

Antennas

In order to complete the channel sounder, two monopole antennas with ground planes were constructed. We used a 1/4 monopole with four bars inclined 45 degrees to simulate the ground plane. The measured SWR at the antenna port is lower than 1.2 in the 1.6-1.7 GHz working band. A photograph of the antenna is shown in Figure 14.

Preliminary Measurements and Conclusion

Using the above described channel sounder, some preliminary measurements of the propagation conditions inside buildings were done. We recorded the received signal during 20 minute periods and using 20 samples/sec. In Figure 15 a typical example of the recorded measurements of the received signal amplitude in two standard cases is shown. The first case is line of sight between the transmitter and receiver and the second when obstructed. In both cases we can see large fluctuations (fading) in the received signal voltage. As was explained in the introduction, these fluctuations are due to the changing phasor sum of the signal components arriving at the receiver antenna via different paths. Presently a complete set of measurements is in process in order to characterize the transition channel in different buildings constructed using different materials.

In conclusion, a simple narrow-band channel sounder, appropriate for characterizing narrow band indoor communications channels, has been described.

Acknowledgements

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RF

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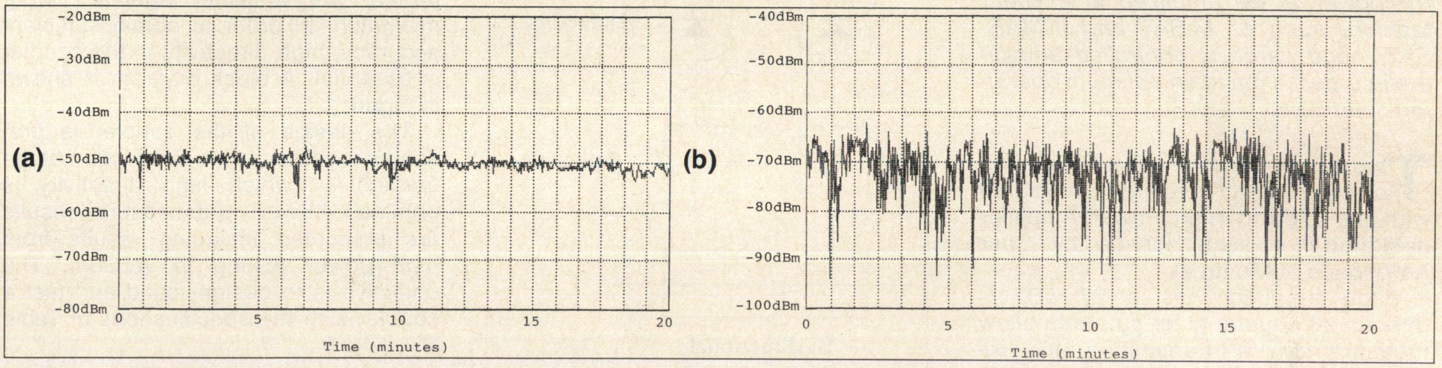


Figure 15. Signal level at the antenna port versus the time, (a) line of sight, (b) obstructed.

High Power Bi-Directional Coupler For Space Applications

By Pascal R. Goux
Ancom Electromagnetics Ltd.

Satellites requiring their transmitted RF power to be amplified in a linear fashion, such as M-SAT and INMARSAT, need carefully controlled amplifiers in order to maintain optimum operating conditions.

This paper describes the RF portion of the control circuitry associated with the L-Band High Power Amplifier designed for such service by Spar Aerospace of Montreal.

Each amplifier has a 27 dB bi-directional coupler at its output to allow measurement of transmitted and reflected signal power levels. The resulting detector signals are used to limit output power, hence power consumption without degrading linearity, and to protect the amplifier against output mismatch. In order to obtain sufficient accuracy, high directivity in the coupler is essential. A block diagram is shown in Figure 1.

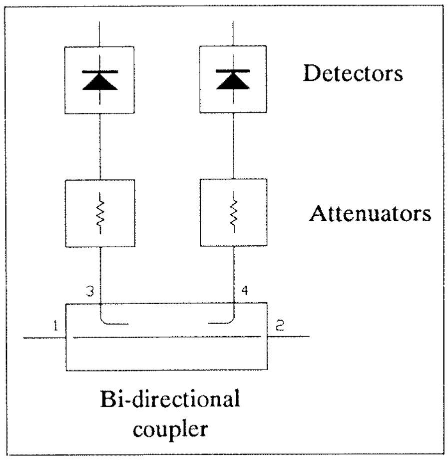


Figure 1. Block diagram.

The design of the coupler is presented together with a description of the method by which high directivity is obtained. Artwork and measured results are presented including results from high power testing in vacuum. The objective is to design, build and test a coupler with the specifications of Table 1.

Choice of Configuration

In order to be compatible with the rest of the circuitry making up the amplifier, the coupler must be fabricated using microstrip techniques on metallized alu-

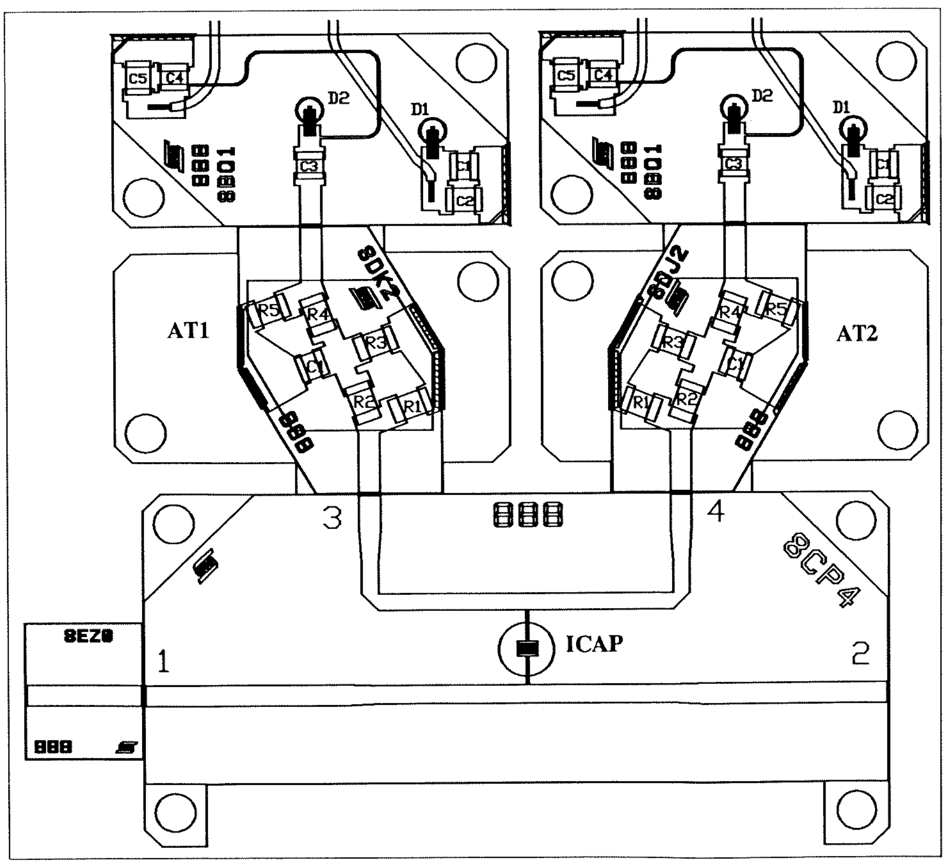
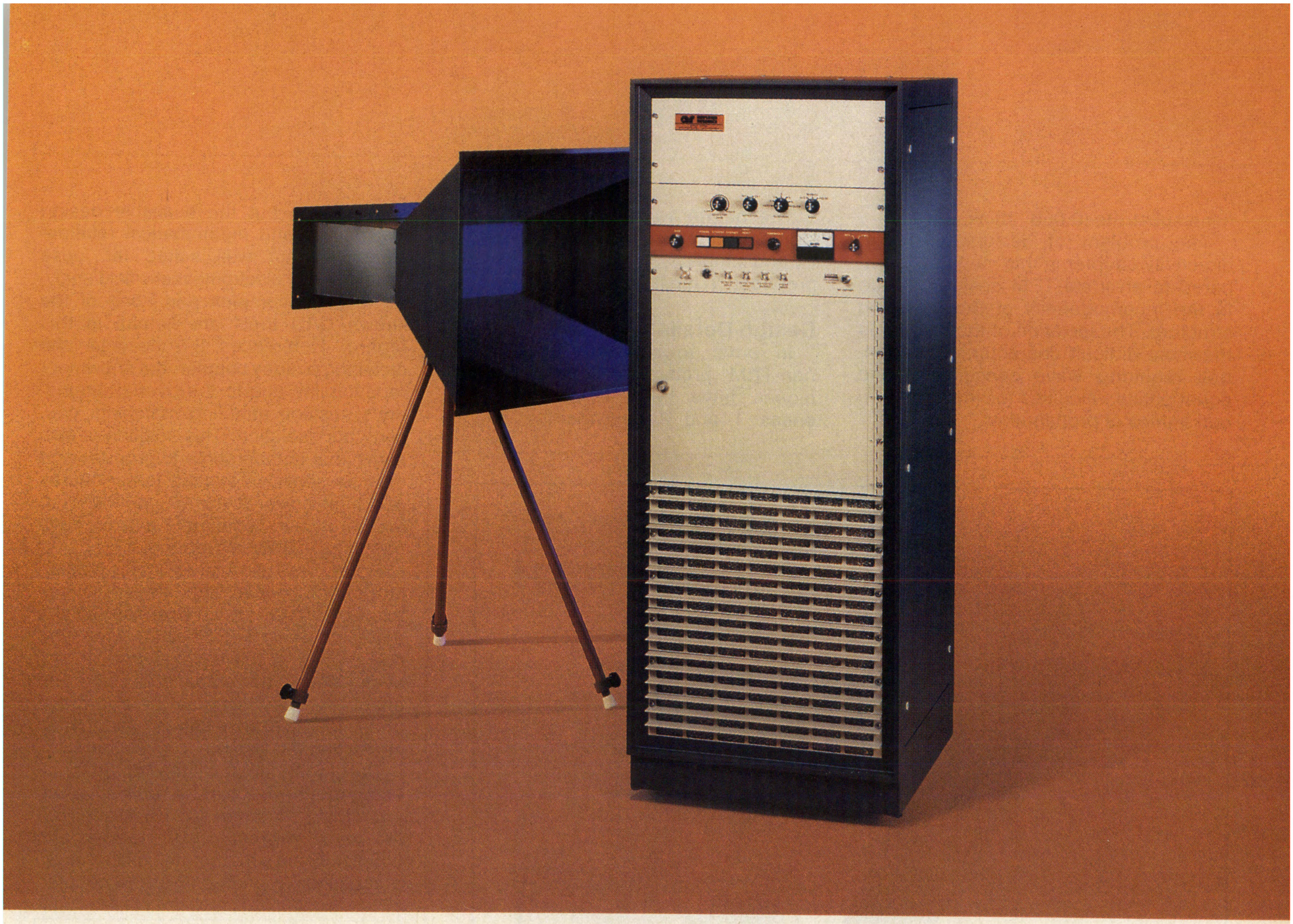


Figure 2. Artwork.

Operating Frequency	1 to 2 GHz
Low Insertion Loss (to minimize impact on amplifier efficiency)	⩽0.15 dB
Low Coupling Value (to minimize insertion loss)	⩽−25 dB
High Directivity (to differentiate forward & reflected signals)	⩾25 dB
High Isolation (to achieve high directivity with low coupling)	⩾50 dB
Return Loss (to guarantee the directivity)	⩾25 dB
Must sustain high CW power and be temperature insensitive	⩾100W −60°C +120°C

NOTE: Must also be bi-directional, symmetrical and easy to fabricate (preferably not to require tuning).

Table 1. Design specifications.



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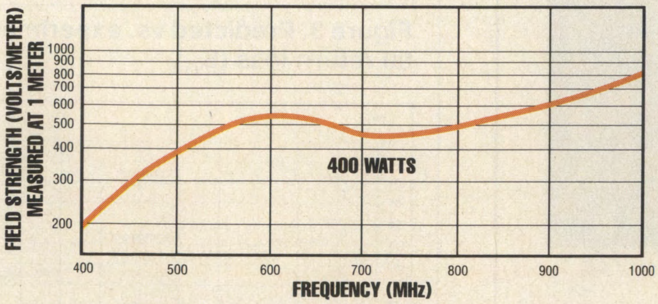

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mina substrate (0.050'' thick, $\varepsilon_r = 9.9$).

The problem of designing a high directivity coupler in this medium, (compared to an ideal TEM design), is due to the inhomogeneity of the dielectric structure. The isolation of the coupler is the most difficult characteristic to control, requiring wave cancellation and equal phase velocities for the even and odd modes of propagation.

The quarter-wave edge coupled design approach is chosen for its symmetry and simplicity, however the non-TEM propagation must be compensated.

Design Details

In order to solve the problem of non-TEM propagation, various techniques have been proposed. References 1 and 2 describe the Wiggly coupled line but the available design information and mathematical relationships are not well suited for loose coupling. The electromagnetic field analysis is quite complex and none of the known CAD tools are helpful in this regard. References 3-5 describe the dielectric overlay coupler, but show poor repeatability due to the overlay boundary's position affecting matching, coupling and isolation. This solution is also somewhat cumbersome to produce and may be sensitive to high temperatures and high power levels. A configuration that is simple, repeatable and requires no tuning would be preferable.

References 6-10 present the design of an edge coupler with lumped capacitors at each end to compensate for the unequal phase velocities of even and odd modes at the isolated port. In the even mode, the capacitors have no effect, but they increase the odd mode phase length (transit time, ε_{reff} odd), thus equalizing the waves.

Due to the small value of the capacitance required, it can be realized as an interdigital structure (ICAP) formed on the same microstrip substrate as the coupler. From a mathematical approach, a capacitor at each end of the coupled

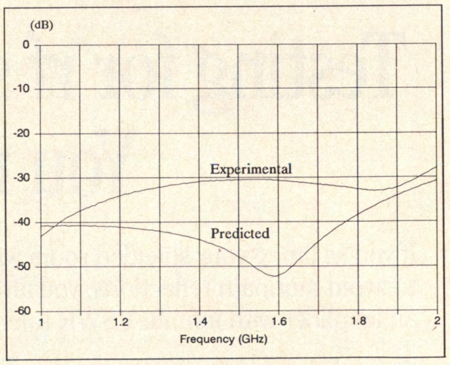

Figure 3. Predicted vs. experimental return loss (S_{11}).

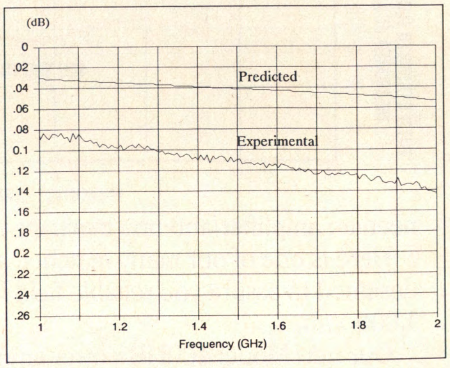

Figure 4. Predicted vs. experimental insertion loss (S_{21}).

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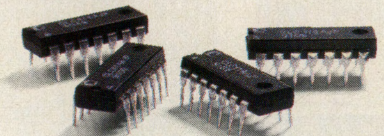

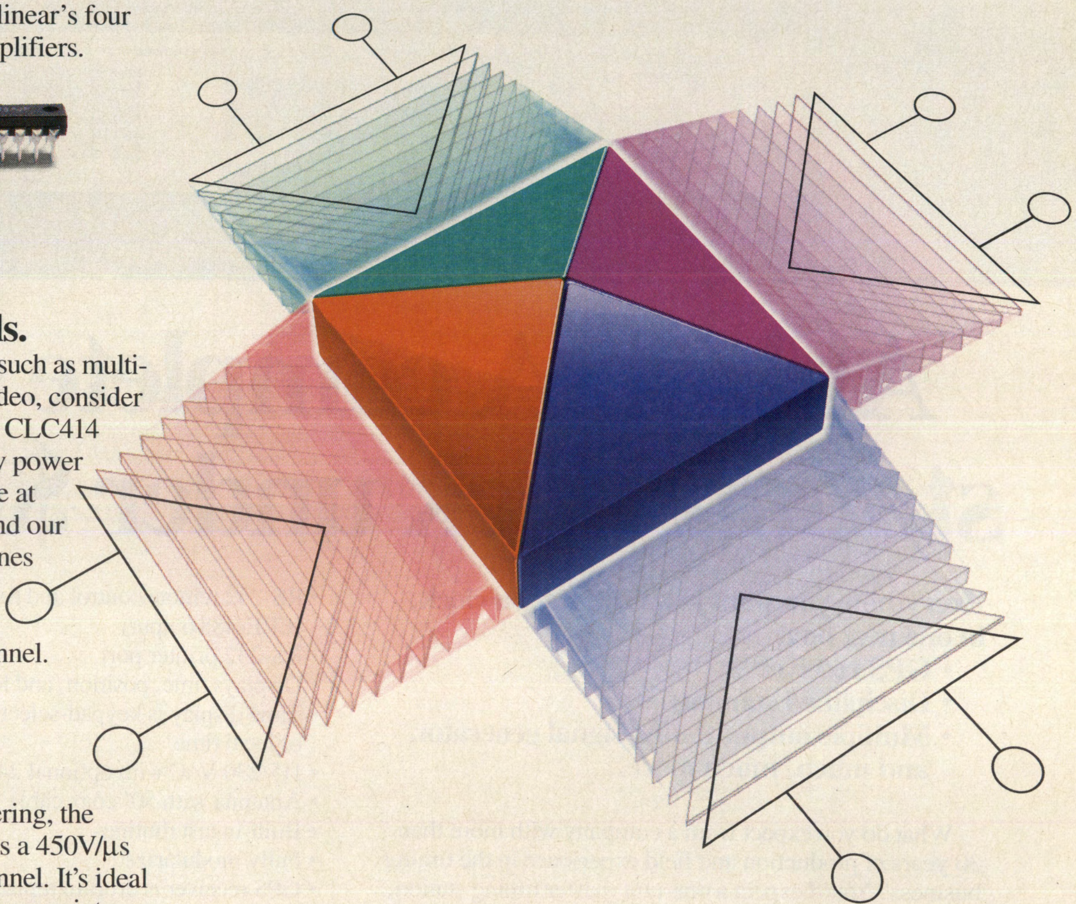

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line is used, (they must be separated by a quarter-wave length in order not to generate scattered waves to all ports). As mentioned in Reference 11, the capacitor at either end is satisfactory, but in order to have a good return loss, the capacitor should be located in the middle of the coupler to preserve the double symmetry of the circuit. The possibility of using just one capacitor located in the middle of the coupler was then investigated.

With this configuration, some characteristics tend to conflict and it is theoretically impossible to obtain both coupling flatness and optimum isolation. The selection of the capacitor requires evaluating the major performance parameters in terms of the intended service. In our case, the slope in frequency response of the coupled port can be conveniently equalized with a capacitor in the middle of the double-pi attenuator. Note that if the ICAP varies, the slope does not change, only the level changes. The design therefore is made for maximum isolation and the capacitor value is calculated to be 0.11 pF. The model of the ICAP was derived from Reference 12. It is important to note that the ICAP has a significant shunt component that can affect the even mode. In order to reduce this undesired shunt capacitance, the ground plane below the ICAP is removed. Figure 2 details the artwork and layout of the coupler, attenuators and detectors. Note the ICAP and compensating capacitors (C1) in the attenuators (AT1 and AT2).

The coupling value was chosen for the best tradeoff between low insertion loss (loose coupling) and high directivity (tight coupling because of fixed isolation). Optimum power levels and slope for the Schottky diode detectors (D2) is then adjusted with the appropriate attenuator. The computed prediction and experimental results are shown in Figures 3-6.

Results

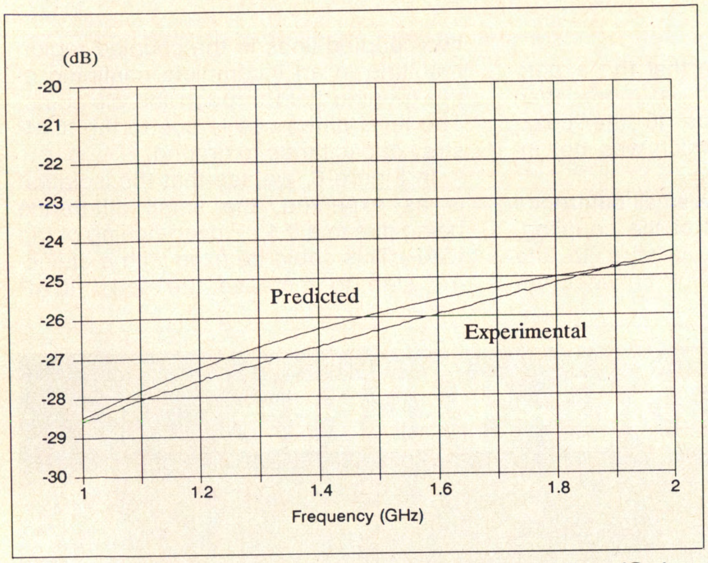
Due to the double symmetry of the coupler (verified experimentally), only four parameters are shown: the return loss (S_{11}) in Figure 3, the insertion loss (S_{21}) in Figure 4, the coupling (S_{31}) in Figure 5 and the isolation (S_{41}) in Figure 6. Comparing these experimental results and computer predictions, we can see some small explainable differences.

In Figure 3, we see that the predicted return loss is better than the one obtained experimentally. This is attributed to the connector-to-microstrip discontinuities.

In Figure 4, we see that the experimental insertion loss is more than expected. This is due to the coax-microstrip transition which was not included in the simulation.

In Figure 5, we see a small difference in the slope and level of the coupling. The slope difference is due to the presence of the capacitor between the two coupled lines in the coupled region resulting in an incomplete confining of the fringing field through the dielectric. The level difference is due to the ICAP stray capacitance to ground.

In Figure 6, we see that the isolation is as expected, the small difference being due to a 0.1 pF over-etching of the ICAP. This could be fixed with a dielectric slab to increase the serial ICAP



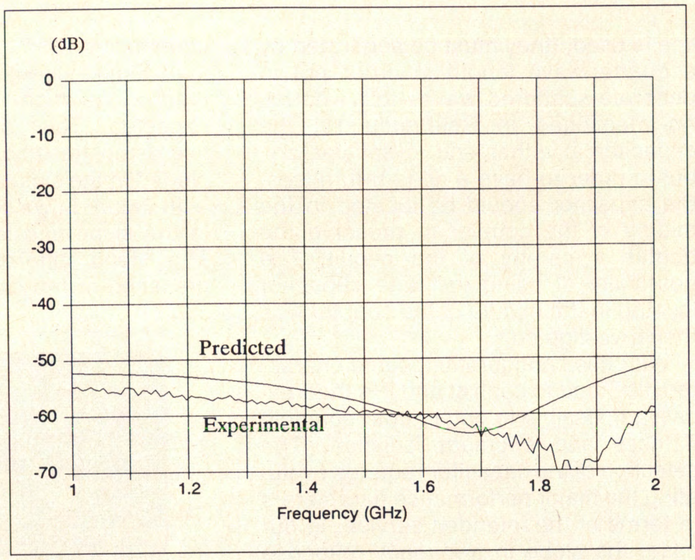


Figure 5. Predicted vs. experimental coupling (S_{31}).

Figure 6. Predicted vs. experimental isolation (S_{41}).

value (note that all results are without tuning). *RF*

Acknowledgement

This paper was written while the author was employed by Spar Aerospace as an RF/Microwave engineer.

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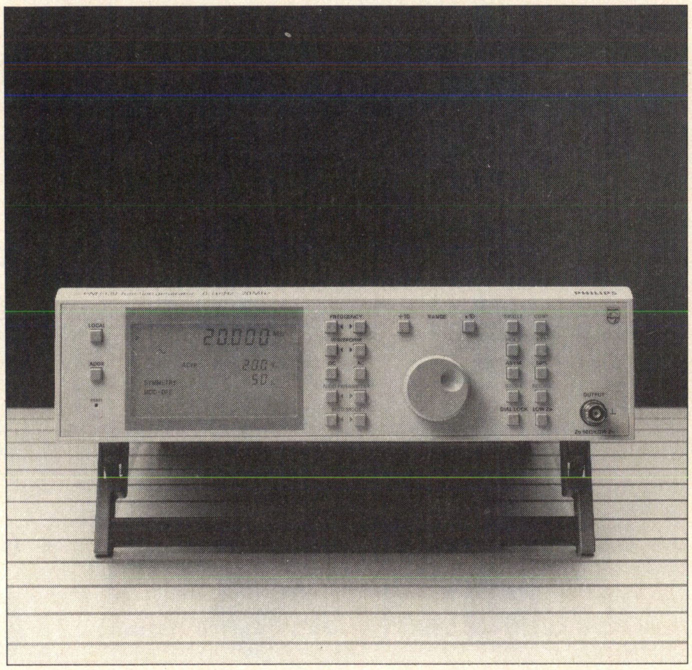


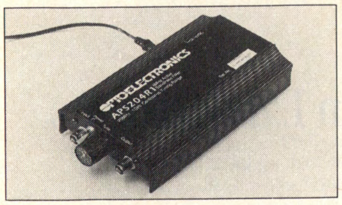
Six-Octave Bandpass Filter

Optolectronics announces a new receiver bandpass filter that separates closely space radio signals. The Model APS-204 has a constant 4 MHz bandwidth, tunable over a 20 MHz to 1000 MHz frequency range.

The unit is an active circuit with no loss and very small size (4 × 1 1/2 × 7 inches). At the heart of the circuit is a 4 MHz bandwidth, 4-pole resonant cavity filter with a Q of 325. A unique double-hetrodyne technique currently being patented routes the signal to the filter and back to the operating frequency.

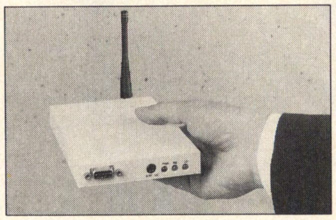
Noise figure is specified at 10 dB, and the third-order intercept is 15 dBm, typical. Power requirements are 12 VDC with 6 watts consumption. Price of the APS-204 is $995.

Optoelectronics Inc.
INFO/CARD #229



RF Modem for Wireless Networking

The RF Modem, using unlicensed spread spectrum tech-



niques defined by FCC Part 15, has been introduced by Proxim. Applications include hospital bedside monitors, point-of-sale devices, portable data recorders, traffic management, and security systems.

Usable for distances up to 500 feet, the RF Modem interfaces to a computer using a standard RS-232 or RS-485 port. Packages and unpackaged versions are available.

Two unpackaged version are available for designing into OEM products. One operates at 121 kbps, while the other operates at 242 kbps.

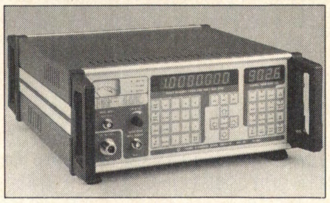
Proxim, Inc.
INFO/CARD #228

Farnell Instruments Now Available in U.S.

Wayne Kerr announces the introduction of the Farnell line of RF test instruments to the U.S. and Canadian markets. Farnell equipment has previously been marketed in the U.K. and the rest of Europe.

The product line includes the following signal generator products: The PSG1000 Synthesized Signal Generator (pictured below) covering 10 kHz to 1 GHz, the ESG Signal Generator which also covers 10 kHz to 1 GHz, the 100 kHz to 2 GHz SSG2000 Signal Generator. Other current product models include the 352C Spectrum Analyzer, covering 300 kHz to 1 GHz, and the AMM2000 Modulation Meter with 250 kHz to 2.4 GHz coverage.

Wayne Kerr Inc.
INFO/CARD #227

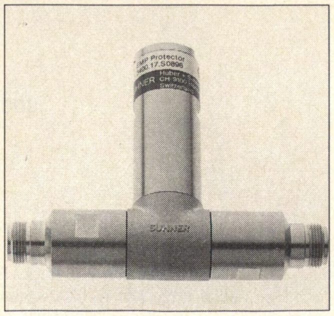


Lightning Protector for Cellular Radios

Huber+Suhner has introduced a 1/4-wave shorted stub specifically designed for cellular radio base stations, with an operating frequency range of 797-1016 MHz. Features include lower residual pulse than other protection methods and no intermodulation effects. The unit is offered with either type N or 7/16 connectors.

Mechanically, the lightning protector features zinc diecast construction, waterproof design for outdoor installation, and easy mounting with standard plumbing brackets or clamps.

Huber+Suhner AG
INFO/CARD #224



Smaller SMT DIP Package

Signetics announces the first two integrated circuit products in their new SSOP small-footprint SMT package. The package is small enough to replace chip-and-wire assembly in some applications. The NE/SA575DK Compandor and the NE/SA605DK FM IF System are available in the 1.5 × 4.5 × 6.75 mm SSOP. In 100s, these device are priced at $2.04 and $3.54, respectively.

Signetics Company
INFO/CARD #226

GPS Precise Positioning Module

Stanford Telecommunications has received National Security Agency endorsement of its STEL-9300 Precise Positioning Service - Security Module (PPS-SM), allowing authorized users to realize the most accurate capabilities of GPS. Anti-Spoof and Selective Availability techniques deny access to the Precise Positioning Service (PPS) to unauthorized users. With this capability, GPS equipment can provide access to PPS for authorized users without making the equipment classified.

Stanford Telecommunications
INFO/CARD #225

Standard Power Amplifier Line

Alpha Industries announces the new CMA Series of Standard Power Amplifiers for terrestrial and satellite microwave communications systems. Models cover 5.9-6.4 GHz with 60 dB linear gain (minimum) and power outputs ranging from 40 to 44 dBm (minimum), utilizing +12 or +15 volt power supplies.

Alpha Industries, Inc.
INFO/CARD #224

Power Measurement Upgrade

High speed CW power measurement capabilities have been added to the 8003 Precision Scalar Analyzer from Wavetek Microwave. These new measurement modes provide power readings at greater than 100 per second over GPIB. The upgrade provides a 10 to 100 times improvement in speed over thermocouple CW power meters. This speed is particularly attractive in automated measurement systems.

Wavetek Microwave, Inc.
INFO/CARD #223

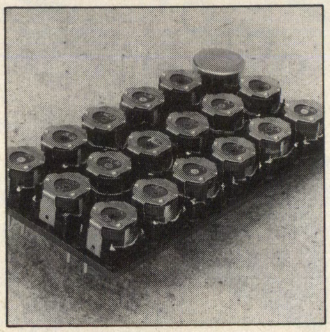
1000 Watt Amplifier

Model BHE4819-1000 from Power Systems Technology offers 1000 watts CW power output in AB linear service over 400-1000 MHz. RF input is 0 dBm, in CW, AM, Pulse, FM, or phase modulation. Protective circuits prevent damage from excessive VSWR, and the unit offers graceful degradation in the event of transistor failure. Applications include EMI testing, EW and ECM, troposcatter, television broadcasting, and laboratory amplification.

Power Systems Technology Inc.
INFO/CARD #222

Video Brickwall Filters

Miniature filters for NTSC and PAL video systems are available from Allen Avionics, offering cut-



off frequencies from 2 to 8.4 MHz. P.C. board or packaged versions are available. The typical ABW-4P20-P has passband flatness of 0.15 dB to 4.2 MHz, less than ± 10 nsec delay distortion, and 1.066 stopband ratio. These filters reduce interference from high frequency subcarriers and other out-of-band signals.

Allen Avionics, Inc.
INFO/CARD #221

ECL Oscillators With Enable/Disable

MF Electronics announces the M1900 ECL oscillators for graphics and other high speed circuits requiring clocks of 10 to 225 MHz. A unique feature is a "tristate" enable/disable feature which effectively disconnect the output. System test, selectable clock frequencies, and noise reduction by shutoff of unused sources are all benefits of this feature.

MF Electronics Corp.
INFO/CARD #220

Multipath and Fading Simulator

The new HP 11757B multipath-fading simulator/signature test set features an internal printer, firmware and error counters to produce permanent records of M-curves. These signatures are used in installation and maintenance of digital radio systems. The HP 11757B is $19,000, and an upgrade to the earlier HP 11757A is $4000.
Hewlett-Packard Co.
INFO/CARD #219

MIL-STD-2000 Capability

Bliley Electric announces that it is qualified and certified to design and manufacture soldered electrical and electronic assemblies to meet all requirements of MIL-STD-2000. This capability is used in custom-designed crystal oscillators, and is also available for contract assembly.
Bliley Electric Co.
INFO/CARD #218

Feedthrough Terminations

Coaxial feedthrough terminations designed to terminate inputs to oscilloscopes, frequency counters, RF voltmeters, etc., have been introduced by Elcom Systems. They are available in 50, 75 and 93 ohms, with BNC, TNC, type N or SMA connectors. Prices start at $17.50 in small quantities.
Elcom Systems, Inc.
INFO/CARD #217

Hybrid Amplifiers

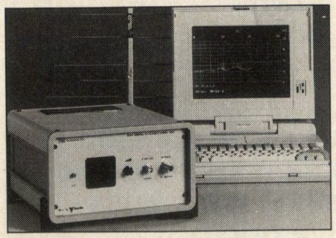
Two new 10 MHz - 1.2 GHz high linearity amplifiers have been introduced by Motorola RF Devices. Models CA5900 and CA5915 are designed for operation in 50 ohm linear systems, offering 28 or 15 volt supply, 15 dB gain, and 1 watt power output at 1 dB compression. Packaging is in the CATV style case 714P-01, and they are also available with a low profile heat sink package. Pricing is $91.30 (U.S. delivery, 100s).
Motorola RF Devices
INFO/CARD #216

3-Way Divider

New three-way power dividers/combiners are available from RLC Electronics. Featuring 5 watt dissipation and either pin or SMA connector configurations, standard units are offered in octave or narrow band ranges from 500 MHz to 12 GHz. Amplitude balance is ±0.5 dB, phase balance is ±3 degrees, and isolation is 18 dB minimum. Prices start at $225.
RLC Electronics, Inc.
INFO/CARD #215

EMI Measuring System

The Vianello PMM-8030 from IBEX Group is a complete emission measuring and monitoring system in the 30 to 1000 MHz band, meeting CISPR, VDE, FCC and EN specifications. External PC control offers ease of use and flexibility. A complete package is available, including receiver, antennas, computer and software.



IBEX Group Inc.
INFO/CARD #214

SMT RF Chokes

Vanguard Electronics is offering surface-mount RF chokes designed and built to MIL-C-15305 or MIL-T-27. The chokes are fully encapsulated; suitable for various soldering techniques. Products are available for the 1 to 500 MHz frequency range with impedances up to 10 kohms. 1000-piece pricing ranges from $3.50 to $4.50 each.
Vanguard Electronics Co.
INFO/CARD #213

EMC/TEMPEST Antennas

Two new models from Antenna Research Associates are designed for shield room testing requirements of MIL-STD 285 and NSA 65-6. Each antenna is a 12-inch tuned loop operating from 1 kHz to 30 MHz. The PLA-130/A is a two-band transmit/receive antenna capable of handling 1 kW input power, and the Model ALA-130/A is a receive antenna containing a low noise preamplifier. Each is individually calibrated.
Antenna Research Assoc., Inc.
INFO/CARD #212

New SMT Filters

Trilithic has introduced two new surface mount filter packages, which extend center frequencies to the 10-2000 MHz range and increase ultimate rejection to 70 dB. Designs available include Chebyshev, Butterworth, Linear Phase, Bessell, Gaussian and Elliptic with −3 dB bandwidths from 2 to 150 percent.
Trilithic, Inc.
INFO/CARD #211

Crystal Products

New products from KS Electronics include a 70.0 MHz VCXO with pullability of over ± 1500 ppm and +10 dBm output. Higher frequencies are also available. Also announced are 70.0 MHz crystal filters with 1.0 MHz bandwidth, with other frequencies available. The company has the capability of manufacturing crystal filters with third order intercept of +30 dBm.
KS Electronics
INFO/CARD #210

Matrix Switch

Watkins-Johnson introduces a series of Matrix Switch Distribution Systems offering high performance and reliability. Technology used varies with the frequency range, and may be reed relay, FET or PIN switching. Designs are available for audio (DC-20 kHz), video (DC-20 MHz), HF (1-30 MHz), VHF (20-500 MHz), V/UHF (20-1000 MHz) and IF (narrow bands 80-200 MHz). Control is via front panel, RS-232

interface, or IEEE-488 bus.
Watkins-Johnson Co.
INFO/CARD #210

+17 dBm Amplifier

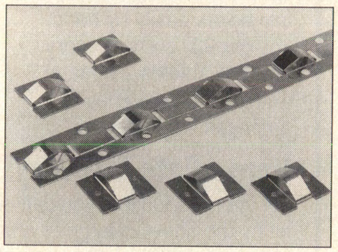
Model TM5150 from Amplifonix offers 19 dB gain, +17 dBm power output at 1 dB compression, and 3.5 dB noise figure over the −55C to +85C temperature range. Package options include 4-pin TO-8, surface mount, and connectorized housings.
Amplifonix
INFO/CARD #209

SPDT Switches

Teledyne Microwave announces type CS-53 coaxial switches with faster switching times; 20 msec for failsafe and 10 msec for latching designs. DC to 6 GHz performance specifications include: 1.25:1 maximum VSWR, 0.4 dB maximum insertion loss and minimum 70 dB isolation. To 18 GHz, these specifications are 1.5:1, 0.5 dB and 60 dB, respectively, and from 18-26.5 GHz, they are 1.8:1, 0.7 dB and 50 dB.
Teledyne Microwave
INFO/CARD #208

Grounding Contacts

The Omni Grounding Series provides a simple means of mounting grounding contacts individually or in series, using pressure sensitive tape or mechanical fasteners. Made from beryl-



lium copper, they compress to 80 percent of their height and allow sliding contact from either end.
Tech-Etch, Inc.
INFO/CARD #207

ESM Antenna

RF Microsystems has added the model 216 antenna to their line, a shaped biconical design with a very broad bandwidth, omnidirectional azimuth coverage and vertical polarization. The antenna operates over 0.5 to 18 GHz with a maximum VSWR of 2.5:1 in a 50 ohm system. Weight is one pound, and the size is approximately 5 inch diameter by 5.25 inch height, plus mounting flange.
RF Microsystems Inc.
INFO/CARD #206

2 GHz Power FETs

ST5000 and ST5001 from Semetex provide 0.5 and 1.0 watts of class A power with 11.5 and 10 dB gain, respectively, at 2 GHz. Applications include broadband multi carrier, low noise, high intercept amplifiers, medium gain power amplifiers, high level VCOs and high performance receiver front ends. 1-99 unit pricing is $45.00 for the ST5000 and $52.00 for the ST5001.
Semetex Corp.
INFO/CARD #205

SMT Connectors

Surface mount connectors in SMA and SMB types are available from Applied Engineering Products. Accurate solder seating with no-hole mounting shortens assembly time. Parts tested withstood straight pull of 24 lb., side loads to 40 lb., and radial torque to 16 in -lb.
Applied Engineering Products
INFO/CARD #204

Switched Filter Modules

Integrated Microwave offers two, three and four channel modules for frequencies up to 3 GHz. GaAs switching offers 150 ns speed and typical switch loss of 4 dB. Highpass, lowpass, bandpass and band-reject responses are available.
Integrated Microwave
INFO/CARD #203

GSM/PCN SAWs

Sawtek has applied low-loss filter techniques for IF filtering requirements of GSM/PCN equip-

ment. Center frequencies between 70 and 250 MHz have been demonstrated, with insertion loss of approx. 6 dB. Group delay variation is an excellent 150 nsec. A simple fabrication technique reduces manufacturing costs.
Sawtek, Inc.
INFO/CARD #202

Multitone Generator

Systematix offers a multitone generator for intermodulation testing, with up to four carriers available, tuned as close as 100 kHz at frequencies of 800-900 MHz or higher. Internal intermod spurs are better than −80 dBc at 45 dBm output. All four tones are independent and front-panel tunable. Modulation and automatic control are optional. Price of the unit is $19,500.
Systematix
INFO/CARD #201

BNC Crimp Connectors

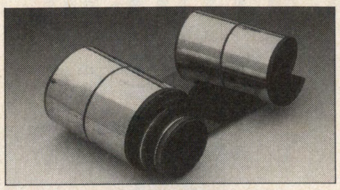
Alpha Wire announces their new BNC Crimp Connector, a nickel plated connector with Delrin dielectric and gold-plated center pin. Rapid assembly time and reliable electrical and mechanical contact are major features. The connector is provided in standard 50 ohm impedance.
Alpha Wire Corp.
INFO/CARD #200

Cell Site Antenna

Model ASP-2892 is a 13 dB gain log periodic array for 60-degree sectorized cell sites. The design covers a bandwidth of 800-960 MHz with only a 15 degree vertical beamwidth. Five degrees of beam tilt is built in, with additional tilt available by adjustment of the mounting clamp. Power rating is 500 W and wind velocity rating is 130 MPH.
The Antenna Specialists Co.
INFO/CARD #199

Solderable Flexible Laminates

Foil-film laminates are offered by Chomerics, to serve as flexible circuits in commercial and military equipment. The laminates can be copper or solder-coated copper, bonded to a variety of dielectric films in continuous rolls



or die-cut pieces. These products are also suitable for grounding and shielding applications.
Chomerics, Inc.
INFO/CARD #198

Package Seam Welder

The Venus II Parallel Seam Sealer from Polaris adjusts to variations in carrier configurations in electronic packages and package placement. Square or rectangular package sup to 6 inches are sealed by resistance welding or solder reflow.
Polaris Electronics Corp.
INFO/CARD #197

Coaxial Resonator Oscillators

T and M Microwave offers the CR 1000 series of coaxial resonator oscillators covering the frequency range of 300 MHz to 4.5 GHz. Nominal power output is +13 dBm, harmonics and spurious are specified at −20 dBc and −80 dBc, respectively, and phase noise at 10 kHz offset is −115 dBc. Size for a typical 840 MHz unit is .475 × 1.0 × 1.5 inches.
T and M Microwave, Inc.
INFO/CARD #196

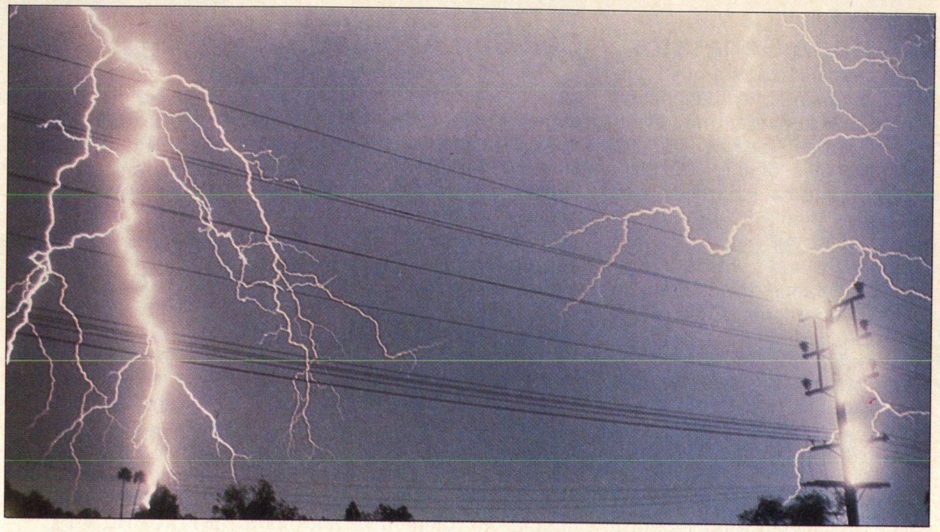
Arbitrary Waveform Generator

The HI-2009A by Hitronetic has been introduced in the U.S. by Kersaint Associates. The unit synthesizes analog signals with a time resolution of 2.2 nsec per point, two megabytes of signal memory, programming flexibility, and a 450 Mpts/sec sample rate. The price starts at $40,000.
Kersaint Associates
INFO/CARD #195

Surge and Transient Considerations

By Gary A. Breed
Editor

Surges and transients on power and signal lines are major EMC problems in electronic equipment of all types. It is common to see these events occur at magnitudes of thousands of volts, with a considerable amount of total energy. The problem is a particular nuisance at radio frequencies, where even small amounts of energy can interfere with the reception of desired signals. Eliminating, or at least limiting the effects of this unwanted energy usually requires a combination of several standard protection components and circuits.



Surges

Surges are by common definition a low frequency phenomenon, whether they occur on power distribution circuits or signal lines. Intuitively, they can be considered as low frequency modulation of the desired electric or electronic energy normally found on the line. Most signal line surges can be traced to lightning, while power lines are also subject to the inductive kick of machinery or sudden changes in load.

Lightning surges (not direct hits) are usually the result of the discharge path being inductively coupled to a power or telecommunications line. Rural locations are particularly susceptible, since there are long, uninterrupted stretches of distribution lines available for coupling. This type of surge can be very substantial, but is largely dissipated by a combination of spark-gap discharge units and metal-oxide varistors (MOVs). However, equipment must be designed to withstand surges that are below the protection threshold of these devices.

Inductive kick and distribution circuit load changes are somewhat more predictable. Unless they are accompanied by switching transients, they tend to be smooth changes in voltage, with a damped swing of voltage on the line. Their effects are almost always limited to power supply variations. Linear power supplies must be designed to have an adequate raw DC voltage and capacitive storage to remain above the regulator's dropout threshold. Switching supplies react much like they would to a change in load. If the supply is near its maximum current output capability, the voltage swing of the surge can modulate the regulated output.

Transients and Noise

Variations with very short time duration or with repetitive spikes are called transients, and they come from a large number of sources. In power distribution, they occur when circuits are switched in and out of the power grid, or when nearby lightning results in energy from close coupling or by direct contact of charged ions to the lines. Physical damage causing arcing or shorts contribute to the problem, as well. On signal lines, the potential effects due to lightning are the same as on power lines. AC line filters will remove most of the RF noise conducted by the lines, but radiation from nearby wiring will still be present, unless shielded by conduit.

AC or DC potentials on signal lines can be another source of trouble. DC and ringing voltages are present on telephone circuits, and DC power is present on CATV distribution coaxial cable, as well the RF signals. If a leakage path is present, or if balanced lines become unbalanced with respect to ground, the resulting currents can cause problems with hum pickup from longitudinal currents coupled from AC power distribution circuits. These currents are normally suppressed by the balanced nature of the signal lines.

Outside sources of trouble are common on signal lines. Failed distribution equipment, faulty customer equipment and physical damage are relatively common. The result may not be restricted to disturbances in the desired signals

Metallic Surge (across wires):	100 A, 800 V, 10 x 560 msec rise/fall
Longitudinal Surge (common mode):	200A, 1500 V, 10 x 160 msec
Surge Protection Devices:	Breakdown voltage greater than 250 V
Leakage Current (tip or ring to ground):	10 mA, 1000 VRMS
On-Hook Impedance (tip or ring to ground):	100 kohms
Signal-to-Noise Ratio (voice grade):	greater than 30 dB, typical
Signal-to-Noise Ratio (digital loops):	depends on data rate
Longitudinal-to-Metallic Balance:	60 dB minimum

Table 1. Some telephone circuit performance specifications.

and voltages. A broken shield on a coaxial line or unbalance due to inadvertent grounding of one side of a telephone line increases susceptibility to RFI from nearby transmitters. Illegal telephone and CATV connections are notorious culprits for increased RFI because they create mismatch, unbalance or compromise shielding integrity.

Noise in the user's equipment is the result of all the above effects, and more. Other sources of impairment on signal lines include crosstalk, distortion and excess attenuation. Line balance, its physical condition, and the status of transmission equipment all affect overall signal transmission performance. Table 1 shows some key performance specifications for telephone line performance, including surge and overvoltage protection standards, as well as signal path performance.

Protection and Prevention

Transients and noise are insidious problems. They often result in intermittent problems that are extremely difficult to diagnose. Methods of dealing with voltage-current surges and transients include filtering techniques to minimize high-frequency components of the disturbances, along with fuses, spark gap, varistor and zener protection against more dramatic spikes.

Standard practices in the construction of these facilities include separation of signal and power distribution conductors whenever possible, metallic sheathing on telephone trunk lines, periodic grounding to minimize the accumulation of a potential on the shield, and static discharge wires on certain power lines. CATV systems, with a completely shielded system, usually include a ground block with a spark-gap arrestor at the entrance to the customer's location. Telephone lines have fuses in both tip and ring circuits, which may be at a distribution pedestal or at the customer service entrance.

Noise, hum, distortion and interference that appear along with the desired signal require different techniques. Most areas of noise performance are dependent on the condition of the transmission equipment and the transmission medium. Designers of user equipment must be certain that they are compatible with the wire or coax system they utilize, and must allow sufficient design margin to be sure that the transmission medium is the weakest link in the system. Minimizing noise and distortion requires control of signal levels to maintain exactly what is specified for the system used.

Above all, familiarity with system requirements, and the nature of surges, transients and noise helps make a good relationship with the power, telephone and CATV companies. The best service always seems to be given to knowledgeable customers. *RF*

A Simple Clock and PSK Carrier Recovery Circuit

By Francois Methot
RadioSysteme, Inc.

Today, analog designers have to interface with the digital world by recovering the clock from their modem (electric or optic) for high speed data transmission.

The interface between analog and digital is critical, especially when it must be done at a low cost. I propose a different approach to obtain PSK demodulator and clock recovery with the same kind of circuit.

The circuit is based on a PLL with a "D" flip-flop as the phase comparator, coupled with a VCXO to avoid locking on undesired frequencies.

The carrier and/or clock is recovered by using the positive edge of the incoming PSK signal or data stream to sample the carrier or clock and create a DC voltage to control the VCXO. Figure 1 depicts the interconnection between the "D" flip-flop and the VCXO.

However, there is a compromise to be made in using this design. The phase detector does not have a linear transfer function versus phase error. The gain of the phase detector ($K\phi$) in frequency is low but $K\phi$ for the phase is very high. Figure 2 shows an example of the transfer function of this design.

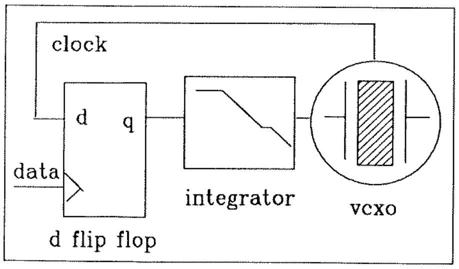


Figure 1. Interconnection between "D" flip-flop and VCXO.

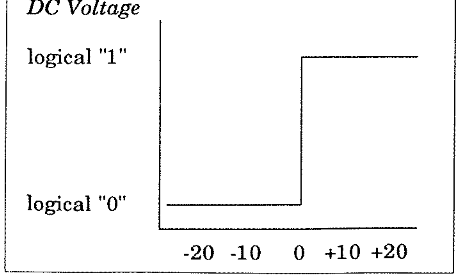


Figure 2. Transfer function from design in Figure 1.

In the absence of data, the output of the "D" flip-flop can be at a logical "0" or "1" depending where the data did the last sampling in the clock, so the VCXO can be at the maximum or the minimum frequency. When the PLL begins the search for the frequency, it requires many samples to converge on the correct frequency, but the search for the phase is very fast.

For a low bit rate, I recommend the use of a CMOS "D" flip-flop because it has the same threshold for a logical "1" or logical "0". Most of the time a passive filter will do, when it is connected on the Q or Q output, depending if it is the positive edge or negative edge that is required at the center of the data bit.

For a high bit rate, an ECL "D" flip-flop must be used because it has only 1 nsec of propagation delay. However, an active filter (integrator) must be implemented to increase the range on the diode varicap in the VCXO circuit.

This circuit can be implemented to demodulate a PSK signal with some compromise. The carrier must be an exact multiple of the bit rate. Moreover, the phase transition of the carrier must be synchronized with the data. These requirements are cost efficient and easily implemented on the demodulator.

When feasible or when given the choice of IF frequency, the carrier to be modulated must be an exact multiple of the data clock to be transmitted. A simple PLL can multiply the clock synchronously and form the carrier to be modulated. Figure 3 shows an example of a simple PSK modulator. Now the PSK carrier can be seen as a data stream of "1", "0", "1" ..., with one extra "0" or one extra "1" at each transition of the data. Refer to Figure 4 for the timing diagram.

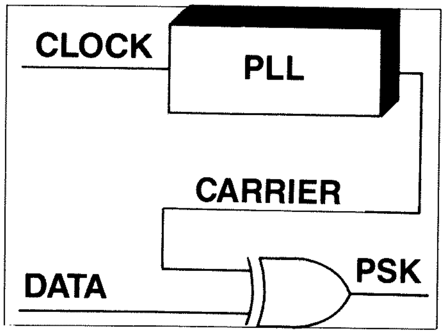


Figure 3. A simple PSK modulator.

A carrier can be regenerated at twice the PSK carrier frequency with the same circuit for the clock recovery and later, by dividing the carrier by two. A simple exclusive "or" can demodulate the PSK signal. Because of the use of the divider, there exists a phase ambiguity of 180 degrees. By using a differential code or NRZI encoding, the demodulated data will always be the same polarity.

In Figure 5, one ECL IC is used for limiting the PSK signal, another IC has the phase comparator and the frequency divider, and a final IC to demodulate the PSK signal and produce the VCXO. An NRZI to NRZ must be added after the low pass filter. This technique, compared to the conventional approach, requires no treatment on the signal (X^2) to rebuild the carrier information.

The component values of the integrator were calculated using Communica-

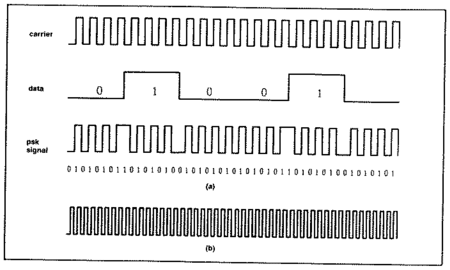


Figure 4. Timing diagram (a) the PSK signal can be seen as a data stream of 1, 0, 1, ... with one extra 0 or 1 at each transition of the data, (b) a carrier can be regenerated at twice the PSK carrier with a simple clock recovery and when it is divided by two, can demodulate the PSK signal.

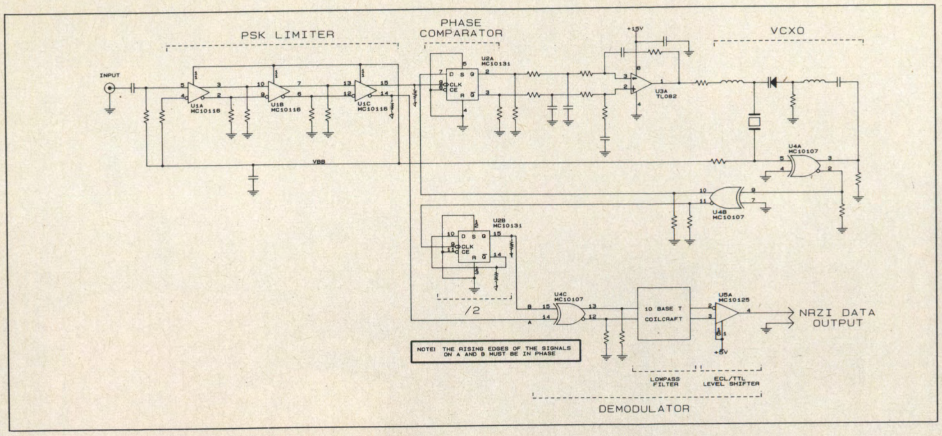


Figure 5. PSK demodulator circuit.

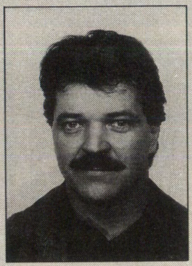
tions Consultion Corporation's (now Compact Software) PLL Design Kit.

Figure 6 shows my RF Design Contest entry: clock recovery based on "D" flip-flop phase detector. Figures 7 and 8 show the performance of the circuit's clock and data recovery.

Conclusion

The circuit shown in Figure 6 uses only one IC to both perform the phase comparator and sample the data. The VCXO was made with a crystal in fundamental at 139.264 MHz to eliminate frequency multiplication and achieve a larger range. A Δ F of 50 kHz was obtained with ±15 kHz of capture range. The data stream comes from the optical detector from a source at −25 dBm. The clock recover still locks up to −31 dBm even if the data is at 10^{-2} of error rate. For the PLL, the parameters are: W_n = 10 kHz, K_{vco} = 1 kHz/volt, K_o = 10 volts/radian, ζ = 0.7. *RF*

About the Author



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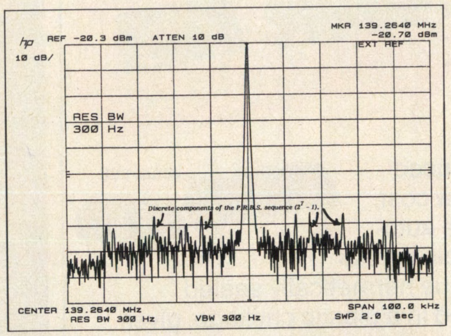


Figure 7. Spectrum analyzer display of the recovered clock signal.

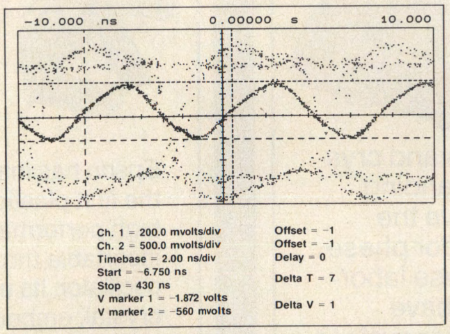


Figure 8. Digital oscilloscope display of clock and recovered data.

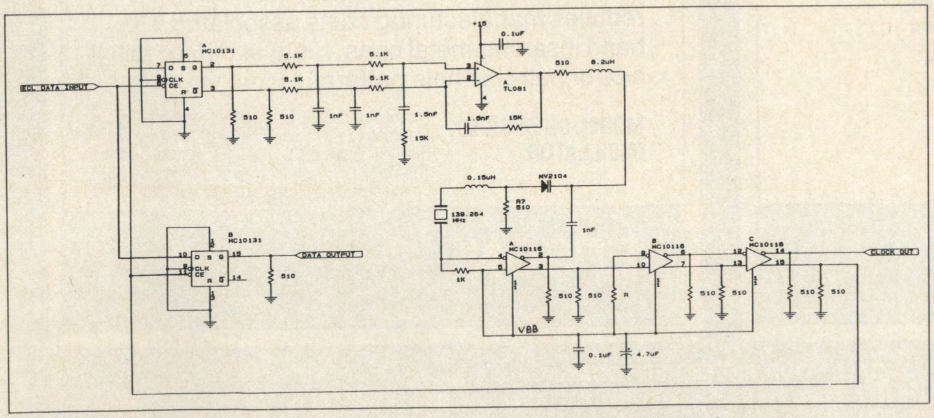


Figure 6. Clock recovery based on "D" flip-flop phase detector.

VHF and UHF Crystal Oscillators — Part II

By Andrzej B. Przedpelski
A.R.F. Products, Inc.

This note contains some theory, more circuits, some suggestions and further considerations of 5th, 7th, and 9th overtone crystal oscillators (up to 630 MHz) using simple easy to build circuits.

I should have started Part I, July of 1990 (1), with some applicable theory, but I got carried away by the ease of constructing and adjusting these high frequency crystal oscillators and their reliable performance. To rectify this omission let's consider the requirements for an oscillator. Barkhausen tells us that for a circuit to oscillate, two conditions have to exist.

- the open loop gain (A( ) x ( ) of Figure 2) has to be more than unity, and
- the phase of this gain has to be 360n, where n is an integer (including 0).

In the circuits described, the gain, A( ) is supplied by the MMIC amplifier and the feedback, ( ) consists of the crystal, in series with a length of transmission line, and the capacitor(s).
Using an ideal non-inverting amplifier (such as Figure 1a) (1), the feedback can have 0 degrees phase shift and no phase shifting network is thus necessary. The crystal can, theoretically, operate in series resonance (presenting only a resistance in the feedback path) and becomes an attenuator in conjunction with the 50 ohm input impedance of the MMIC amplifier. If this attenuation is lower than the gain of the amplifier, the circuit will oscillate. The oscillations start slowly (relatively speaking) by amplifying circuit noise at the resonant frequency until the amplifier saturates (gain of the amplifier becomes equal to the feedback attenuation).
In the inverting amplifier configuration (such as Figure 1b) (1), the feedback has to provide the (theoretically) required 180 degree phase shift to allow oscillations to take place.
In practice, of course, the non-inverting amplifier has some phase shift at the operating frequency, and the inverting amplifier phase shift is usually somewhat less than 180 degrees. Thus, the feedback circuit has to compensate for these small phase shift discrepancies.
The inverting amplifier oscillator configurations are generally preferred, since they have less tendency to oscillate at undesired frequencies. Any "simple" undesired feedback usually does not have the required 180 degree phase shift.
To provide the needed feedback phase shift, different methods can be used. The crystal itself can provide some phase shift. However, it is better to use it in a series resonance mode to reduce losses through the crystal. Thus, a phase shifting network providing almost 180 degree phase shift has to be used. At lower frequencies, a pi or tee phase shifter can be used. However, at the higher frequencies, accurate inductors are difficult to construct, and spurious reactances cause additional uncontrolled phase shifts. To overcome this, and to provide very simple implementation of reliable circuits, a transmission line can be used to provide the main required feedback phase shift. This approach is feasible at the higher frequencies and provides easily reproducible phase shifts. To demonstrate this, the circuit shown in Figure 3 was built. While feasible and very simple it was not considered very practical since the length of the transmission line was difficult to optimize. A more feasible method using variable capacitors to fine-adjust the feedback phase shift was used in the more practical circuits. This method had the additional advantage of allowing some "pulling" of the oscillator frequency.
Now back to some more circuits

Circuit # 1 (350 MHz fifth overtone)

The circuits presented in Part I used hybrid type RF amplifiers. To reduce size still more and allow the use of surface mounted components, monolithic RF amplifiers were then tried. A typical inverting type was used (Avantek MSA-0735). Two possible circuit configurations were tried (Figures 4 and 5). Both gave an output of 6-8 dBm with either a 3.6 inch or 6 inch length of transmission line. The capacitors had to be retuned, of course. The circuit of Figure 5 seemed to have lower harmonic content (35-40 dB below fundamental for second and third harmonics). While monolithic amplifiers usually have the disadvantage of requiring input and output capacitors, one capacitor can be omitted, since the crystal provides DC isolation. These amplifiers also require a biasing resistor. In the case of the circuits shown, this was actually taken advantage of to provide a circuit which worked reliably directly from an airborne power source of 22-30 VDC. This was accomplished by the use of the high value (1000 ohm) biasing resistor.

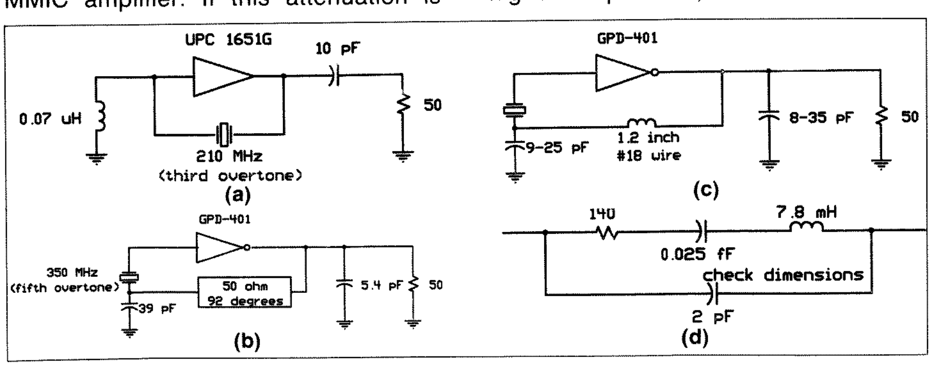


Figure 1. Figures from Part I: (a) third overtone circuit, (b) 350 MHz fifth overtone, (c) overtone oscillator with L/C phase shift, (d) approximate crystal equivalent circuit at fifth overtone.

Circuit # 2 (490 MHz seventh overtone)

The basic circuit of Figure 5 could be easily modified to provide reliable operation on the seventh overtone of the crystal. The same crystal was used with an 8.2 or 7.2 inch length of transmission line, as shown in Figure 6. An output of 6-8 dBm was obtained, but the harmonic content was high (−10 to −15 dB). Otherwise, the circuit performed reliably.

Comments

Output was always observed on a spectrum analyzer to ensure that the proper overtone operation was obtained. Power output was measured at the desired output frequency.

All circuits were checked to ensure that no mode switching could be induced. DC power was turned on instantaneously or was brought up gradually while output was observed on the spectrum analyzer. The oscillators always started at the desired frequency.

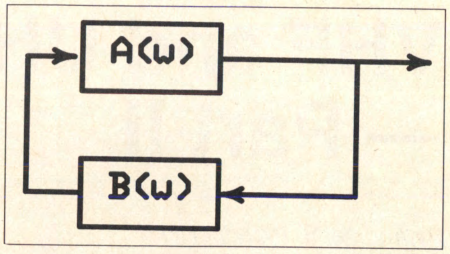


Figure 2. Oscillator equivalent circuit.

50 ohm coaxial transmission lines were used, since they are readily available and easily changed. Once the circuit is finalized, the transmission line can be part of the PC board. There is no apparent reason why other impedances cannot be used. However, only 50 ohm coax cables were used.

An amplifier gain of about 12 dB at the desired frequency seems adequate. Higher gain amplifiers would allow a higher loss in the feedback circuit. This could reduce the crystal drive and improve stability. However, too much gain may make the circuit more susceptible to oscillations at undesired frequencies.

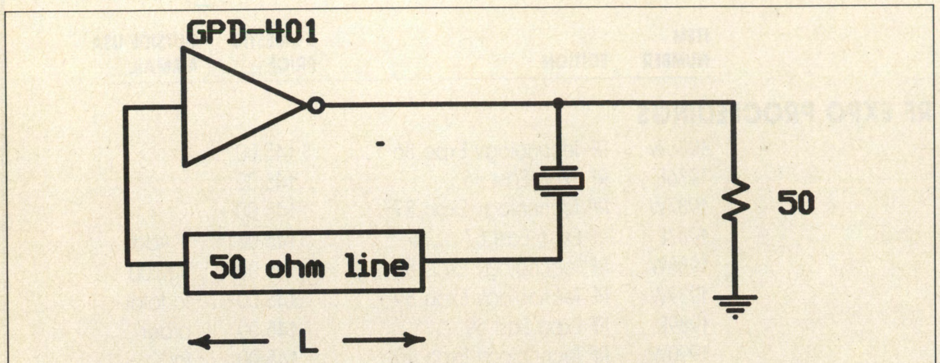
The phase shift in the amplifier changes with frequency. The typical inverting amplifier actually becomes non-inverting at higher frequencies. This occurs at about 4.5 GHz for the MSA-0735 used in the tests. The amplifier still has gain at this frequency, however, and this should be considered.

The crystal resistance does not seem to be critical. The crystal used had a resistance of about 140 ohms at the fifth harmonic (Figure 10). Higher resistance crystals can be used, if the amplifier gain is adequate.

The feedback path loss has to be less than the amplifier gain to sustain oscillations. The input impedance to the feedback path should be higher than 50 ohms to allow most of the amplifier power output to appear as useful power in the load. This is not unreasonable, since the feedback path can have appreciable insertion loss.

A quick temperature check was made using the circuit of Figure 5. The frequency vs. temperature curve was smooth and no mode jumping was observed.

The circuits were very reproducible.



Line Length L, inches	Frequency MHz	Overtone	Power Output dBm	Notes
7.2''	350	5	—	some spuriouses
5.9''	490	7	4	clean
3.0''	630	9	6	clean
0.5''	770	11	5	unreliable

Figure 3. The simplest circuit.

The same circuit was built using direct interconnections (no chassis), using a ground plane or a PC board with all surface mount components. The performance was essentially the same.

Increasing feedback loss (reducing drive to the amplifier) reduces harmonic content (as expected). This reduces power output somewhat. For a power output reduction of 1-2 dB, the harmonic content can be reduced by 15-25 dB.

The initial tuning procedure is quite simple. Using a spectrum analyzer (to make sure that the right overtone operation is achieved) set one capacitor and tune the other (if two are used). If proper operation is not obtained, retune the first capacitor and repeat tuning with the other. Once the proper frequency is obtained, the output can be maximized using both capacitors.

The crystal can oscillate at the fundamental or an odd overtone. Circuit #10 (Figure 1c) (1), for instance could be made to oscillate at the third, fifth and seventh overtone by retuning the two capacitors. At these different frequencies, the phase shift through the amplifier is different. While at very low frequencies it is the theoretical 180 degrees, it decreases with frequency. The crystal also looks different at these frequencies. Mr. David Smith of Innovative Frequency Control Products, Inc. kindly supplied typical equivalent circuit values for the crystal used (see Figure 1d) (1). Series R is about the same for the overtones, but is about 1/6 this value for the fundamental operation. The motional inductance (series L) is about the same for the fundamental and overtones. The motional capacitance varies to provide the proper series resonance.

While operation only up to the 9th overtone (630 MHz) was confirmed, there were indications that the 11th overtone was also feasible. The transmission line length became very short and was difficult to optimize. However sporadic operation at 730 MHz was obtained using the Figure 3 circuit and a transmission length of about 0.5 inch. As results of Figure 3 indicate, there is some spurious phase shift caused by

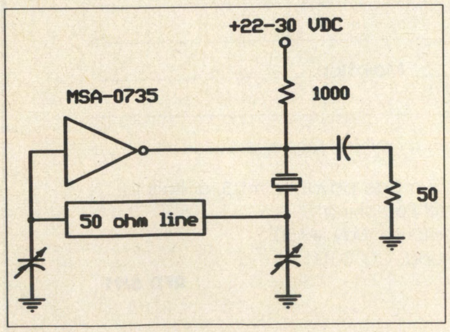


Figure 4. 350 MHz fifth overtone oscillator.

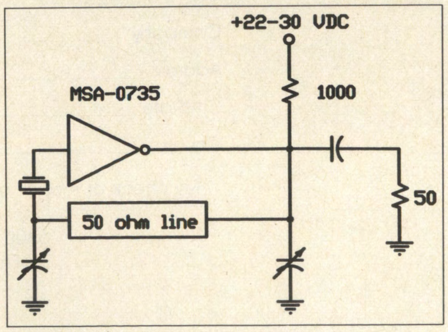


Figure 5. 350 MHz fifth overtone oscillator.

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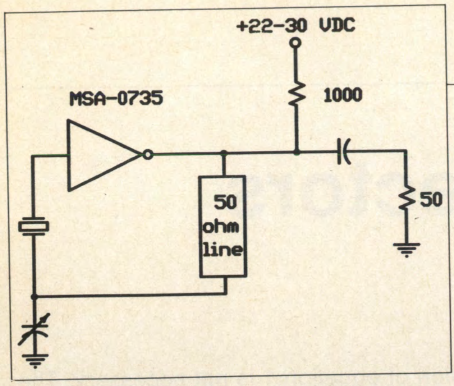

Figure 6. 490 MHz seventh overtone oscillator.

the stray inductances and capacitances in the circuit. With the breadboard used it was also somewhat difficult to accommodate the very short transmission line lengths. A better layout may make the 11th overtone operation quite feasible.

Another problem at the higher overtones is the stray pin-to-pin crystal capacity. While small (about 2 pF) its reactance at the higher frequencies becomes comparable or smaller than the crystal resistance. It may then provide the necessary feedback path to sustain oscillation at the wrong frequency. No effort was made to "tune-out" this reactance, as is sometimes done at lower frequencies to obtain true series resonant operation and to reduce the risk of spurious oscillations.

While the above treatment of UHF crystal oscillators is by no means complete, the main effort was directed toward the simplest low cost reproducible designs using non-critical components. It was also meant as a starting point for further investigations using this general overall approach. The non-inverting amplifier circuits were also not adequately explored. One reason was that only one device was available at the time of the test; several manufacturers make suitable low cost inverting types.

RF

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1. Andrej B. Przedpelski, "VHF and UHF Crystal Oscillators," *RF Design*, July 1990, pp. 63-65.

About the Author

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Design of Quadrature Detectors

by Edward A. Richley
Xerox Palo Alto Research Center

The quadrature detector is a very common circuit used for the demodulation of both wide band and narrowband FM. However, there is a general misconception about its method of operation, and little information on its proper design. After careful analysis, it can be seen that the engineering trade-offs involved in the design of a quadrature network are easy to understand and very interesting. This article will describe such an analysis, and will show the important considerations for proper design.

The basic single-tuned quadrature detector network is shown in Figure 1. This network is typically used with FM receiver ICs such as the Motorola MC3357 (and related parts), Signetics NE604, or the RCA CA3189. The network has one input and one output. The input generally comes from the output of a limiting IF strip. This input, V_{IN} and the output of the network, V_{OUT}, are connected to the inputs of an analog multiplier.

The basic purpose of the quadrature network is to produce two signals which have a nominal phase difference of 90 degrees (hence the term "quadrature detector"), but with a variation in phase which depends on instantaneous frequency. The multiplier then serves as a phase detector, and can detect the lead or lag in phase caused by the interaction of the frequency modulated signal with the network.

In some versions of the circuit, the coupling capacitor, C_1 of Figure 1, is replaced by an inductor. The operation of the circuit is conceptually the same in both cases. Many texts give an explanation for the operation of this circuit which is based on the following assumption (1,2):

$$X_1(\omega_0) >> R_L \tag{1}$$

where X_1 is the magnitude of the reactance of the coupling component, either a capacitor or an inductor, and ω_0 is the center frequency of the frequency modulated signal. ω_0 is also the resonant frequency of the network formed by L_L and C_L. With the assumption of equation 1, the nominal 90 degree phase shift is provided by the reactance of X_1 in series with a much smaller resistance, R_L. The subsequent phase lag or lead about the nominal will result from changes in the received signal frequency which cause the reactance of the R-L-C circuit to swing inductive or capacitive.

A quick look at some application notes, however, shows that the condition of equation 1 is often not met in practice. In fact, it is often strongly violated (3,4). Thus, the question arises as to how the nominal 90 degree phase shift is provided, and how the phase shift varies with frequency. In order to address this point of confusion, the network of Figure 1 must be examined more closely.

It is instructive to look at the network of Figure 1 in an alternate form. The dashed portion of Figure 2 can be transformed into a Norton equivalent, as shown in Figure 3. The resultant parallel capacitors combine as a single element of value $(C_1 + C_L)$. From Figure 3, it is evident that the nominal 90 degree phase shift can be guaranteed at the frequency ω_0 by simply making the combination of L_L and $C_L + C_1$ resonant at ω_0. Thus, the condition of equation 1 is not necessary. The size of C_1 relative to C_L does not affect the phase characteristics of the circuit at all, since the two capacitors are effectively in parallel. Their relative sizes only affect the amplitude of the output, and, as shall be seen later, the input impedance of the network. Thus, the condition of equation 1 is not necessary at all.

The Transfer Function

The circuit of Figure 1 has a transfer function, $H(j\omega)$, defined as the ratio of the output voltage to the input voltage, when both are expressed in phasor notation:

$$H(j\omega) = \frac{V_{OUT}(j\omega)}{V_{IN}(j\omega)} \tag{2}$$

From the equivalent circuit of Figure 3, it is easy to show that $H(j\omega)$ is determined by the impedance of the R-L-C circuit, and is given by:

$$H(j\omega) = -\frac{\omega^2 C_1 L_L R_L}{R_L(1 - \omega^2 L_L(C_1 + C_L)) + j\omega L_L} \tag{3}$$

There are four main concerns regarding the design of the quadrature network. These are as follows:

1. $|H(j\omega_0)|$ should be as large as possible.
2. The variation in $|H(j\omega)|$ over the desired range of frequency deviation should be as low as possible.
3. The phase variation in $H(j\omega)$ over the desired range of frequency deviation should be as linear as possible.
4. The input impedance of the network must be sufficient so as to not load the driving circuit.

In order to simplify the investigation of each of these requirements, it is useful to define some quantities. The first of these is the loaded Q of the network:

$$Q = \frac{R_L}{\omega_0 L_L} \tag{4}$$

In this analysis, any load resistance due to the phase detector input is lumped into the resistance R_L.

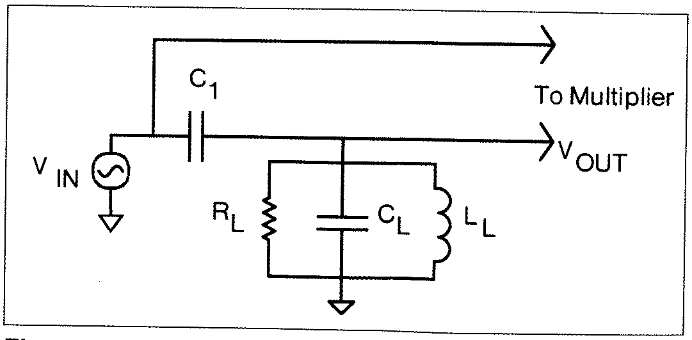

Figure 1. Basic single tuned quadrature detector network.

Another useful quantity is the center frequency, ω_0. As described earlier, component values are chosen for resonance at ω_0, so that:

$$\omega_0^2 = \frac{1}{L_L(C_1 + C_L)} \tag{5}$$

Finally, the frequency, ω, can be normalized to ω_0 by defining a new quantity, γ, as follows:

$$\gamma = \frac{\omega}{\omega_0} \tag{6}$$

In terms of these quantities, the transfer function is given by:

$$H(j\omega) = -\frac{C_1}{C_1 + C_L} \frac{\gamma^2 Q}{(1-\gamma^2)Q + j\gamma} \tag{7}$$

From equation 7 it follows that the magnitude of the transfer function at the center frequency ($\gamma = 1$) is given by:

$$|H(j\omega_0)| = \frac{C_1 Q}{C_1 + C_L} \tag{8}$$

From equation 8 it is clear that for greatest magnitude of $H(j\omega_0)$, both Q and C_1 should be as large as possible. However, once C_1 is made comparable to C_L there is not much to be gained from increasing it further.

It is also useful to define the normalized frequency, γ, in terms of the deviation, ξ:

$$\gamma = 1 + \xi \tag{9}$$

In terms of ξ, the transfer function is :

$$H(j\omega) = -\frac{C_1}{C_1 + C_L} Q \frac{(1+\xi)^2}{(-\xi(2+\xi))Q + j(1+\xi)} \tag{10}$$

The Amplitude Response

In order to examine the flatness of the amplitude response with frequency, it is helpful to look at the magnitude squared of the transfer function:

$$|H(j\omega)|^2 = \left(\frac{C_1}{C_1 + C_L}\right)^2 Q^2 \frac{(1+\xi)^4}{Q^2\xi^2(2+\xi)^2 + (1+\xi)^2} \tag{11}$$

This expression can be expanded in terms of ξ about the center frequency, $\xi = 0$:

$$|H(j\omega)|^2 = \left(\frac{C_1}{C_1 + C_L}\right)^2 Q^2(1 + 2\xi + (1 - 4Q^2)\xi^2 + \ldots) \tag{12}$$

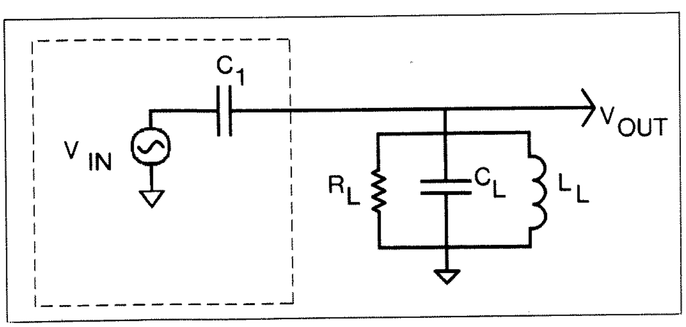

Figure 2. The quadrature network redrawn.

From equation 12 it is clear that in order to prevent AM distortion from becoming induced on the detector, it is necessary that:

$$|\xi| << \frac{1}{2} \tag{13}$$

and that:

$$4Q^2|\xi|^2 << 1 \tag{14}$$

In general, $|\gamma|$ is quite small (0.01 is typical) so that both of these requirements will be met for modest values of Q.

The Phase Response

The phase response can be examined in a similar manner. Equation 10 leads to the following expression for ϕ, the phase of the transfer function, $H(j\omega)$:

$$\phi = \arctan\left(\frac{(1+\xi)}{\xi(2+\xi)Q}\right) \tag{15}$$

Since the nominal value of ϕ is known to be 90 degrees ($\pi/2$), it is most useful to look at the deviation in ϕ about this mean. This deviation, θ, is given by:

$$\theta = \phi - \frac{\pi}{2} = -\arctan\left(\frac{\xi(2+\xi)Q}{(1+\xi)}\right) \tag{16}$$

The linearity of the phase response of the network can be evaluated by looking at the quantity $\partial\theta/\partial\xi$. From equation 16, it follows that:

$$\frac{\partial\theta}{\partial\xi} = -Q \frac{2 + 2\xi + \xi^2}{1 + 2\xi + (4Q^2+1)\xi^2 + 4Q^2\xi^3 + Q^2\xi^4} \tag{17}$$

This expression can be expanded in terms of ξ to give:

$$\frac{\partial\theta}{\partial\xi} = -2Q(1 - \xi + \frac{1}{2}(3 - 8Q^2)\xi^2 + \ldots) \tag{18}$$

In order to maintain a linear phase response, it is important for the constant term in equation 18 to be the dominant term. Thus, the requirement for phase linearity can be expressed as follows:

$$|\xi| << 1 \tag{19}$$

and

$$4Q^2|\xi|^2 << 1 \tag{20}$$

These requirements can be seen to be almost identical to the requirements of equations 13 and 14.

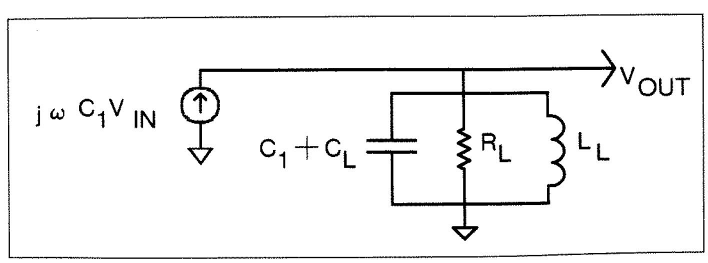

Figure 3. Norton equivalent of the quadrature network.

The Input Impedance

The remaining requirement has to do with the input impedance of the network. The magnitude of the impedance of the network of Figure 1, at the center frequency, ω_0, as seen by V_{IN}, is:

$$|Z_{IN}(\omega_0)| = \omega_0 L_L \frac{\sqrt{Q^2 + \left(\frac{C_1 + C_L}{C_1}\right)^2}}{1 + Q^2\left(\frac{C_1}{C_1 + C_L}\right)^2} \tag{21}$$

For even modest values of Q, and for values of C_1 which approach C_L (which is encouraged from equation 8), $|Z_{IN}(\omega_0)|$ approaches the following form:

$$|Z_{IN}(\omega_0)| \rightarrow \frac{\omega_0 L_L}{Q} \tag{22}$$

Clearly, a large enough value of Q and a small enough value of L_L could lead to an input impedance which is too low to be driven from the IF output. This consideration places a limit on the size of Q or the size of C_1 relative to C_L.

Design Procedure

Consideration of all of these constraints leads to a reasonable approach to the design of these single tuned quadrature networks. As with most engineering problems, there are trade-offs to be made. In general, the following procedure should suffice:

1. Choose frequency deviation or center frequency such that $|\xi| << 1/2$.
2. Choose Q as large as possible, without violating the requirement: $4Q^2 |\xi|^2 << 1$.
3. Choose the largest value for L_L which can be obtained within the limits of physical size, self-resonance, and the effects of its unloaded quality factor on Q.
4. If $(\omega_0 L_L/Q)$ is an acceptable value for input impedance, choose a value for C_1 which about is half the capacitance needed for resonance with L_L at ω_0. Let C_L be the remaining capacitance needed for resonance.
5. If $(\omega_0 L_L/Q)$ is not acceptable, choose a value for C_1 which provides an acceptable value for $|Z_{IN}(\omega_0)|$, from equation 21. Choose a value for C_L which completes the requirement for resonance with L_L at ω_0 (equation 5).

This procedure should work well for most designs. It will provide the largest possible output level without sacrificing other parameters. Equations 8, 9, 12, 18, and 21 provide some quantitative insight into the behavior of signal tuned quadrature detectors. Although this analysis has been carried out for capacitive coupling, it must be emphasized that a similar exercise could be performed for a circuit using an inductor in place of C_1.

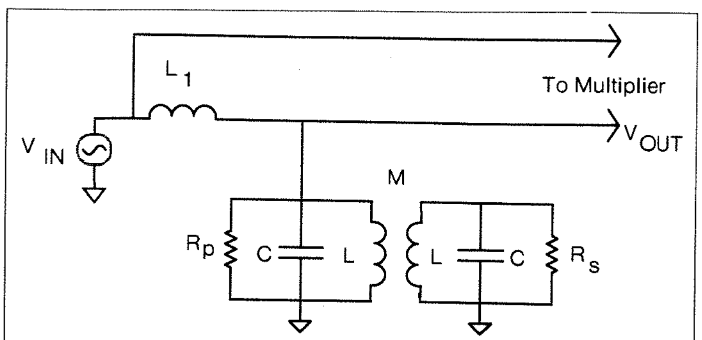

Figure 4. The double tuned quadrature network.

The Double Tuned Quadrature Detector

There is a higher performance version of the quadrature network which is often described in application notes (4,5). This is the double-tuned network which is shown in Figure 4. Much better linearity is claimed for this network.

This network uses a weakly coupled transformer having a tuned primary and a tuned secondary. In most implementations, both windings are identical. In Figure 4, both windings are shown to have an inductance, L. The amount of mutual inductance, M, is always a fraction of L. Thus:

$$M = kL, \qquad k < 1 \tag{23}$$

where k is known as the coupling coefficient.

Instructions in various application notes indicate that the coupling coefficient, k, should be at least 70 percent of critical coupling, k_{cr}, where k_{cr} is defined in terms of the primary Q, Q_p, and the secondary Q, Q_s, as follows:

$$k_{cr} = \frac{1}{\sqrt{Q_p Q_s}} \tag{24}$$

Instructions also indicate that the primary is to be tuned for maximum signal, while the secondary is to be tuned for maximum linearity.

An analysis of the double tuned network will now be presented. This analysis will focus on the phase response of the network, as that is the important advantage of a double tuned network when compared with the single tuned network. An analysis of the amplitude response and the input impedance can be derived easily from this phase response analysis, and so will not be presented.

Some Approximations

The double tuned network is significantly more complicated and difficult to analyze than the single tuned network. Consequently, some assumptions and approximations will be used. These approximations are extremely plausible and serve to greatly enhance the understanding of the analysis. In fact, the analysis of the single tuned network would be greatly simplified by these approximations and would yield exactly the same result as equation 20.

The first assumption is simply that the maximum frequency deviation is very small. Thus, it is assumed that:

$$|\xi| << 1 \tag{25}$$

In comparison with terms of order 1, ξ can be ignored. Of course, one must be very careful to not ignore ξ when subtraction of large numbers will result in a term of the order ξ.

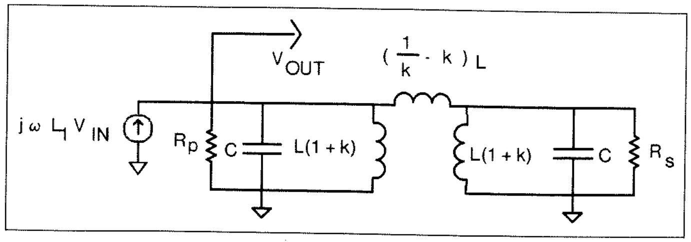

Figure 5. Equivalent of the double tuned quadrature network.

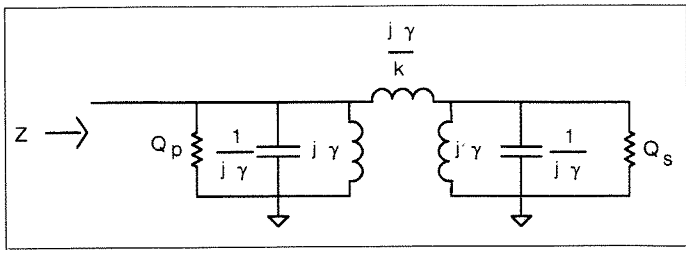

Figure 6. Approximation of the double tuned quadrature network.

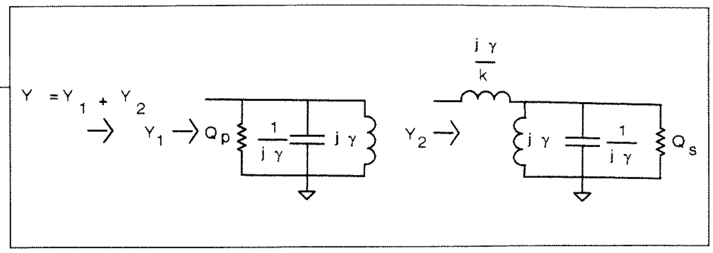

Figure 7. Simplified double tuned quadrature network.

Furthermore, it is assumed that the Q of both primary and secondary are large:

$$Q_s >> 1, \qquad Q_p >> 1 \tag{26}$$

From these assumptions, and the definition of k_{cr}, it is clear that:

$$k << 1 \tag{27}$$

Equivalent Circuit

An equivalent circuit can be made for the double tuned network of Figure 4. This is shown in Figure 5. This equivalent will yield exactly the same output as the original circuit. The transfer function is determined once again by the input impedance of this network.

Only an analysis of the phase response of the double tuned network will be given here. Clearly the variation in the phase of the transfer function, about the 90 degree point, is given by the phase of the input impedance of the network. Just as with the single tuned network, the coupling element provides the initial 90 degrees of phase shift while the tuned network provides the deviation. Thus, the phase of interest is simply the phase of the input impedance.

Figure 6 shows this input impedance as Z, with the impedance of all elements normalized to $\omega_0 L$. Simplification is provided by the use of equation 27. This simplified approximate network clearly consists of two networks in parallel. Figure 7 shows these two networks distinctly. Clearly, the admittance of the combined network Y, is simply the sum of the two admittances, Y_1 and Y_2. The admittance, Y, is just as useful for determining the phase response of the network as the impedance, Z, because Y and Z are reciprocals of each other. Thus, the variation component of phase response is simply the negative of the phase of Y.

Using equation 25, it is easy to show that, to a very good approximation sufficiently close to the center frequency, the admittance, Y_1, is given by:

$$Y_1 = \frac{1}{Q_p} + 2j\xi \tag{28}$$

Similarly, the admittance, Y_2, can be well approximated by:

$$Y_2 = k\,\frac{1 + 2jQ_s\xi}{Q_s(k - 2\xi) + j} \tag{29}$$

These forms for the admittances, Y_1 and Y_2, assume that the tuned circuits of Figure 7 are tuned for resonance at $\xi = 0$. This is not necessarily the case. In fact, some degree of mistuning will yield the desired results. This mistuning can be represented by replacing ξ with $(\xi - \xi_1)$ in equation 28 and $(\xi - \xi_2)$ in equation 29. Thus, the total admittance, Y, is given by:

$$Y = \frac{1}{Q_p} + 2j(\xi - \xi_1) + k\,\frac{1 + 2jQ_s(\xi - \xi_2)}{Q_s(k - 2(\xi - \xi_2)) + j} \tag{30}$$

This detuning accounts for several effects. Slight differences in winding inductance, the loading effects of the coupling element, L_1, and the approximation made by replacing $L(1 + k)$ if Figure 5 with L in Figure 6 are a few of these effects. In general, the quantities ξ_1 and ξ_2 allow two more degrees of freedom for optimizing the transfer function.

The Phase Response

The trouble with the single tuned network is that once it has been tuned for nominal 90 degree phase shift, the transfer function is completely determined. Basically, the phase shift in a single tuned network is determined to be $\arctan(2Q\xi)$. There is no such limitation in the double tuned network. If the argument of the arctan( ) function could be made to more closely resemble the tan( ) function, a more linear phase relationship would be possible. This is exactly what the double tuned circuit allows. In order to see this more clearly, it is important to look at the Taylor expansion of the function tan(ax):

$$\tan(ax) = ax + \frac{a^3x^3}{3} + \frac{2a^5x^5}{15} + \ldots \tag{31}$$

where a is some constant. For the purpose of this analysis, it is clearly desirable to have:

$$\frac{\mathrm{Im}(Y)}{\mathrm{Re}(Y)} = \tan(a\xi) \tag{32}$$

In practice, the best that can be obtained is an approximation to equation 32 for some range of ξ. If this relation can be approximated, the arctan function performed by the phase detector would more closely reflect a linear relation with input frequency. Several requirements on Y are immediately clear:

Im(Y) must vanish as ξ goes to zero. The ratio Im(Y)/Re(Y) must have no even order terms. The coefficients of $(a\xi)$ should match those in the expansion of equation 31 through as many terms as possible.

The first two of the above requirements can be met by proper choice of ξ_1 and ξ_2. Although not easy to derive, it is easy to verify that the choice of $\xi_1 = \xi_2 = -k/2$ satisfies both. Inserting this condition into equation 30 leads to:

$$\frac{\mathrm{Im}(Y)}{\mathrm{Re}(Y)} = 2\xi Q_p \left(\frac{(1 - k^2Q_s^2)}{(1 + k^2Q_sQ_p)}\right)\left(\frac{\left(1 + \dfrac{4Q_s^2}{1 - k^2Q_s^2}\,\xi^2\right)}{\left(1 + \dfrac{4Q_s^2}{1 + k^2Q_sQ_p}\,\xi^2\right)}\right) \tag{33}$$

This expression clearly has no terms in even powers of ξ. It also goes to zero as $\xi \to 0$. The polynomial quotient can be expanded to give the complete power series:

$$\frac{\mathrm{Im}(Y)}{\mathrm{Re}(Y)} = 2\xi Q_p\,\frac{(1 - k^2Q_s^2)}{(1 + k^2Q_sQ_p)}\left[1 + 4Q_s^2\,\frac{k^2Q_s(Q_s + Q_p)}{1 + k^2Q_s(Q_p - Q_s) - k^4Q_s^3Q_p}\,\xi^2 + \ldots\right] \tag{34}$$

The coefficient of the third order term in the Taylor expansion of tan(ax) can be matched by making:

$$\frac{1}{3}\,4Q_p^2\,\frac{(1 - k^2Q_s^2)^2}{(1 + k^2Q_sQ_p)^2} = 4Q_s^2\,\frac{k^2Q_s(Q_s + Q_p)}{1 + k^2Q_s(Q_p - Q_s) - k^4Q_s^3Q_p} \tag{35}$$

This defines a relation between k, Q_s, and Q_p. For example, if $Q_s = Q_p = Q$, the requirement of equation 35 can be found to be simply that $kQ = 0.327$. For $Q_s \neq Q_p$, the opportunity to make more terms in the expansion coincide, at least approximately, is available.

The data sheet for the RCA CA3189 (4) indicates that a preferred circuit uses $Q_s = 20$, $Q_p = 55$, and k is at least 70 percent of k_{cr}. Application of equation 35 leads to a value of 0.0229 for k. This is roughly 76 percent of k_{cr}, precisely as indicated by the data sheet. Although this choice does not give a good match for the term in ξ^5, it gives a much better coefficient for that term than could be obtained from the choice of $Q_s = Q_p = Q$.

In general, for quadrature network design, the phase response, amplitude response, and input impedance are the important considerations. Of these quantities, the one most often overlooked is input impedance. It is clear from this analysis that as the output impedance of the limiting IF is lowered, its drive capability is increased. With proper design, a greater output signal from the quadrature network can be provided. Finally, the double tuned network can be used to greatly improve the linearity of the demodulator. Proper design of the double tuned network can be a bit tedious, as several terms in the power expansion of equation 34 should be carefully evaluated. However, it is fairly easy to get a good match through the third order term.

Although the quadrature detector is a very popular device for the demodulation of FM, the literature contains very little information on its operation or design. This analysis provides the fundamental relations needed for the proper design of quadrature networks. The relations derived here provide the RF engineer with a means for optimizing the design of quadrature detectors. *RF*

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CAD of Single and Tandem Lange Couplers

Timothy O'Connell and Patrick Murphy
University College Cork, Ireland

In recent times the 90 degree hybrid has become one of the most important passive elements to be used in microwave integrated circuits (MICs), because of its important function in balanced mixers, balanced amplifiers, and other circuits. Although there are many types of hybrids available that provide equal power division with 90 degree phase shift, the Lange coupler has been the most widely used because of its superior broadband performance. This article describes a CAD program which calculates the design parameters of Lange couplers. Since the main disadvantage of Lange couplers is their difficulty in fabrication due to the very small dimensions of the interdigitated fingers, the concept of combining couplers in tandem is presented, and this option is incorporated into the CAD program.

Since its introduction by Julius Lange in 1966 the Lange coupler (1) has been the most widely used 90 degree hybrid at microwave frequencies because of its very wideband −3 dB coupling over octave bandwidths. The structure of the device, shown in Figure 1 is relatively simple, with alternate interdigitated fingers coupled together using bond wires. For this structure:

S is the spacing between interdigitated fingers.

W is the width of the interdigitated fingers.

L is the width of the coupler; 1/4 wavelength at the center frequency.

h is the substrate thickness of the microstrip.

To date, the accepted design approaches are those reported by Presser (2) and Osmani (3), which are similar in nature and are suitable for CAD implementation. Their design approach may be summarized as follows:

The interdigitated directional coupler may be viewed as a multiconductor transmission line of K (K even) elements, not including the ground plane. However, coupling between non-adjacent sections may be neglected with little error. Using this simplification, Ou (4) derived the following set of equations for the even and odd mode impedances for such an array of parallel coupled lines.

$$Z_{oo} = Z_o \left(\frac{1-C}{1+C}\right)^{1/2} \frac{(K-1)(1+q)}{(C+q)+(K-1)(1-C)} \quad (1)$$

$$Z_{oe} = Z_{oo} \frac{C+q}{(K-1)(1-C)} \quad (2)$$

where,

K = even number of strips

Z_0 = terminating impedance (usually 50 ohms)

C = voltage coupling coefficient

$q = [C^2 + (1 - C^2)(K - 1)^2]^{1/2}$

Next, the shape ratios are calculated from the synthesis equations of Akhtarzadm et al (5):

$$S/h = \frac{2}{\pi} \cosh^{-1}\left[\frac{\cosh\left(\frac{\pi}{2}(W/h)_{se}\right) + \cosh\left(\frac{\pi}{2}(W/h)'_{so}\right) - 2}{\cosh\left(\frac{\pi}{2}(W/h)'_{so}\right) - \cosh\left(\frac{\pi}{2}(W/h)_{se}\right)}\right] \quad (3)$$

$$W/h = \frac{1}{\pi} \cosh^{-1}\left[\frac{g+1}{2} \cosh\left((W/h)_{se}\frac{\pi}{2}\right) + \frac{g-1}{2}\right] - \frac{S}{2h} \quad (4)$$

where,

$$(W/h)'_{so} = 0.78(W/h)_{so} + 0.1(W/h)_{se} \quad (5)$$

and $(W/h)_{se,so}$ are given by Wheeler's equations for wide and narrow microstrip lines:

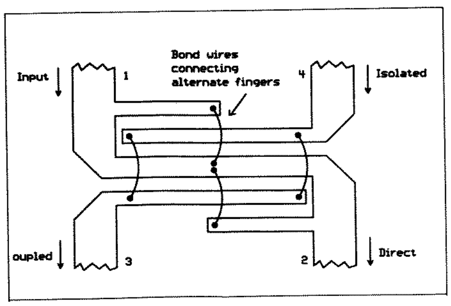

Figure 1. Diagram of a Lange coupler.

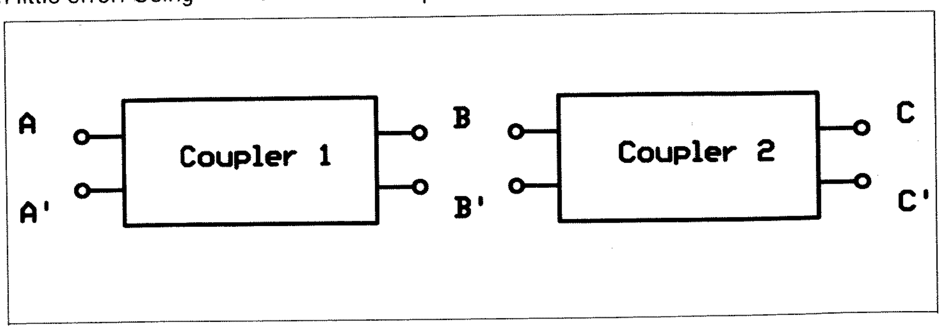

Figure 2. Couplers connected in tandem.

$$(W/h)_{se,so} = \frac{8}{p}\left[\frac{p(7 + 4/\varepsilon_r)}{11.0} + \frac{11/\varepsilon_r}{0.81}\right]^{1/2} \tag{6}$$

where,

$$p = \left[e^{\left(\frac{Z_{oo,oe}}{84.8}\sqrt{\varepsilon_r + 1}\right)} - 1\right] \tag{7}$$

Tandem Lange Couplers

Although the design of Lange couplers is relatively straightforward using the above equations, their main disadvantage lies in the fact that the dimensions of the finger width (W) and spacing (S) are often so small that fabrication of the circuit is difficult. Typical dimensions for Lange couplers are of the order of 10 μm, and this can pose difficulties unless the fabrication process has very good tolerances. A novel solution is to combine two Lange couplers in tandem, so that the overall coupling coefficient is higher than the individual coupling coefficients. Therefore, since the dimensions are inversely related to the value of the coupling coefficients, the etching constraints are greatly reduced.

Consider two couplers connected in tandem as shown in Figure 2. Assuming the signal level at input A is 1, and is 0 and A′, and that the coupling coefficient of each section is k, the signal level at the other ports are:

$$B = k \tag{8}$$

$$B' = j\sqrt{1 - k^2} \tag{9}$$

$$C = k^2 - (1 - k^2) \tag{10}$$

$$C' = j2k\sqrt{1 - k^2} \tag{11}$$

Here we can see that the 90 degree phase shift between C and C′ is still preserved. Now for −3 dB power division, signals at C and C′ must be equal. Therefore,

$$K^2 - (1 - k^2) = 2k\sqrt{1 - k^2} \tag{12}$$

Defining $l = k^2$ yields:

$$8l^2 - 8l + 1 = 0 \tag{13}$$

By solving this equation and discarding one of the solutions, we get k = −8.34 dB. Thus, to achieve an overall coupling factor of −3 dB, each of the two couplers connected in tandem must have a coupling factor of −8.34 dB.

Therefore, two Lange couplers with coupling factors of −8.34 dB can be combined to make a −3 dB coupler with the same overall characteristics as a single coupler. The more loosely coupled individual sections will relax the etching constraints by increasing the dimensions of the coupled fingers. Figure 3 shows two Lange couplers connected in tandem.

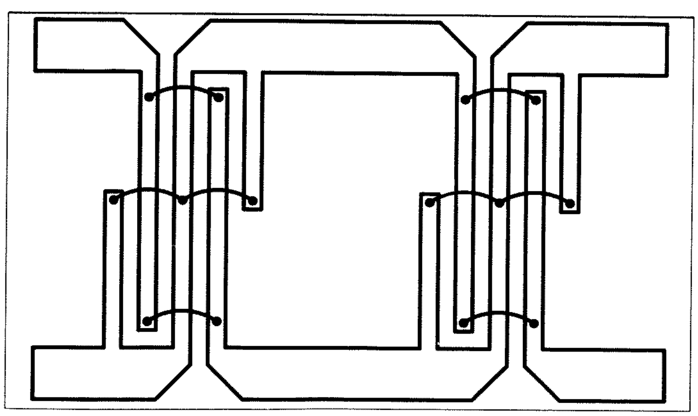

Figure 3. Tandem Lange coupler arrangement.

The CAD Program

A FORTRAN CAD program has been written which enables the user to design Lange couplers using the equations above. Single or tandem configurations can be selected, depending on the dimensions obtained and the quality of the etching process available. Figure 4 shows the simulated Touchstone® response of the single and tandem couplers. While the response of the single coupler is marginally better, in some applications this slight deviation is much less than the discrepancy in results (with a single coupler) due to the etching tolerances of the process available.

Finally, the CAD program gives the user the option of producing an output file directly in the format of a Touchstone input file. This allows great flexibility in incorporating the Lange coupler into a larger circuit, or allows the mask of the circuit to be generated directly from software such as MICAD® or GAS-STATION® .

The operation of the program TLANGE is described by the flowchart in Figure 5. First, the design parameters are entered; substrate dielectric constant, substrate thickness, center frequency, and overall coupling coefficient. If a tandem arrangement is selected, the program then calculates the coupling coefficient of each section. Next, the dimensions of the Lange coupler are calculated using the Osmani/Presser method, and finally, the results are produced either directly to the screen, or in the format of a Touchstone input file. Figure 6 shows a typical output of the TLANGE program. Here a six finger tandem Lange coupler with a center frequency of 4 GHz has been designed on 5880 RT-Duroid® (0.79 mm substrate thickness).

The program is available from the RF Design Software Service, including a compiled version and source code listing. See page 30 for more information. **RF**

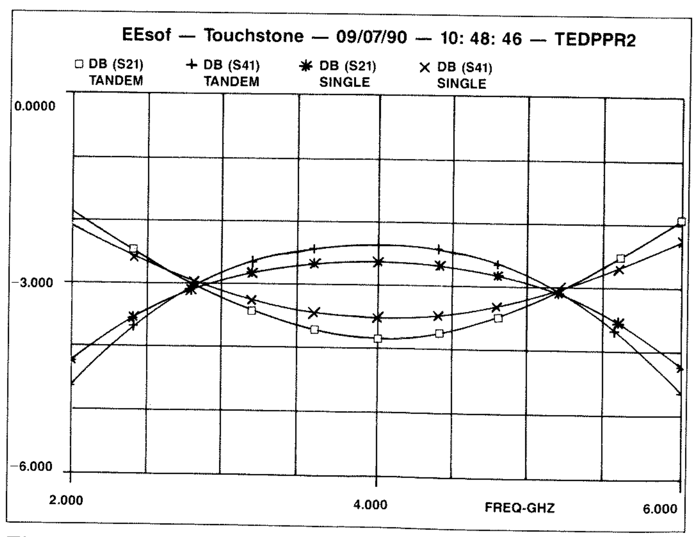

Figure 4. Simulated results of a single and tandem Lange coupler.

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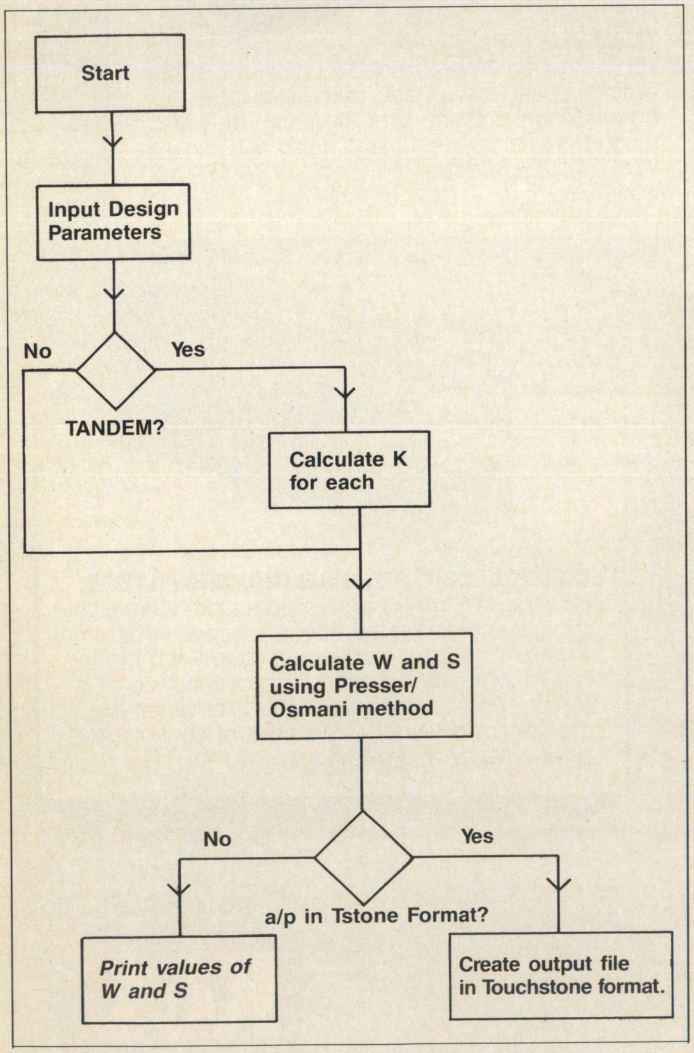

Figure 5. Flowchart of the TLANGE program.

```
DIM
FREQ GHZ
RES OH
IND NH
CAP PF
LNG MM
TIME PS
COND /OH
ANG DEG

CKT
 MSUB   ER=   2.20 H=0.79 T=.005 RHO=1 RGH=0
 MLANG6 1 2 3 4 W=.253 S=.406   L=13.500
 MLANG6 2 6 4 7 W=.253 S=.406   L=13.500
DEF4P 1 6 3 7 LNGE

OUT
LNGE DB[S21] GR1
LNGE DB[S41] GR1

FREQ
 SWEEP 2.5 4.0 5.5

GRID
RANGE  2.5 4.0 5.5
GR1 0 -20 2
```

Figure 6. Touchstone file output from the TLANGE program.

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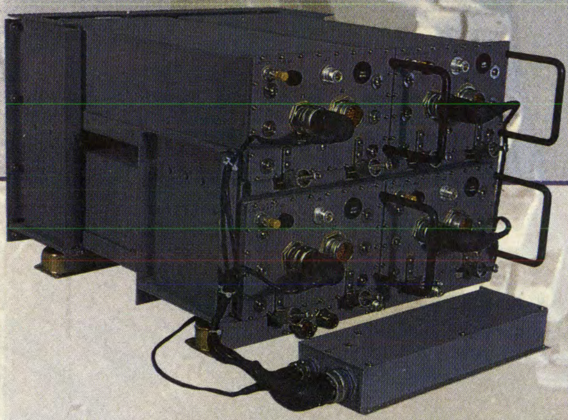

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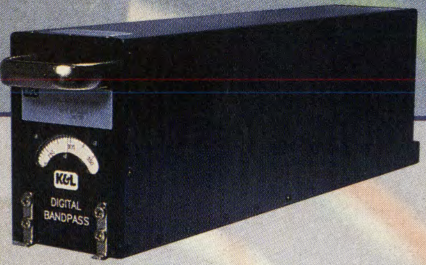

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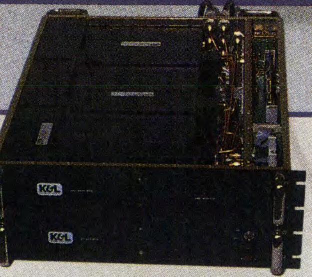

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A Clear Signal for RF Filters

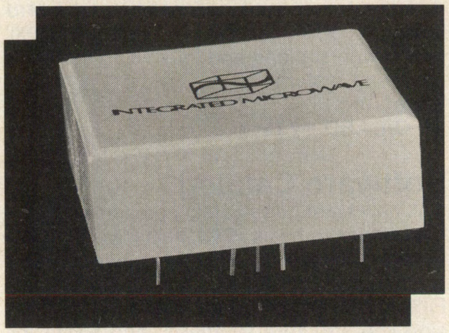
By Liane G. Pomfret
Associate Editor

The market for RF filters is stable and growing with minor variations that can be linked to the current recession. As usual, the commercial market is enjoying greater success than the military market. Despite a worldwide economic slump, some of the best business is being generated internationally and has buoyed relatively flat domestic sales. Surface mount technology is also contributing to continued profits. With a host of new applications and new technology, there's little to stand in the way of a successful future for RF filters.

It comes as no great surprise that the military market for RF filters has suffered because of congressional cutbacks. While the cutbacks have not been disastrous, they have been noticeable and companies involved in military have felt the pinch. For companies who focus on the commercial market, business is much better. The recent explosion of commercial RF applications has opened up new business opportunities. The list of applications — wireless communications, PCNs, mobile and land mobile, data transfer, medical systems, and satellites — continues to grow. While much of this technology was developed for the military, it has found success within the commercial arena, making it doubly profitable for the manufacturers.

Not all of the technology evolved from military contracts. Some developments came from the commercial sector. Surface mount and miniaturization of RF filters are the results of requests for smaller commercial devices — the 2 pound cellular phone as opposed to the 7 pound cellular phone. Like much of the RF industry, surface mount technology is now working its way into the filter industry. Yet the military still lags behind. Arie Verhagen, Vice President of Marketing for Murata-Erie comments, "The industry as a whole, has to do their homework to bring surface mount to the military's attention." Qualifying new technology for military specifications is a long, expensive process and the industry is reluctant to invest time and money without assurances that the technology will be accepted.

In the commercial sector, suppliers have been reducing their prices while maintaining or improving quality. As a result, manufacturers have more latitude to experiment with manufacturing techniques and materials. Steve Sodaro, Vice President of Marketing and Sales at TTE notes, "Components now are not as costly. It affords us the opportunity to do miniaturization. We're always trying to miniaturize without sacrificing performance." However, while a large firm has the resources to experiment with technologies such as surface mount, smaller companies are often restricted because of the lack of available packaging technology within their company. Packaging technology is available from outside vendors but cost and availability are a stumbling block to in-house development.



A switched filter module from Integrated Microwave.

The communications industry offers a large number of uses for filters. Voice communications, data transfer and acquisition, LANs, and voice over data are all demanding more sophisticated filtering techniques. "The applications for filters in general are increasing, mainly due to enhancements in telecommunications, LANs, and various radio systems including cellular and land mobile beyond the classic military market," says Dave Distler, Director of Sales and Marketing for Trilithic. Manufacturers are demanding extremely sophisticated filters capable of high reliability, high performance filtering.

Every company has its own area of specialty and responses will vary regarding the active areas for filter products. However, the general consensus is that communications equipment is the hot area right now. Randy Rhea, President of Eagleware notes, "We see a definite increase in interest in filters for data communications systems and that often results in a greater emphasis in the delay characteristics of filters." Al Lowenstein, National Sales Manager, for Motorola's RF filter products components division, sees interest for their products in other areas as well. "The new activity is in the 915 MHz range, especially spread spectrum, new generation portable telephones, and remote data acquisition, meter reading and things along that line." He adds, "The other thing that is taking off is GPS and the satellite industry in general." Wavetek's focus has been high reliability communications such as those found in avionics communications systems. Even with the slowdown in the domestic military market, Sara Mussman, Components Business Unit Manager for Wavetek comments, "I see a significant amount of business out there."

A lot of this business is coming from the international sector. Mussman notes, "Any country that is aggressively pursuing an upgrade will be a good source of business." For George Szentirmai, President of DGS, business has been fairly flat. "The only thing is that our international sales are pretty strong especially in Europe." His company has had a great deal of interest from eastern european countries, but since there's no money available, there's been no sales. Werner Mueller, Vice President of KVG, North America comments, "I would hope that due to the reunification of Europe and due to the communication lack in eastern Europe, that there will be an increase in the European market." Toyocom's market situation is different. They currently hold less than 10 percent of the U.S. market and are looking to expand within the United States according to Dennis Reifel, Marketing Manager at Toyocom.

With the majority of the RF filter business on the upswing, the outlook for the future is optimistic. The advent of surface mount technology and constant improvements in price and performance make filters an excellent prospect for profit in the coming year. ***RF***

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INFO/CARD #232

Graphics-Based Analysis

DADiSP is a visually-oriented graphics software package designed for scientific and technical applications. Among the 200 data processing functions are signal arithmetic and editing, waveform generation, FFTs, and peak-finding. Extended memory support is an option, as is an IEEE-488 bus driver and an software module to interface with popular data acquisition boards. Versions are available for a wide range of PCs, workstations and larger computers.
DSP Development Corp.
INFO/CARD #231

SPICE Library for Linear ICs

SPICE macromodels are now available for nearly every Burr-Brown op amp, difference amp, and instrumentation amplifier. Two levels are offered: PSpice® Parts™ and Enhance Parts™ macromodels. A 5 1/4 inch diskette with more than 75 models is free from Burr-Brown.
Burr-Brown
INFO/CARD #230

Thermal Analysis

Version 1.33 of SAUNA, a 3-D thermal annalysis package has been released by Tatum Labs. Larger models are possible; up to 700 nodes and 2750 thermal resistors. Reduced air density due to altitude is also accoutned for. SAUNA runs on IBM PC compatible and Macintosh II and SE/30 computers.
Tatum Labs, Inc.
INFO/CARD #229

RF literature

RF/Microwave Product Guide

Avantek offers a 48-page Product Selection Guide to give engineers a quick overview of the company's standard products. Summary specifications are provided for transistors, integrated circuits, amplifiers, signal processing and control products, LNAs, uplink amplifiers, cellular base amplifiers, and YIG oscillators. Most items are in stock at Avantek distributors.
Avantek, Inc.
INFO/CARD #250

Ceramic Filter Catalog

Murata Erie announces a design manual and catalog for their line of ceramic filters for communications. Single and multiple element filters are listed, covering standard frequencies such as 455 kHz, 3.58 MHz, 4.5 MHz and 10.7 MHz. Ceramic discriminators and signal detectors are also included, as are surface mount filters. Application information includes theory of operation and circuit design considerations.
Murata Erie North America
INFO/CARD #249

Frequency Control Products

Comprehensive data on products available from Piezo Technology, Inc. is included in a new 96-page handbook. Monolithic and discrete crystal filters, LC filters, cavity bandpass filters, combline filters, interdigital filters, crystal and non-crystal oscillators, precision crystal resonators, RF subassemblies, and custom hybrid circuits are listed.
Piezo Technology, Inc.
INFO/CARD #248

RF Data Book

Motorola's RF Products Division has released their new RF Data Book, DL110/D Rev. 4, a two-volume edition covering power FETs, power bipolar and small-signal transistors, plus RF power modules. 84 new products have been included since the previous version, along with new applications information.
Motorola, Inc.
INFO/CARD #247

Coaxial Switches

A new Standard Products Brochure is available from Dow-Key Microwave, listing standard electromechanical coaxial switches and relays in configurations from SPDT to SP12T. Transfer switches, bypass switches and low cost RF relays are also included. Photographs, specifications and schematic diagrams are provided for these products.
Dow-Key Microwave Corp.
INFO/CARD #246

1991 Test Instruments

Tektronix' 1991 customer catalog is now available, highlighting more than 3000 products in three broad areas: electronic test and measurement instruments, professional broadcast equipment, and computer peripherals. New products include communication signal analyzers, VXIbus products, high resolution display monitors, HDTV test equipment and low cost broadcast spectrum analyzers.
Tektronix, Inc.
INFO/CARD #245

Coatings Brochure

A line of coating and ink systems is described in a new brochure from Emerson & Cuming. One and two-component systems are presented, used in making membrane switches, molded and flexible circuits, polymer thick film circuitry, and for modifying circuit boards. The coatings provide protection against electrical contact during component assembly.
Emerson & Cuming, Inc.
INFO/CARD #244

Frequency Chart Update

The RF Products and Linear Integrated Circuits Divisions of Motorola have published a new U.S. frequency allocation chart covering the RF spectrum up to 4000 MHz. Added to the previous version's allocation categories is a segment for Part 15 unlicensed devices. The chart is free, and is suitable for convenient wall mounting.
Motorola, Inc.
INFO/CARD #243

SAW Filter Note

A new Application Note discussing SAW coupled-resonator (CR) filters is available from RF Monolithics. SAW CR filters are suitable for many narrow-band applications from 70 MHz to 1500 MHz, and can be built with fractional bandwidths from 0.03 to 0.5 percent. The filters exhibit low loss (3 dB typical) under matched conditions. The note covers theory, operational performance limits, and a comparison with other filter technologies.
RF Monolithics, Inc.
INFO/CARD #242

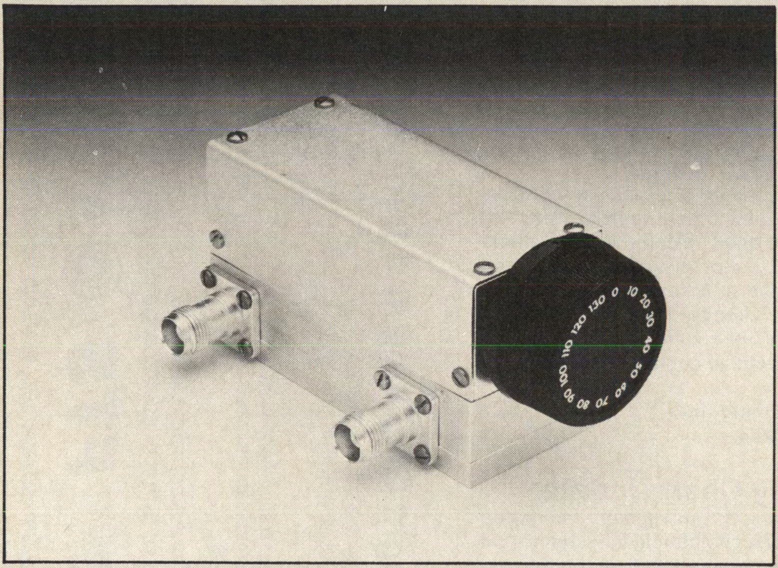
Reconditioned Equipment

A catalog featuring reconditioned Philips/Fluke test and measurement equipment is announced by Rubytron. Oscilloscopes, synthesizers, frequency counters/timers, sweepers, function generators and pulse generators are among the equipment types available. All instruments are calibrated and warranted.
Rubytron Instruments
INFO/CARD #241

Distributor Catalog

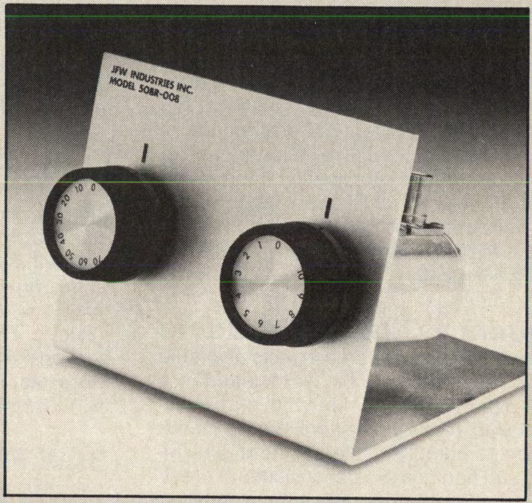
Tektronix 1991 Distributor Products Catalog contains data on analog and digital oscilloscopes, frequency counters, digital multimeters and function generators. These products are offered through authorized distributors.
RAG Electronics, Inc.
INFO/CARD #240

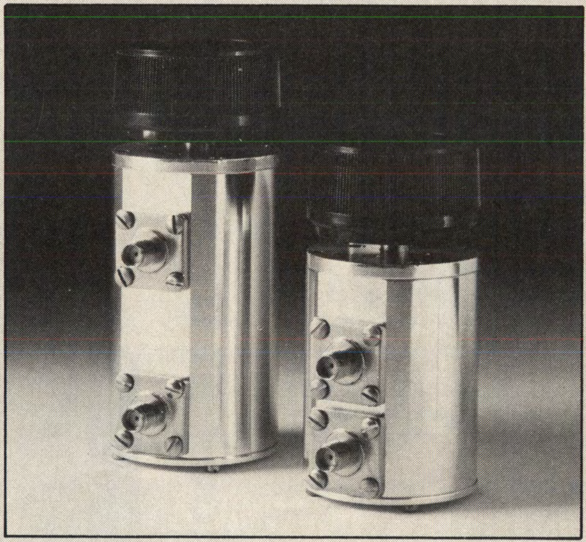
JFW Rotary Attenuators



50R-002
Frequency Range: DC-1000 MHz
Attenuation Range: 0-130 in 10 dB steps
Connectors: BNC, N, TNC or SMA

50BR-008
Frequency Range: DC-2000 MHz
Attenuation Range: 0-80 dB in 1 dB steps
Connectors: BNC female (standard) SMA and N available upon request



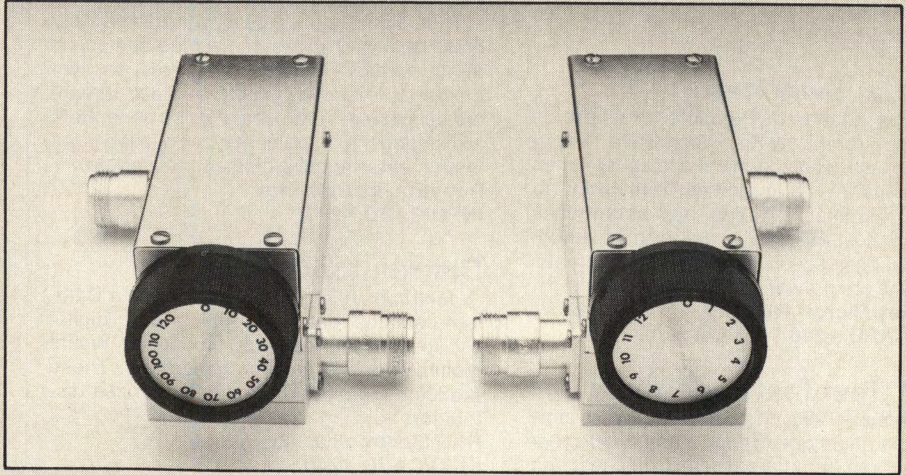


50R-029 (left)
Frequency Range: DC-2000 MHz
Attenuation Range: 0-70 dB in 10 dB steps
Connectors: SMA female (standard) BNC available upon request

50R-019 (right)
Frequency Range: DC-2000 MHz
Attenuation Range: 0-10 dB in 1 dB steps
Connectors: SMA female (standard) BNC available upon request

50R-079 (left)
Frequency Range: DC-1000 MHz
Attenuation Range: 0-120 dB
Connectors: BNC female or "N" female

50R-080 (right)
Frequency Range: DC-1000 MHz
Attenuation Range: 0-12 dB
Connectors: BNC female or "N" female



JFW Industries, Inc.
5134 Commerce Square Dr.
Indianapolis, Indiana 46237
(317) 887-1340

INFO/CARD 69

Advertising Index

RF engineering opportunities

RF ENGINEERS

GEO-CENTERS, INC., a high technology, growth-oriented research and development company specializing in engineering and high-tech services for the U.S. Department of Defense, has the following immediate, full-time openings at our central New Jersey location:

Senior RF Engineer — to perform R&D on high stability frequency standards and clocks, including oscillator design and specifications; and resonator and oscillator testing. A PhD in EE or a related field and CAD/ATE experience are desirable. Must be a self-starter who is capable of independent and creative work in an R&D environment.

RF Engineer — to perform R&D on high stability frequency standards and clocks, including oscillator design, fabrication and specification; and resonator and oscillator testing. A BS in EE or a related field and CAD/ATE experience are desirable. Experience in data analysis, modern computational methods and good oral and written communication skills are necessary.

GEO-CENTERS, INC. offers competitive salaries, an outstanding benefits package and unlimited growth potential. To explore these positions, please forward your resume with salary requirements to: GEO-CENTERS, INC., Human Resources Dept. - RF, 762 Rt. 15 South, Lake Hopatcong, NJ 07849. An Equal Opportunity/ Affirmative Action Employer. U.S. citizenship or permanent residency required.

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RADIO FREQUENCY ENGINEERS

When you're creating the world's most advanced cellular network, you need the most advanced Radio Frequency Engineers.

Among your varying responsibilities, you will evaluate subscriber traffic, present and future RF coverage and growth data to forecast system expansion and to provide optimal service for our customers. You will design the frequency plan, conduct testing on potential installation areas, assist the real estate group on potential cell and enhancer locations, and remain current on new technology and industry advancements.

The individuals we're seeking will have a Bachelor's degree in Electrical Engineering and two or more years of experience in the cellular or radio design industry.

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